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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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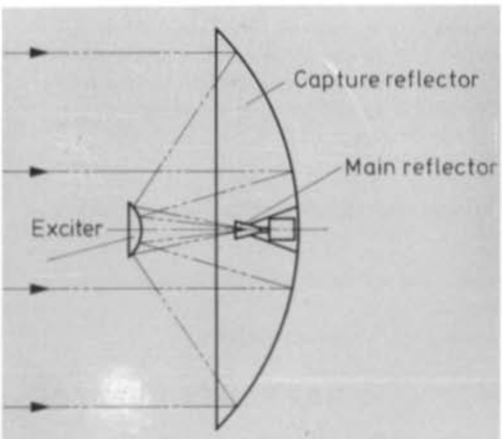

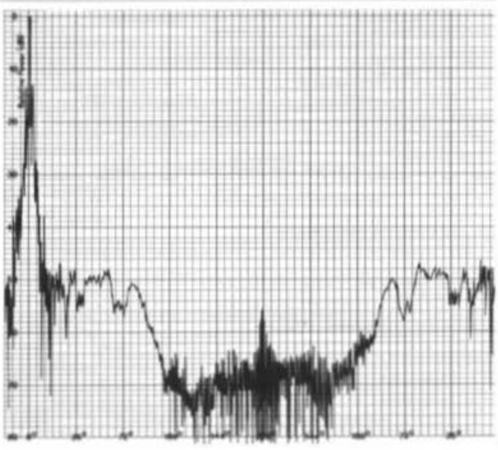

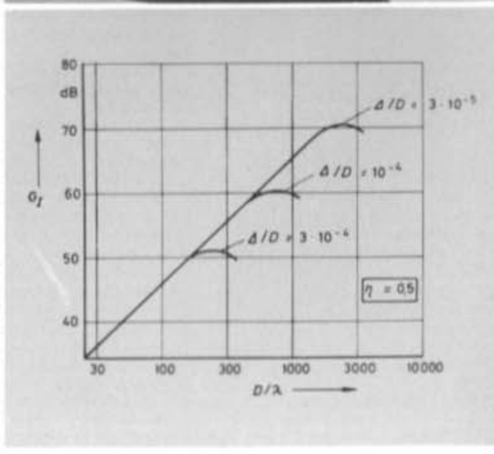

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THE MOST IMPORTANT FEATURES AND CHARACTERISTICS OF GHz ANTENNAS

by H. Berner, VDE/NTG

More and more telecommunications activity is moving to the microwave bands, mainly due to the overloading of the frequency range under 1 GHz, but also due to the high directivity and gain that can be obtained using relativily compact arrays. Beamwidths of 1° and less can easily be realized. It is known that the lower the beamwidth, the higher the gain of the antenna. This means tha low output power levels will be able to provide sufficient signal strength over greater distances. Further details regarding propagation at frequencies in excess of 1 GHz are to be dealt with separately.

1. MICROWAVE ANTENNAS

This article is to be limited to antenna designs for the frequency range of 3 GHz to 30 GHz, mainly because it is in excess of 3 GHz where the antennas differ from those known in VHF/UHF technology, and since very little activity is actually being made at frequencies in excess of 30 GHz. The range of frequencies between 3 and 30 GHz is also called the centimetric wave range. It is in this range where there are such a large number of different types of antennas, especially in the sphere of aperture antennas.

Aperture Antennas

Horn Antennas	Reflector Antennas	Discrete Antennas
Pyramidal horns	Parabolic antennas	Slotted-line radiators
Sectoral horns	Shell-shaped antennas	Multiple-element arrays
Corrugated horns	Hog-horn antennas	Printed circuit antennas
Conical horns	Cassegrain antennas	Phase-controlled arrays
Double conical horns	Cylindrical parabolics	

Logarithmic-periodic Antennas	Spiral Antennas	End-on Arrays	Focussed Arrays
LP V-Antennas	--	--	--
LP Dipole Antennas	--	--	--

Table 1: Microwave antennas for the frequency range over 3 GHz

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Dipl.Ing. Hellmut Berner is a staff-member of Standard Elektrik Lorenz AG

2. SPECIFIC CHARACTERISTICS

2.1. Microwave antennas are large with respect to the wavelength. This means that generally geometric and optical laws must be used to describe the radiation mechanics.

2.2. Linear bodies such as dipoles are no longer used for radiation, but reflective surfaces or specially shaped waveguides (horns).

2.3. Low beamwidths can be achieved with handy dimensions. This means a high directivity in both planes, which ensures that the interference to other services on the same frequency and ground reflection are kept to a minimum.

We are to differentiate between »gain« and »directivity«. Both of these designations are used in the NTG recommendations 1301 of 1969. The main difference is that the same power is fed to the test antenna and reference antenna in the gain measurement, whereas it is the measurement of the effective radiated power that is defined in the directivity measurement. The relationship between gain G and directivity D is:

$$G = \eta \times D$$

Where the antenna efficiency is: $\eta \leqq 1$.

The gain therefore includes the losses in the antenna and any required matching networks.

The directivity can be obtained from the measured radiation diagram by means of integration, and does not include the losses.

A halfwave dipole or isotropic radiator are often given as reference antenna. All gain and directivity values given in this article are referred to an isotropic radiator and are given in a logarithmic scale (dB).

3. MAIN TYPES OF ANTENNAS AND THEIR CHARACTERISTICS

3.1. Aperture Antennas or Radiating Surfaces

3.1.1. Horn Antennas

The aperture of the horn can be classed as radiating surface and the main radiating characteristics can be calculated from this. The surface should be assumed to be covered with an infinite number of infinitely small elementary radiators. The radiation diagram results from the interference of the many individual radiations. Directivity of a pyramidal horn (see **Fig. 1**) can be calculated approximately as follows:

$$D \approx 10 \log\left(\frac{10\ a \times b}{\lambda^2}\right) \text{ dB}$$

Where »a« and »b« are the lengths of the aperture surfaces.

In practice, a value of between 15 dB and 25 dB results. Horn antennas of this size exhibit low sidelobes. However, if the horn and thus the directivity are increased, this will also increase the sidelobes. If the beamwidths of the horn are sufficiently small, e.g. 10°, it is possible to match a horn antenna in a frequency range of 1 : 1.5. Special types of horns allow a frequency range of 1 : 10 with a VSWR of 2.5 : 1.

In addition to the previously mentioned pyramidal horn, one also uses conical, sectoral, corrugated and double-conical horns. Since horn antennas are relatively easy to manufacture, they are very popular as individual antennas, standard-gain antennas, as driving elements of reflector antennas, or as individual elements in phased arrays (see 3.1.3).

3.1.2. Reflector Antennas

Reflector type antennas represent the largest group of aperture antennas (see **Table 2**). Operation can be compared with that of an automobile headlight and need not be discussed in detail here. Since the reflecting surface must be large with respect to the wavelength, as in optics, these antennas are usually only used in the centimeter and millimeter wave ranges.

Type	Principle	Use		
		Single narrow lobe	Formed lobe	Multiple lobe
FRONT-FED PARABOLIC ANTENNA		Microwave links Satellite comm. Radio astronomy Radar	Radar	Satellite comm. Satellite antennas Radar
SHELL-SHAPED ANTENNA		Microwave links	—	—
HOG-HORN ANTENNA		Microwave links Satellite comm.	—	—
CYLINDRICAL PARABOLIC		Radar	—	—
CASSEGRAIN ANTENNA		Microwave links Satellite comm. Radio astronomy	—	Radar
GREGORIAN ANTENNA				

Table 2: Types of reflector antennas

Reflector type antennas exhibit a simple and robust construction, allow simple feed systems and provide a good front-to-back ratio. The simplest and most popular type is therefore the front-fed parabolic (see **Fig. 2**).

The parabolic reflector is energized by a horn mounted at the focal point. The microwave energy is fed to the horn via a bent waveguide. The directivity is dependent on the following relationship and is usually in the order of 10 dB to 60 dB :

$$(D/\lambda)^2$$

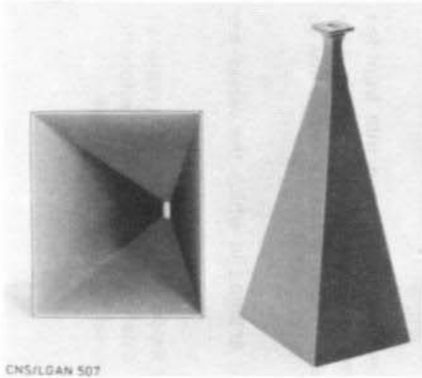

Fig. 1: Horn radiator

Fig. 2: Front-fed parabolic antenna

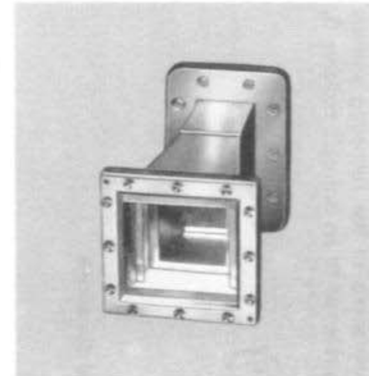

Fig. 4:
Rectangular horn radiator for 4 GHz

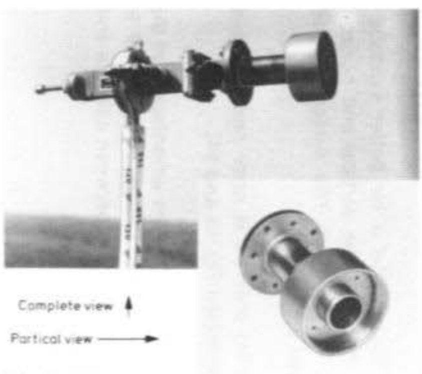

Fig. 6:
Laboratory model of a
coaxial multimode radiator

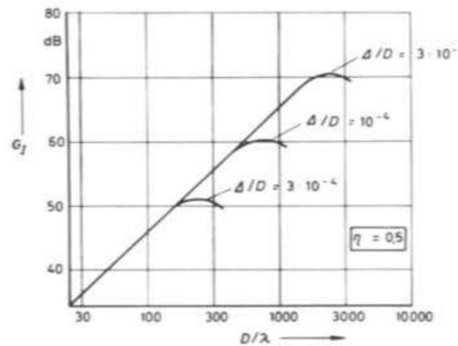

Fig. 3:
Gain of a parabolic antenna with deviations
from the ideal parabolic shape

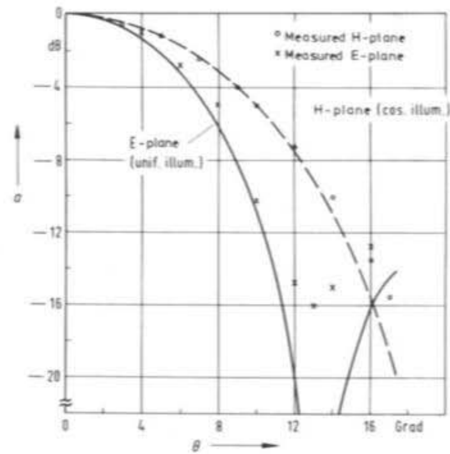

Fig. 5:
Radiation pattern of a
square horn radiator

With a given wavelength, the directivity will increase on increasing the diameter. However, in practice there are limiting factors since inavoidable tolerances of the reflecting surface reduce the directivity, as can be seen in **Fig. 3**. The gain referred to an isotropic radiator G_I was calculated according to the following equation:

$$G_I = \eta \left(\frac{\pi D}{\lambda}\right)^2 e^{-159\left(\frac{\Delta}{D}\frac{D}{\lambda}\right)^2}$$

The maximum gain $G_{I\,max} = \dfrac{\eta}{16\, e\left(\frac{\Delta}{D}\right)^2}$

Where: η = Surface area used
Δ = Square mean value of the dimensional deviations
D = Reflector diameter
λ = Wavelength

It is mainly horn antennas with a square, rectangular or circular cross-section that are usually used to feed the antenna (**Fig. 4**).

In spite of the square aperture, the radiation characteristics in the E- and H-planes are different due to the different current ratios in each plan (see **Fig. 5**).

Since the sidelobe suppression and the directivity are mainly dependent on the distribution of the energy from the horn to the reflecting surface, methods of optimizing this energy distribution were sought. The central laboratories of the German PTT therefore developed a so-called coaxial radiator (**Fig. 6**), which excites several wave modes within its aperture plane and thus obtains an approximately symmetrical energy distribution on the reflector.

A corrugated horn (see **Fig. 8 and 9**) also provides a virtually symmetrical energy distribution in all planes.

However, there are other methods of obtaining a favourable energy distribution.

Shell-shaped (**Fig. 10**) and hog-horn antennas (**Fig. 11 and 12**) are constructed in a similar manner to front-fed parabolics. The difference is that no circular parabolic reflector is used but only a section of such a reflector.

These antennas exhibit a far better sidelobe suppression than in the case with front-fed parabolics. However, the vertical and horizontal diagrams are different due to their non-symmetrical construction. The enclosed construction makes such antennas insensitive to environmental influences.

The cassegrain antenna is once again a truely parabolic antenna in which the waves are deflected twice (see **Fig. 13 and 14**).

This type of antenna has the advantage that the long bent feeder of the parabolic is replaced by a short waveguide. This is especially of interest for antennas used in radio astonomy where the waveguide loss between antenna and receiver should be as low as possible.

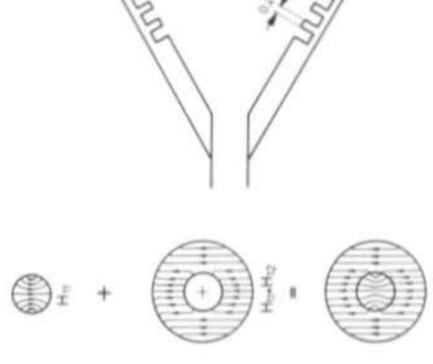

Fig. 7:
Coaxial multimode radiator (FTZ, W. Germany)

Fig. 8:
Corrugated horn

Fig. 9:
Laboratory model of a corrugated horn

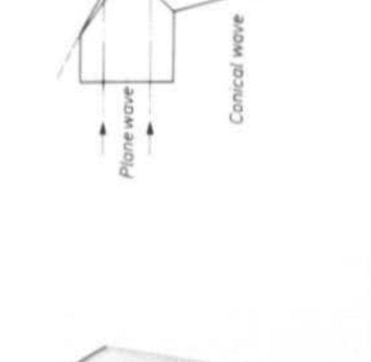

Fig. 10:
A shell-shaped antenna

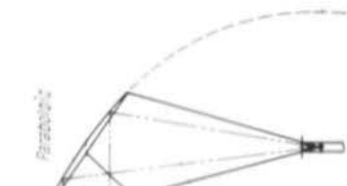

Fig. 11:
Principle of a horn-parabolic (hog-horn)

Fig. 12:
Hog-horn antenna

A truely parabolic construction also has the advantage that two waves can be transmitted and received when using cross polarisation. It may be a disadvantage that the sub-reflector covers a part of the reflector surface, which means that the first sidelobes are only suppressed by 13 dB to 17 dB (see **Fig. 15**) compared to 18 dB to 23 dB with other types.

The previously mentioned reflector antennas with directivity factors from 20 dB to 60 dB are used for microwave links, space communications, radio astronomy, satellite ground stations (**Fig. 17**), and for radar applications.

The cylindrical parabolic antenna should also be mentioned whose reflective surface is only parabolic in one plane (see Fig. 17).

This type of antenna possesses entirely different beamwidths in the horizontal and vertical planes, as is often required for radar applications.
Example: 2.4° horizontal and 6.5° vertical beamwidth.

A slotted waveguide (**Fig. 18**) is normally used for energizing such antennas. This is a straight piece of waveguide into which slots have been cut on the side facing the reflecting surface. These slots radiate energy towards the reflector. It is possible by altering the geometry of the slots to vary the energy distribution along the length of the waveguide so as to achieve a sidelobe attenuation of 20 dB and more.

3.1.3. Antennas with Discrete Elements

As in the lower frequency ranges such as VHF, it is also possible in the microwave range to combine a large number of individual elements to form phased arrays. Virtually any radiation pattern can be obtained in this manner. One of the horn types, or slotted lines are often used as individuel elements. **Fig. 19** shows such a phased array comprising several slotted waveguides.

The directivity and sidelobe attenuation is dependent on the size of the array and on the distribution of the energy. The array shown in the photograph is approximately 500 mm by 500 mm and the - 3 dB beamwidths in the H- and E-planes amounts to about 5°. The sidelobe attenuation is at least 20 dB (see **Fig. 20**).

It is possible by altering the phase relationships between the individual elements to change the direction of the radiation pattern. A very special, but interesting example is given in Fig. 21.

The energy distribution to the individual radiating elements is not made with the aid of lines but in free space by radiation between the primary radiator and a number of collector antennas. A phase-shifter is provided between each of these collector antennas and the radiating emitter antennas. It is possible to shift the beam in both planes by correct adjustment of the phase relationships (**Fig. 22**).

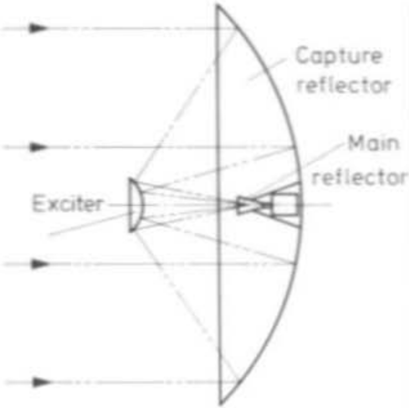

Fig. 13:
Principle of a Cassegrain parabolic antenna

Fig. 14:
A 3 m Cassegrain parabolic

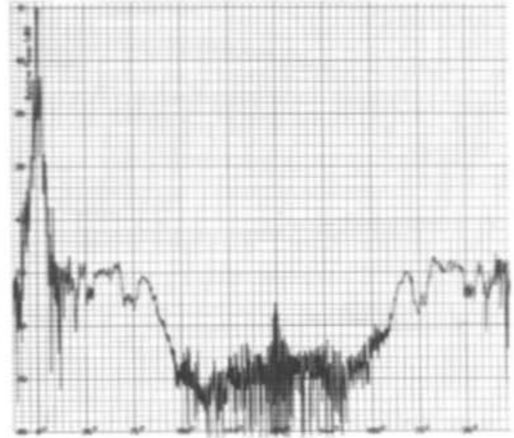

Fig. 15:
Radiation pattern for a Cassegrain parabolic with 3 m dia., f = 6 GHz

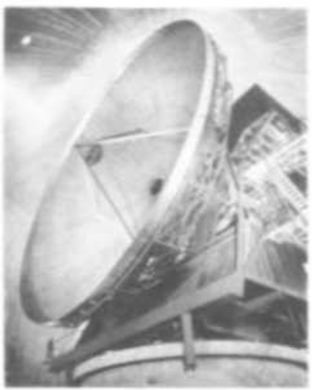

Fig. 16:
Antenna 1 of the satellite groundstation Raisting

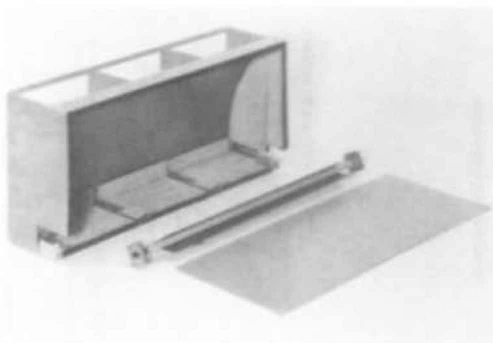

Fig. 17:
Cylindrical parabolic antenna

Fig. 18:
Slotted waveguide R 100 (X-band)

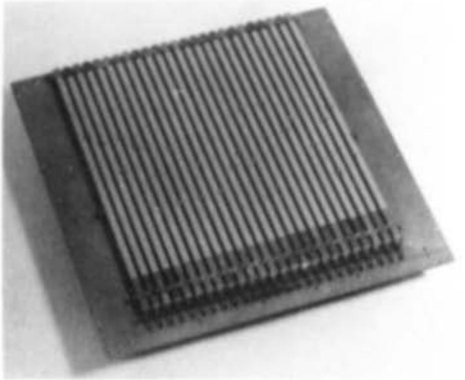

Fig. 19:
Phased array with several slotted waveguides

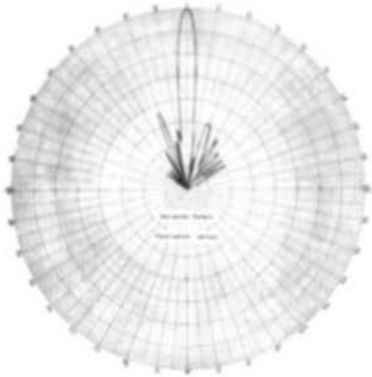

Fig. 20:
Radiation pattern of a radar antenna

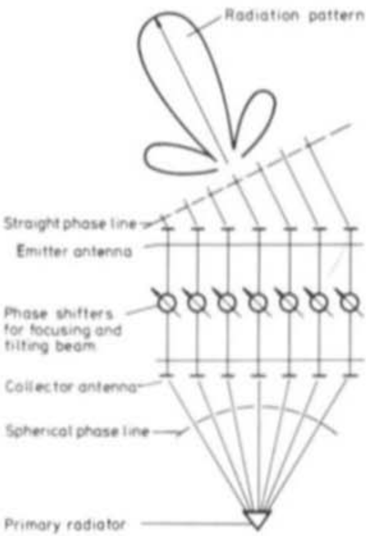

Fig. 21:
Principle of a phased array with »optical-type« feed

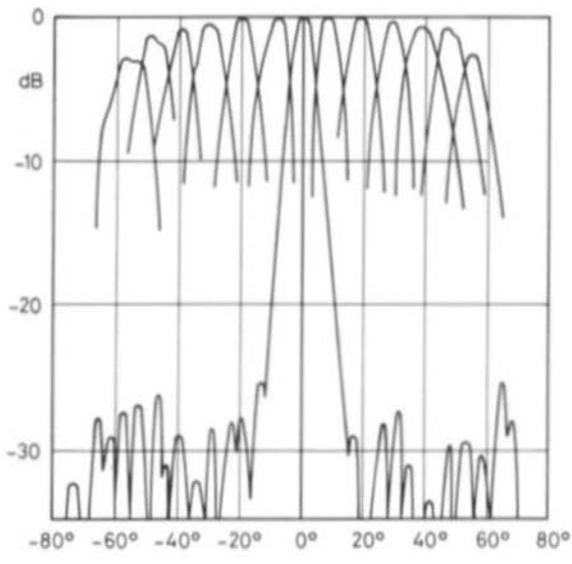

Fig. 22:
Beam deflection of a phased array in the azimuth plane

A coming trend is the use of printed striplines in phased arrays. In this case, the individual radiating elements are replaced by a complete phased array in the form of a printed circuit board. The advantage of this is that the manufacture of large numbers of these antennas is simple and inexpensive, and the antennas are very flat and lightweight. The disadvantage is that the losses are relatively high. When using a PTFE base material, a loss of approximately 2 dB per meter will present at X-band frequencies, whereas with waveguides only approximately 0.1 dB/m will be exhibited. A line length of 1 m is not sufficient even for small antennas, which means that losses in the order of 3 dB to 9 dB are exhibited by printed, phased arrays.

