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A Publication
for the Radio-Amateur

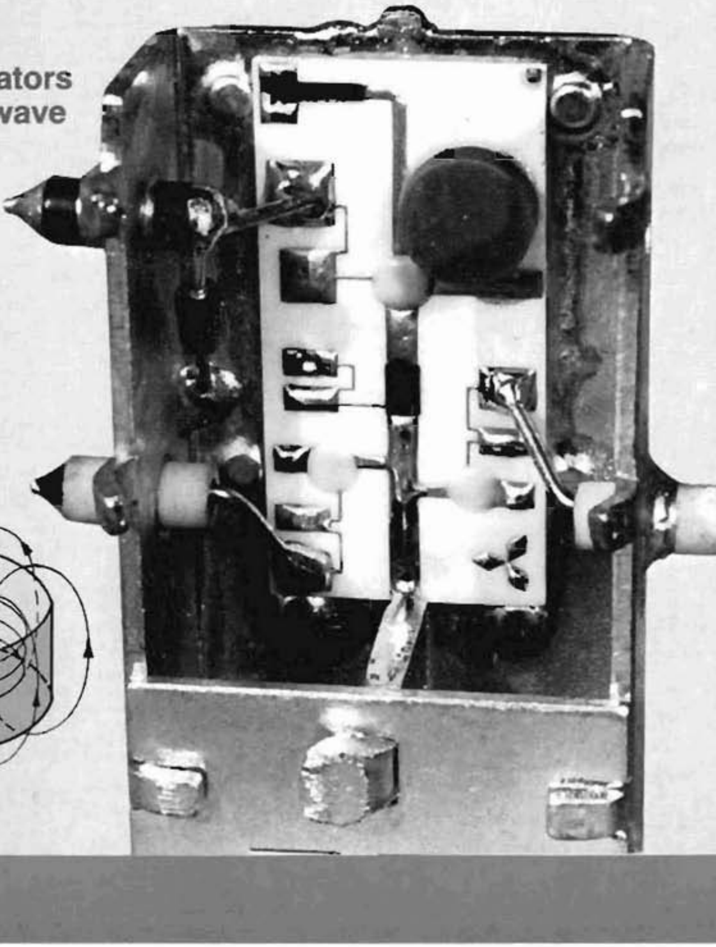
Especially Covering VHF,
UHF and Microwaves

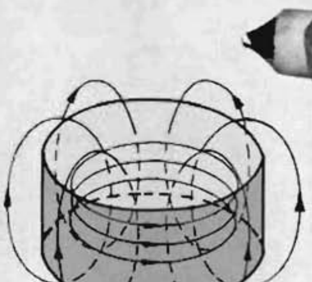
VHF communications

Volume No. 15 · Winter · 4/1983 · DM 6,50

Dielectric Resonators

can replace cavity resonators in the microwave range





VHF communications

A Publication for the Radio Amateur
Especially Covering VHF, UHF, and Microwaves

Volume No. 15 · Winter · Edition 4/1983

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Contents

In the Focus

It's only in very few countries that radio amateurs are lucky enough to find good, inexpensive measuring equipment on the surplus market Germany is not one of them, therefore we keep developing our own measuring equipment. Since the situation will probably be similar in your country and the experimenting radio amateur is always looking for measuring equipment, you will certainly be interested in the following:

In this edition you will find some little measuring aids – mini effort, maxi effect – and a sensitive power meter having a lowest range of 100 μW (– 10 dBm) f.s.d. which will be usable up to 10 GHz when properly constructed. Its critical miniature parts are not especially inexpensive, but they are easily available

On my desk are descriptions of an RF millivolt meter (100 kHz – 1.3 GHz, 1 mV – 10 V), a dummy-load wattmeter (5/50 W: up to 1.3 GHz), and a digital colour TV-pattern generator. Furthermore, DJ3RV is preparing an article describing a time base which receives and evaluates the frequency/time standard for Central Europe on 77.5 kHz. All frequencies of a transceiver, for instance, can be derived from this source to enable coherent communications, such as CCW or CRTTY. In addition, time and date can be displayed to the second, with all eventual changes being carried out automatically.

Last, but not least I can announce a microwave noise source using a Philips noise diode, for our Automatic Noise-Figure Measuring System (VHF COMMUNICATIONS 1/83 and 2/83)

So far as measuring equipment is concerned in the near future, I think there is lots of reasons to keep on reading VHF COMMUNICATIONS

Have a good 1984!

Your's

Robert E. Lentz

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Tel West Germany 9133-855. For Representatives see cover page 2

Jochen Jirmann, DB 1 NV, and Friedrich Krug, DJ 3 RV

The Dielectric Resonator

A Miniature Component for Realizing Stable Microwave Oscillators and Microwave Filters

For a few years now, the leading component manufacturers have offered a microwave component that now becomes interesting to radio amateurs due to the reduction in prices: This is the dielectric resonator. The following article is to describe this component and to show what it can be used for.

Only two types of transmit or receive oscillators were used for amateur radio communications in the 10 GHz range: The mechanically, or varactor-tuned Gunn oscillator with cavity resonator for simple, portable equipment, and the crystal-controlled varactor multiplier for home stations. A cavity resonator constructed using amateur methods often does not work satisfactorily, since high Q-values are difficult to obtain, and since the mechanical construction required for frequency tuning and temperature stabilization are very difficult. Also the trend to miniaturization is turning against such types of resonators.

The dielectric resonators, and ready-to-operate components equipped with them, allow one to construct a good transceiver for the 3 cm band relatively easily.

1. THE DIELECTRIC RESONATOR

In its simplest form, a dielectric resonator is a cylindrical disk made from a dielectric having a very high dielectric number ε_r. The electromagnetic fields form standing waves in this dielectric. As is the case with all cavity resonators, these waves represent the frequency-determining resonances, which are dependent on the geometric dimensions, the relative dielectric number ε_r, and the relative permeability number μ_r.

Although a multitude of standing waves, and thus resonances, can form within a resonator we are only going to consider a certain fundamental oscillation caused by the type of excitation in the following article. In the case of a cylindrical disk whose diameter D is approximately twice as large as the height H, the fundamental oscillation will form a wave which is similar to the H_{011}-mode (TE_{011}) of round cavity resonators. The field lines are shown in **Figure 1**.

In the case of a metallic cavity resonator, the

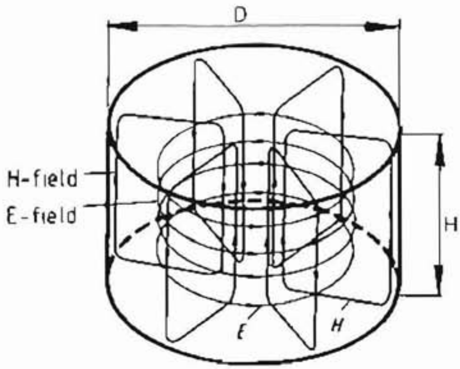

Fig. 1a:
Model of the electrical and magnetic fields of a H_{011}-resonance of a metallic cavity resonator

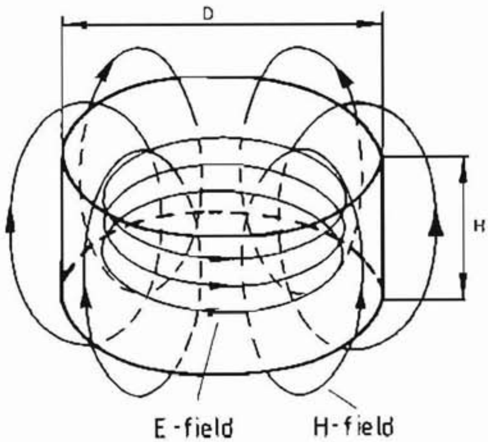

Fig. 1b:
Model of the electrical and magnetic fields of the H_{011} resonance of a dielectric resonator.

conductive walls and the induced currents in the walls form the limitation for the internal fields. Practically speaking, no fields are present outside, and the resonator must be provided with coupling holes, or striplines through the wall in order to be connected to the rest of the circuit. This makes it difficult to use such resonators in stripline circuits.

In the case of a dielectric resonator, the electrical field is concentrated inside the disk due to the high dielectric number. The higher the dielectric number, the better. For this reason, the materials used for dielectric microwave resonators have dielectric numbers between 5 and 150. Since the permeability number is $\mu_r = 1$, the magnetic field will extend outside the resonator in contrast to a metallic cavity type, which allows the dielectric resonator to be coupled to lines relatively easily

The greatest advantage of the dielectric resonator is, however, its compact dimensions. In the case of the H_{011}-resonance of an airfilled resonator, the diameter D corresponds approximately to a wavelength, which means D = 3 cm in the 3 cm band, and H = 1.5 cm. In a dielectric , the velocity factor of the electromagnetic waves is reduced by the factor $1/\sqrt{\varepsilon_r}$, which means that the dimensions of the dielectric resonator can be reduced by the same factor With a dielectric of $\varepsilon_r = 38.5$, as used by us, the resonators for the 10 GHz band will have a diameter of 5 mm and a height of 2 mm.

The quality of such a resonator is determined by its Q, and the temperature stability of its resonant frequency. Both magnitudes are determined by the material, but also by the coupling to the circuit.
Due to the low losses in the dielectric, the non-load Q of available dielectric resonators for the 3 cm band amounts to approximately Q = 5000. The effective operating-Q will be lower due to the coupling to the circuit.

The temperature coefficient of the material is practically negligible since it is $\pm$ 1 x 10^{-6}/K. However, it is possible to adjust it in relatively wide ranges during manufacture, so that a compensation with the temperature response of the circuit is possible. In the case of a negative temperature coefficient of the material, it is possible to compensate for the usual positive temperature response of the circuit so that a virtually temperature-independent oscillator frequency results.

These possibilities are very exciting, but only those radio amateurs can carry out their own experimentations who have access to the corresponding laboratory equipment.

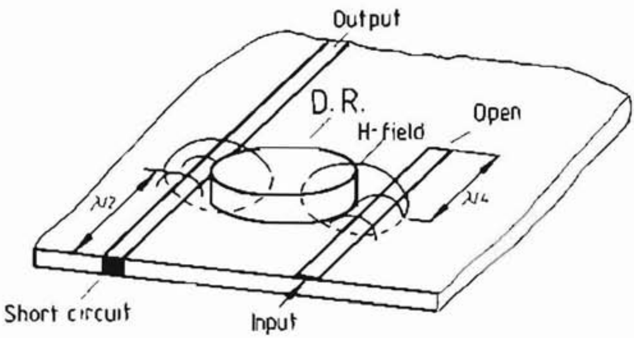

Fig. 2:
Coupling of the magnetic fields of striplines with those of the dielectric resonator

1.1. Coupling the Resonator to the Circuit

Due to the fact that fields – especially the magnetic field – are also present outside the resonator, it is possible for it to be coupled easily to a stripline. As can be seen in **Figure 2**, the resonator is mounted in the vicinity of a stripline so that the magnetic fields can couple between line and resonator. The degree of coupling can be adjusted with the aid of the spacing between line and resonator, which is most easily made with the aid of a PTFE-foil between resonator and the PC-board material of the stripline. The foil will also increase the spacing of the resonator from the metallic ground surface of the stripline, which reduces the heavy current losses due to the resonator field, and will improve the Q.

The coupling to the line is made best at a point of maximum current, which is λ/4 from the end of an open, or λ/2 from the end of a short-circuited line, since the magnetic field is strongest at this position. The matching to the line is made by altering the degree of coupling.

To construct high-quality filters, several dielectric disks can be combined together that are directly coupled to another in a series circuit (1), (2).

1.2. Tuning the Resonant Frequency

The resonant frequency of the dielectric disk is altered on mounting it onto a stripline circuit, since the metallic edge of the ground surface of the stripline PC-board material and the case will limit the fields. This effect is used for tuning the resonator. **Figure 3** shows a drawing of the installation of a dielectric resonator in a metallic cavity. With the aid of the tuning screw, it is possible to reduce the free space over the resonator, which increases the resonant frequency. This method allows one to tune over the whole 3 cm band without difficulties.

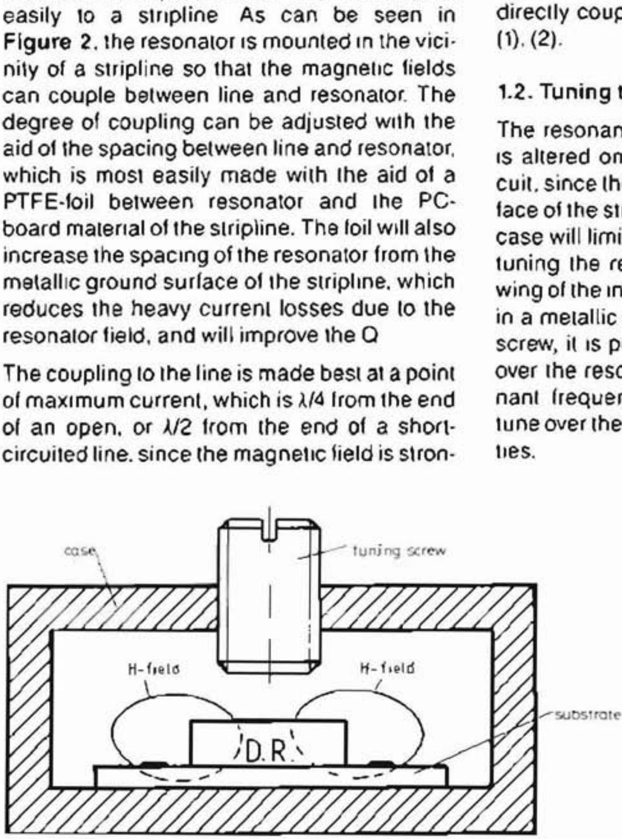

Fig. 3:
Mechanical tuning of the resonant frequency

The disadvantage is that the additional heavy current losses in the tuning screw reduce the operating Q. It is therefore advisable to polish the end of the screw and silver-plate it. A further disadvantage was found in that the temperature coefficient was dependent on the position of the screw. Theoretically speaking, it is most certainly possible to provide a compensation over a wider tuning range, however, this is so difficult in practice that it was not attempted by the authors.

2. DIELECTRIC STABILIZED OSCILLATOR (DSO)

The dielectric resonator with its stable temperature characteristics and its high Q allows the construction of compact, microwave oscillators having a high stability, good efficiency, and low cost. If a GaAs-FET is used as active component, this will result not only in relatively high efficiency but also in a low dependence of the frequency on the operating voltage and load. It is easily possible to achieve a mechanical tuning over a range of more than 500 MHz in the 3 cm band.

Such oscillators have been offered as modules by Mitsubishi since 1981. They are used as security systems at 10.525 GHz and are available at interestingly low prices (3). The doppler module FO-DP 12 KF and the receiver module FO-UP 11 KF were examined to establish their suitability for amateur communications in the 10 GHz band. DB 1 NV constructed a transceiver with the former module that will be described in a later edition of VHF COMMUNICATIONS.

In the case of both modules, the whole circuit

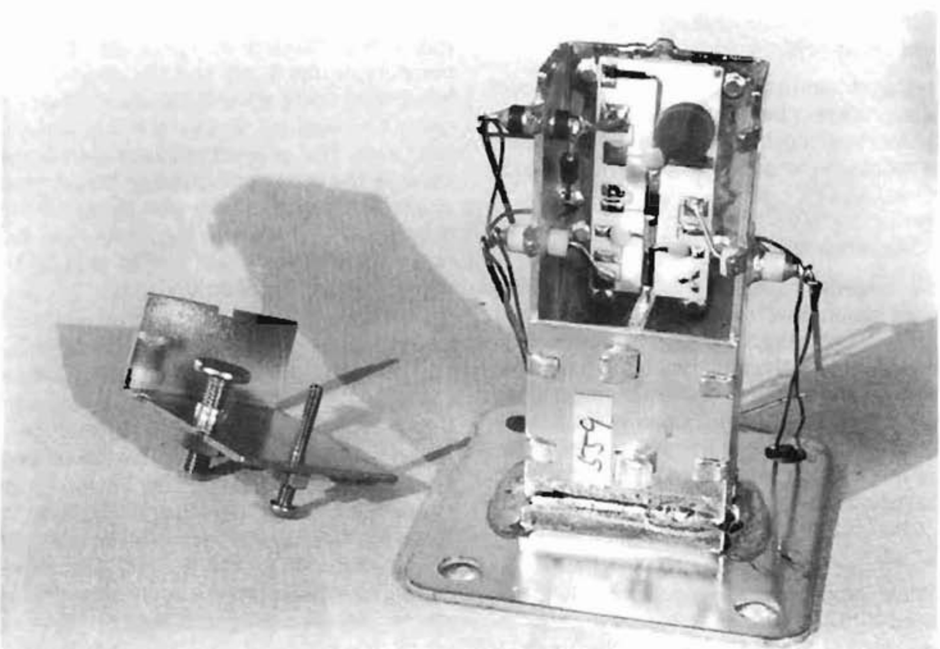

Fig. 4: The construction of a DSO manufactured by Mitsubishi

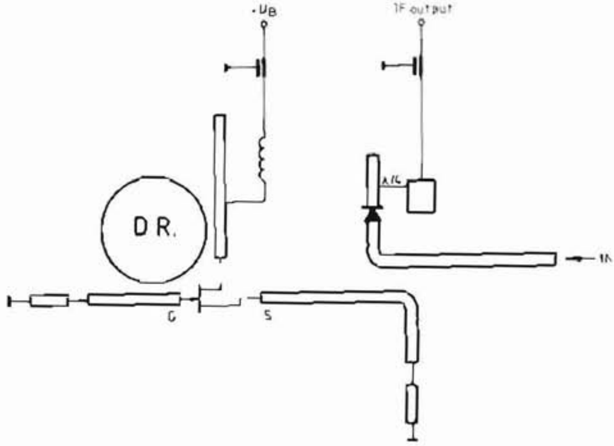

Fig. 5:
Principle circuit diagram of the receiver module FO-UP 11 KF

is realized on a ceramic substrate using a microstripline technology and is directly mounted into a waveguide with flange (**Figure 4**). The operating voltages are fed into the circuit via feedthrough capacitors, as is the IF-signal. A screw is provided for tuning, however, it should be replaced by a micrometer thread if the module is not to be used for a fixed frequency.

Further screws or threaded holes are provided for alignment of the mixer for minimum noise, and for matching, or – in the case of the receiver module – for suppression of the oscillator signal.

2.1. Receiver Module FO-UP 11 KF

The receiver comprises a Schottky diode mixer together with a dielectrically-stabilized FET-oscillator. As can be seen in the circuit given in **Figure 5**, the transistor is fed back from the drain line via the dielectric resonator to the gate line. A directional coupler at the source couples out the oscillator signal and feeds it to the mixer diode, which is also provided with the receive signal. The IF-signal is passed via a $\lambda/4$-choke for microwave frequencies to the output.

In its original state, the receiver module was tuned to 10.465 GHz which results in a receive frequency of 10.525 GHz when using an intermediate frequency of 60 MHz. The tuning screw allowed a frequency variation of the oscillator from 10.15 GHz to 11.4 GHz.

The mixer current was in the range of 2 to 3 mA, and the current drain of the oscillator amounted to 50 mA at U_B = 6 V, which is far less than when using the corresponding Gunn oscillator.

When using an input frequency of 10.360 GHz and an IF of 100 MHz, the conversion loss was measured to be 9 dB, and the double-sideband noise figure of the mixer was 10 dB when using a subsequent receiver with a noise figure of 2 dB. The radiated oscillator level at the input of the mixer amounted to 80 μW and could be reduced to zero with the aid of the tuning screw. This screw also influences the noise figure of the mixer and is aligned for minimum noise in the factory.

In a second module, the oscillator frequency was aligned to 10.514 GHz, and was provided with a subsequent IF-amplifier equipped with a GaAs-dual-gate-FET, similar to that described in (4). **Figure 6** shows the circuit and **Figure 7** gives the gain and DSB noise figure for an intermediate frequency of 135 MHz to 155 MHz. This corresponds to an input signal range of 10.359 to 10.379 GHz. The mixer current for the lowest noise figure amounted to 1.5 mA, and the operating voltage of the receiver module was 6.2 V together with a total current drain including the stabilization, and the IF-amplifier, of 67 mA.