Fig. 23 shows the principle of a printed slot antenna as an example. Slots are etched into the front of the board, and form the radiating elements. The slots are coupled to a stripline feeder, which is tilted to ensure a slightly different coupling to the slots according to the required energy distribution. The base material is glassfibre reinforced PTFE (Teflon). Due to the inavoidable tolerances of the etching process, printed antennas are only used upto approximately 35 GHz.

3.2. Spiral Antennas

Spiral antennas are in a class of their own, and are usually also manufactured in printed circuit technology. In the case of **Fig. 24**, two spirals are wound on a flat surface. The two inner ends are fed symmetrically, and this type can be compared with a bent dipole configuration. Due to the varying spacing between the two spirals a resonance condition for any frequency will appear at various points on the spiral, which means that the antenna can be used over a very wide frequency range (1 : 10 to 1 : 40). The polarisation is linear at low frequencies and becomes more and more circular towards higher frequencies. The radiation is perpendicular to the plane of the spiral, and the - 3 dB beamwidth of the antenna shown in **Fig. 25** is in the order of 90°. Such antennas are used nearly exclusively for aeronautical applications, where its extreme wideband characteristics can be utilized to the full.

The spiral can also be wound on a cone so that a semi-circular radiation pattern results.

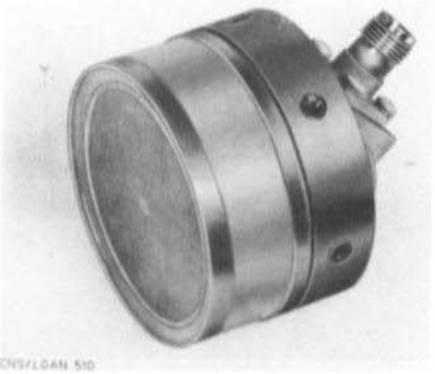

Fig. 25: Spiral antenna with directional characteristics

3.3. Log Periodic Antennas

Finally log periodic antennas are to be discussed. In this type of antenna, several fed dipoles are mounted in either one plane (**Fig. 26**) similar to Yagi-antennas, or in a »V« arrangement where the two front-ends meet (**Fig. 27**).

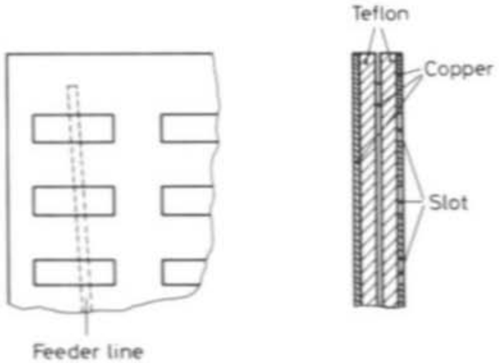

Fig. 23:
Slot antenna in printed circuit technology

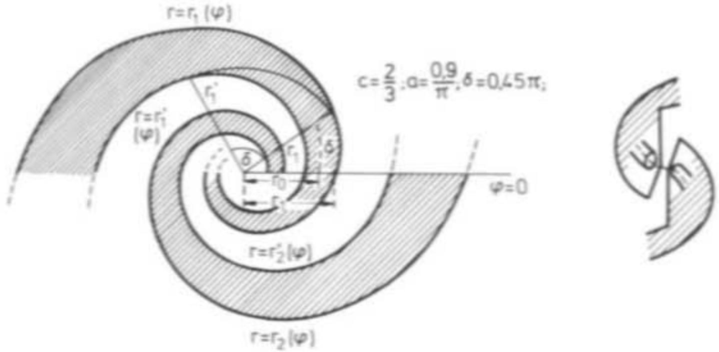

Fig. 24:
Principle of a spiral antenna right view: feeder connection

Fig. 26:
Flat log-periodic antenna

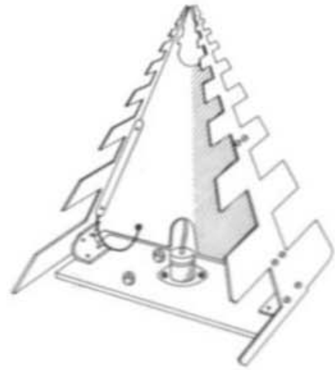

Fig. 27:
Pyramid-shaped log-periodic antenna

The lengths and the spacings of the dipole elements are related by the value τ:

$$\tau = \frac{l_{(n+1)}}{l_n} = \frac{h_{(n+1)}}{h_n} < 1$$

If the operating frequency is varied continuously, there will always be dipoles that are approximately resonant at the actual frequency selected. The electrical characteristics therefore change periodically with the logarithm of the frequency, which led to the designation of these antennas that also possess a wide frequency range of (1 : 10 to 1 : 20). The directivity is limited since only a portion of the antenna is resonant at any one frequency. The average gain is in the order of 7 dB. The dipole dimensions for the higher frequencies will be very small, which is the reason why such antennas are only manufactured for frequencies upto about 12 GHz.

Log periodic antennas are either used on their own as antenna or as exciter for parabolic antennas.

In addition to the types of antennas described there are other types such as focused and end-on antenna arrays. However, they are very few and are not of great importance here.

A TRANSMIT CONVERTER FOR 432 MHz WITH SCHOTTKY RING MIXER

by F. Weingärtner, DJ 6 ZZ

A transverter is to be described that converts 10 m SSB signals to the 70 cm band. In a similar manner to its 2 m brother (1), the 2 m transverter DJ 6 ZZ 005, the excellent characteristics of this transverter are mainly dependent on the use of the inexpensive Schottky ring mixer IE-500. An output power of 2 W is provided at 432 MHz when using an operating voltage of 12 V. This means that this module is suitable for both portable and home operation.

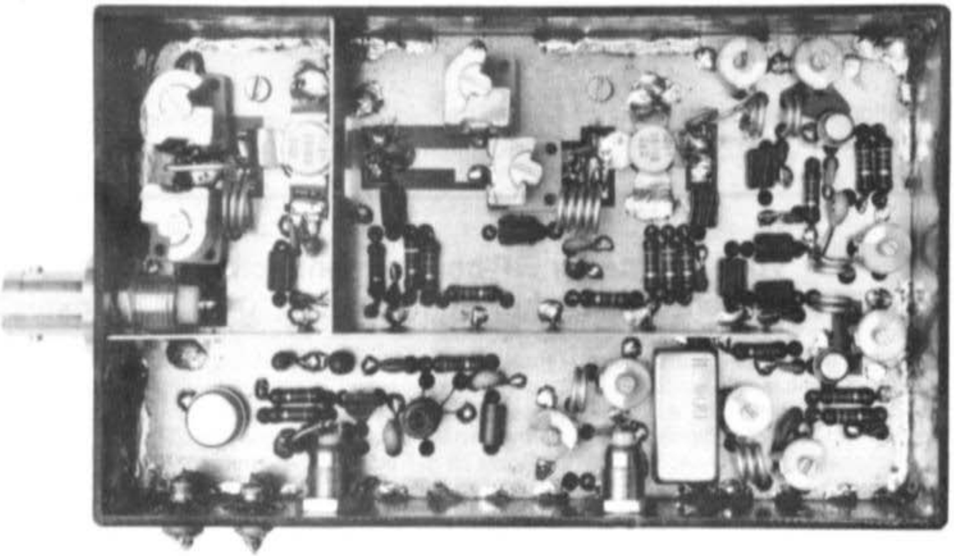

Fig. 1: Prototype of the 10 m / 70 cm transmit converter

1.THE RING MIXER

According to the data sheets, a conversion loss of 7.dB is present when converting to a frequency of 432 MHz in the case of the SRA-1, and 9 dB in the case of the IE-500.
For price reasons, the first prototypes used the (2 dB inferior) ring mixer IE-500.

Whereas the usability of the standard version of this ring mixer seems to be favorable at 150 MHz, it seemed, at first, questionable whether this mixer could be utilized in the planned transverter. The required signal of 432 MHz is relatively near to the upper frequency limit, where the conversion loss increases rapidly. However, the matching to 404 MHz oscillator module and the linear amplifier chain for 432 MHz with the aid of tuneable Pi-links ensured the usability of the ring mixer IE-500. As can be seen in the measured values, spurious signals are well suppressed. Of course, the bandpass filter coupling and the screened construction also play an important part. **Fig. 1** shows the author's third prototype.

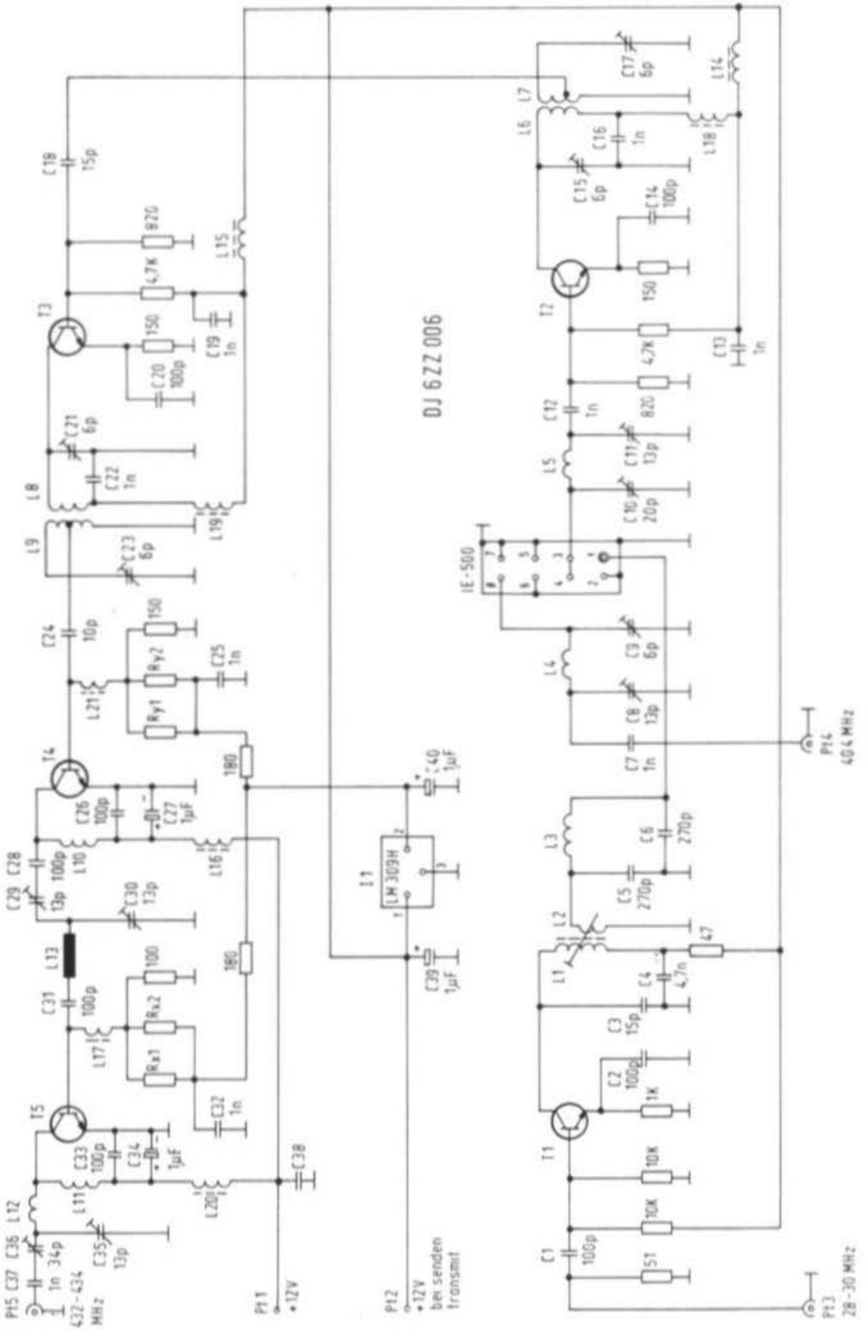

Fig. 2: Circuit diagram of the transmit converter 28 MHz / 432 MHz

2. CIRCUIT DESCRIPTION

The circuit diagram given in **Fig. 2** shows that the described transmit converter does not possess its own local oscillator chain. The required local oscillator frequency of 404 MHz is obtained more favorably from a separate module, e.g. using the local oscillator module DJ 4 LB 003 (2). This can also be used for the local oscillator frequency of the receive converter in order to ensure complete transceive operation. A local oscillator signal of approximately 700 mW into 50Ω (≙ 10 dBm) is required for the transmit converter DJ 6 ZZ 006.

The 28 MHz signal to be converted is fed to connection Pt 3 and should not exceed 100 mV under linear conditions. The signal is amplified, filtered to remove any harmonics and fed to the ring mixer. The required conversion product is filtered out with the aid of a Pi-filter, correctly matched, and fed to the first linear amplifier stage equipped with transistor T 2. Since this stage just compensates for the conversion loss, three further linear amplifier stages equipped with transistors T 3 to T 5 are required in order to obtain the output power of 2 W. Experiments have shown that transistors are required for T 2 and T 3 that possess a very high gain at a power level in the order of several tens of mW. Transistors BF 224 and BFX 62 were not sufficiently good for these stages. The transistor BFX 59 used is a low-power transistor used for UHF masthead amplifiers. The stages equipped with T 2, T 3 and T 4 are coupled together with the aid of bandpass filters, in order to ensure that all unwanted conversion products are not passed to the power transistors T 4 and T 5.

An integrated 5 V voltage stabilizer which feeds the base voltage dividers is used for stabilizing the operating points of driver and output transistors (T 4, T 5).

The collectors of T 4 and T 5 are always connected to the operating voltage. They are blocked in the receive mode with the aid of the base resistors. In the transmit mode, an operating voltage of + 12 V is connected to Pt 2. This ensures that the 5 V stabilizer and the other stages are provided with the required operating voltages.

2.1. Special Components

T 1:	BF 224
T 2, T 3:	BFX 59 (Siemens)
T 4:	C 1-12 (CTC)
T 5:	C 3-12 (CTC)
I 1:	LM 309 (National Semiconductor), SG 309 T (Silicon General)
I 2:	Ring mixer IE-500
L 1:	18 turns of enamelled copper wire wound on 0.3 mm dia. coilformer with green core
L 2:	3 turns of wire as for L 1 on the core side above L 1
L 3:	12 turns of enamelled copper wire wound on a 3 mm former, self-supporting

All other coils from silver-plated copper wire 1 mm dia.

L 4:	1.5 turns inner dia. 5 mm
L 5:	2.5 turns inner dia. 4 mm
L 6, L 8:	1.5 turns inner dia. 4 mm
L 7:	1.5 turns inner dia. 4 mm coil tap approx. 6 mm from cold end

L 9:	1.5 mm turns inner dia. 4 mm coil tap approx. 0.5 turn from cold end
L 10:	4.5 turns inner dia. 5 mm
L 11:	2.5 turns inner dia. 6 mm
L 12:	wire of 37 mm in length, bend according to Fig. 4b
All chokes:	Ferrite beads with several turns of 0.3 mm dia. enamelled copper wire
L 14 ... L 17:	5 turns
L 18 ... L 20:	6 turns
L 21:	7 turns

C 8, C 11: 12 pF plastic foil capacitors 7 mm dia. (Valvo, Dau)
C 9, C 15, C 17, C 21, C 23: 6 pF plastic foil capacitors 7 mm dia.
C 10: 22 pF plastic foil capacitors 7 mm dia.
C 29, C 30, C 35: 13 pF air spaced trimmer (Tronser) with 2 connections
C 36: 34 pF air spaced trimmer (Tronser) with 2 connections

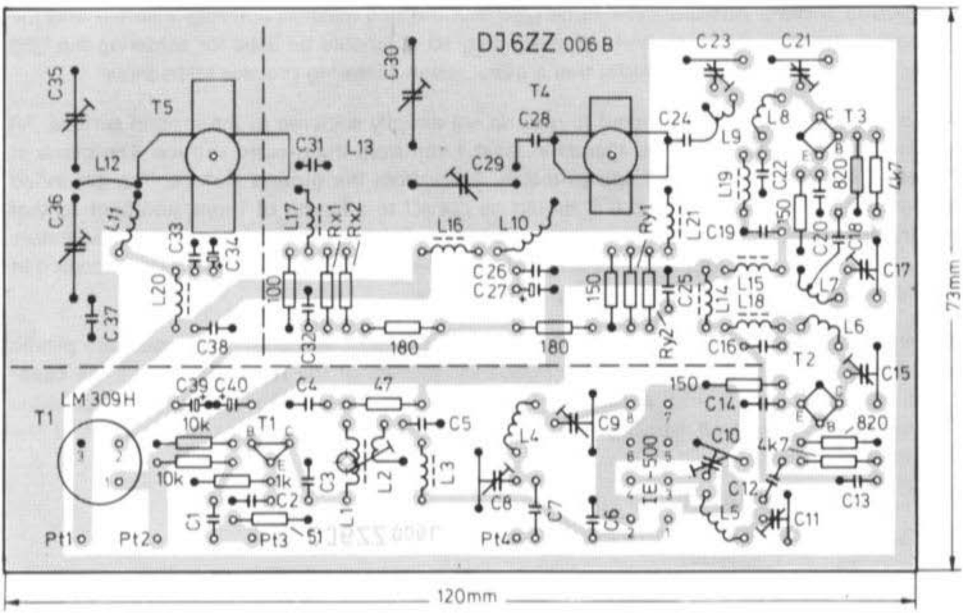

Fig. 3: Component locations and conductor lanes on the lower side of PC-board DJ 6 ZZ 006

3. CONSTRUCTION

All components are accommodated on the double-coated PC-board DJ 6 ZZ 006 whose dimensions are 123 mm x 73 mm. The component locations and the conductor lanes on the lower side of this board are shown in **Fig. 3**. Screening panels of 25 mm in height are required which are firstly soldered together according to the dashed lines and are soldered to the ground surface of the board after all components have been mounted into place. A small cutout should be provided on the small screening plate for the base feedthrough (**Fig. 4a**).

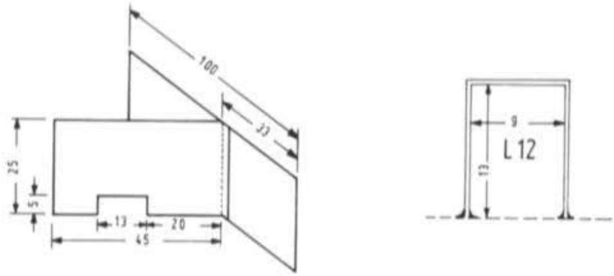

Fig. 4: a) Screening panels; b) inductance L 12

After drilling the holes from the conductor side, the copper coating is drilled out using a larger drill to approximately 2 to 3 mm dia. around the component connections.

The Schottky ring mixer is placed onto the board and the two ends of the case soldered to the ground surface. Attention should be paid that the ring mixer is correctly inserted into the PC-board. A hotter soldering iron (approximately 60 W) should be used for soldering the ring mixer into place, in order to ensure that a quick, clean soldering process takes place.