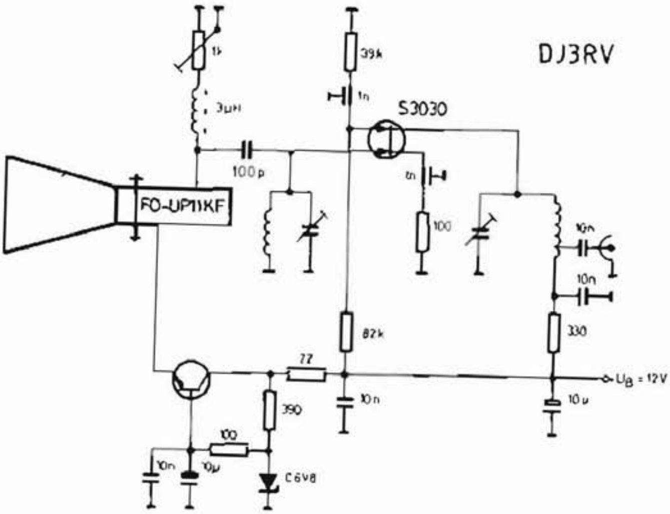

Fig. 6:
Receiver module with amplifier for a 144 MHz IF

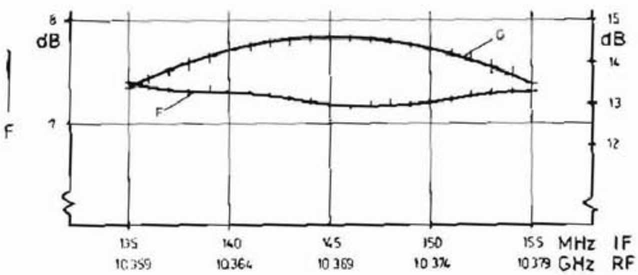

Fig. 7:
Noise figure and gain of the 10.3 GHz receiver module as shown in Figure 6

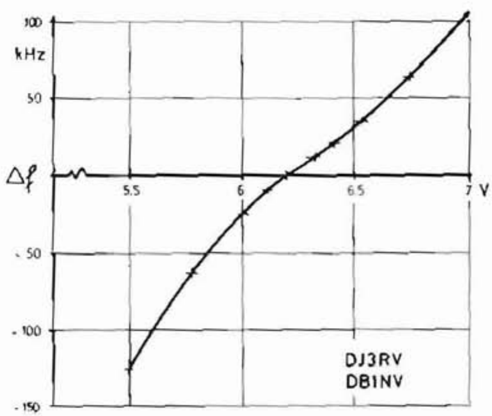

Fig. 8:
Frequency deviation as a function of operating voltage fluctuations

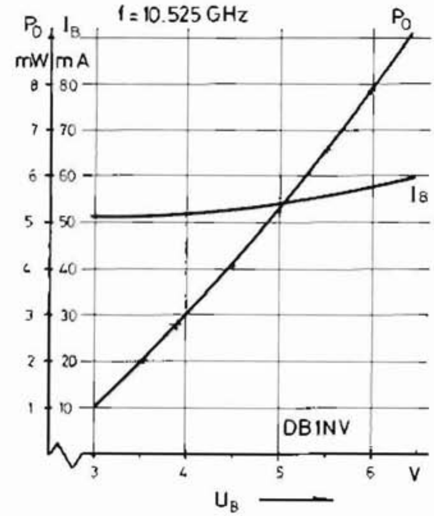

Fig. 9:
Output power and current drain as a function of the operating voltage

In order to determine the frequency stability as a function of the operating voltage, the voltage stabilizer circuit was disconnected and the receiver module was directly driven. As can be seen in **Figure 8**, the measured module was considerably better than the value of 2 MHz/V given by the manufacturer. The diagram shows a slightly positive coefficient, and the deviation can easily be compensated for with the aid of an AFC in the subsequent receiver.

During this measurement, the matching screw was aligned for minimum noise and it was determined by chance that the frequency behaviour became worse to the value of approx. –400 kHz/V when aligning for minimum oscillator radiation, and this led to negative coefficients. This is a further reason why one should not align the matching screw at random.

The typical temperature drift is given by the manufacturer as 10 MHz in a range of –30°C to +70°C which corresponds to an instability

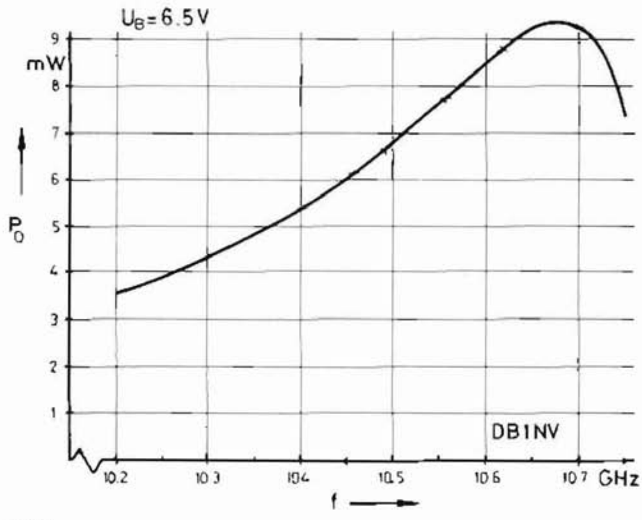

Fig. 10:
Output power as a function of frequency

of 100 kHz/°C. Measurements made in a test chamber at 0°C and +40°C showed a frequency variation of −1.2 MHz, which shows that this value is considerably better than specified for the original frequency. If, on the other hand, one detunes the oscillator frequency, this will have a negative effect on the temperature behaviour! Exact measurements were not carried out, since the oscillator tuning screw was first to be replaced by a micrometer thread.

2.2. The Doppler Module FO-DP 12 KF

The doppler module was designed for determining moving objects in security systems. It operates according to the principle of a straight-through mixer. The transmit signal is used simultaneously for converting the receive signal, in the same manner as when using the wellknown "Gunnplexer". The doppler module is to be used for the same type of operation, and it is interesting to know the output power, frequency stability, and the tuning range characteristics.

Figure 9 gives the output power and current drain as a function of the operating voltage. A further module was used at the original frequency and at the recommended operating voltage of U_B = 6.5 V recommended by the manufacturer. It provided an output power of 13 mW at a current drain of 44 mA.

One will notice the high efficiency of the oscillator with respect to Gunn diodes; the power requirements are approximately only one third that of a Gunn oscillator of the same power.

Unfortunately, the output power is very dependent on the output frequency, as can be seen in **Figure 10**. In the amateur band in the order of 10.35 GHz, at least 4 mW RF-output power was available, which could be increased to 7 mW when using a matching transformer between module and antenna. Attention should be paid that the mixer diode current does not drop considerably, since the conversion loss will then increase to impermisssible levels. For this reason, the manufacturer recommends that the mixer diodes are operated with a bias current of approximately 1 mA.

The stability of the output frequency with respect to operating voltage fluctuations is better than 200 kHz/V; the frequency can be shifted by a maximum of 800 kHz by load reflections. The frequency drift as result of temperature is negligible when compared with resonator-controlled Gunnplexers.

The high frequency stability of the DSO makes it virtually impossible to use the conventional type of modulation or fine tuning with the aid of the operating voltage. However, it allows the AFC-operation to be made at IF-level, which simplifies the circuit.

As can be seen in the photograph given in Figure 4, the doppler module is provided with two mixer diodes, which can be used for modulation. One feeds in a constant current and the AF-signal is coupled in capacitively.

Since the doppler module is used in the transceiver constructed by DB1NV, as has been previously mentioned, the operation can be studied there in more detail.

3. REFERENCES

For those readers that want to read more on dielectric resonators, we would like to recommend the following articles (1) and (2):

(1) J.K. Plourde, C.-L. Ren:
Application of Dielectric Resonators in Microwave Components
IEEE Trans. MTT, Vol. 29, No. 8, P 754-770

(2) K. Pöbl, G. Wolfram:
Dielektrische Resonatoren, neue Bauelemente der Mikrowellentechnik
Siemens Components 20 (1982)
Edition 1, pages 14–18

(3) Microwave GaAs-FET's Modules
Stabilized Oscillators and Sensor Modules
Mitsubishi Electric Corp., Tokyo

Erich Stadler, DG 7 GK

Determining the Antenna Gain in the GHz Range

It is possible in the microwave range to measure antenna gain with a few, simple aids (transmitter, directional coupler, diode probe, voltmeter), and when one has two identical antennas available. The described method can be carried out indoors.

The free-space attenuation a_{fs} can be exactly calculated for line-of-sight propagation in the VHF, UHF, and GHz range:

$$a_{fs/dB} = 20 \lg (4 \pi d/\lambda) \qquad (1)$$

where:

d = Distance between transmitter and receiver

λ = Wavelength of the transmit frequency

This equation is, however, only valid when an imaginary isotropic radiator is assumed. It shows the attenuation that is present due to the "waste" of transmit energy in all directions, and by not using it in the optimum means of concentrating the energy using a directional antenna.

Of course, radio amateurs use antennas that possess a certain gain (even a dipole has 2.2 dB gain over an isotropic radiator). In this case, one will obtain a different free-space attenuation a_{tr} between transmitter and receiver that takes the gain of the transmit antenna – $G_{1/dB}$ – and the gain of the receive antenna – $G_{2/dB}$ – into consideration which can be calculated as follows:

$$a_{tr/dB} = 20 \lg (4 \pi d/\lambda) - G_1 - G_2 \qquad (2)$$

This loss can, in contrast to a_{fs}, be actually measured. The attenuation a_{fs} cannot be measured in practice, but only calculated, since the isotropic radiator is not realizable in practice. The attenuation a_{tr}, on the other hand, is obtained by measuring the tansmit power P_{tx} and the receive power P_{rx} and by use of logarithms:

$$a_{tr/dB} = 10 \lg (P_{tx}/P_{rx}) \qquad (3)$$

The principle of the antenna gain measurement is as follows:

The power levels of P_{tx} and P_{rx} are measured and one calculates then the attenuation a_{tr} according to equation (3). The first expression in (2) corresponds to the free-space loss a_{fs} in the case of isotropic radiators, and can be calculated according to (1). This leaves the sum of G_1 and G_2 if one converts equation (2):

$$G_1 + G_2 = a_{fs} - a_{tr} \qquad (4)$$

It is now important that the transmit and receive antennas are identical! In this case, G_1 is equal to G_2, and one will only need to divide the result of equation (4) by half to obtain the gain of one of the antennas.

A horn radiator was now measured according to the described method at 8 GHz and the result was as follows:

At 8 GHz, λ is 3.75 cm. The spacing between the two horn radiators amounted to d = 128 cm. The transmit power was measured to be 32 mW, and the receive energy was 0.32 mW According to equation (4) the following will result:

$$G_1 + G_2 = 52.6 \text{ dB} - 20 \text{ dB} = 32.6 \text{ dB}$$

This means that each of the two antennas exhibits a gain of 16.3 dB.

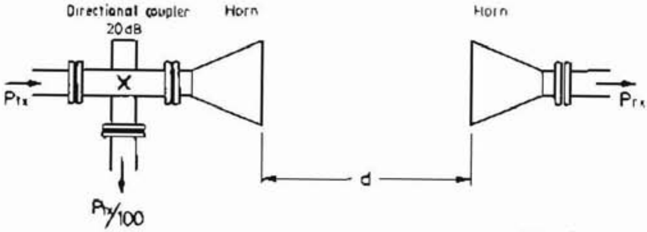

Fig. 1: The distance d is selected so that $P_{rx} = P_{tx}/100$ in other words, that a_{tr} is exactly 20 dB.

The described measurement can also be carried out without absolute power measurement in a more elegant manner if a directional coupler is available. It is used to determine the transmit power. If, for instance, the directional coupler possesses a coupling attenuation of 20 dB, it is possible to select a_{tr} = 20 dB directly by placing the antennas only so far from another that the signal at the coupling output and at the output of the receive antenna are exactly equal. This is shown in **Figure** 1. The signals can be measured one after another using the same diode probe (test demodulator) and using the same operating point! This means that no power meter is required!

The accuracy of the measurement is dependent on the exact knowledge of the couple attenuation of the directional coupler. The distance 'd' is not given, but measured after determining the correct attenuation between transmitter and receiver. The given result coincides well with the data-sheet specifications of the antenna manufacturer. The gain measured in this manner is referred to an isotropic radiator, and not against a dipole, as is usually the case with professional, amateur, and other antennas.

Werner Hanschke, DC0RZ

A 30 MHz FM-Receiver for SHF Receive Systems

This FM-receiver was designed as IF-strip for 10 GHz and 24 GHz stations. It can be used with the well-known microwave receive systems (1), (2) with straight-through mixer, or the well-known Gunnplexer to form wideband, simple superhet receivers with a variable oscillator. The IF-strip is equipped with only three integrated circuits, two dual-gate MOSFETs, as well as a transistor for voltage stabilization (see Figure 1).

1. CIRCUIT DESCRIPTION

As can be seen in **Figure 2**, the RF-input stage is equipped with a low-noise dual-gate MOSFET, type BF 961, whose input and output circuit are dampened with 330 Ω resistors in order to avoid self-oscillation.

This is followed by the well-known integrated

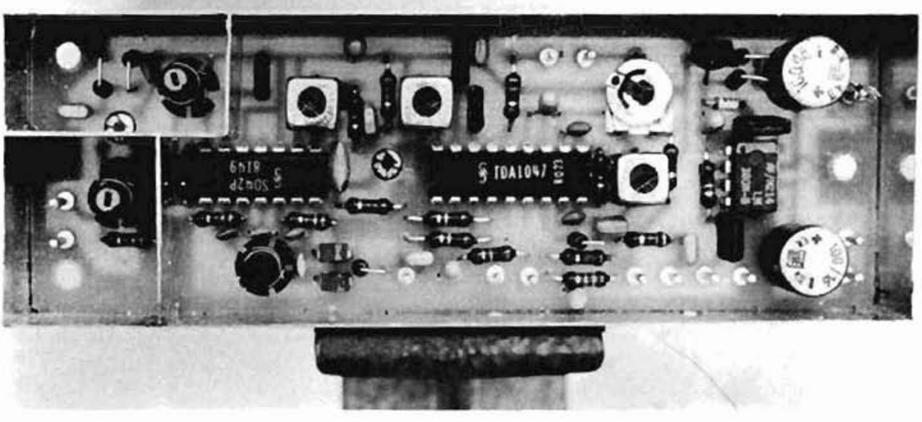

Fig. 1: Variable wideband FM-receiver (30 ± 1 MHz) for microwave transceivers

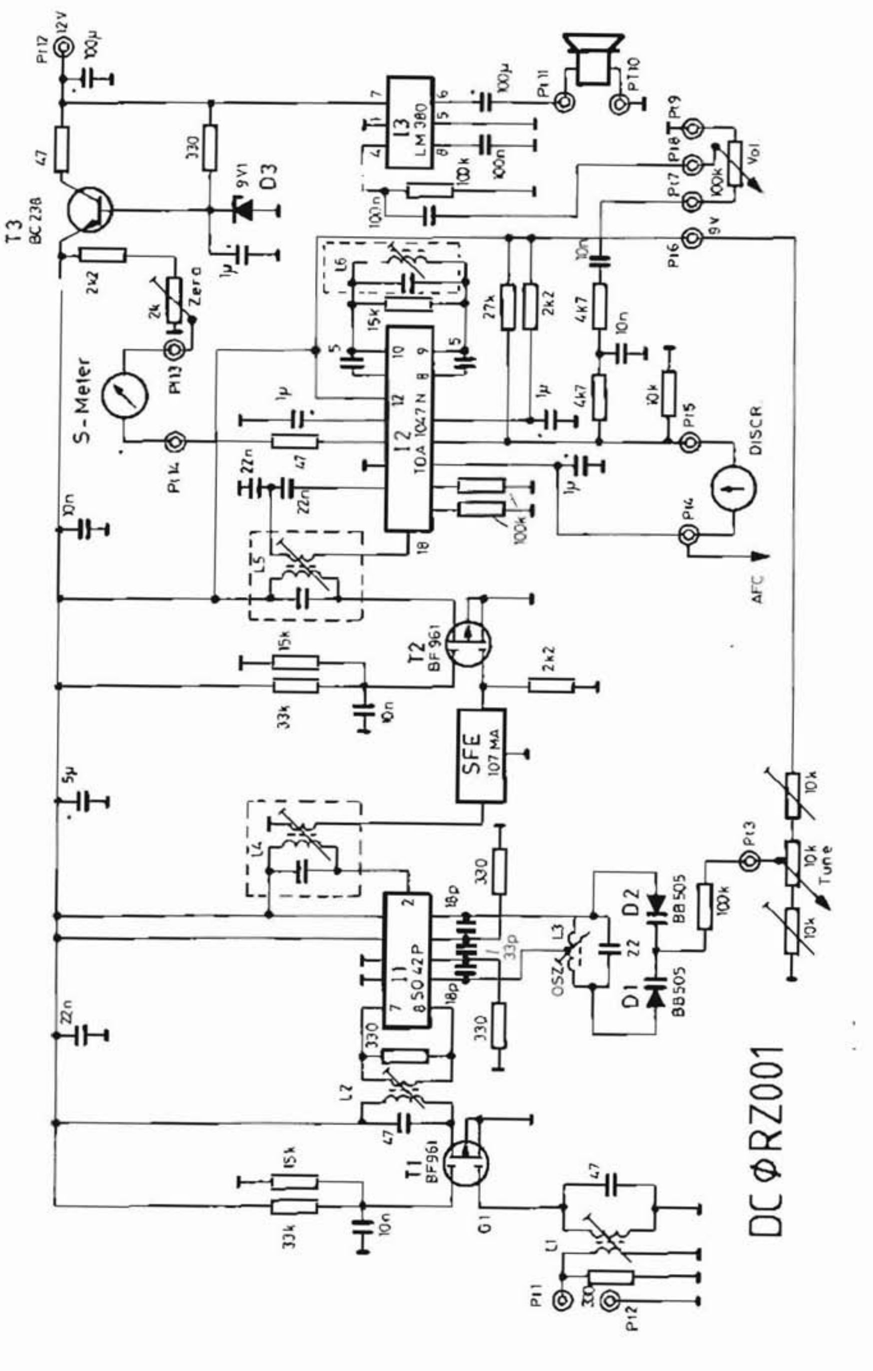

Fig. 2: Circuit diagram including the external components required with the 30 MHz FM-receiver.

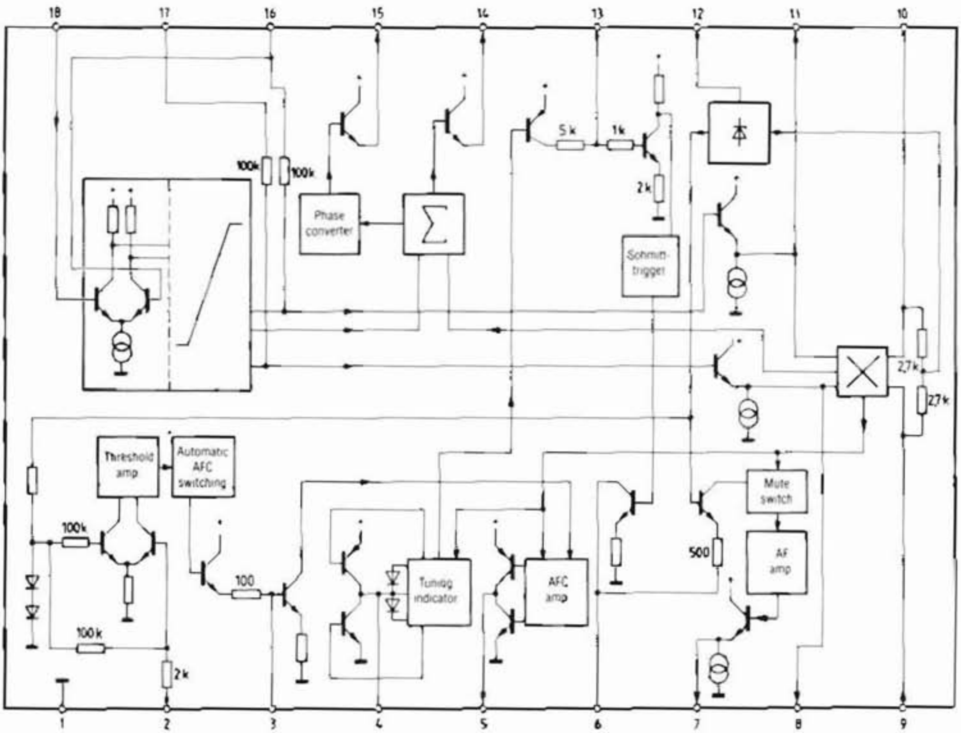

Fig. 3: Block diagram (TDA 1047)

push-pull mixer S042P together with a voltage-controlled oscillator. The frequency of this oscillator is 10.7 MHz over the receive frequency and it can be tuned by ±1 MHz. A crystal was not used at this position in order to allow an electronic fine tuning. This is used in addition to the tuning of the microwave oscillator.

The intermediate frequency signal is passed via a resonant circuit and a ceramic filter to an IF-preamplifier, which is also equipped with the low-noise BF 961. The subsequent integrated IF-amplifier and demodulator type TDA 1047 has several special features which are to be mentioned briefly here.

The TDA 1047 (see **Figure 3)** possesses outputs for an S-meter, a discriminator-meter, an automatic frequency control (AFC), and for a squelch in addition to the required FM-demodulator. All these facilities are utilized with the exception of the squelch. Since experience has shown that weak signals are often not heard when the squelch circuit is closed, this was not used in our application. Instead of this, the possibility was used to suppress the noise by approximately 20 dB automatically. The gain and thus the noise only increase to their full level on receiving a signal.

As is always the case with integrated FM-IF circuits, a DC-voltage is also present at the S-meter output even under zero signal conditions. This means that an opposite voltage must be used to compensate this if the S-meter is to indicate zero. This was achieved using the trimmer potentiometer connected to P1 13 and designated "zero".