All components that are connected to ground are directly soldered to the ground surface. All self-supporting coils should be spaced at least 1 mm from the ground surface. The cases of T 2 and T 3 should be spaced approximately 2 mm from the ground surface. The grounded connections of trimmer C 8 and C 9 should be cut off to a length of 1 mm, and bent so that the trimmers are located directly on the PC-board. The connections of trimmer capacitors C 11, C 15, C 17, C 21, and C 23 are only bent on one side since one connection is located in a hole, without conductor lane, to the ground surface.

All trimmer capacitors with the exception of C 29, C 30, C 35, and C 36 (air spaced) are plastic foil trimmers which guarantee a good frequency stability of the resonant circuits. All capacitors are disk types and should be soldered into place with the shortest possible connections, but without causing a short-circuit. The number of turns in the case of the chokes is not critical.

4. COOLING OF THE POWER TRANSISTORS

The driver transistor T 4 (C 1-12) is soldered to a cylinder-headed, 4 mm diameter brass screw. After this, T 4 and the output transistor T 5 are screwed to a common aluminium heat sink. An aluminium block with the dimensions of 25 mm x 85 mm x 5 mm has been found sufficient in continuous operation. This heat sink can be seen on the lower side of the PC-board in **Fig. 5**.

5. CASE

The first prototypes were tested without case and were found to be sensitive to surrounding influences. For this reason, the described prototype was enclosed in a 35 mm high case made from double-coated PC-board material. However, this did not completely cure all these neighbouring effects. This was solved after soldering in the intermediate panels, soldered into place every 15 mm. The supply voltages are fed into the case via feedthrough capacitors, and the RF signals via coaxial sockets.

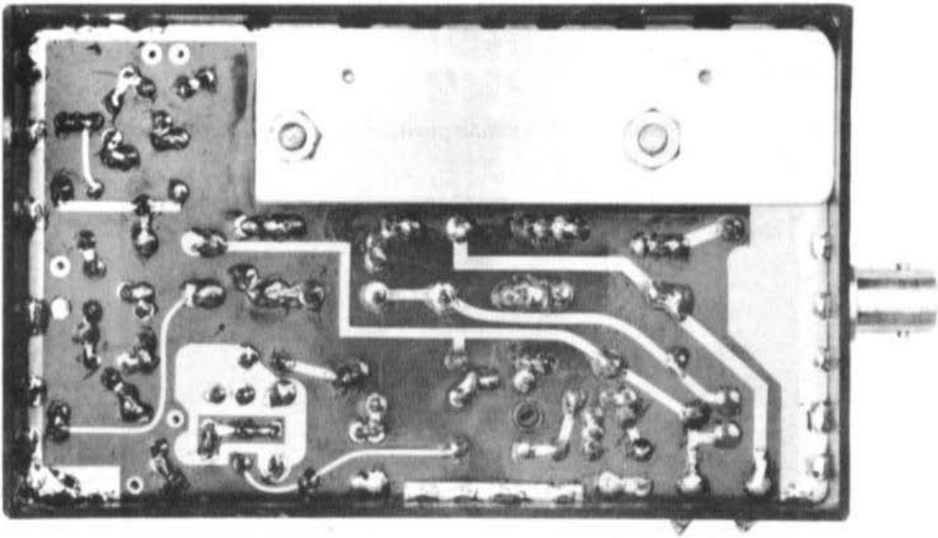

Fig. 5: Lower side of the prototype DJ 6 ZZ 006

6. ALIGNMENT AND INTERCONNECTION

a) Select minimum capacitance for all air-spaced and foil trimmers.

b) Connect an operating voltage of + 12 V to Pt 1 via a mA-meter (500 mA).

c) Connect an operating voltage of + 12 V to Pt 2.

A voltage of + 5 V should be measured at connection 2 of I 1. The value of resistors $R_{X\ 1,2}$ and $R_{Y\ 1,2}$ should be established experimentally to obtain quiescent currents of 30 mA for T 4 and 60 mA for T 5. Since it is sometimes difficult to find the actually required resistance values, and since the transistor characteristics can fluctuate, holes are provided for two parallel resistors each. Orientation values for these resistors are 330 Ω for $R_{X\ 1}$, and 430 Ω for $R_{Y\ 1}$. A voltage of 1.5 V (≙ 10 mA quiescent current) should be measured at the emitter resistors of T 2 and T 3.

d) Place a signal of approximately 80 mV to Pt 3 and a local oscillator signal of approximately 700 mV at Pt 4. Connect a terminating resistor of 50 Ω to Pt 5 and connect to a power meter, or reflectometer.

e) Align L 1 / L 2 and all trimmer capacitors except C 9 alternately to the maximum output signal. This should be done while monitoring the required signal on a receiver at 432 MHz, since it could be possible to align the transverter to 404 MHz or even to 376 MHz. The coil tap on coils L 7 and L 9 should be increased from the cold end towards the hot end during the alignment process, until no increase in output power is achieved.

f) If a receiver is available which allows the local oscillator frequency to be monitored at full output power, it is possible for trimmer capacitor C 9 to be aligned for minimum local oscillator signal. This position should be in the vicinity of minimum capacitance. If this is not possible, C 9 should remain at its point of minimum capacitance.

7. MEASURED VALUES

After correct alignment, the transverter module provides an output power of 2 W on 432 MHz with an input power of 100 mV.

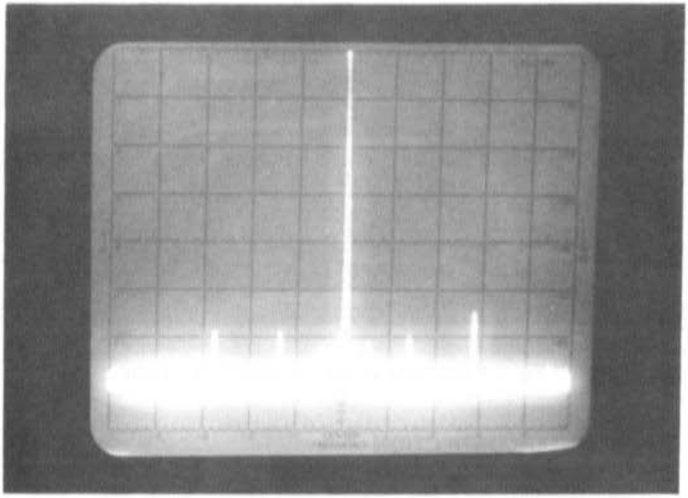

Fig. 6: Output spectrum of the transverter DJ 6 ZZ 006 between 330 and 530 MHz (Vertical 10 dB / div.; horizontal: 20 MHz / div.)

It will be seen in **Fig. 6** that the local oscillator signal (second signal from left) is suppressed by more than 55 dB with respect to the peak output of the required signal. The image rejection (376 MHz) is in the same order. The conversion product with the first harmonic of the carrier (second signal from the right) is also suppressed by 55 dB as is the second harmonic of the carrier (right-hand signal).

Whereas all conversion products are reduced together with the drive signal, the local oscillator signal will increase slightly.

The suppression of the first harmonic of the output signal of an average of 35 dB is relatively poor. This means that a bandpass filter should be used, especially when the output signal is to be fed to a linear amplifier.

This was followed by a two-tone test in order to establish the intermodulation characteristics. These two carriers were fed to the transverter at a frequency spacing of 1.6 kHz and at a power level of 30 mV. The result can be seen in **Fig. 7**. The signal falls off by more than 55 dB at a frequency spacing of 5 kHz from the carrier. In practical operation, it was found that no signal could be established more than 6 kHz from the nominal frequency even when

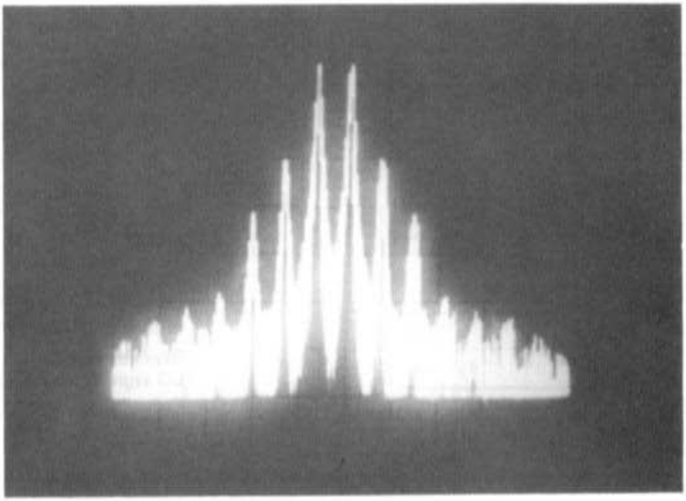

Fig. 7: Intermodulation spectrum of the transverter with two-tone signals (30 mV / 1.6 kHz spacing; vert.: 10 dB/d., horiz.: 2 kHz/d.)

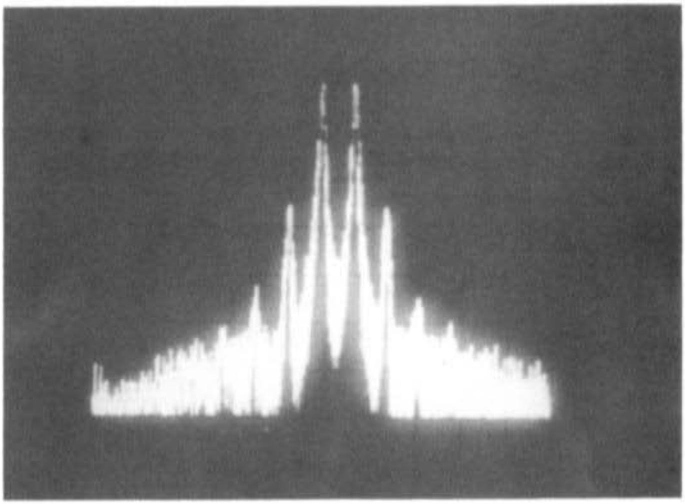

Fig. 8: Intermodulation spectrum of DJ 6 ZZ 006 with two-tone signals (15 mV / 1.6 kHz spacing; vert.: 10 dB / div., horiz.: 2 kHz/div.)

the required signal was more than 60 dB over the noise. Of course, the signal improves still more at lower drive levels (15 mV each): the third order intermodulation products are more than 26 dB weaker than each of the two-tone signals **(Fig. 8)**.

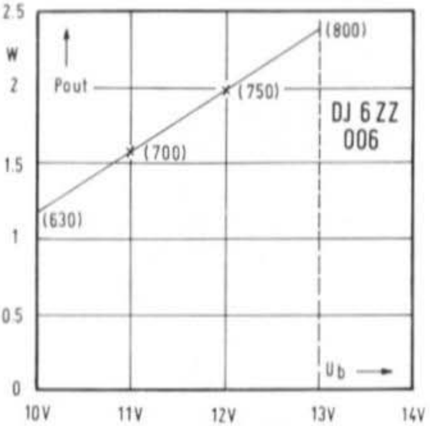

Fig. 9:
Output power as a function of the operating voltage.
Bracket values:
LO signal (404 MHz) in mV

Fig. 9 shows the dependance of the clean output power from the operating voltage. The drive level during this measurement was kept constant at 100 mV (single tone). The operating voltage of the local oscillator module DJ 4 LB 003 was not changed. The output voltage values in mV are given in brackets.

If a greater drive signal than 90 to 100 mV is fed to the transverter, the output signal will be distorted, and will become wider. This shows that the ring mixer is being overdriven.

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DESIGN OF TRANSISTOR FREQUENCY MULTIPLIERS

by Dr. Ing. H. Schierholt, DL 3 ZU

The described design information was obtained during experiments on low-power transistor multipliers. Of course, the information given cannot be generalized, but should provide some useful information when working in conjunction with frequency multipliers in the frequency range between 10 MHz and approximately 150 MHz.

1. EXPERIMENTAL CIRCUIT

The circuit diagram of a frequency multiplier with the associated measuring equipment is given in **Fig. 1**. The test equipment comprises a tube voltmeter URI (Rhode & Schwarz) with RF probe, a measuring receiver and transmitter ASV (Rhode & Schwarz) for 30 - 300 MHz, as well as a calibrated attenuator for Z = 60 Ω of up to 70 dB. A fast switching transistor (f_T approx. 400 MHz) with large reactive capacitance (similar to BSY 18 or 2 N 708) was used on purpose since an experimental circuit that is able to operate using such a transistor would also be uncritical with respect to modifications of the mechanical construction and transistor type. However, this does not mean that such transistor types should be used for frequency multipliers instead of true RF types such as the 2 N 918.

All resonant circuits were accommodated in screening cans (coil sets type D 3 with core and two caps). However, it was found that the design rules were also valid for simple screened coils with core but without caps.

In order to make it suitable for various types of drive, inductance L 2 (40 % of the turns on L 1) was provided with a center tap and built up in a strictly, balanced manner. This allowed a push-pull doubler to be obtained. In addition to this, drive via a capacitive tap was also tried.

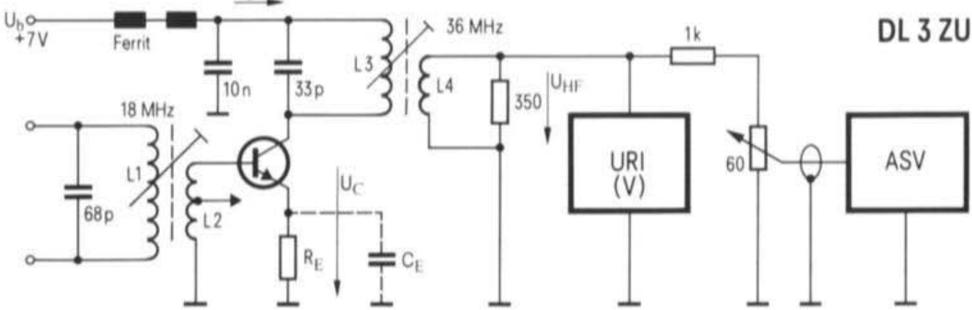


Fig. 1: Experimental circuit of a frequency multiplier

The drive circuit not shown here was constructed from a 9 MHz crystal-controlled oscillator with buffer, a phase modulator (without AF-signal), and a doubler from 9 to 18 MHz. This doubler was not fully driven at the collector from the somewhat weak signal of the phase modulator, which means that the formation of higher harmonics is very low and will not interfere with the measurement, as will be explained later. At the 36 MHz output of such a circuit, virtually all harmonics of 18 MHz can be expected, as well as harmonics of 9 MHz at a lower level.

2. THE MEASUREMENTS

A conventional doubler stage was firstly constructed. The base was provided with 0 V bias voltage and was driven by the whole winding L 2 (40 % of the turns of L 1). The emitter resistor R_E that was selected (68 Ω) was bypassed as usual using a capacitor. The value was C_E = 680 pF (series resonance at approx. 18 MHz).

This circuit seemed to be stable, but the tuning of the collector circuit comprising L 3 for maximum voltage on the VTVM resulted in a different adjustment thán when aligning for maximum output at 36 MHz on the measuring receiver. This indicated strong harmonics and spurious signals. The measuring receiver indicated mainly that non-harmonic spurious signals occured, although the RF drive voltage in conjunction with the threshold voltage of the transistor should have resulted in a suitable current flow angle for stable frequency doubling. Different values of the RF drive voltage, different taps, and different base bias voltages did not improve the instable operation.

However, operation was satisfactory as soon as the emitter capacitor C_E was removed. It was now possible by suitable selection of an emitter resistor to obtain a favorable current flow angle and a favorable collector drive without formation of non-harmonic spurii. The harmonic rejection could also be classed as being good.

Various other types of drive were used: capacitive coupling to the 18 MHz circuit comprising L 1, various base bias voltages (even with diode stabilization), as well as a push-pull doubler. However, none of them represented any particular advantages over the simple circuit given in Fig. 1, that requires a very low number of components.

3. RESULTS

As has been previously mentioned, no emitter capacitor should be provided. It was then found that the output matching mainly determines the suppression of harmonic and spurious signals, as well as the efficiency of the doubler. This means that the load resistance of the collector circuit, including its loading by the subsequent stage must be designed so that the collector voltage is driven just up to the limit characteristics but not overdriven.

In the described case, the RF voltage at the collector should not exceed 4 to max. 5 V (measured with the URI), otherwise strong harmonics could be measured up to 300 MHz.

3.1. Measured Values

The circuit given in Fig. 1 with L 2 = 40 % of L 1, without bias voltage, and without emitter capacitor provided the following measured values:
(All values referred to the level of the required frequency 36 MHz)

27 MHz:	not measurable
45 MHz:	not measurable
54 MHz:	- 34 dB
72 MHz:	- 36 dB
90 MHz:	- 39 dB
108 MHz:	- 39 dB

Measuring conditions:

I_C	= 2.3 mA
u_{HF}	= 1.15 V (at 350 Ω)
u_C	= 5 V (effective value)
R_E	= 150Ω

Higher harmonics could not be determined.

4. REALIZATION

The required optimum output matching cannot always be achieved in practice. Since the loading of the collector circuit by the next stage is difficult to calculate, it is assumed to be given and the emitter resistor is changed so that the optimum drive of the collector voltage is obtained. For example, an emitter resistor of 150 Ω was found to be favourable in the experimental circuit for the 350 Ω load resistor. On reducing the load resistance to 110 Ω, it was possible to reduce the emitter resistor to 68 Ω.

The measurement of the RF voltage at the collector causes a certain problem if this successful method is to be used. This must be made with a low-capacitive oscilloscope or VTVM probe. When measuring in conjunction with a voltmeter, the RMS value of u_C should be somewhat less than $0.7 \times U_b$. In the case of an oscilloscope, the peak-to-peak value of the collector voltage should be somewhat less than $2 \times U_b$. The resonant circuit should be tuned for maximum voltage with the aid of the probe.

During the experiments (36 MHz), the 2 pF probe of the URI was used. It was even possible to tune in a 7 pF oscilloscope probe, but a 10 pF probe caused difficulties.

If the above measuring equipment is not available, the following method can be used:
The collector current of the subsequent class B or class C stage represents a reliable RF voltmeter, which although it cannot indicate any absolute values, will not falsify the measuring set-up since the original load conditions are prevalent. It is advisable to dampen the collector circuit of this »indicating circuit« with approximately 100 Ω so that no internal characteristics (e.g. instability) of this stage is able to falsify the measurement.

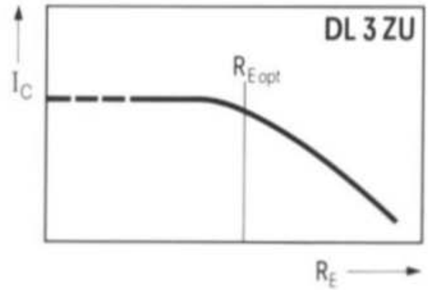


Fig. 2:
Collector current of the subsequent stage as a function of the emitter resistor of the multiplier

The collector current of this stage corresponds to the output voltage of the doubler to be aligned. It is dependent on the emitter resistance R_E of the doubler according to the same principle as shown in **Fig. 2**. When the collector of the doubler is fully driven, the current will not continue to increase to any extent. However, the limit is not very sharp, and the most favorable emitter resistance is to be found before reaching this limit.

Since each subsequent stage has an effect on the drive of the previous stage, it is advisable for the alignment to be made from the lowest towards the highest frequency of a multiplier chain, at the same time correcting the value of the emitter resistances from the output back to the oscillator. The final alignment is, however, not especially critical.

5. NOTES

The measuring set-up in Fig. 1 is somewhat primitive, which means that the measured values cannot be classed as being extremely accurate. However, the difference between the output spectrum when using the same circuit with overdriven collector is considerable.

ESTIMATING THE SIGNAL-TO-NOISE RATIO OF AN ATV-LINK

by R. Lentz, DL 3 WR

The following short article allows the si ıal-to-noise ratio of an amateur TV link to be estimated based on the signal strength of a SSB signal. A comparison between SSB and ATV signal strength allows a more accurate estimation of whether such communications would be possible, and with which signal-to-noise ratio than when calculating the signal-to-noise ratio from the path loss.

The path loss calculation under non-line-of-sight conditions, which are usually prevalent in amateur radio, can only establish the order of magnitude. On the other hand, the comparison to a known SSB signal strength on the same frequency band also includes both the specific conditions at both locations, and also the actual conditions at that moment.

1. PREREQUISITES

SSB, and TV communications (CCIR/B) are to take place in the same frequency band.

The same antenna and antenna cables should be used, or the difference in gain or loss should be taken into consideration during the calculation.

The transmit power (PEP in the SSB mode, or synchronizing level in the ATV mode) should be approximately equal. Larger variations should be taken into consideration during calculation.

The noise figures of the receivers should be approximately equal. Larger variations should also be taken into consideration.

2. FUNDAMENTAL DIFFERENCE DUE TO THE INCREASE OF THE BANDWIDTH

The sensitivity limits of a receive system (1) can be determined according to the following equation when the same antenna and cables are used:

$$P_n = F \times k \times T_o \times B$$

with:
- P_n = noise power of the receiver (W)
- F = noise factor (not in dB)
- k = Boltzmanns constant = 1.38×10^{-23} Ws/K
- T_o = ambient temperature = 290 K
- B = bandwidth of the IF module

If the noise factor F is the same for both modes, it is the bandwidth B that is the determining factor.