The AF-voltage from the FM-demodulator comprising L 6 is passed via a low-pass filter (4k7 – 10n – 4k7) and a high-pass filter (10n – 100 kΩ) to the integrated AF-amplifier LM 380. The output of this integrated circuit is designed for a 8 Ω loudspeaker.

2. COMPONENT DETAILS

T 1, T 2: BF 961 (Siemens)
T 3: BC 238 B, BC 413, BC 550
I 1: S042P (Siemens)
I 2: TDA 1047 N (Siemens)
I 3: LM 380N-8 (National Semiconductors)
D 1, D 2: BB 105 G, BB 505 G (Siemens)
D 3: C9V1 zener diode
L 1, L 2: 11 + 2 turns of 0.4 mm dia enamelled copper wire, wound on a special coil former Sp 3.5/14.6-2348 C with 3.5 mm violet core
L 3: 2 + 4 turns, otherwise as L 1, L 2
L 4, L 5, L 6: Japanese 10.7 MHz resonant circuits, 7 mm × 7 mm, colour code green
F 1: Ceramic filter SFE 10.7 MA
All resistors for 10 mm spacing, 5% tolerance
Ceramic capacitors: Values up to 10 nF 2.5 mm spacing
22 nF and 100 nF: 5 mm spacing
Metal box: 41 × 121 × 28 mm

3. CONSTRUCTION DETAILS

Figure 4 shows the single-coated PC-board DC0RZ 001 which was designed for accommodating this FM-receiver. The dimensions are 39 mm x 119 mm and the board is mounted into the mentioned metal box with a spacing of 5 to 7 mm from the base plate. A straight and a bent piece of tin plate form two screening panels for the 30 MHz input circuit. They are approximately 18 mm high and can be easily seen in the photograph. They should be installed after drilling, but before mounting the components onto the PC-board. 4.5 mm holes are drilled into the board for accommodating the two dual-gate MOSFETs. The integrated circuits are soldered onto the board without sockets.

4. ALIGNMENT, MEASURED VALUES

The alignment is very simple! Firstly adjust the S-meter with the aid of the 2 kΩ trimmer potentiometer so that a reading of approximately S 1 is indicated. After this, align the resonant circuits comprising L 4 and L 5 for maximum noise in the loudspeaker, after which the discriminator inductance L 6 is aligned so that the discriminator-meter indicates zero under non-signal conditions.

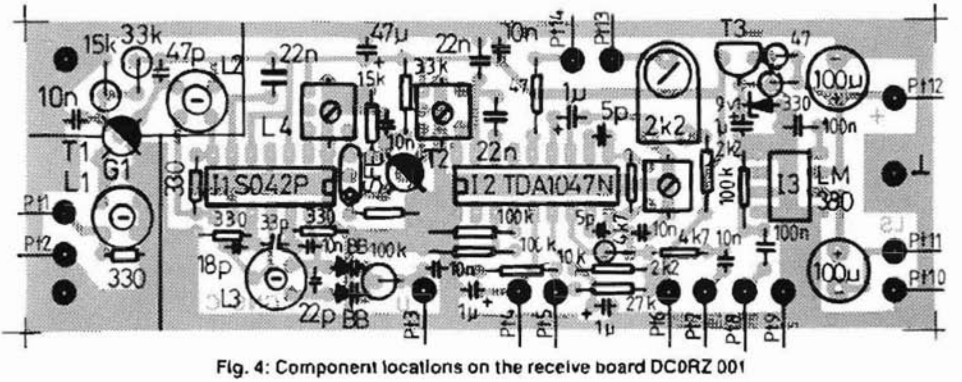

Fig. 4: Component locations on the receive board DC0RZ 001

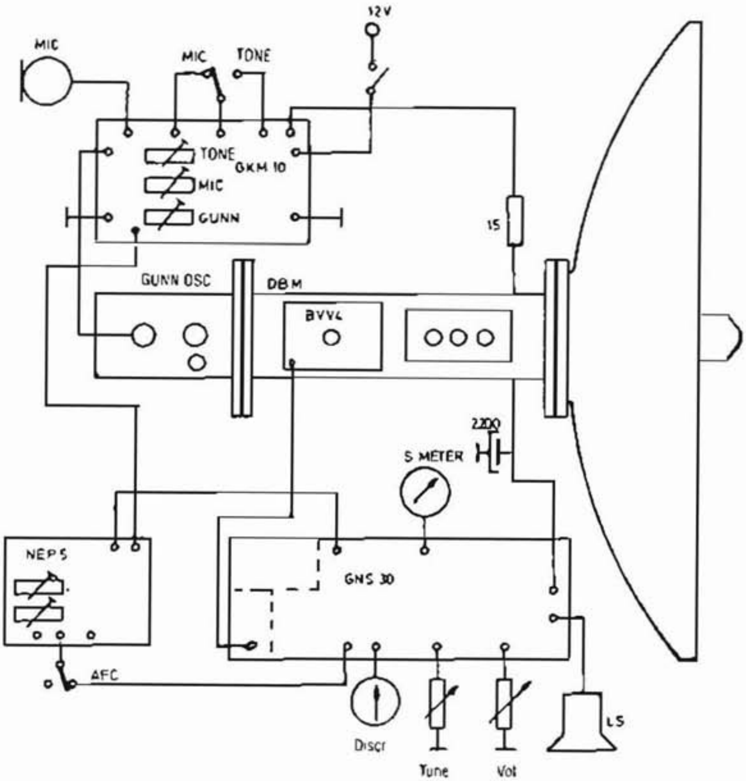

Fig. 5: A transceiver for 10 or 24 GHz using modules designed by DC0RZ

The following average values were measured on a number of receivers:

Operating voltage range:	12 – 15 V
Operating current:	50 mA
Sensitivity: For 20 dB (S+N)/N:	0.5 µV
Frequency range, variable:	29 – 31 MHz
IF-bandwidth (–3 dB):	150 kHz
Automatic noise suppression:	–20 dB
S-meter:	100 to 500 µA
Discriminator-meter:	± 50 to ± 200 µA
AFC-output:	≤ +4 V
AF-output power:	1 W into 8 Ω

One will now require a frequency-modulated signal at the input – preferably from a signal generator. The external tuning potentiometer is then adjusted to its center position and the oscillator is aligned at L 3 to 40.7 MHz (signal generator to 30 MHz). Align the oscillator to 41.7 MHz (signal generator to 31 MHz), the input circuit comprising L 1 should be aligned for maximum S-meter reading, and the intermediate circuit comprising L 2 for maximum at 39.7 (29) MHz.

The two limiting resistors of the tuning potentiometer are adjusted so that the potentiometer is just at its stops at 29 and 31 MHz respectively. These can be replaced by fixed resistors subsequent to the alignment, if required. The whole RF-IF alignment should be repeated several times to optimize the alignment.

Finally **Figure 5** shows how the described receiver module can be used in a 10 GHz or 24 GHz transceiver. The other modules originate from the same author and their designations are as follows:

GKM 10: Gunn-supply, microphone amplifier, and dot generator.
DBM: Straight-through mixer
BVV 4: 30 MHz preamplifier
NEP 5. AFC tuning unit
GNS 30: 30 MHz FM receiver

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(1) J. Reithofer, DL6MH.
A Transceiver for the 10 GHz Band
VHF COMMUNICATIONS 11,
Edition 4/1979, Pages 208–215

(2) J. Reithofer, DL6MH:
A Straight-Through Mixer for 24 GHz
VHF COMMUNICATIONS 14,
Edition 2/1982, Pages 99–105

Hans Wessels, PA 2 HWG

A 6 cm Preamplifier Equipped with the MGF 1400, and a Push-Pull Mixer for Transmit and Receive

I had wanted to design something for the 6 cm band for several years now, since this band is relatively unknown to radio amateurs. The reason why it did not come to this was because no measuring equipment was available for this frequency range in the order of 5.7 GHz. Recently, I was attached to a department of my company (unfortunately only for a short period of time) and had all the measuring equipment necessary for measurements in the frequency range between 1 and 20 GHz. Of course, such a chance could not be ignored!

1. THE PREAMPLIFIER

Reproducible results are only obtainable when using microstripline technology. For this reason, the circuit was calculated on a TI-59 calculator and fed with the S-parameters of the GaAs-FET MGF 1400. The article described in (1) shows how this can be carried out.

The design of the board using computer graphics and tables, as well as the required preparation and photography was only a matter of hours. All calculations were made for 0.79 mm thick PTFE material with $\varepsilon_r = 2.55$.

1.1. Construction

The upper and the lower side of the board are connected together on the long sides of the PC-board shown in **Figure 1** using strips of copper foil. Solder should be only used thinly.

Since a PTFE-board of the thickness used does not possess a considerable mechanical stability, it is soldered to a 0.5 mm thick brass plate having the same dimensions. This can be carried out well using a household iron. Finally, the holes for the source bypass capacitors are drilled, so that they can be grounded with the connections as short as possible. This can be carried out best with the aid of a dentist drill or with a small fine file. Further details are given in **Figure 2**.

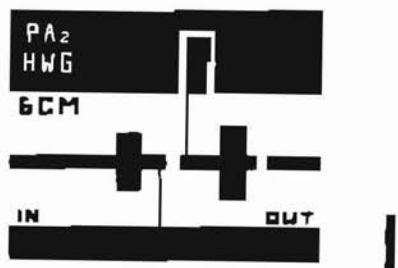

Fig. 1:
Double-coated RT/duroid PC-board of a MGF 1400 preamplifier for the 6 cm band

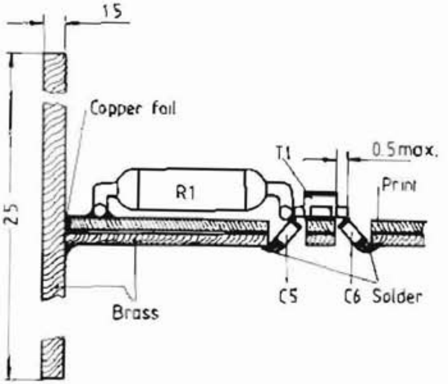

Fig. 2:
Critical component installation in the vicinity of the preamplifier transistor

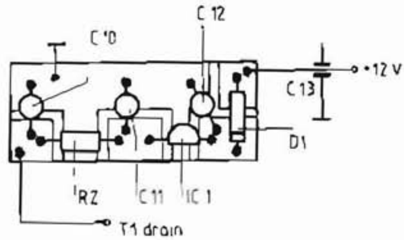

Fig. 3:
The small supply board can be made simply, as required

The holes for the SHF-connectors and the feedthrough capacitor are now drilled into the 25 mm wide strip of 1.5 mm brass plate, after which it is bent and soldered around the PC-board.