A bandwidth of 5.75 MHz (2) has been established for ATV, and 2.4 kHz (crystal filter) for SSB. The following, higher power requirements for ATV can be calculated as follows:

$$\frac{B_{ATV}}{B_{SSB}} = \frac{5750\ kHz}{2.4\ kHz} = 2396 \triangleq 33.8\ dB$$

This means that an ATV signal-to-noise ratio of 0 dB is to be expected when the SSB signal strength is approximately 34 dB over the noise. At a SSB S/N ratio of more than 60 dB, the ATV signal-to-noise ratio will be approx. 30 dB. A noise-free ATV signal is obtained at approx. 40 dB, which can be expected when a SSB S/N ratio of approx. 75 dB is available.

2.1. Difficult conditions

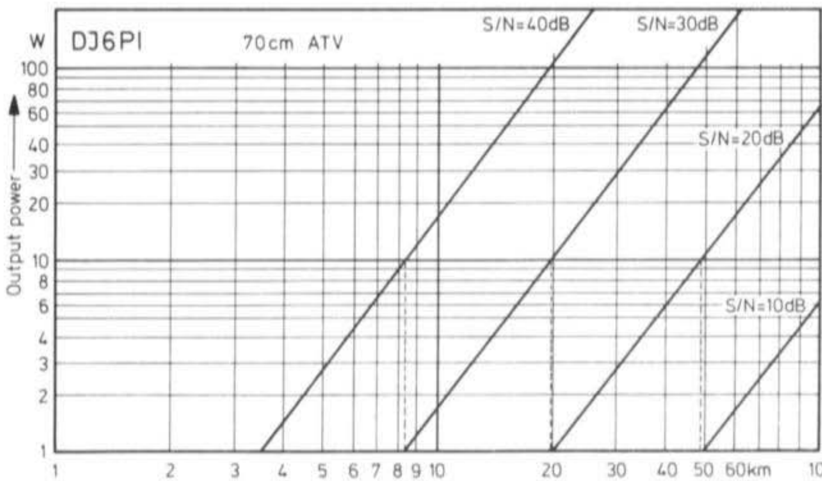
Whereas the previous calculation only considers the bandwidth difference and thus only the major disadvantage of ATV, further difficulties must be taken into consideration in practice:

If an old, modified tuner with a noise figure of 10 dB is used for reception, whereas a good converter with approx. 4 dB is used in the SSB mode, the ATV link will have a further disadvantage of 6 dB bringing the total up to 40 dB.

If the SSB link is made with a peak power of 400 W, and only 10 W are available for the ATV transmission, a further 16 dB disadvantage will be present for the ATV link.

Whereas a low-noise voice link is available at a signal-to-noise ratio of 20 dB, a noise-free TV picture will not be available until the signal-to-noise ratio is in the order of 40 dB (3), (4).

These considerations show that all possibilities should be used to the full for long distance (wideband) television links.



Range of an ATV station as a function of the transmit power

3. RANGE UNDER LINE-OF-SIGHT CONDITIONS

For completion, a diagram taken from (3) has been included to indicate the range of an ATV station as a function of the transmit power under the following conditions:

Line-of-sight or virtual line-of-sight conditions over hilly countriyside and suburban areas, but not over-the-horizon communications.

The following conditions are assumed to be present at both stations: 15 dB antenna gain, 2.5 dB cable loss, 7.8 dB noise figure, frequency range: 70 cm band. The range of a 10 W transmitter with various noise ratios as parameter is given in the form of dashed lines. It will be seen that 10 W will only have a range of 20 km inspite of the high total antenna gain of 30 dB, even when a slightly noisy picture (S/N = 30 dB) is acceptable.

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Modification of the STE Receiver ARAC 102 for Reception of the OSCAR Satellites in the 10 m and 2 m Band

by AMSAT Newsletter 1/1976

The receiver ARAC 102 manufactured by STE in Milan is a small, complete receiver with built-in loudspeaker. The receiver is designed to operate from a voltage of 12 V and is designed for AM, FM, and SSB/CW reception in the frequency ranges from 28 to 30 MHz, and 144 to 146 MHz. The receiver is further equipped with an automatic noise limiter, and squelch.

The tuning range of 2 MHz is somewhat too wide, when the receiver is only to be used for reception of the satellites OSCAR 6 and OSCAR 7. However, the 28 - 30 MHz receive module (AR 10) can be easily modified so that the tuning range is limited to 29.38 to 29.55 MHz, or 145.8 MHz to 146.00 MHz. The following modifications are required:

Three wire bridges designated with A, B, and C are located on the lefthand side of the triple variable capacitor of module AR 10. These bridges should be disconnected and a 5 pF capacitor soldered into these positions. This can usually be achieved without removing the board. The three capacitors should be ceramic disk types.

Trimmer capacitors C 3, C 8, and C 9 located adjacent to these bridges, and inductance L 6 (between C 9 and the crystal) are aligned for maximum gain across the new tuning range of 29.38 to 29.55 MHz.

The 38.6667 MHz crystal provided in the 2 m converter module AC 2 should be exchanged for a HC-25/U, series resonance crystal of 38.833 MHz. This ensures that a frequency of 145.9 MHz is converted to 29.4 MHz.

A UNIVERSAL CONVERTER FOR HF AND VHF

by J. Kestler, DK 1 OF

A large number of amateurs only licenced for operation on the VHF bands and up possess high quality transceivers or receivers for the frequency range of 144 to 146 MHz, which are usually equipped for several modes (AM, FM, SSB, and CW). The following article is to describe a receive converter which allows complete coverage of the frequency range of 50 kHz to 30 MHz in sections of 2 MHz in bandwidth and converting them to the 2 m band. This allows the excellent characteristics of modern VHF equipment such as high selectivity, good large-signal handling capabilities and high stability, as well as often digital readout to be used to cover the large, continuous range from longwaves to the highest shortwave frequency. The quality of reception is mainly dependent on the VHF equipment, and of course on the antenna used.

1. CONCEPT

In order to reduce the mechanical work to a minimum, no tuned input circuits are used. This solves a large number of problems that would be encountered with respect to range switching. As can be seen in the block diagram shown in **Fig. 1**, the input signal from the antenna is fed via a lowpass filter (cut-off frequency: 32 MHz), and fed directly to the mixer which is equipped with a Schottky ring mixer. The output frequency is passed via the subsequent bandpass filter and is fed to the 2 m receiver. This filter also ensures that the mixer is loaded with its nominal impedance. This is very important with respect to optimizing the large-signal handling capabilities (1), (5).

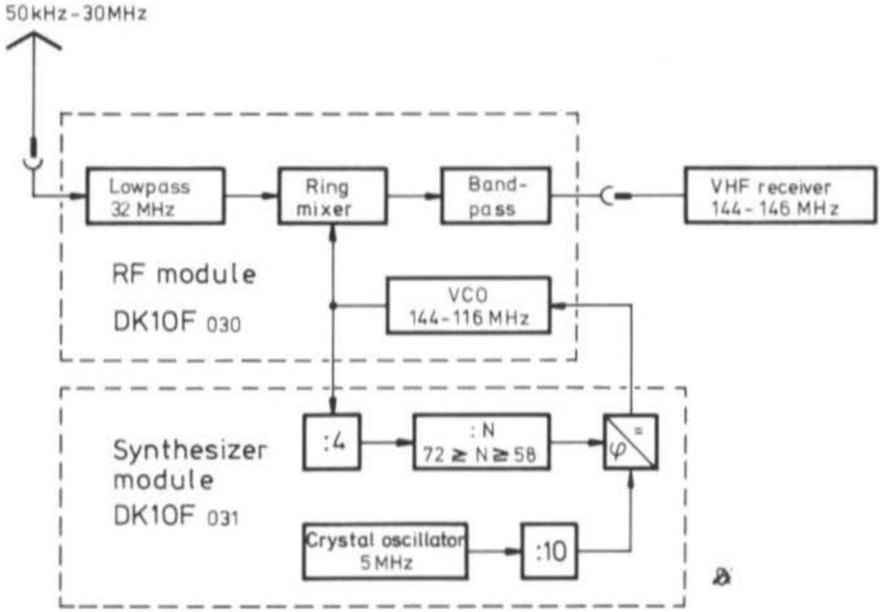


Fig. 1: Block diagram of the HF receive converter for 2 m receivers

The sensitivity threshold of the receive system is thus mainly dependent on the noise figure of the VHF receiver (usually in the order of 3 dB), and from the conversion loss of the ring mixer including matching losses which amount to a total of approximately 8 dB. This results in a total noise figure of approximately 11 dB. This value is completely satisfactory for the short wave range, since antenna noise, atmospheric and electrical interference are usually higher (2).

In order to increase sensitivity (at the cost of reducing the large-signal handling capabilities), it would be possible to use a tuned preamplifier for the upper frequency range (20 to 30 MHz).

The local oscillator of the converter must be variable in steps of 2 MHz in the range of 144 to 116 MHz. If a conventional crystal oscillator were to be used, a total of 15 crystals would be required. With the aid of the PLL-technology used here (3), (4), it is possible for only one single crystal to be used. The main oscillator (VCO) oscillates directly at the final frequency required for the mixer. The oscillator signal is then divided by four and finally fed to an adjustable frequency divider which divides it by factor N. The frequency range selection is made in this manner. The output signal of the N-divider is fed to the phase detector where it is compared to a frequency of exactly 500 kHz (crystal-controlled frequency of 5 MHz divided by 10). The output DC voltage of the phase comparator is used to tune the VCO with the aid of its varactor diode to the exact frequency required (selected by N).

2. CIRCUIT DESCRIPTION

2.1. RF circuit

A detailed circuit diagram of the RF circuit is given in **Fig. 2**. A series circuit of two RF chokes is connected across the antenna input (Pt 301) to protect the input circuit from static charges (thunderstorm). This is followed by a four-stage lowpass filter (π-circuit) comprising inductances L 301 to L 304. This filter has a cutoff frequency of 32 MHz. In order to guarantee good filter characteristics up in to the UHF-range, the capacitances are divided over two capacitors of 100 pF each. In this manner, the unwanted effect of the connection inductance is kept as small as possible. Ceramic disk capacitors should be used here which should be soldered into place with the shortest possible connections. The measured attenuation curve of the lowpass filter is given in **Fig. 3**.

The input filter is followed by a matching link comprising L 305, R 301 and C 301. This link has the task of terminating the input of the Schottky ring mixer for frequencies in excess of the passband range of the input lowpass filter.

It should also be mentioned that the RF signal to be received is not fed to the RF input of the mixer, but to the IF connection, since it is only this input that can be used down to very low frequencies. The two other connections possess a lower frequency limit of 5 MHz due to the built-in transformers.

The resulting sum frequency is fed from the RF connection of the Schottky mixer via a band-pass filter (L 306, C 302, and R 302, L 307, C 303) to the output of Pt 302. The component values for the matching links were calculated according to the equations given in (5).

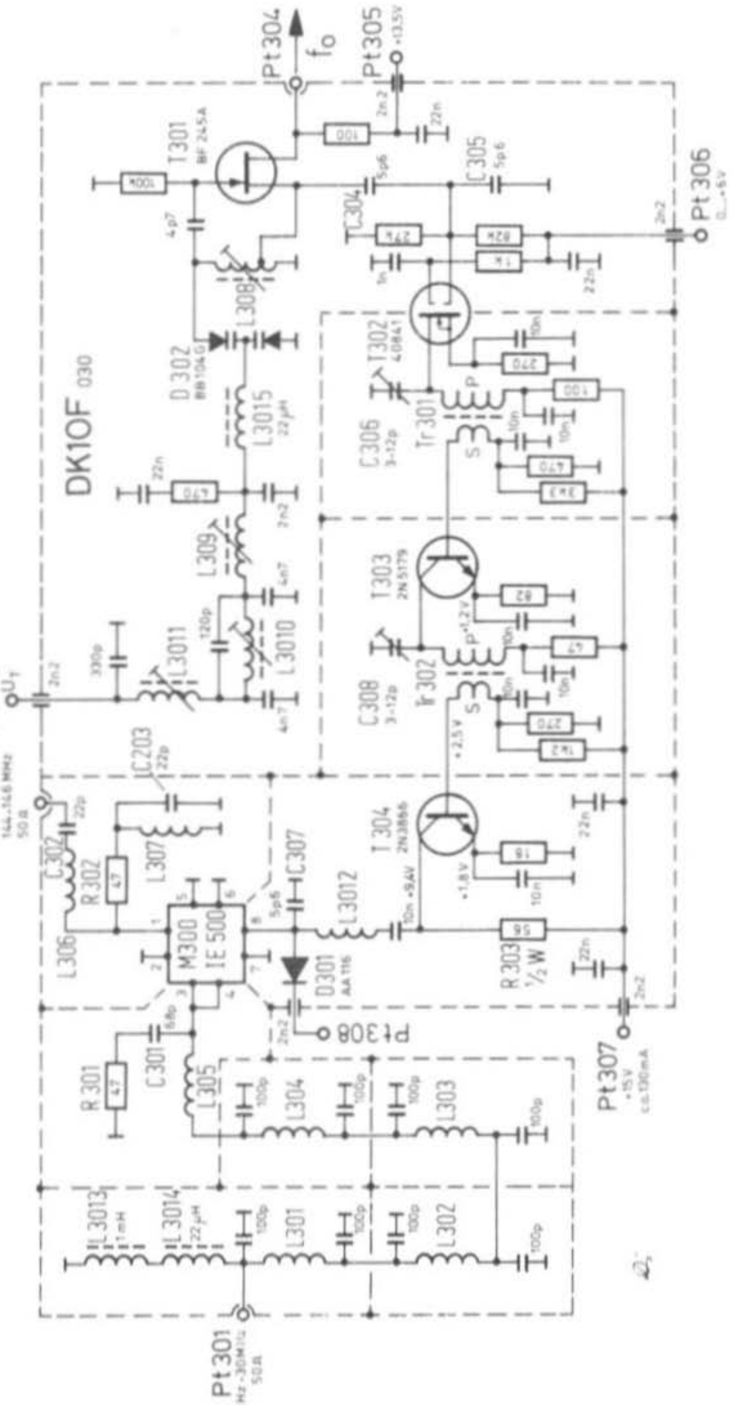


Fig. 2: Circuit diagram of the HF-converter (RF-module DK 1 OF 030)

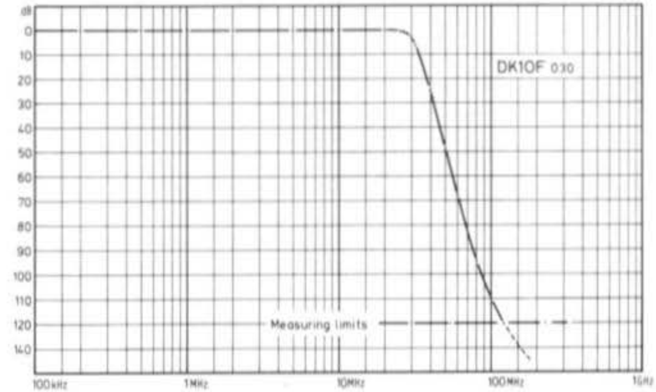


Fig. 3: Attenuation curve of the input filter

The VCO is equipped with transistor T 301, and is variable with the aid of varactor diode D 302 in the range of 116 to 144 MHz. Some of the RF output power is tapped off at the drain circuit and fed via Pt 304 to the synthesizer circuit. The tuning voltage fed to Pt 303, must be passed via a multiple stage filter link (L 309, L 3010, L 3011), to ensure that no residual phase comparator frequency is present. The buffer stage T 302 is connected via the voltage divider C 304 / C 305. The gain of this MOSFET stage can be varied with the aid of the control voltage fed via Pt 306, which ensures a simple automatic level control of the oscillator power.

Transformer Tr 301 is provided in the drain circuit of the buffer stage. The primary of this transformer is tuned to resonance with the aid of trimmer C 106. This is followed by the driver equipped with transistor T 303, and a further transformer (Tr 302) for coupling to the output stage of the local oscillator (T 304).

The output stage T 304 operates in class A (collector DC current approx. 100 mA). This stage is able to provide an output power of 10 mW at low distortion to the ring mixer. The matching link comprising R 303, L 3012 and C 307 with a cutoff frequency in excess of 500 MHz, ensures the required 50 Ω termination. Diode D 301 generates a DC-voltage proportional to the amplitude of the oscillator signal, which is fed via Pt 308 to the control amplifier in module DK 1 OF 032 **(Fig. 7)**.

The power supply for the oscillator amplifier is connected via Pt 307. The VCO is provided with an extra carefully filtered operating voltage via connection Pt 305.

2.2. Synthesizer

Figures 4 to 6 show the circuit diagram of the synthesizer. Connection Pt 311 is provided with the oscillator signal taken from Pt 304 of the RF-module. This is followed by a buffer stage (T 311) equipped with a MOSFET. This stage is provided to isolate the digital circuitry from the VCO. Due to the relatively large bandwidth required for the oscillator frequency (116 to 144 MHz), the buffer is connected via a transformer (Tr 311) to the amplifier stage T 312. The primary winding forms a resonance circuit at approximately 130 MHz together with the output capacitance of the MOSFET. Due to the loading of the input impedance of T 312, the required bandwidth of virtually 30 MHz is obtained without difficulties.

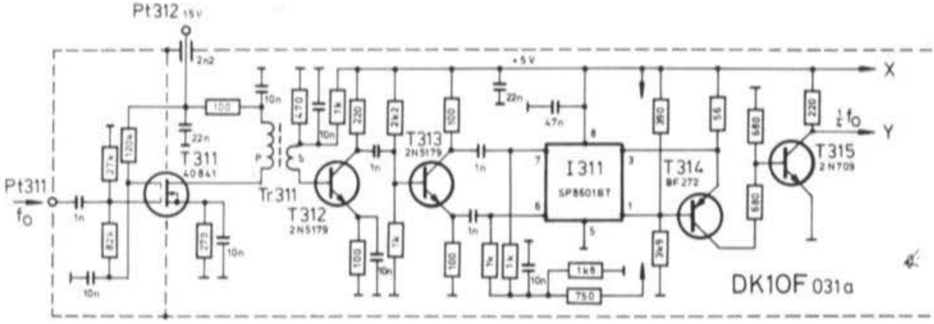


Fig. 4: Circuit diagram of the HF-converter
(Buffer, pre-divider, level converter)

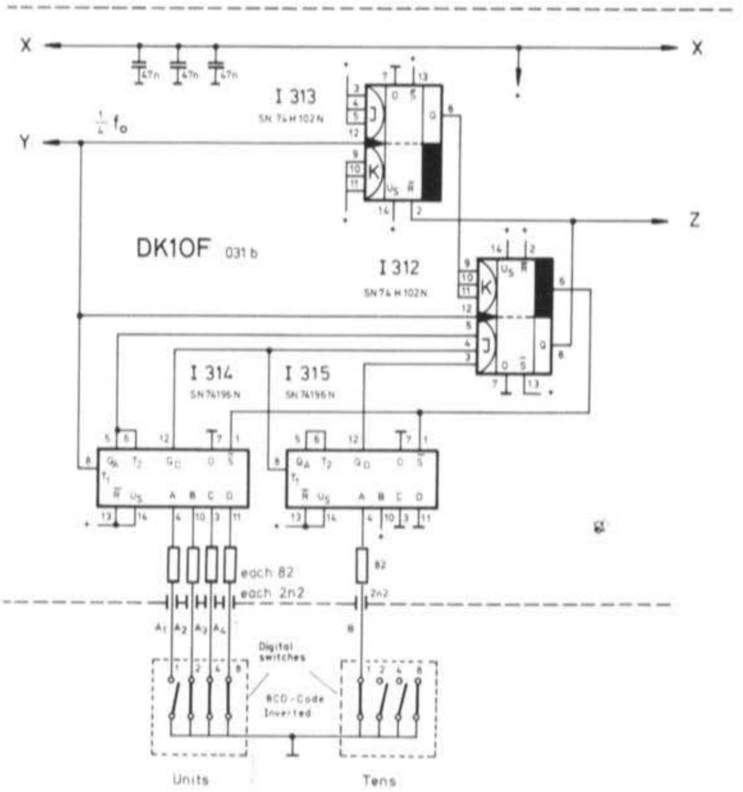


Fig. 5:
Circuit diagram of the HF-converter
Variable divider with digital switches)

The subsequent circuit comprising T 313 as phase reversal circuit, I 311 (ECL-divide by 4), and transistors T 314 and T 315 (level conversion ECL/TTL) was described in detail in (4). A TTL-compatible signal of one fourth of the oscillator frequency is available at point Y. It is fed to the variable frequency divider, whose circuit diagram is given in **Fig. 5**. This comprises a two-stage decade counter (I 314 and I 315) which is adjustable using the digital switches via inputs A 1 to A 4 and B. When the tens-position (I 315) has reached the value 8 and the units-position (I 314) has attained the value 9, the J-input of the flipflop I 312 will be released. The next input pulse will place it into its working position and reset flipflop I 313. At the same time, the decade counter will be switched off and charged again with its commencement value. The following clock pulse switches I 313 so that the K-input of I 312 is released. The next input pulse switches I 312 back to its rest position which switches the counter on again so that a new cycle can commence. The divided oscillator frequency is taken from the output of I 312 and fed to the phase comparator circuit via point Z.

The following table shows the relationship between channel selection and the oscillator, or receive frequency of the converter.

Selection	B	A_4	A_3	A_2	A_1	Oscillator frequency (MHz)	Receive frequency with IF = 144 - 146 MHz
00	L	L	L	L	L	144	0 - 2
01	L	L	L	L	H	142	2 - 4
02	L	L	L	H	L	140	4 - 6
03	L	L	L	H	H	138	6 - 8
04	L	L	H	L	L	136	8 - 10
05	L	L	H	L	H	134	10 - 12
06	L	L	H	H	L	132	12 - 14
07	L	L	H	H	H	130	14 - 16
08	L	H	L	L	L	128	16 - 18
09	L	H	L	L	H	126	18 - 20
10	H	L	L	L	L	124	20 - 22
11	H	L	L	L	H	122	22 - 24
12	H	L	L	H	L	120	24 - 26
13	H	L	L	H	H	118	26 - 28
14	H	L	H	L	L	116	28 - 30

Fig. 6 shows the remaining stages of the synthesizer. A 5 MHz crystal controlled oscillator comprising T 316 with subsequent pulse shaper (T 317) provides the reference signal, which is divided by 10 in I 317 and compared to the divided oscillator frequency in the phase detector I 316. Transistor T 318 then amplifies the DC-voltage proportional to the phase difference between the two frequencies, after which it is fed to the VCO as tuning voltage via L 311 and Pt 315. The following operating voltages are required: + 13.5 V (3 mA), and + 5 V (approx. 250 mA).

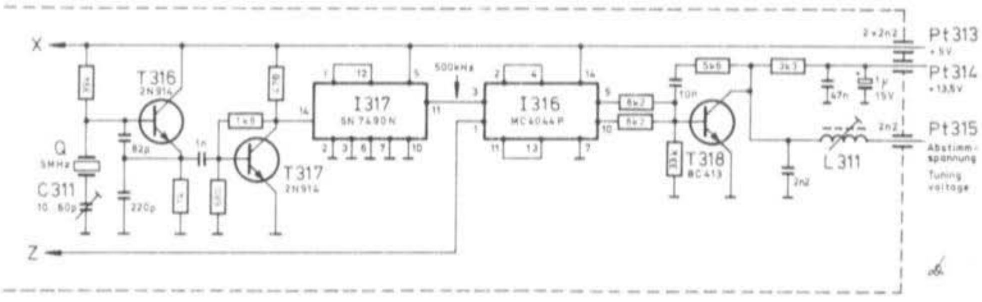


Fig. 6: Circuit diagram of the HF-converter (Reference oscillator, phase comparator)

2.3. Power supply

The circuit diagram of the power supply is given in **Fig. 7**. The power transformer possesses two 12 V windings, and such transformers are easily available on the market. Two DC-voltages are obtained with the aid of the bridge rectifier G 321: one is fed via Pt 324 to the voltage stabilizer I 322 and is stabilized to + 15 V; the other voltage is taken from the center tap of the transformer and feeds the stabilizer I 323 which is provided for the 5 V supply of the TTL-circuits. The supply voltage for the VCO and the DC-amplifier can be taken via a filter link of 180 Ω / 3.3 µF subsequent to the phase comparator circuit (Pt 326).

However, since this voltage should be well filtered, the 3.3 µF capacitor should not be an electrolytic type, but have a plastic dielectric.

The rest of the circuit shown in Fig. 7 comprises a control amplifier for the oscillator output power. The DC-voltage provided by diode D 301 (Fig. 2), which is proportional to the AC-voltage of the oscillator at connection 8 of the Schottky mixer, is fed via connections Pt 308 and Pt 329 to the gate of T 321. The RC-combination of 1 MΩ / 15 kΩ / 1 µF ensures the stability of the control circuit. Transistor T 321 is provided as source follower and ensures that the DC-voltage potential is increased to approximately + 5 V so that the following operational amplifier I 321 can be driven correctly. The nominal value of the diode voltage can be adjusted with the aid of trimmer resistor R 321, which also allows any variation of the pinch-off voltage of T 321 to be adjusted. Integrated circuit I 321 is not provided with feedback via resistors, but operates at full gain for the control loop. The saturated output voltage of I 321 is in the order of 2 V, and is compensated with the aid of the threshold voltage of the three diodes in series at the output, so that 0 V can be obtained at Pt 327. Connection Pt 327 is connected to Pt 306 which ensures that the gate bias voltage of the MOSFET of the T 302 is controlled so that the rectified oscillator voltage at Pt 308 always possesses its most favorable value.

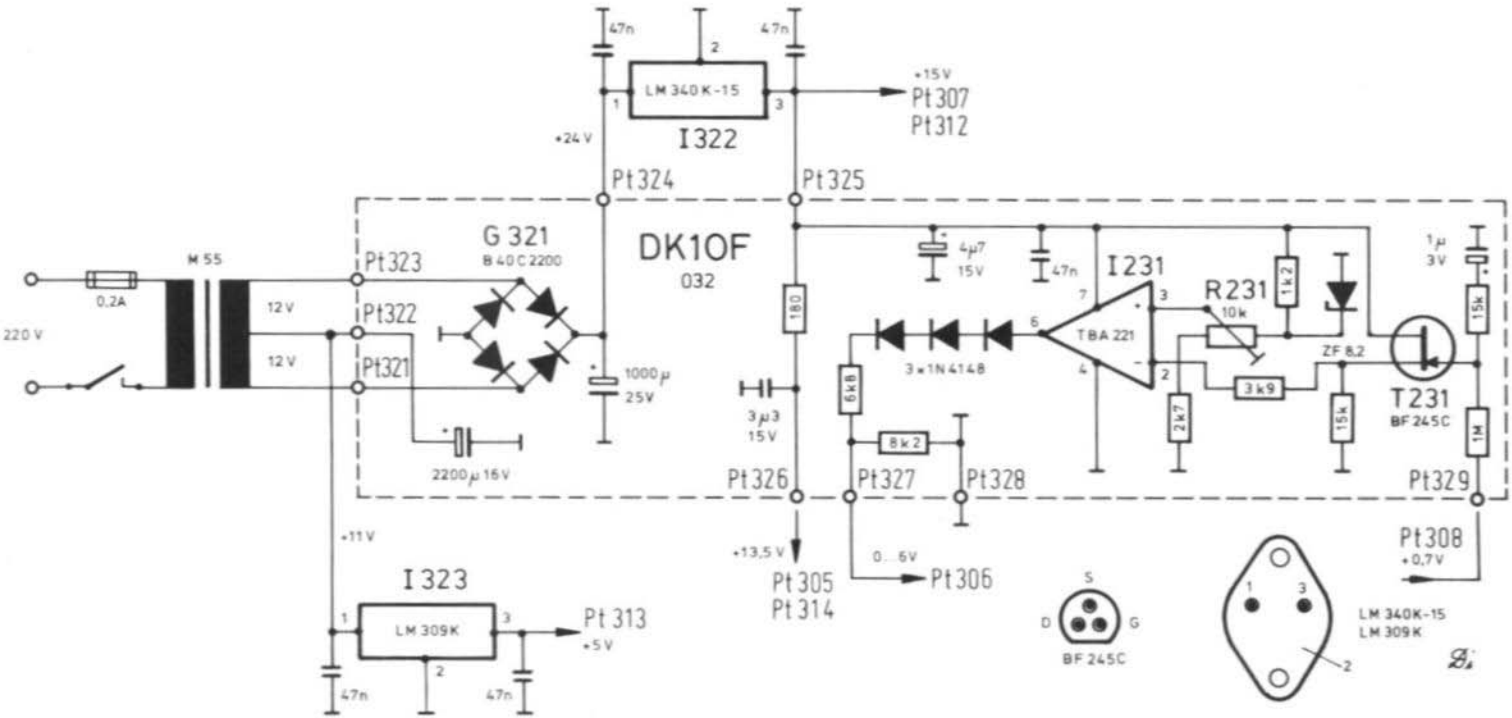


Fig. 7: Circuit diagram of the power supply with control amplifier for the SW-converter

3. CONSTRUCTION

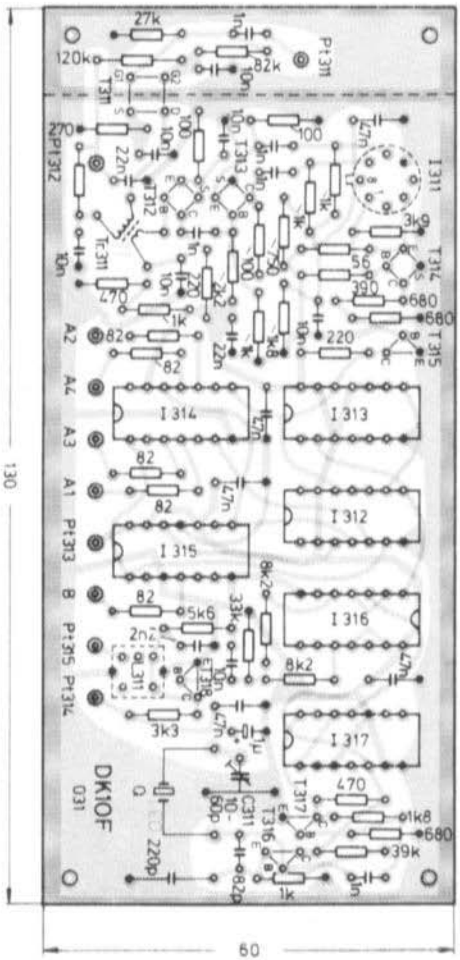
PC-boards DK 1 OF 030, 031, and 032 have been designed for construction of the converter. Boards 030 and 032 are single-coated, whereas 031 is double-coated and provided with through-contacts. **Figures 8, 9 and 10** give the component locations on the PC-boards. **Figures 11 and 12** show photographs of the author's prototype. The RF and synthesizer modules are provided with approximately 30 mm high screening panels made from 0.5 mm thick brass or tin plate. The feedthrough capacitors and coaxial sockets for Pt 301 and Pt 302 are soldered into place. The spacing between the lower side of the board and the lower edge of the screening panel should amount to approx. 5 mm.

It is most favorable for the RF module (030) to be mounted on one side of the metal chassis, and the synthesizer module (031) on the other side. It is also advisable for Pt 304 and Pt 311 to be located virtually adjacent to another. The interconnection between these two points can either be made using a thin coaxial cable (approx. 5 cm long) or using an insulated wire, which must be within the screening. In this case, a capacitor of approx. 5 pF should be connected in parallel to Pt 311.

3.1. Special components

RF-module DK 1 OF 030

M 300:	Schottky ring mixer IE-500
T 301:	BF 245 A (TI) or W 245 A (Siliconix)
T 302:	40841 or 40673 (RCA)
T 303:	2 N 5179 (RCA), also BF 224 (TI) applicable
T 304	2 N 3866 (different producers) with cooling fins
D 301:	AA 116 or similar Ge-diode
D 302:	BB 104 G (green), different producers
Tr 301, Tr 302:	primary 3 turns, secondary 1 turn of 0.4 mm enamelled copper wire in potted core, 9 dia. x 7 or 11 dia. x 7, material K 12, A_L = 16; (B 65531 - L 0016 - A 012); directly glued to the PC-board; coilformer with 2 chambers, or use insulated wire for the secondary;
L 309, L 3010, L 3011:	IF circuits 455 kHz, capacitor removed, yellow marking;
L 301 - L 304:	15 turns of 1 mm silver-plated copper wire, self-supporting, 6 mm inner dia.
L 305:	6 turns, other details as L 301
L 306, L 307:	3 turns, other details as L 301
L 308:	4 turns of 1 mm silver-plated copper wire, turns spaced 2 mm, tap 0.75 turns from the ground end; Trolitul former 6 mm dia. with VHF core (brown)
L 3012:	1.5 turns, other details as L 301
L 3013:	approx. 1 mH, 100 turns of 0.15 mm enamelled copper wire in potted core 14 dia. x 8 mm, material M 25, A_L value 100 (B 65541 - K 0100 - A 025) or 79 turns for material N 22, A_L-value 160 (B 65541 - K 0160 - A 022)
L 3014, L 3015:	ferrite choke, 22 µH, spacing 10 mm
C 306, C 308:	plastic foil or ceramic disk trimmers 10 mm dia., 3 to 12 pF



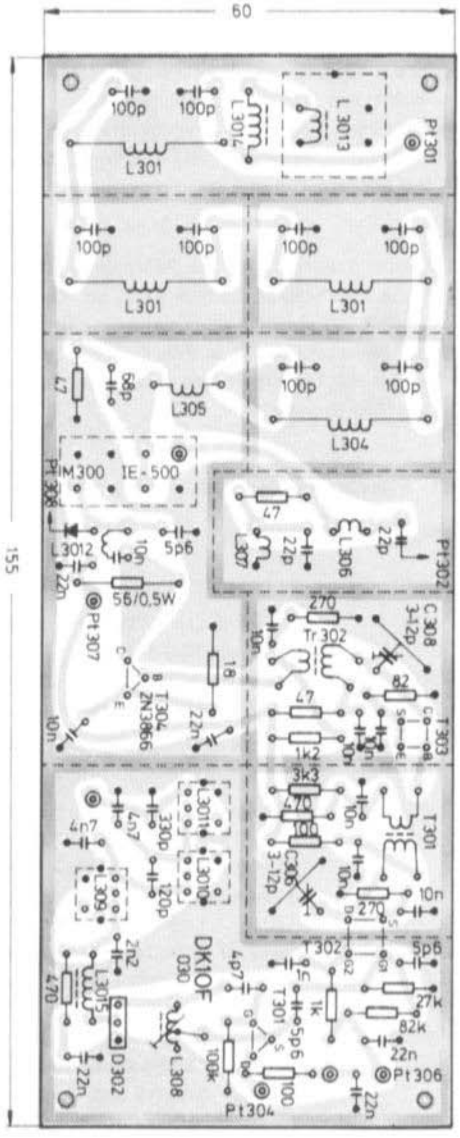


Fig. 8: Component locations on PC-board DK 1 OF 030

Fig. 9: Component locations on PC-board DK 1 OF 031

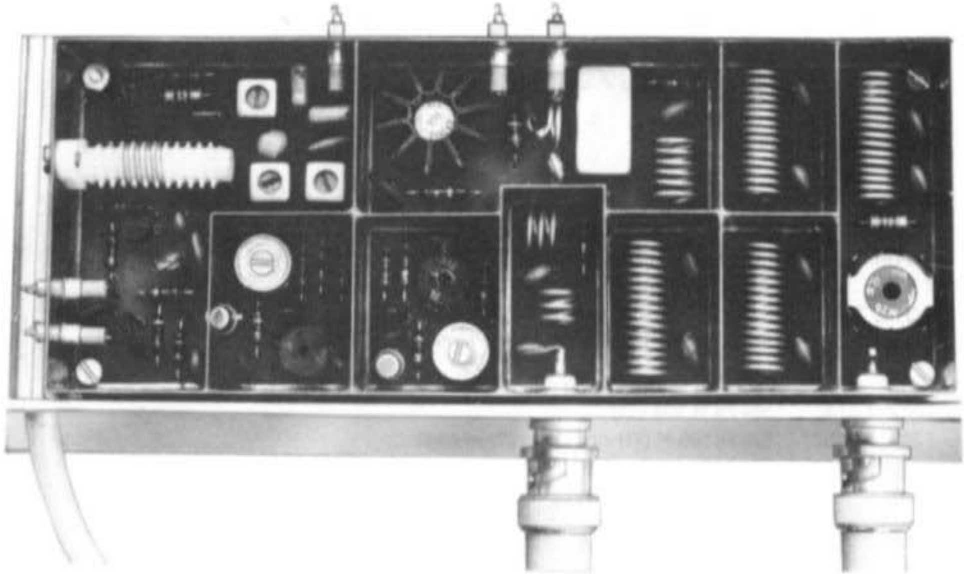


Fig. 11: Author's prototype of the SW-converter (RF-module DK 1 OF 030)

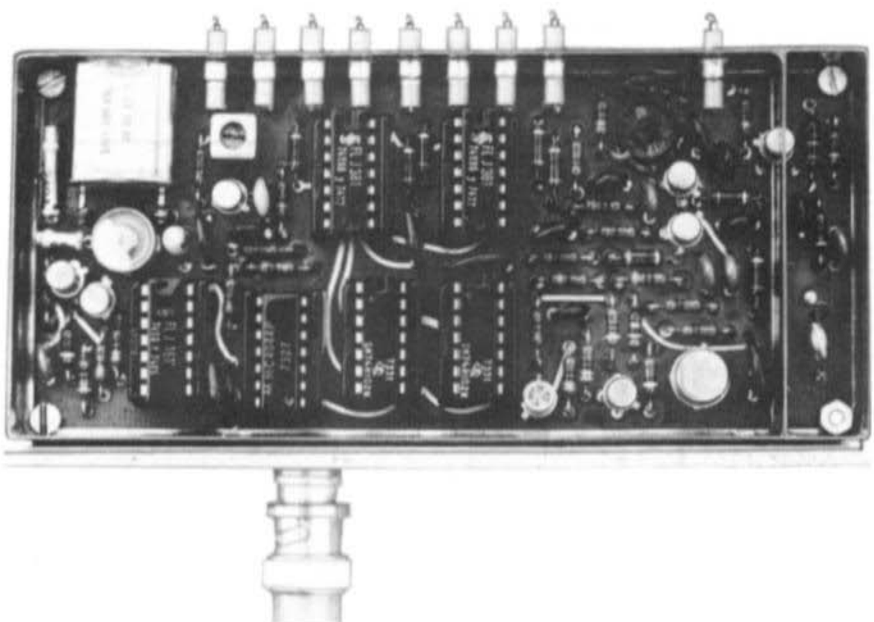


Fig. 12: Author's prototype of the receive converter (synthesizer module DK 1 OF 031)

Feed-through capacitors: approx. 2.2 nF / 30 V, for solder mounting
All other capacitors: ceramic disk type
A spacing of at least 10 mm is available for the resistors.

Synthesizer DK 1 OF 031

T 311:	40841 or 40673 (RCA)
T 312, T 313:	2 N 5179 (RCA)
T 314:	BF 272 or BSX 29 (SGS)
T 315:	2 N 709 (different manufacturers)
T 316, T 317:	2 N 914 or other silicon NPN transistors
T 318:	BC 413, BC 109 (Siemens) or similar Si-NPN transistors
I 311:	SP 8601 BT (formerly SP 601 B) Plessey
I 312, I 313:	SN 74 H 102 (TI)
I 314, I 315:	SN 74196 N (TI) or FLJ ... (Siemens)
I 317:	SN 7490 N (TI) or FLJ 161 (Siemens)
I 316:	MC 4044 P (Motorola)

All IS except I 311 with 14 pin connectors

Q:	crystal 5 MHz ± 2 x 10^{-5}, parallel resonance, load approx.30 pF, holder HC-6/U
L 311:	like L 309
Tr 311:	like Tr 301, except primary 4 windings
C 311:	foil trimmer or ceramic disk trimmer 10 mm dia., 10 to 60 pF

Feed-through capacitors: approx. 2.2 nF/30 V, for solder mounting
All bypass and coupling capacitors: ceramic disk types
For the resistors a spacing of 10 mm is available.

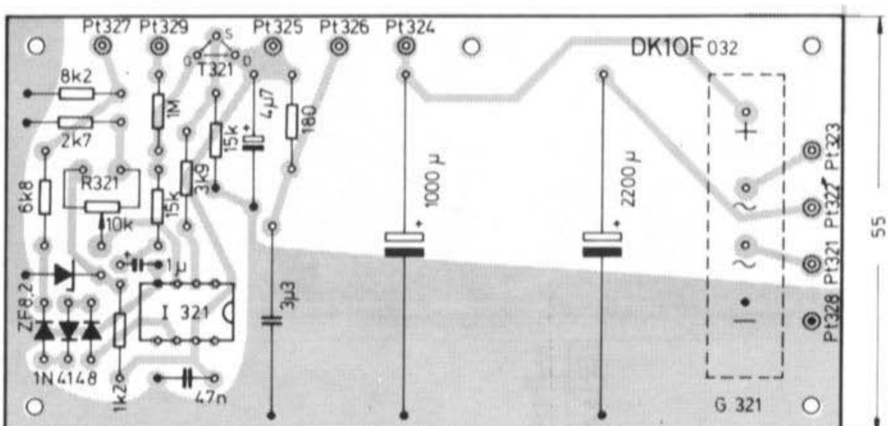


Fig. 10: Component locations on PC-board DK 1 OF 032

Power supply DK 1 OF 032

T 321:	BF 245 C (TI) or W 245 C (Siliconix)
I 321:	TBA 221 B (Siemens) or 741 CM (different manufacturers)
I 322:	LM 340 K - 15 (NS) or SG 8715 C (Silicon General)
I 323:	LM 309 K (NS), also other manufacturers
G 321:	Silicon-bridge rectifiers B 40 C 2200 or B 80 C 2200

Power transformer

Select the suitable transformer available on the market.