It is important that the holes for the center pins of the connectors are exactly 2.3 times the diameter of the pins themselves, so that these transitions maintain the 50 Ω impedance. The connectors are now mounted into place, and their center pins are soldered. Finally, solder in the source bypass capacitors to ground so that their other side is at the same height as the upper side of the board. Attention should be paid that the spacing to the case of the MGF-1400 is a maximum of 0.5 mm. This is indicated in **Figure 2**.

One end of the source resistor is now soldered to one of the source bypass capacitors, and its other end to ground. After this, the two drain

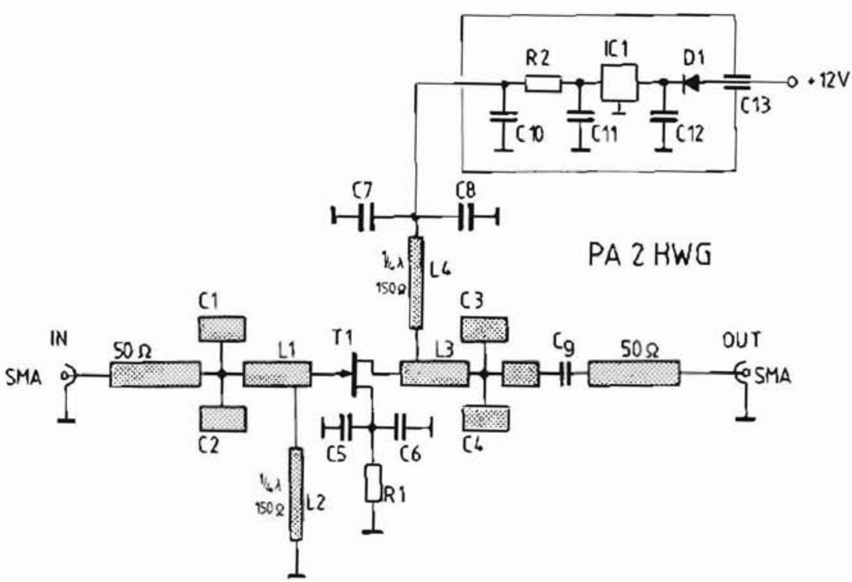

Fig. 4: Circuit diagram of the 5.7 GHz preamplifier

decoupling capacitors are mounted This is made directly after the transition of the $\lambda/4 - 150\ \Omega$ line to a wider line.

This is followed by soldering in the feed-through capacitor for the operating voltage, and finally soldering the supply board according to **Fig. 3** on the lower side to the brass plate. One then only requires a connection between the supply board and the drain connection point, after which T 1 is soldered into place. The preamplifier is completed after providing a brass or tin plate cover.

1.2. Circuit

The circuit diagram is given in **Figure 4** The circuit generates its own bias voltage, which means that it is only necessary for the values of R 1 and R 2 to be determined experimentally Resistor R 1 is selected so that a drain current of 10 mA is achieved; R 2 determines V_{DS} This voltage should amount to 2.5 to 3 V. Orientation values for these two resistors are given in Section 1.3

1.3. Components

T 1.	MGF 1400 (Mitsubishi)
D 1	1N4001 (various manufacturers)
I 1:	78L05 (various manufacturers)
L 1 – L 4:	Printed inductances
C 1 – C 4:	Printed inductances
C 5 – C 9:	150 pF multilayer capacitor (Philips type 2222 851 13151)
C 10:	2.2 µF/25 V tantalum cap.
C 11, C 12:	1 µF/25 V tantalum cap.
C 13:	1 nF feedthrough capacitor, solder mounting
R 1:	approx. 100 Ω (see text)
R 2.	approx. 120 Ω (see text)

1.4. Alignment and Operation

No SHF-alignment is required. The adjustment of current and voltage have already been described.

The described amplifier can be used both as receive preamplifier in front of the mixer which is to be described later, or as transmit amplifier subsequent to the mixer. Of course,

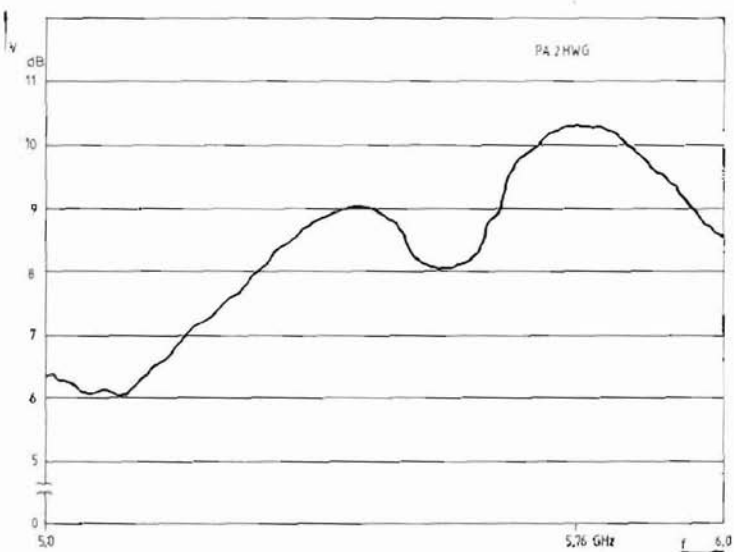

Fig. 5: The gain in the 6 cm amateur band amounts to approximately 10 dB

the receive preamplifier should be mounted as near to the antenna as possible

In the transmit mode, it is advantageous to use two such amplifiers in series, where by the drain current of the second amplifier can be increased to 30 mA when using a drain source voltage V_{DS} of 3 V.

In this case, the output power is in the order of 20 mW. It can be calculated as follows:

Maximum IF-input power:	0 dBm
Conversion losses:	−7 dBm
Gain of the two stages:	+20 dB
Transmit power max.	13 dBm

This is quite a power for this frequency, however, it is now possible using a single GaAs-FET to achieve 3.5 W in this range! MSC manufactures such transistors, but they cost more than DM 2000,– in June 1982, and this is too expensive for amateur applications. The antenna still represents the cheapest amplifier, and a relatively small parabolic antenna will provide at least 20 dB in this frequency range, which provides an effective radiated power of 2 W when used in conjunction with the described amplifier. This allows communication to quite a range

1.5. Measured Values:

The frequency response of the amplifier is given in Figure 5: the gain amounts to at least 10 dB at the required frequency. At the same time, the return loss was also measured; it amounted to more than 15 dB (VSWR < 1.4) at the input and output.

The noise figure amounted to 4 dB.

2. THE MIXER

The most popular mixer applied for the 6 cm band is a single-diode arrangement mounted in a waveguide. The described mixer differs from this in three points:

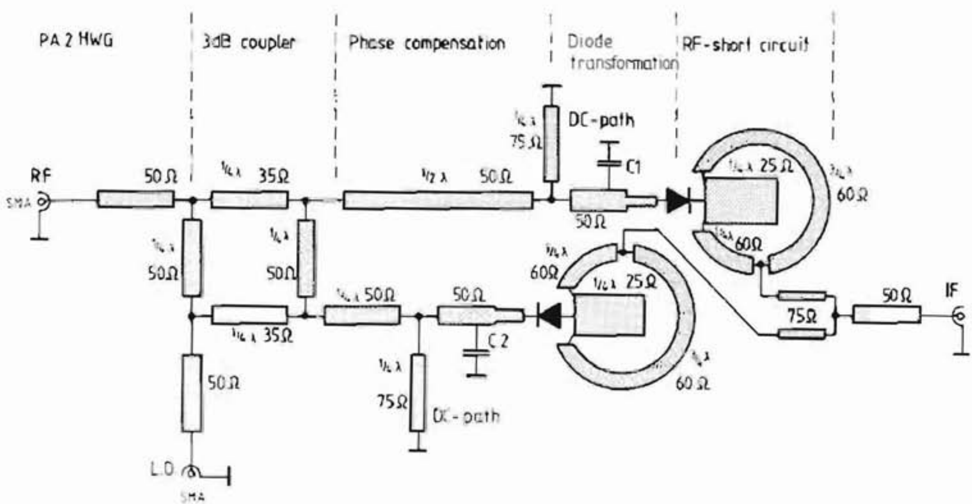

Fig. 6: Push-pull mixer for the 6 cm band in microstrip technology

Firstly, a push-pull mixer is to be used;

Secondly, the mixer is built up on a PTFE-board of 0.79 mm thickness with an ε_r of 2.55, which reduces the mechanical problems to a minimum;

Thirdly, the mixer is suitable both for receive and for transmit applications and for all modulation modes.

2.1. Principle of Operation

The principle of operation is known from similar configurations for the 23 cm and 13 cm band. It is provided with a 3 dB coupler by which the diodes are driven with the frequencies to be converted (signal and oscillator frequencies).

The coupler is made in the form of a ring instead of the conventional square shape; the square shape showed considerable matching problems, that were not, or only very difficult to solve. The round version, on the other hand, does not exhibit these problems.

One point which must not be overlooked using this type of mixer is that of the phase which the SHF-signal must have with respect to the oscillator signal at the two diodes. The oscillator signal must have a phase shift of 180° at this point, and the SHF-signal 0°. **Figure 6** shows how an extra phase-shift of 90° is achieved between the output of the coupler and the DC-bypass.

2.2. Construction Details

Figure 7 shows the PC-board for the mixer module in the scale of 1:1. Generally speaking, the same is valid for the mixer as for the pre-amplifier. The upper and lower side of the board are provided with through-contacts on the long edges using a strip of copper foil. Since the ground surface of the upper side of this board is relatively narrow, the copper foil is not bent around, but it is left straight and the interconnection between PC-board ground and the foil is made with the aid of a little solder.

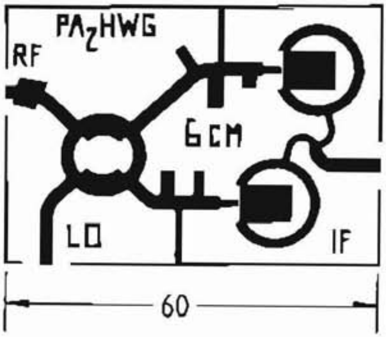

Fig. 7:
RT/duroid PC-board for the mixer circuit given in Figure 6 using the given diodes

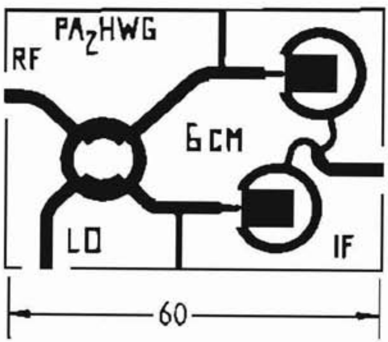

Fig. 8:
Basic board (before alignment) for diodes differing from those given

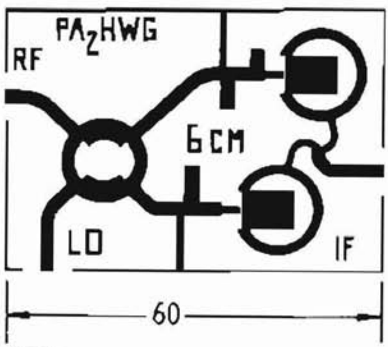

Fig. 9:
Converter board for other diodes after alignment

It is now possible for the board to be soldered onto a brass support, and afterwards mounted in a frame of brass plate into which the three connectors are mounted. This method is similar to that described for the amplifier.