4. ALIGNMENT

It is advisable for the power supply to be checked before connecting it to the other modules.

This is followed by connecting the RF-module. Pt 306 is temporarily connected via a resistor of 150 kΩ to Pt 307 where it is connected to the operating voltage of + 15 V. Pt 305 is provided with a filtered operating voltage from Pt 326. A variable DC-voltage of 0 V to approx. + 13 V is connected to Pt 303 which can be made with the aid of a variable resistor of approx. 10 kΩ connected between Pt 305 and ground. If a frequency counter is available, it can be connected to Pt 305. With the aid of the potentiometer connected to Pt 303, a voltage of 12 V should be selected and the VCO frequency should be aligned to approximately 144 MHz with the aid of the core of L 308. The lowest VCO-frequency of 116 MHz should be obtained with the aid of a tuning voltage of approximately 3 V. Values of between 2.5 and 4 V are permissible. **Fig. 13** shows the tuning curve measured on the author's prototype.

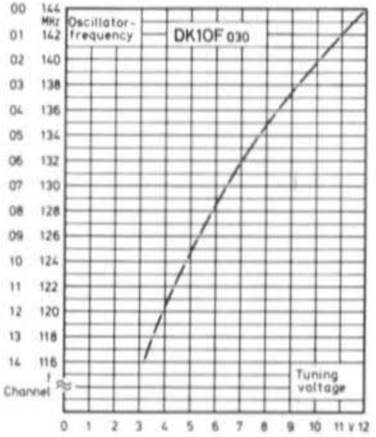


Fig. 13:
Frequency of the VCO as a function of the tuning voltage

This is followed by aligning the local oscillator amplifier. It is advisable to firstly check the DC-operating points of the transistors T 302, T 303, and T 304 with the VCO inoperative (disconnect Pt 305). A voltage drop of approx. 1.5 to 2.5 V must be present across the source resistor of T 302. The emitter voltage values of T 303 and T 304 are given in Fi. 2. The VCO is connected again and the tuning voltage (Pt 303) adjusted to 12 V so that the oscillator frequency is in the order of 144 MHz. A high impedance voltmeter (range approx. 3 V, $R \geqq 100\ k\Omega$) is connected between Pt 308 and ground, and capacitors C 306 and C 308 are aligned for maximum reading. A value of approximately 0.9 to 1.4 V should be availabe. If required, it is possible to alter the voltage divider C 304 / C 305. The indicated voltage should be within the given range also at the lower frequency limit (f_o = 116 MHz).

The digital part of the circuit is now brought into operation. Connection Pt 311 is connected to Pt 304, as well as Pt 312 to Pt 314 with the corresponding connections on the power supply. After switching on, the crystal oscillator (counter to the collector of T 317) and the subsequent divide-by-ten dividers are checked (counter or oscilloscope to pin 11 of I 317). The cores of inductances L 311, L 309, and L 3011 are inserted completely, after which an interconnection is made between Pt 315 and Pt 303 and a voltmeter (range $\geqq$ 15 V) connected between Pt 315 and ground. The phase control circuit should now lock in. The voltage reading should change to that given in Fig. 13 when altering the position of the digital switches.

Finally, the operational amplifier is aligned. The resistor temporarily connected to Pt 306 is removed and this point connected to Pt 327. Pt 308 is connected to Pt 329. Connect the voltmeter between Pt 308 and ground. A voltage of 0.7 V is adjusted with the aid of R 321. This value must be constant over the whole tuning range from channel 00 to channel 14.

The 2 m receiver is now connected to Pt 302, and Pt 301 terminated with 50 Ω. Select channel 00 and tune the receiver to 144 MHz in the SSB or CW mode. A clean heterodyne should be heard. If the accuracy of the 2 m receiver is sufficient (e.g. digital read-out), it is possible for the crystal oscillator to be aligned to zero beat with the aid of C 311.

The receiver is now tuned to 144.5 MHz. The spurious oscillation present at this point can be aligned for minimum by adjusting the core of L 3010. In the case of the author's prototype, this spurious signal could be reduced to approx. 10 dB over the noise level. Of course, this will only be the case if the harmonics of the 500 kHz signal (from the phase comparator) are not induced into the 2 m receiver via a different path. It is advisable to use a well-screened construction for the converter.

5. FURTHER INFORMATION

Especially the large signal handling capabilities of shortwave receivers can be easily checked during the evening hours on the 40 m band. Although the described converter does not possess any preselectivity, it was found to be considerably better than several shortwave receivers including a tubed double-conversion superhet with three tuned input circuits. The author's 2 m station described partially in (6) and (7) was used as 2 m receiver, and a W 3 DZZ antenna was used. Experiments made with other 2 m transceivers were also positive. Under these conditions, a 10 dB attenuated connector between the converter and the receiver brought a noticible improvement of the signal-to-noise ratio.

Since this converter comprises a passive mixer circuit, it is possible for the signal path to be reversed so that a 2 m signal injected to Pt 302 will be converted to the shortwave range. However, the drive power should not be higher than 1 mW in order to ensure that the linear range of the ring mixer is not exceeded. In this case, an output power of approximately 0.2 mW will be available at Pt 301. A wideband amplifier can be used to increase this to the required output power since the image frequency is completely suppressed with the aid of the input lowpass filter. The editors would be interested to receive further information regarding a suitable power amplifier when the converter is used in this mode.

If a 10 m receiver (28 to 30 MHz) is available, it is also possible for the described module to be also used as VHF receive converter in the frequency range of 86 to 116 MHz and from 144 to 174 MHz. The associated channels and receive frequencies are given in **table 2**:

Adjustment	f_0 (MHz)	f_0 - IF (MHz)	f_0 + IF (MHz)
00	144	116 - 114	172 - 174
01	142	114 - 112	170 - 172
02	140	112 - 110	168 - 170
03	138	110 - 108	166 - 168
04	136	108 - 106	164 - 166
05	134	106 - 104	162 - 164
06	132	104 - 102	160 - 162
07	130	102 - 100	158 - 160
08	128	100 - 98	156 - 158
09	126	98 - 96	154 - 156
10	124	96 - 94	152 - 154
11	122	94 - 92	150 - 152
12	120	92 - 90	148 - 150
13	118	90 - 88	146 - 148
14	116	88 - 86	144 - 146

Image rejection is best made using a tuned VHF preamplifier, which is also advisable for improving the sensitivity at VHF. The author has also tested the large-signal handling capabilities of the converter when converting from 145 MHz to 29 MHz. When not using a HF preamplifier, it is possible to transmit 600 kHz from the receive frequency with an output power of 400 W and to monitor one's own transmission via a repeater using a second antenna. The two antennas are decoupled by approximately 60 dB, and the FM repeater signal was only approximately 30 dB over the noise.

Of course, it is also possible for the described converter to be used as transmit converter from shortwave to the 2 m band. The maximum drive level for AM or SSB should not exceed 1 mW PEP. The conversion loss is also in the order of 7 to 8 dB.

The various operating modes of the converter are to be given more clearly in the following **table 3**:

Table 3	Receive converter	Transmit converter
Input frequency Output frequency	50 kHz - 30 MHz 144 - 146 MHz	144 - 146 MHz 50 kHz - 30 MHz
Input frequency Output frequency	86 - 116 MHz 28 - 30 MHz	- -
Input frequency Output frequency	144 - 174 MHz 28 - 30 MHz	28 - 30 MHz 144 - 174 MHz

6. REFERENCES

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A SECOND VERSION OF THE MODULAR AF-AMPLIFIER AND VOLTAGE STABILIZER

by D.E. Schmitzer, DJ 4 BG

The first version of the AF and voltage stabilizer module DJ 4 BG 007 used an integrated circuit type PA 237, which is no longer manufactured. An interchangeable replacement module has now been developed, and experiments were made in order to reduce the low intrinsic noise of the previous module. The voltage stabilizer circuit used in the DJ 4 BG 007 module has not been changed.

1. DEVELOPMENT

An integrated circuit was required to replace the PA 237 that was also able to work over a wide voltage range. Furthermore, it should be manufactured by several different companies so that deliveries could be guaranteed for a considerable time. These demands were met by the TBA 800. This IC also offers the additional advantage of a self-balancing circuit so that exactly half of the voltage will appear at the output, independent of the operating voltage. This ensures that the drive range remains at optimum.

Although the data sheets for the TBA 800 listed better intrinsic noise characteristics than the PA 237, the author was surprised to find that virtually the same noise was present as with the original module (the original circuit had only been modified where absolutely necessary). Since the cause of this could not be the actual integrated circuit, the fault had to lay with the external components.

After consideration, the cause could only be the resistors of the AF-filter at the input of the integrated circuit. This was found to be true by measuring the intrinsic noise level of the circuit as a function of the source impedance of the integrated circuit. The results of this measurements are given in **Figure 1**, curve a. All noise measurements were made with the wideband AF-Millivoltmeter type hp 403 B, which possesses a measuring range of 1 Hz to 1 MHz.

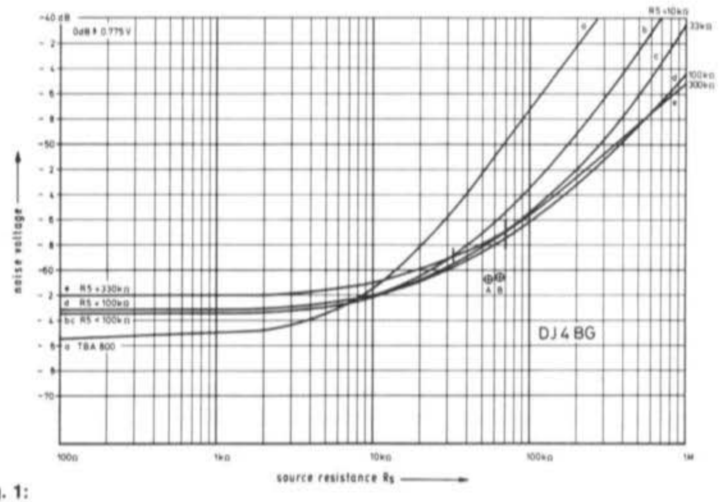


Fig. 1:
Noise voltage at the output of the TBA 800 as a function of the source impedance Z
Curve a: TBA 800 alone – curves b - e : with transistor T 1

The large increase of the intrinsic noise when exceeding a source impedance of approximately 3 kΩ was very surprising, since the IC possesses a relatively very high input impedance (1 MΩ) by using a Darlington circuit. When using the original circuit possessing a source impedance in the order of 70 kΩ (DJ 4 BG 007), the intrinsic noise will be approximately 8 dB worse than when directly driving the IC at low impedance. Due to the AF-filter, module DJ 4 BG 017 possesses a somewhat lower noise level than is indicated in Figure 1. The noise level of the complete module is also indicated in Figure 1. The noise level of the complete module is indicated by the value A with a short-circuit at the input, and by value B on the open circuit conditions. The reason for this improvement is capacitor C 4, which keeps the base of T 1 at low impedance for the higher frequencies.

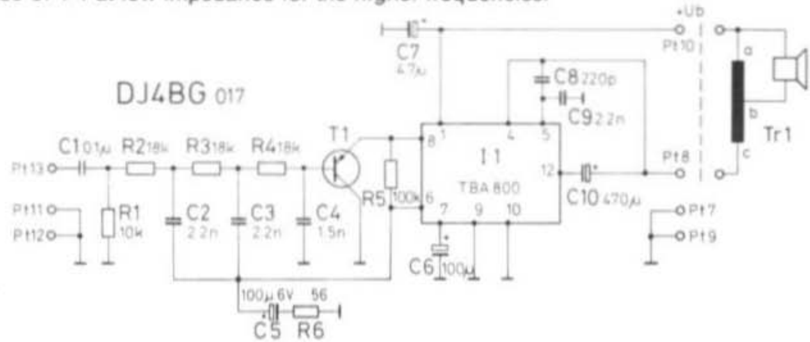


Fig. 2: Circuit diagram of the new AF amplifier equipped with the TBA 800

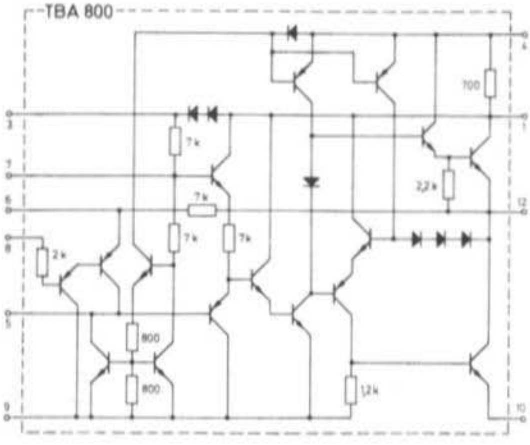
2. MODIFIED CIRCUIT

After some experiments, the circuit given in **Figure 2** was finally found. It offered a very low intrinsic noise level inspite of the provision of the AF-filter. A low-noise transistor is connected in front of the TBA 800. This allows the use of a relatively high-impedance AFfilter, which does not increase the intrinsic noise to any great degree, and is able to drive TBA 800 at such a low impedance that IC hardly increases the intrinsic noise level. The noise characteristics are influenced by resistor R 5 that determines the operating current of transistor T 1. If the value of R 5 is high, or not provided, the source impedance provided to the TBA 800 will not be of very low impedance and will considerably increase the total noise level. However, the increase of the noise towards higher source impedances is low (Figure 1, curve e). If, on the other hand, the value of R 5 is relatively low, the IC will be provided with a low source impedance. At low source impedances, the total intrinsic noise is hardly higher than when using the TBA 800 directly connected to a very low source resistor; however, the improvement of the noise characteristics at higher source impedances is no longer so prevalant. This effect is shown by curves b - e in Figure 1.

A disadvantage of the provision of T 1 is the shift of the output DC voltage at connection 12 of I 1. This is due to the fact that the drive range has been decreased.

However, this can be compensated for in a simple manner: the internal circuit of the TBA 800 **(Figure 3)** is provided with two diode paths between connections 1 and 3 that are usually short-circuited for most applications. If connection 3 remains unconnected, these two diodes will compensate for the voltage shift at connection 12 that is caused by transistor T 1, and will ensure that the drive range is optimum again.

Fig. 3:
Internal circuit of the TBA 800



3. FREQUENCY RESPONSE

The frequency response of the modified AF-circuit **(Figure 4)** does not differ greatly from that of module DJ 4 BG 007. The lower cut-off frequency is mainly determined by R 1 and C 1, whereas R 6 and C 5 exhibit a far lower cut-off frequency in the selected circuit. When the AF-filter is only to be used in the SSB-mode, it would be possible for the value C 5 to be decreased still further (down to 22 µF), which would result in a better suppression of hum voltages. However, this would make it difficult to zero in the signal in the CW-mode; if the lower cut-off frequency of the module is to be reduced still further for this reason, it is possible for C 1 to be increased to 0.47 µF.

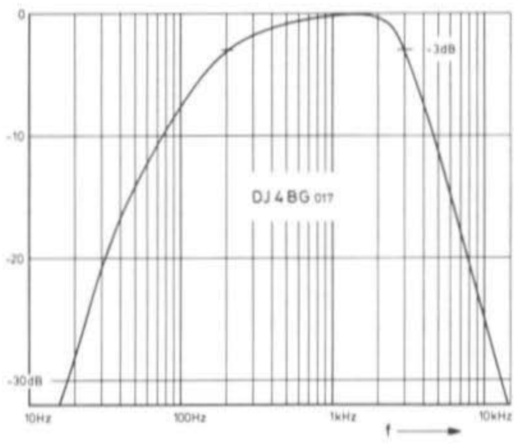


Fig. 4:
Frequency response of the AF-amplifier

The far higher gain of the TBA 800 with respect to the PA 237, and the possibility of using a higher feedback results in a more favorable operation of the AF-filter. This means that the highest possible skirt slope of the low-pass filter is really obtained which means that a higher attenuation of higher frequencies is obtained with this module at the same cut-off frequency as with module DJ 4 BG 007. The difference amounts to approx. 7 dB at 10 kHz.

Capacitor C 6 is important for suppressing low frequency (hum) voltage components, e.g. from the operating voltage of the module. If required, the capacitance values can be increased if sufficient room is available.

4. OUTPUT POWER

Since the integrated circuit TBA 800 possesses considerably more efficient output transistors than was the case with the PA 237, it is possible to directly drive a loudspeaker if the impedance is not too low. Of course, this depends on the operating voltage and required output power. With higher operating voltages, low impedance loudspeakers and low output powers, it will be necessary to use the given output transformer as was the case with DJ 4 BG 007.

The output power levels provided by the circuit for low level applications are to be given in two different ways:

Figure 5 shows the output power at various operating voltages as a function of the load impedance, whereas **Figure 6** gives the output power at various load impedances as a function of the operating voltage. The given curves are somewhat lower than the values given in the data sheets of the TBA 800, which are given for a distortion factor of 10 %.

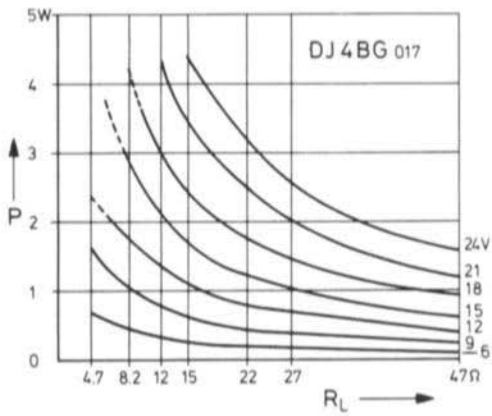


Fig. 5:
Output power as a function of the load resistor
Parameter: operating voltage (6 - 24 V)

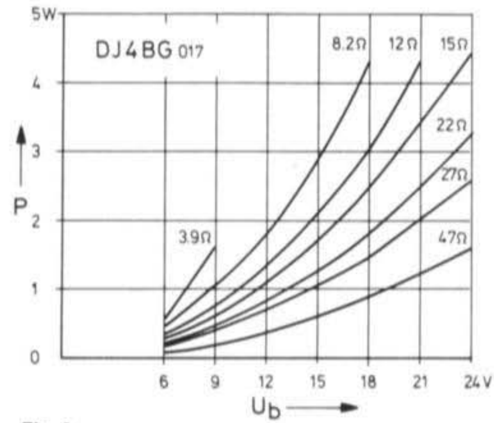


Fig. 6:
Output power as a function of the operating voltage
Parameter: load resistor (3.9 - 47 Ω)

Since the author was not able to carry out a harmonic distortion measurement, the slight limiting of the output signal when observed on an oscilloscope was used as criterium for the maximum drive level. The harmonic distortion at this level is far lower than the 10 %.

Since the available heatsink area provided in module DJ 4 BG 017 is not really large and since the modules are enclosed in an aluminium box, the heat dissipation is not very favorable. For this reason, the circuit should not be used at an output power in excess of 2.5 W. Of course, the provision of additional cooling of the TBA 800, as well as better air flow (holes in the aluminium box) could provide a better heat dissipation, but it is felt that this should not be necessary since the output power of the new module is already more than twice that of module DJ 4 BG 007.

Finally, the curves given in Figure 5 and 6 give some further measuring values regarding the current gain at full drive, and the efficiency at several voltages and impedances of interest for practical operation.

A) Without output transformer

U_b V	R_L Ω	I_{max} mA	P W	η %
12	8.2	192	1.8	77
12	15	122	1.1	76
12	27	76	0.7	76
18	15	180	2.44	75
18	27	112	1.45	72
18	47	72	0.9	70
9	3.9	250	1.6	70
9	8.2	148	1.06	78
9	15	90	0.62	76

B) With output transformer (see 1)

U_b V	R_L Ω	I_{max} mA	P W	η %
12	4	93	0.85	75
18	4	145	1.8	70
12	8.2	52	0.44	70
18	8.2	78	1.02	73

5. TENDENCY TO SELF-OSCILLATION AT RF FREQUENCIES

A slight tendency to RF-oscillation was observed during the power measurements to establish the values given in Figures 5 and 6. These measurements were made with a pure ohmic load and operating voltages in excess of 12 V. This tendency to oscillation did not appear even when using relatively low inductive loads (such as loudspeaker or loudspeaker and output transformer), which means that when using a loudspeaker, no neutralization will be required. However, if such oscillation should occur in some cases, it will only be necessary to provide a low-impedance coil between the module and load. It was found during the power measurements using a purely resistive load that 10 turns of 0.5 mm dia. enamelled copper wire wound on a 100 Ω resistor will be sufficient.

Attempts made to neutralize the TBA 800 so that no tendency to oscillation was exhibited at purely resistive, or even slightly capacitive loads were unfortunately without success. However, this is not important since a slightly inductive load (loudspeaker) is always connected under normal practical conditions.

6. VOLTAGE STABILIZER CIRCUIT

As it has been previously mentioned, it was not necessary to make any modifications to the voltage stabilizer circuit. For this reason, the original circuit was used, but the locations on the PC-board are slightly different. In order to avoid repeating already published information, interested readers can find further details in the diagrams given in (1).

7. PC-BOARD DJ 4 BG 017

It was possible for the new AF-amplifier and voltage stabilizer circuit to be accommodated on a PC-board of 65 mm x 90 mm. This board has been designated DJ 4 BG 017. The module is completely interchangeable with the previous AF-module DJ 4 BG 007. The PC-board and component locations are given in **Figure 7**, and a photograph of the author's prototype is given in **Figure 8**.

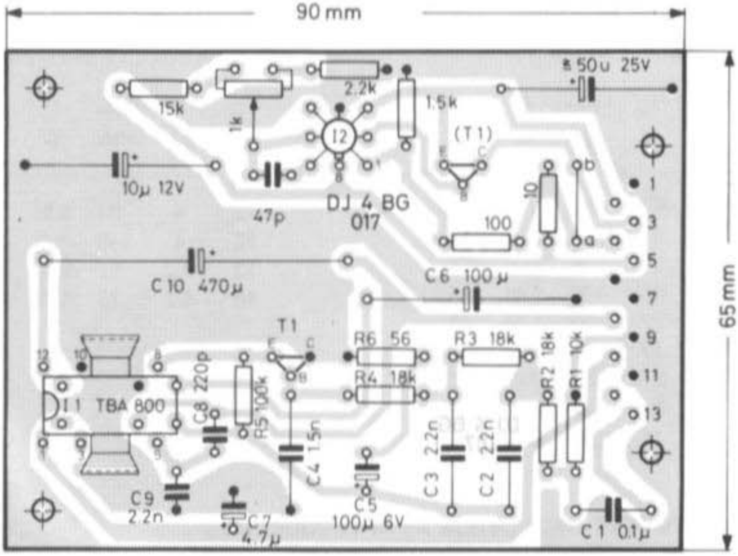


Fig. 7: Component locations on PC-board DJ 4 BG 017

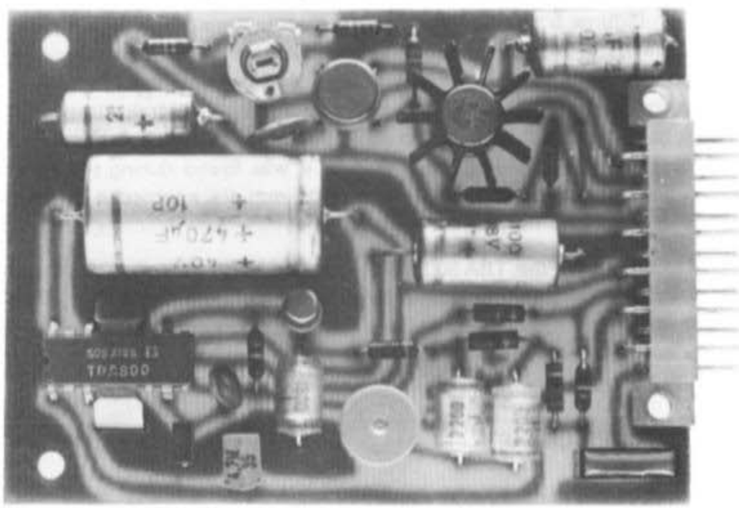


Fig. 8: Photograph of the author's prototype

8. COMPONENT DETAILS

I 1: TBA 800 (Philips, ITT, SGS, Telefunken, EPC)
C 2, C 3, C 4: plastic foil capacitors, low tolerance (5 % or less)
Output transformer: Type EK 25 as for DJ 4 BG 007 (1).
T 1: low noise silicon PNP with a high current gain, e.g. BC 179 C, BC 415 B

9. REFERENCES

(1) D.E. Schmitzer: An integrated AF amplifier and voltage stabilizer
VHF COMMUNICATIONS 4, Edition 1/1972, pages 34 - 39

Data sheets for the integrated circuit TBA 800

A PRECISION DIGITAL MULTIMETER
Part 2: Input amplifier and power supply

by J. Kestler, DK 1 OF

6. INPUT MODULE

The circuit diagram of the input module is shown in **Fig. 9.** The signal to be measured is fed from the input socket of the unit (»IN«) via the mode switch S 1a to the input voltage divider that is equipped with precision resistors, and is frequency compensated using parallel capacitors. In order not to feed the high input voltage in the 2 kV range along the PC-board, this voltage divider (1 MΩ / 1 kΩ) is mounted separately. The output of the range switch S 2a is connected to a second wafer of the mode switch S 1b. This is provided to pass the input signal to the amplifier for AC voltage and DC voltage measurements.

The DC-voltage to be measured is passed via a protective resistor of 100 kΩ to the first amplifier stage equipped with the integrated operational amplifier I 271. A source follower using the two systems of the double FET T 271 is connected before each input of the amplifier. The input impedance of this stage amounts to several hundred MΩ which means that the input voltage divider is practically not loaded. The two, anti-phase zener diodes at the input of the FET protect it against high voltages. The offset voltage of the operational amplifier can be zeroed with the aid of the trimmer resistor R 271. The gain factor or the amplifier is determined by the feedback resistors (10 kΩ / 500 kΩ) to a value of 51.

Since the A/D-converter requires a negative input voltage, a reversal stage equipped with the amplifier I 272 has been provided. The gain factor of this stage amounts to - 1, and the exact value can be adjusted with the aid of R 272. R 273 is used for alignment of the offset voltage.

The reversal stage is bypassed when the relay is energized (shown in its rest position). This is the case when a negative input voltage is to be measured. In order to ensure that the relay does not switch back and forth at an input voltage corresponding to zero, the comparator (I 273) is provided with a low bias voltage at its non-inverting input which ensures that the driver transistor for the relay (T 272) remains blocked.

The output of the DC preamplifier is fed via resistor R 277 and a third wafer on the mode switch (S 1c) to the A/D-converter. The calibration of the unit in the DC-mode is made with the aid of resistor R 277.

6.2. Preamplifier for AC measurements

The input stage (T 273, I 274) is built up in a similar manner to that of the DC-preamplifier stage. In this case, however, the amplifier uses an integrated circuit type LM 301 A, which possesses a considerably higher bandwidth than the internally frequency-compensated TBA 221. The voltage gain of this stage amounts to 25 (1 + 24 k / 1 k).

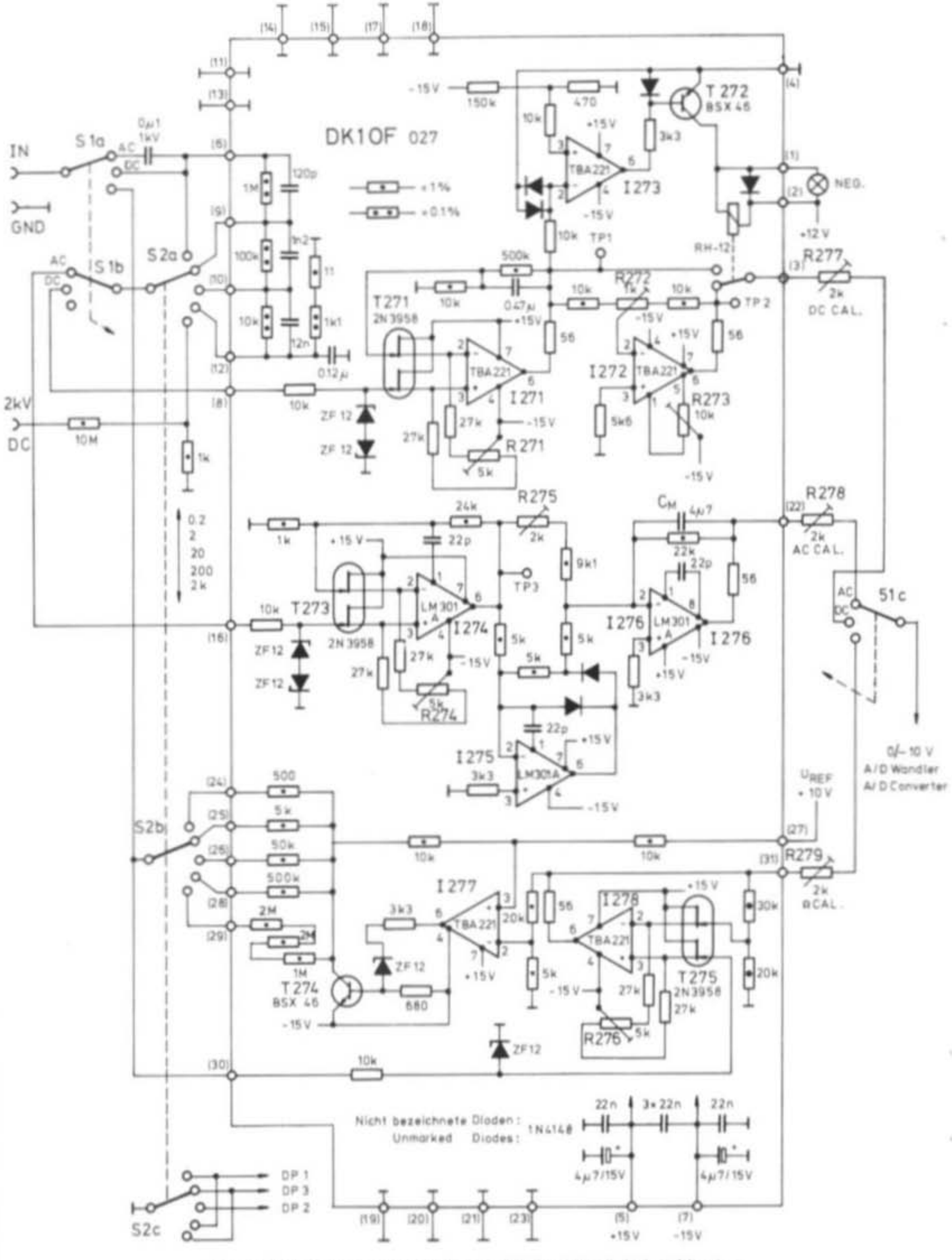


Fig. 9: Circuit diagram of the input circuit of the digital multimeter

The next stage equipped with the integrated amplifiers I 275 and I 276 forms a fullwave rectifier with subsequent filtering of the DC-voltage, and the operation is to be explained with the aid of the basic circuit given in **Fig. 10.**

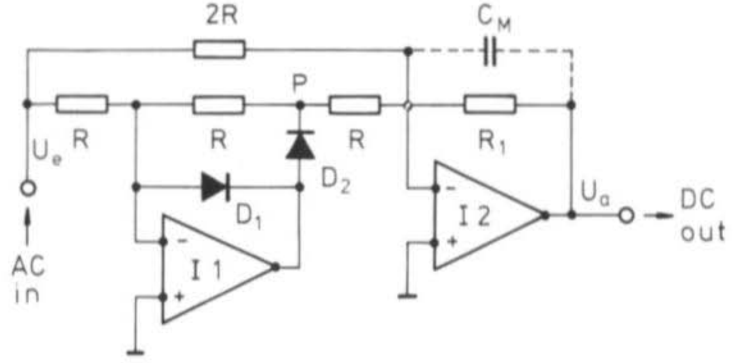


Fig. 10: Principle of the rectifier

The capacitor C_M given in dashed lines is firstly assumed not to be present. A squarewave AC-voltage is assumed to be present at the input of the circuit. During the positive halfwave of the input signal, amplifier I_1 receives full feedback via diode D_1. This means that its gain is zero and its output voltage corresponds to the threshold voltage of D_1. This threshold voltage is compensated for in diode D_2 so that 0 V can be measured at point P. The output voltage of I_2 therefore amounts to

$$Ua = \frac{R_1}{2\,R} \times Ue \quad \text{(positive halfwave)}$$

If the input voltage is negative, diode D_1 will be blocked, and the inverted input signal will be present at point P. The output voltage of the second amplifier in this case is

$$Ua = \frac{R_1}{R} \times Ue - \frac{R_1}{2\,R} \times Ue = + \frac{R_1}{2\,R} \times Ue \quad \text{(negative halfwave)}$$

It will be seen that the resulting gain factor is equal in both cases but the sign is different.

Positive input voltages will be inverted, but not negative. This means that this is a true full-wave rectification. The advantage of this circuit is that the threshold voltages of diodes D_1 and D_2 will have no affect on the resulting DC-voltage as long as they have identical characteristics. The circuit operates practically linearly down to input voltages in the order of several mV.

Capacitor C_M has the task of reducing the gain factor I_2 in the case of AC-voltages, in other words to form a mean value of the output voltage. This means that the digital multimeter described here indicates the mean value (in a mathematical sense). However, it will be calibrated in RMS values, as is usually the case.

6.3. Resistance measurements

The task of the following circuit is to form a DC-voltage that is proportional to the resistance to be measured. This is made in the most simple manner by allowing a known current to flow via a resistor and to measure the voltage drop across it. The current source is formed by I 277 and T 274 (Fig. 9). The integrated circuit I 278 represents the preamplifier whose input impedance must be large with respect to the highest resistance value to be measured (2 MΩ). For this reason, a preamplifier stage equipped with a double FET (T 275) is also used here.

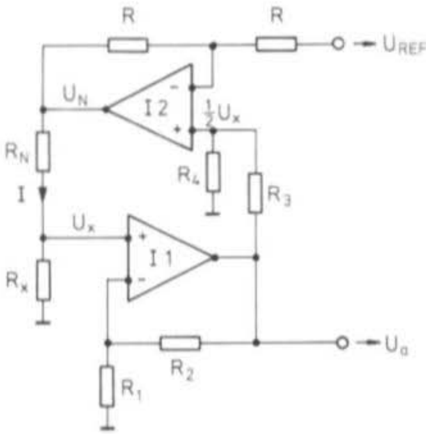


Fig. 11: Principle of the resistance measurement

A circuit showing the principle of operation of the resistance measuring circuit is given in **Fig. 11.** The measuring current I flows via resistors R_N and R_X and causes a voltage drop U_X across the resistance to be measured R_X. This voltage is amplified in I_1 amd fed to the output. The output voltage of the circuit thus corresponds to

$$U_a = I \times R_X \times \left(1 + \frac{R_2}{R_1}\right)$$

The voltage divider (R_3/R_4) is designed so that exactly half the measuring voltage ($U_X/2$) results at the non-inverting input of I_2. This is the case when:

$$\frac{R_3}{R_4} = 1 + \frac{2\,R_2}{R_1} \quad \text{is valid.}$$

Since the measuring current I also flows via R_N, the output voltage of I_2 will amount to:

$$U_N = I\,(R_X + R_N) = U_X + I \times R_N$$

This voltage is also equal to:

$$U_N = 2 \times \frac{U_X}{2} - U_{REF}$$

The following therefore results for the measuring current: $I = -\dfrac{U_{REF}}{R_N}$

This means that the current is independent of R_X, thus constant and wholy determined by the reference voltage U_{REF} and the nominal resistance R_N.

The following results for the output voltage of the circuit:

$$U_a = - R_x \times \frac{U_{REF}}{R_N} \times \left(1 + \frac{R_2}{R_1}\right)$$

The various measuring ranges can be easily obtained by switching the value of R_N.

The last wafer of the range switch (S 2c) is used for switching the decimal points (see section 3.).

6.4. Construction

The PC-board DK 1 OF 027 has been designed for accommodation of the input circuit. The dimensions of the board are 100 mm x 120 mm. It is double-coated and possesses through-contacts. The connections can be made with the aid of a 31-pin connector. The numbers given in the circuit diagram (Fig. 1) correspond to the pin numbers. The component locations are given in **Fig. 12**, and **Fig. 13** shows a photograph of the author's prototype.

6.5. Components

I 271, I 272, I 273, I 277, I 278:	TBA 221 B (Siemens) or LM 741 CM (NS)
I 274, I 275, I 276:	LM 301 A (NS)
T 271, T 273, T 275:	2 N 3958 dual-FET (Siliconix)
T 272, T 274:	BSX 46, 2 N 2219 or similar NPN transistor
Relay:	RH-12 or RSD-12 (national)
Trimmer potentiometer:	spacing 5/10 mm

CM: plastic foil capacitor 4.7 µF / 25 V (Siemens B 32110 - D 3475 - M 000)
S 1: multi-position switch 3 x 3 contacts (mode)
S 2: multi-position switch 3 x 5 contacts (measuring range)

6.6 Interconnection and alignment

The DC-voltage amplifier is aligned firstly. After connecting the supply voltages (± 15 V, + 12 V), a warmup period of approx. 15 min. should be allowed until the operational amplifiers have obtained their operating temperature. The input (connection 8) should then be connected to ground and a sensitive voltmeter (approx. 100 mV) connected between TP 1 and ground. This voltage is now aligned to zero with the aid of R 271. After this, the same should be carried out with respect to TP 2 and R 273.

A voltage of approximately + 150 mV is now connected to the input which can be taken with the aid of a voltage divider from a small battery. After this, the output voltage (connection 3) is measured. The relay should actuate on reversing the polarity of the input voltage, and the previously indicated voltage value is aligned with the aid of R 272.

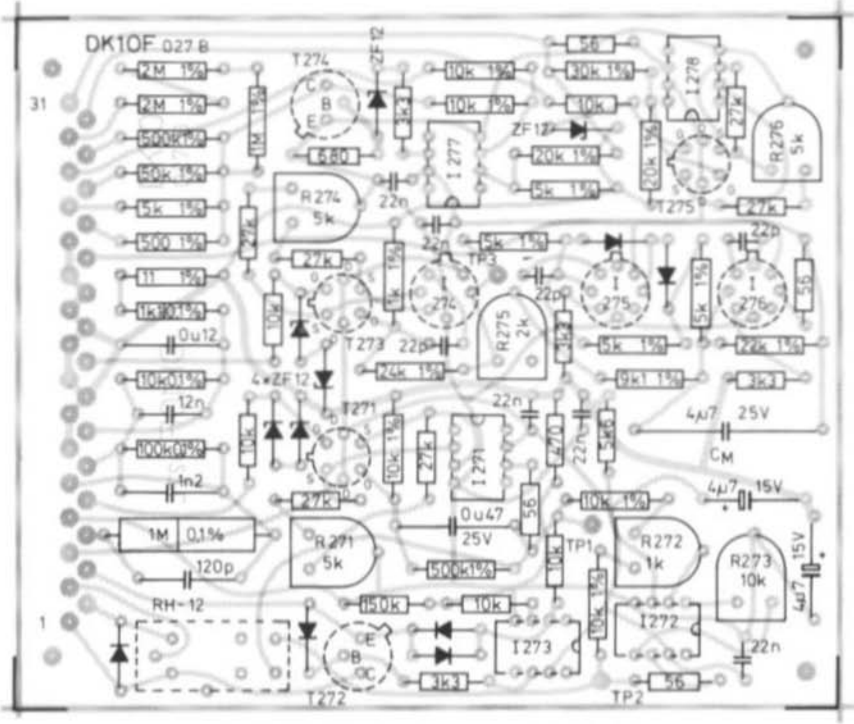


Fig. 12: PC-board DK 1 OF 027 (input circuit)

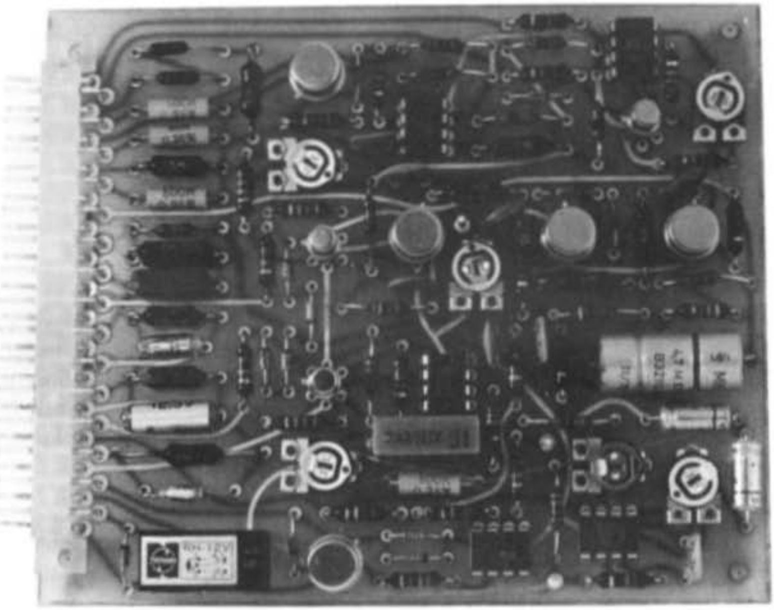


Fig. 13: Prototype of the input circuit

After grounding the input of the AC amplifier (connection 16) the voltage at TP 3 is zeroed with the aid of R 274. A DC-voltage is now connected to the input that results in a voltage of approx. - 10 V at the output (connection 22). R 275 is now aligned so that this value remains constant on reversing the polarity of the input voltage.

The resistance measuring circuit is aligned finally. The output voltage (connection 31) is zeroed with the aid of R 276 with the input (connection 30) grounded.