2.2.1. Mounting the Diodes

The two beam-lead diodes type HP 5082-2767 (Hewlett Packard) or BAS 22 (Philips) are, unfortunately, easily damaged by static charge. In order to avoid this, the following sequence should be adhered to:

- The RF- and IF- port of the diodes should be shorted with a piece of wire before soldering into place.

- Each diode is placed into position with the aid of a wet cocktail stick onto which the diode is held. It is best brought into place with the aid of a magnifying glass (approximately 10 times).

- The soldering is carried out at the lowest possible temperature and within a short period of time. A 6 W miniature soldering iron is suitable for this.

- Indium solder is suitable that has a melting point of approximately 140°C. Any residual flux should be removed with nail-varnish remover.

2.3. Alignment

No alignment is required if the diodes given in the previous section can be obtained and the PC-board is as given in **Figure 7**. If other diodes are used, it is then necessary to use the PC-board shown in **Figure 8** and the alignment is carried out as follows:

Since the whole module operates in 50 Ω technology, it is necessary for the diodes to be transformed to 50 Ω. This is relatively simple, but can add several dB to the conversion loss. Several pieces of copper foil whose dimensions are 2 x 1.5 mm, 3 x 2 mm; 4 x 2 mm; and 5 x 2 mm are cut and are used for alignment. While receiving a test signal, one of these pieces of copper foil should be placed on the 50 Ω stripline between DC-voltage blocking and the connection point of the diode. With the aid of a wooden cocktail stick, the piece of copper foil is placed onto the stripline and shifted carefully back and forth until the most optimum signal is received at IF-level. If this is the case, the same experiment is carried out with a larger and a smaller piece of foil which are also carefully moved back and forth between DC-blocking and matching point of the diode. The piece of foil providing the best result is then finally soldered to the most optimum position with a minimum of solder.

The same method is repeated for the second diode. After this has been carried out, the PC-board can, for instance, look like **Figure 9.**

Since the impedance of the diode is also dependent on the oscillator power, this must remain constant during the alignment, and should not differ greatly during operation later. An orientation value for the oscillator power is 5 to 10 mW

2.3.1. SHF-Shortcircuit

If we are to transform one side of the diodes to 50 Ω, this is only possible when the other side possesses a shortcircuit to ground at the required frequency. This SHF-shortcircuit can be provided easily using an open $\lambda/4$ line, which converts the "infinite" impedance at its open end to an impedance of "zero" at the other end.

It is now only necessary for the IF-signal to be coupled out, without destroying the SHF-shortcircuit. This can be carried out using an open 1 λ-ring where the output coupling point on one side is $\lambda/4$ and 3/4 λ on the other side with respect to the RF-signal, which means that the RF-shortcircuit is not altered.

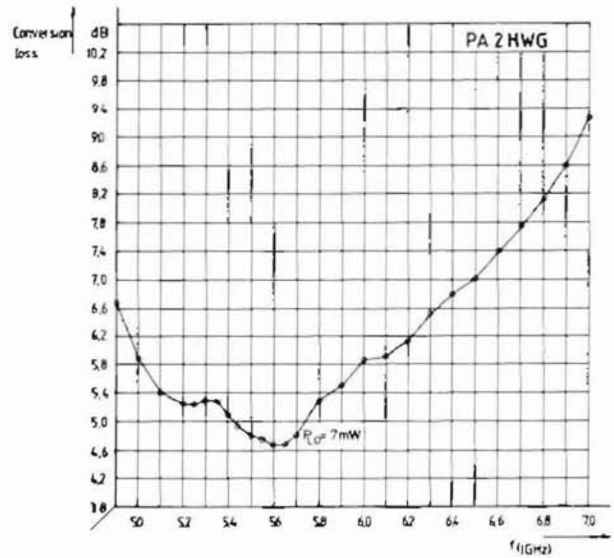

Fig. 10:
Conversion loss as a function of frequency with an optimum oscillator power of 7 mW

2.4. Selecting the Intermediate Frequency

For many reasons, only the 70 cm band is suitable for use as intermediate frequency. The mixer is relatively wideband, however, the next higher amateur band (23 cm) is already spaced so far that the oscillator signal is not within the passband range of the coupler. The attenuation of the oscillator frequency will then be so high that a considerably higher power is required.

A lower frequency such as the 2-m-band would place too high a demand on the input filter which is required in order to suppress the image frequency.

The manufacture of a good input, or output filter (transmit operation) was described by DC8EC in (4). It is only necessary for it to be recalculated for the 6 cm band.

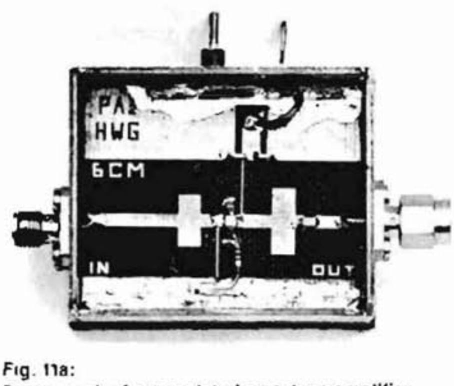

Fig. 11a:
Photograph of a completed prototype amplifier

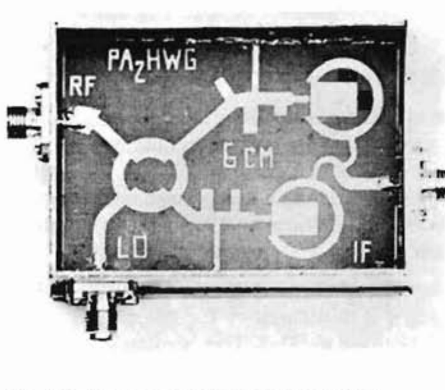

Fig. 11b: Photograph of a completed mixer

2.5. Measured Values

The conversion loss is shown in Figure 10 The return loss at all three ports is better than 18 dB, and the isolation between SHF and the oscillator port is better than 25 dB.

3. MEASURING EQUIPMENT USED

HP 8690 B Sweep Oscillator
HP 8693 B Plug-in 4–8 GHz
HP 11692 D Double Directional Coupler
HP 11664 A Detector
HP 8755 A Swept Amplitude Analyzer
HP 8750 A Storage Normalizer
HP 8555 A Spectrum Analyzer
Magnetic AB model 117 Noise-Test Setup
Magnetic AB model 125 B Noise Source

4. REFERENCES

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VHF COMMUNICATION 12, Edition 1/1980, Pages 2–13

(2) D V Morgan, M. J. Howes: Microwaves solid-state devices and applications
Published by Peter Peregrinus LTd

(3) Microwave Engineer's Handbook Vo 1+2
Theodore S Saad

(4) F. Schmehr, DC8EC: Coupled Microstriplines as Filter
VHF COMMUNICATIONS 13, Edition 3/1981, Pages 144–147

Gerhard Schmitt, DJ5AP

A Linear Transmit Converter for the 13 cm Band

A fully-transistorized 2320 MHz transmit converter is to be described that satisfies the following demands required by the author:

- **Low cost**
- **Use of readily available components**
- **No expensive power varactors**
- **Sufficient output power for portable operation: at least 100 mW**
- **Operating voltage 12 to 14 V, low current drain**
- **High suppression of all unwanted conversion products**
- **Use of an intermediate frequency of 144-146 MHz**
- **Simple construction without lathing or milling**

The author returned to a conventional construction after several attempts with different types of transmit mixers, especially power mixers. He then developed a transverter using metal chambers. Only the transmit converter shown in the photograph given in **Figure 1** is to be described, since suitable local oscillator modules for 2160 or 2176 MHz, and receive converters have already been described by other authors (1), (2), (4), (5), (7) and (10). The constructional and alignment information, as well as the number of references are given in a rather extensive manner, in order to enable new radio amateurs to become active on this interesting amateur radio band.

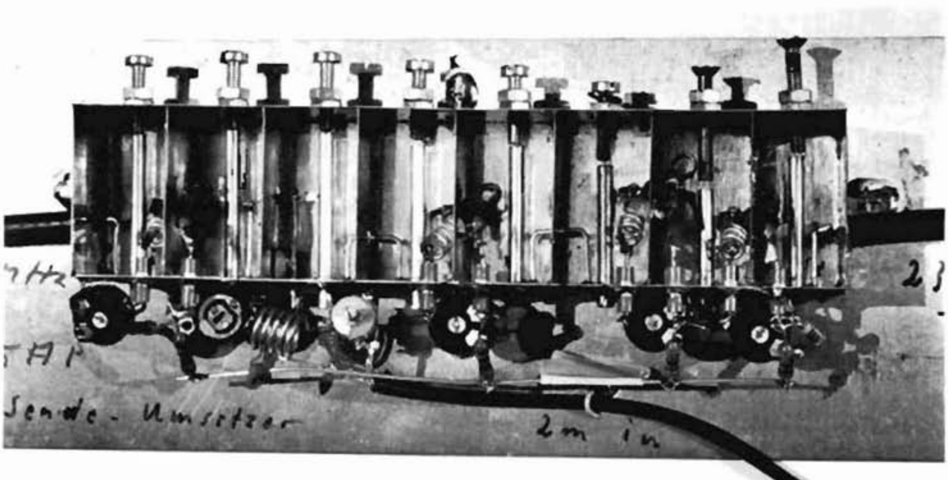


Fig. 1: Photograph of the transmit converter. It is soldered to the base plate of PC-board material with the aid of a few solder tags.

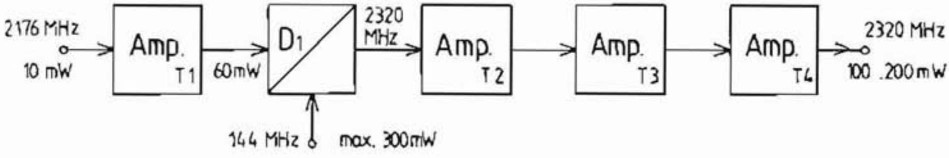


Fig. 2: Block diagram with nominal values of frequency and power

1. CONCEPT

The block diagram shown in Figure 2 shows that the local oscillator signal from the external oscillator module at 2176 MHz is firstly amplified up to a level of 50 to 60 mW. This is followed by the mixer equipped with the Schottky diode D 1. The 144 MHz drive signal is fed to the mixer diode via a 2 m resonant circuit. The mixer is followed by three, selective amplifier stages for the required frequency of between 2320 and 2322 MHz. Approximately 100 to 200 mW of SHF-power are available at the output.