Resistors R 277, R 278 and R 279 determine the »full scale« of the measuring system in each mode. These are adjusted after completing the unit with the aid of a standard voltage generator or already calibrated digital multimeter. It is advisable for these trimmer resistors to be accessible from outside the case, e.g. for them to be mounted on the rear panel.

The capacitance values shown in parallel with the resistors of the input voltage divider in Fig. 9 are only given for orientation. These values can differ according to the wiring of the measuring range switch. If the highest possible accuracy of the described multimeter is to be used to the full, it will be necessary for an individual alignment of these capacitors to be made. This is made in the most simple manner with the aid of an oscilloscope connected to TP 3. After completion, a clean squarewave signal of approx. 2 kHz which can be taken from a TTL flipflop, is fed to the input. The values of the parallel capacitors are then varied in all measuring ranges in order to obtain a clean squarewave signal at TP 3.

7. POWER SUPPLY

7.1. Circuit diagram of the power supply

The circuit diagram of the power supply is given in **Fig. 14**. A voltage of + 15 V required for the operational amplifiers is obtained from the upper transformer winding, is rectified with the aid of G 1 and stabilized in the fixed-voltage stabilizer I 281. The same winding also generates a voltage of + 12 V, which is taken from the center tap. This is used for operation of the relay and the control lamps, and therefore does not require any stabilization. The + 5 V required for supplying the TTL-integrated circuits is also obtained from this voltage with the aid of stabilizer I 291.

A voltage of - 15 V is generated from the center transformer winding using a second fixed-voltage stabilizer. This voltage is also used for operation of the operational amplifiers. The voltage doubler circuit comprising diodes D 281 and D 282 supplies a negative voltage of approx. - 40 V, which is stabilized to - 20 V with the aid of transistor T 281. This voltage is required for the A/D converter.

The lower winding of the transformer provides the plate voltage for the indicator tubes after rectification with the aid of D 291. The appropriate charge capacitor is to be found on PC-board DK 1 OF 026.

Connection 10 of module 029 is connected to the input of the A/D converter. If the voltage level connected at this point is less than - 10 V (adjustable with R 291), the output of the comparator I 292 is released (open collector), so that the multivibrator equipped with amplifier I 293 can oscillate and the pilot lamp connected to connection 26 will blink periodically. This is the out-of-range indication mentioned in the introduction of this article.

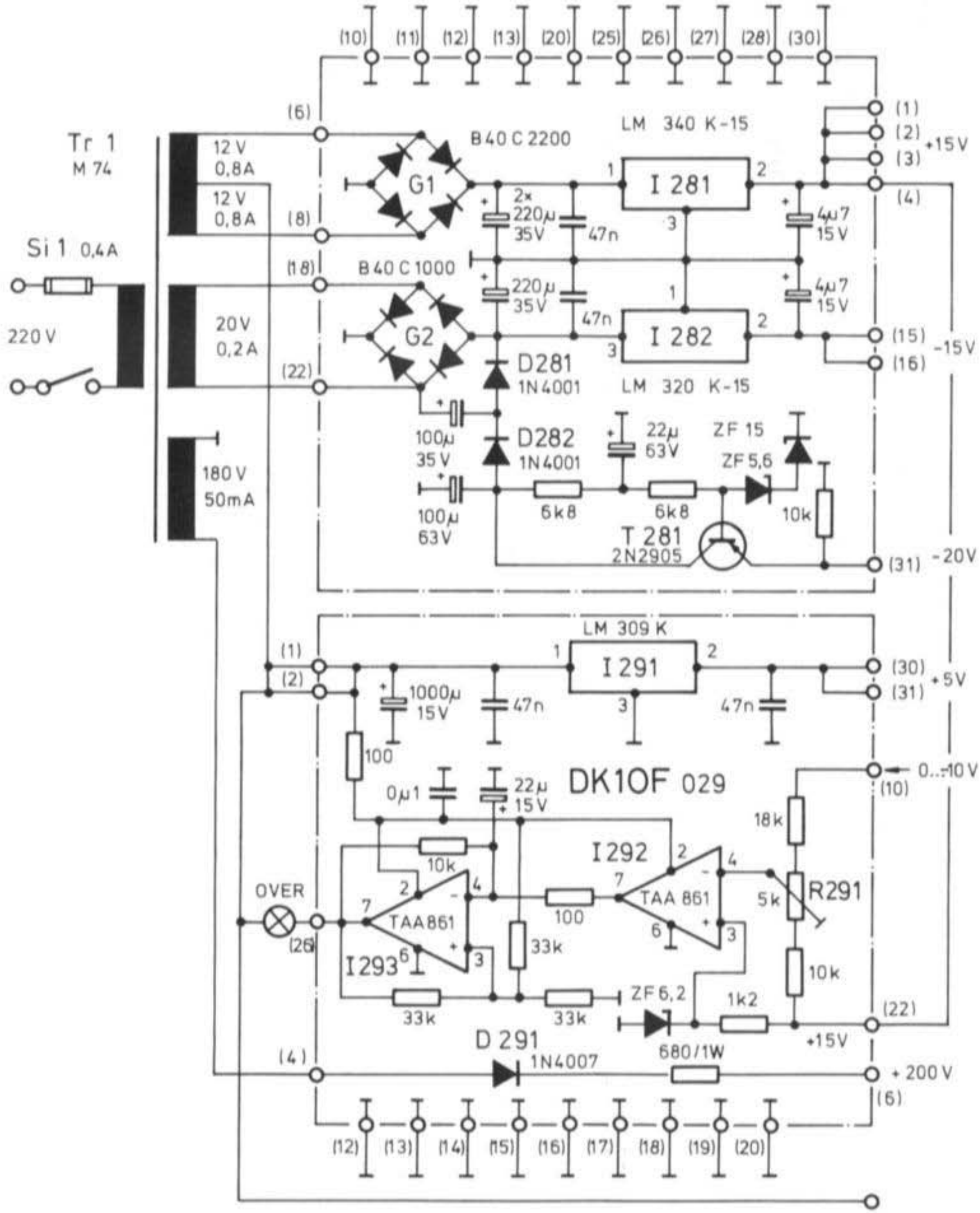


Fig. 14: Power supply circuit for the digital multimeter

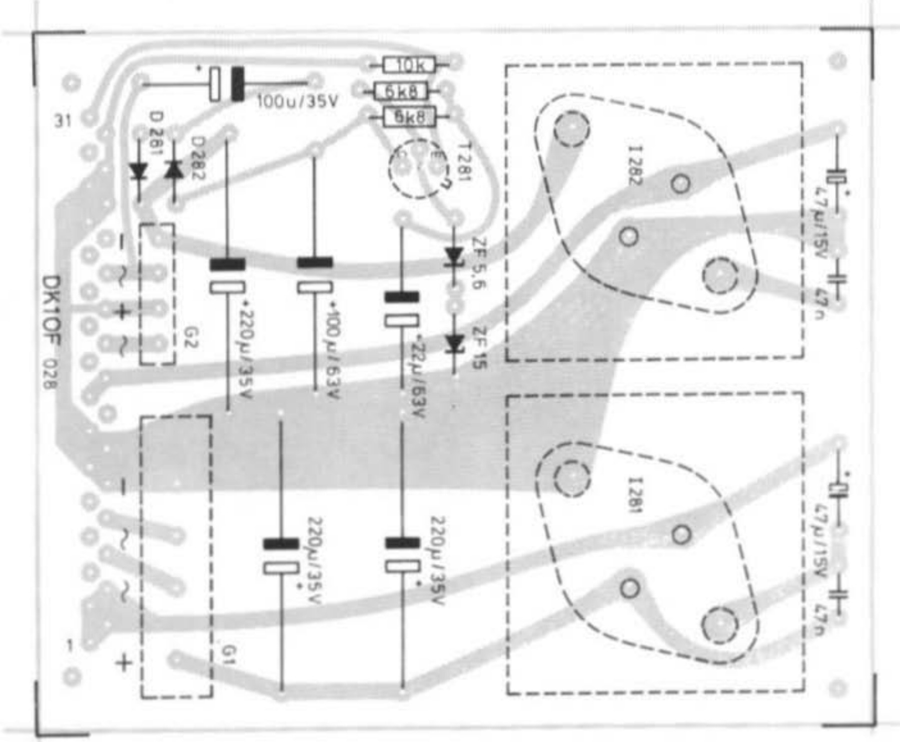


Fig. 15: PC-board DK 1 OF 028 (power supply)

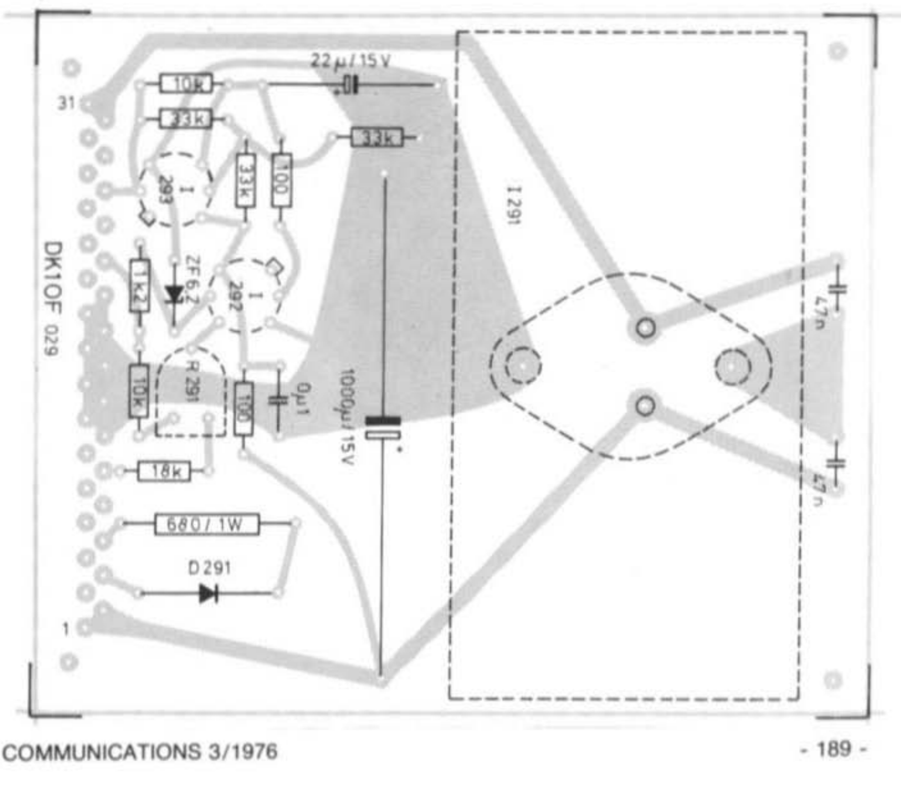
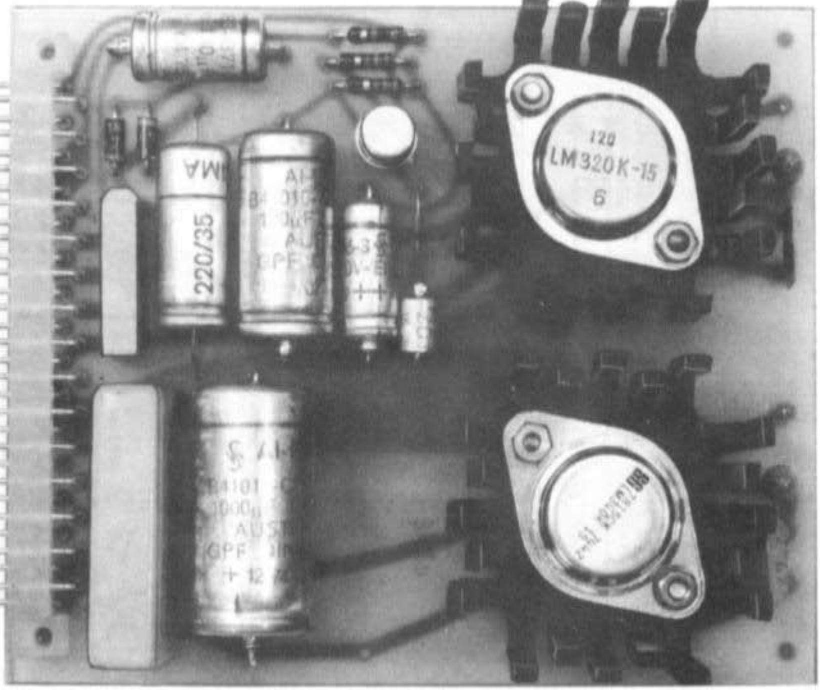


Fig. 16: PC-board DK 1 OF 029 (power supply)



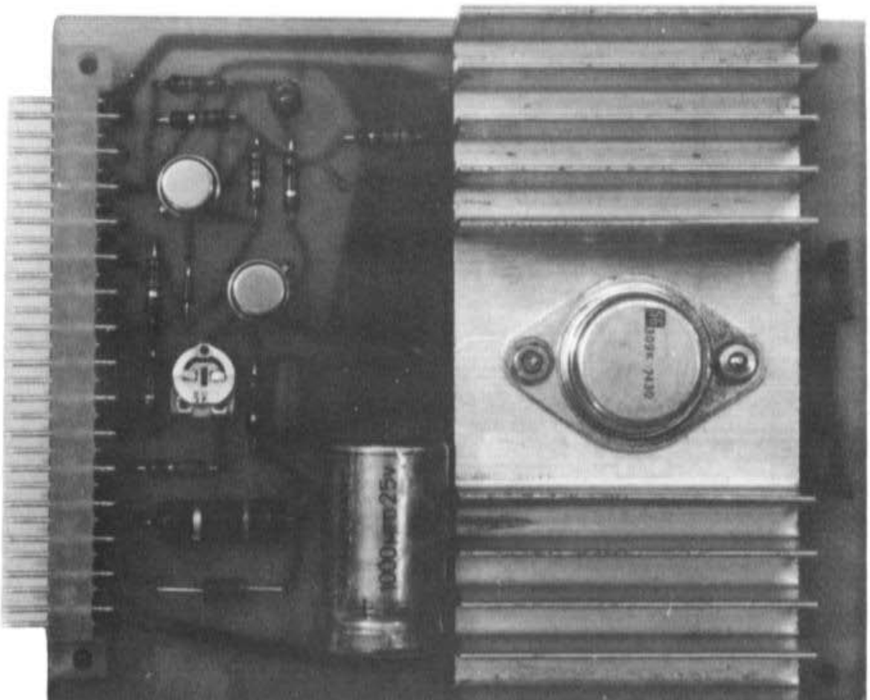


Fig. 17 / Fig. 18: Prototypes of the two power supply boards

7.2. Construction

PC-boards DK 1 OF 028 and 029 have been designed for accommodation of the power supply. Both are 100 mm x 120 mm, single-coated and can also be provided with 31 pin connectors. The component locations are given in **Fig. 15 and 16, and Fig. 17 and 18** show photographs of the prototype boards.

Since such measuring equipment is often in continuous operation, a core type M 74 should be used for the power transformer, although size M 64 would be sufficient for intermediate operation.
The following table gives the number of turns and the wire diameters in the case of a M 74 transformer core:

Primary winding:	1210 turns of 0.32 mm dia.
Screening:	1 x copper foil 0.05 mm
Secondary winding I:	2 x 66 turns of 0.55 mm dia.
Secondary winding II:	112 turns of 0.32 mm dia.
Secondary winding III:	1020 turns of 0.15 mm dia.

7.3. Components

I 281:	Voltage stabilizer LM 340 K - 15 (NS)
I 282:	Voltage stabilizer LM 320 K - 15 (NS)
I 291:	Voltage stabilizer LM 309 K (NS)
I 292, I 293:	Operational amplifier TAA 861 or TAA 761 (Siemens)
T 281:	2 N 2905 or similar PNP-silicon transistor
G 1:	Bridge rectifier B 40 C 2200 (Siemens)
G 2:	Bridge rectifier B 40 C 1000 (Siemens)
D 281, D 282:	Silicon rectifier diode 1 N 4001 or similar
D 291:	Silicon rectifier diode 1 N 4007 or similar
R 291:	Trimmer resistor, spacing 10/5 mm

8. FINAL NOTES

No current measuring ranges were provided on the described multimeter in order not to complicate the range switching. Experience has shown that 95 % of practical measurements are concerned with voltage and resistance. In addition to this, any current can be measured with the aid of a suitable resistor in the form of a voltage drop.

Although three-and-a-half digit digital multimeters are available in the range of 600 to 1000 DM on the market, it can still be worthwhile constructing the described multimeter. On comparing the described unit with the various specifications of the available equipment with respect to measuring accuracy and lowest measuring ranges, it will be seen that the DK 1 OF multimeter is virtually one order of magnitude better than those whose data sheets are available to the editorial staff. Prerequisite is, however, that the described metal-film resistors, of which four pieces should have a tolerance of 0.1 %, are used, and that the completed unit is aligned with the aid of a very good digital voltmeter after construction.

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 3/1976 of VHF COMMUNICATIONS

DJ 4 BG 017	**NEW AF AMPLIFIER and VOLTAGE STABILIZER**		**Ed. 3/1976**
PC-board	DJ 4 BG 017	(with printed plan)	DM 12.—
Semiconductors A	DJ 4 BG 017	(AF-amplifier: 1 IC, 1 transistor)	DM 23.—
Semiconductors B	DJ 4 BG 017	(Voltage stabilizer: 1 IC, 1 transistor)	DM 19.—
Transformer EK 25 – as long as stock lasts – special reduced price :			DM 4.50
DJ 6 ZZ 006	**28 MHz / 432 MHz TRANSMIT CONVERTER with SCHOTTKY RING MIXER**		**Ed. 3/1976**
PC-board	DJ 6 ZZ 006	(double-coated, without through contacts)	DM 19.—
Semiconductors	DJ 6 ZZ 006	(5 transistors, 1 IC, 1 ring mixer)	DM 150.—
Minikit	DJ 6 ZZ 006	(12 trimmer, 2 feedthrough capacitors, 1 coilset with core, 10 ferrite beads)	DM 26.—
Kit	**DJ 6 ZZ 006**	**with above parts**	**DM 190.—**
DK 1 OF 025 - 029	**DIGITAL MULTIMETER**		
DK 1 OF 025	**A/D-CONVERTER**		**Ed. 3/1976**
PC-board	DK 1 OF 025	(double-coated, through-contacts)	DM 30.—
Minikit	DK 1 OF 025	(11 ICs, 9 transistors, 2 diodes, 3 resistors 1%)	DM 89.50
Kit	**DK 1 OF 025**	**with above parts**	**DM 115.—**
DK 1 OF 026	**DECODER and INDICATOR MODULE**		**Ed. 3/1976**
PC-board	DK 1 OF 026	(double coated, through contacts)	DM 17.—
Indicator tubes	DK 1 OF 026	(4 Nixie tubes ZM 1330)	DM 70.—
Semiconductors	DK 1 OF 026	(5 ICs, 1 transistor)	DM 31.—
Kit	**DK 1 OF 026**	**with above parts**	**DM 115.—**
DK 1 OF 027	**INPUT MODULE**		**Ed. 3/1976**
PC-board	DK 1 OF 027	(double coated, through contacts)	DM 30.—
Semiconductors	DK 1 OF 027	(8 ICs, 5 transistors, 12 diodes)	DM 94.—
Minikit	DK 1 OF 027	(1 relay, 27 resistors 1%, 4 resistors 0.1 %)	DM 58.—
Kit	**DK 1 OF 027**	**with above parts**	**DM 175.—**
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PC-board	DK 1 OF 028	(with printed plan)	DM 15.—
Semiconductors	DK 1 OF 028	(2 ICs, 2 rect., 1 transistor, 4 diodes)	DM 65.—
Kit	**DK 1 OF 028**	**with above parts**	**DM 79.—**
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PC-board	DK 1 OF 029	(with printed plan)	DM 15.—
Semiconductors	DK 1 OF 029	(3 ICs, 2 diodes)	DM 32.—
Kit	**DK 1 OF 029**	**with above parts**	**DM 46.—**
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Semiconductors	DK 1 OF 030	(1 ring mixer, 4 transistors, 2 diodes)	DM 64.—
Minikit 1	DK 1 OF 030	(2 chokes, 3 potted cores, 3 coilsets, 1 coilformer with core)	DM 34.—
Minikit 2	DK 1 OF 030 0	(5 feedthrough, 2 trimmer, 34 ceramic caps., 16 resistors, 1 cooling fins, 2 m silver-pl.wire)	DM 34.—
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DK 1 OF 031			
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PC-board	DK 1 OF 031	(double coated, through contacts)	DM 20.—
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Minikit	DK 1 OF 031	(9 feedthrough, 1 trimmer, 22 ceramic, and 1 tantalium capacitor, 38 resistors, 1 coilset, 6 IC sockets, 1 potted core)	DM 54.—
Crystal	5.000 MHz	HC-6/U	DM 26.—
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Semiconductors	DK 1 OF 032	(3 ICs, 1 transistor, 4 diodes, 1 rectifier)	DM 52.—
Minikit	DK 1 OF 032	(4 electrolytics, 1 plastic foil, 1 ceramic capacitor, 9 resistors, 1 trimmer pot.)	DM 18.—
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