2. CIRCUIT DESCRIPTION

As can be seen in Figure 3, all four transistors are in a common-emitter circuit, and use a variable, positive bias voltage (6). This allows the most favorable operating point to be adjusted individually

The mixer diode D 1 is provided with its local oscillator signal from L 2 via the coupling inductance LC 2. The 2 m drive signal is fed via a resonant circuit comprising L 9 and a 13 pF trimmer which transforms it up, and then via a SHF-choke to the coupling inductance LC 3. A ceramic disk capacitor without connection leads (approx. 8 pF) bypass LC 3 for SHF frequencies, but represents only a very small load at a frequency of 144 MHz. The DC-path

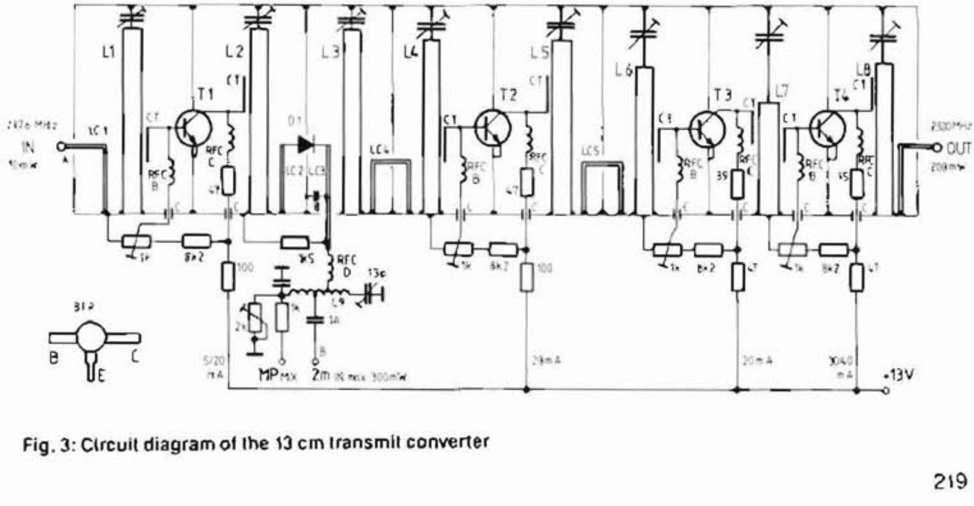


Fig. 3: Circuit diagram of the 13 cm transmit converter

for the mixer diode is closed via a fixed, and a trimmer resistor, the mixer diode current can be measured as the voltage across a choke resistor of 1 kΩ, when using a small meter

The mixer is followed by three linear amplifier stages. A high selectivity is provided by the two bandpass filters, and two individual circuits using coaxial chamber techniques.

2.1. Component Details

- T 1 – T 3: BFR 34 A, or BFR 90, BFR 91
- T 4: BFR 96, BFR 96 S
- D 1: HP 5082-2800 or other Schottky diode
- L 1 to L 8: λ/4 resonant circuits from 3 mm dia. brass or copper tubing, approximately 4 mm should be added (see 3.) to the values given in **Figure 4.**
- L 9: 5 turns of 1 mm dia. silver-plated copper wire wound on a 6 mm former, coil length pulled out to 10 mm, self-supporting. Coil tap for the 1 nF capacitor 1.5 turns; coil tap to choke: 2.5 turns from the cold end.
- LC 1, LC 6: Input and output coupling link from 1 mm dia. silver-plated copper wire with a spacing of 1 mm from L 1 and L 8, resp.
- LC 2: as LC 1, but length 13 mm
- LC 3: as LC 1, but length 26 mm (6 mm protrude out of the chamber), spacing from L 3: 4 mm
- LC 4, LC 5: as LC 1, bent as Fig. 4, spacing to L 3, L 4, or L 5, L 6: 1.5 mm
- CC: Coupling tab made from copper foil, 4 mm × 8 mm, bent at a length of 3 mm
- RFC B: Base choke (4 pcs.) 3 turns of approx. 0.8 mm dia. silver-plated copper wire, wound on a 3 mm former, coil length pulled out to 5 mm, self-supporting.
- RFC C: Collector choke (4 pcs.), one connection wire of the collector resistor of approx. 25 mm in length is wound around a 3 mm former. The 2.5 turns made in this way are pulled out slightly.
- RFC D: Diode choke, 4 turns of approx. 0.5 mm dia. silver-plated copper wire wound on a 3 mm former, pulled out to a length of 5 mm, self-supporting.
- Trimmers for L 9: Plastic foil trimmer 13 pF (Philips: yellow) or similar
- Bypass capacitor for LC 3: Ceramic disk capacitor without connection leads of approx 8 – 10 pF (preferably use the thicker version)

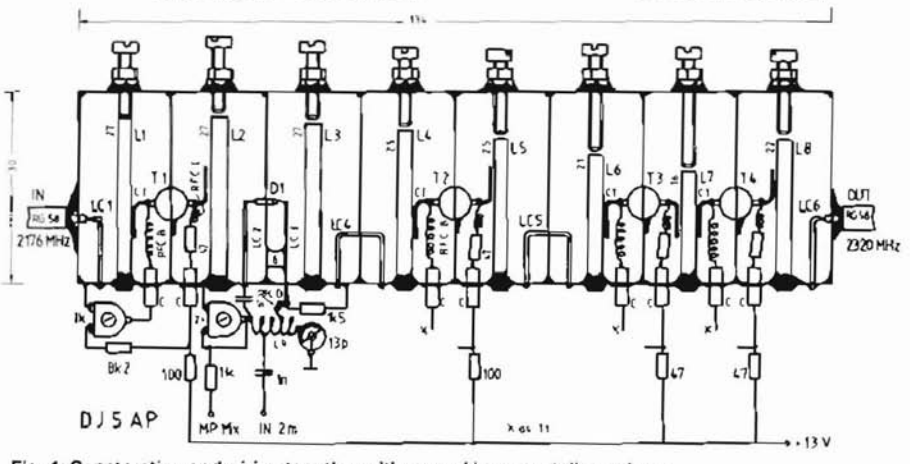


Fig. 4: Construction and wiring together with several important dimensions

Capacitors C: 8 ceramic feedthrough capacitors, approx. 1 nF, for solder mounting, small type (8 mm long, 3 mm dia.)

Tuning capacitors: 8 brass screws M3; five pieces 10 mm long, 3 pieces 20 mm long, each provided with two nuts.

3. CONSTRUCTION

The author used 0.3 mm to 0.4 mm thick metal plate for construction of the first four prototypes of this transmit converter. Several strips of 30 mm and 25 mm in width should be prepared in a metal workshop. Tin-plate is inexpensive and is easy to work with (especially to solder). They can usually be obtained readily from hardware stores.

Of course, it is also possible to use 0.5 mm thick brass plate or thin, double-coated PC-board material. Copper plate is not advisable, since it is not easy to solder it due to its high heat conductivity.

The strips of metal plate are cut using shears to the dimensions given in **Figure 5**. The required holes on the outside and intermediate panels are firstly tapped, after which they are drilled using 3.5 and 3 mm drills, respectively. The cutouts for the transistors in the intermediate panels 2, 5, 7, and 8 are made with a 3 mm drill and then extended to the required dimensions with the aid of a suitable file.

After this, the rear panel (part 10) is drilled with the holes for the tuning screws. This is done by laying this part on part 11 so that they coincide exactly, after which they are fixed together with the aid of two strips of adhesive tape. Based on the previous holes in the rear panel (part 10), the holes for the line circuits L 1 to L 8 are now made. This ensures that the hot ends of the $\lambda/4$ line circuits are directly adjacent to the tuning screws after assembly.

Part 10 is now removed from part 11 and the other holes are then drilled into part 11 for the feedthrough capacitors, and for the coupling line to the mixer diode.

The base plate, part 12, is not provided with any holes.

Before soldering together, clear marks must be made on the inside of part 10 or 11 to indicate where the intermediate panels are to be soldered into place.

All outer and intermediate panels are now soldered into place with a minimum of solder. The guide nuts (M 3, brass) are soldered onto part 10. After this, the coaxial inner conductors of L 1 to L 8 must be soldered into the appropriate holes in part 11 in such a manner that the outside ends protrude by approximately 3 mm. This allows one a little reserve if the lines are found to be too short during the alignment process.

This is followed by making the coupling links according to details given in Section 2.1., and soldering them into place. After this, the feedthrough capacitors, and finally the transistors are installed. Before mounting, the connections of all transistors should be shortened to a length of 2 mm. Attention should be paid that the emitter connections are as short as possible when soldered to the intermediate panels

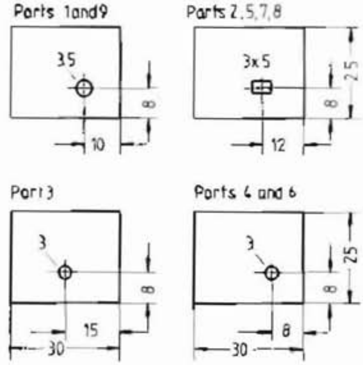
This is followed by installing the mixer diode (shorten these connections on both sides to 3 mm), and the coupling tabs. Finally, the 144 MHz drive circuit and the bias voltage supply should be completed. The ground connections of the 5 trimmer potentiometers are soldered using short, thick pieces of wire, to the outside of part 11. The rest of the wiring is shown in Figure 4.

4. ALIGNMENT

Firstly set all base-bias trimmer potentiometers to ground stop. Transistor T 1 will receive its operating voltage of between 12 and 14 V via the dropper resistor of 100 Ω. The current drain of T 1 is set to 5 mA.

The oscillator module is now connected to point "A" and switched on. If sufficient output power is provided by the oscillator module at 2176 MHz, a collector current of 15 to 20 mA should result for T 1 after L 1 has been tuned to resonance. If the collector current of T 1 should only increase to 8 to 10 mA, this will mean that the output power of the oscillator module will not be sufficient, and that one will not be able to obtain the full 100 to 200 mW output power of the transmit converter.

A high-impedance multimeter is now connected to test point "MP" of the mixer diode, after which the tuning screw of L 2 is aligned for resonance at 2176 MHz. If the output power of the oscillator module is optimum, a maximum of 2.5 V will be measured here when the 2 kΩ trimmer potentiometer of the mixer diode is set to its highest value.



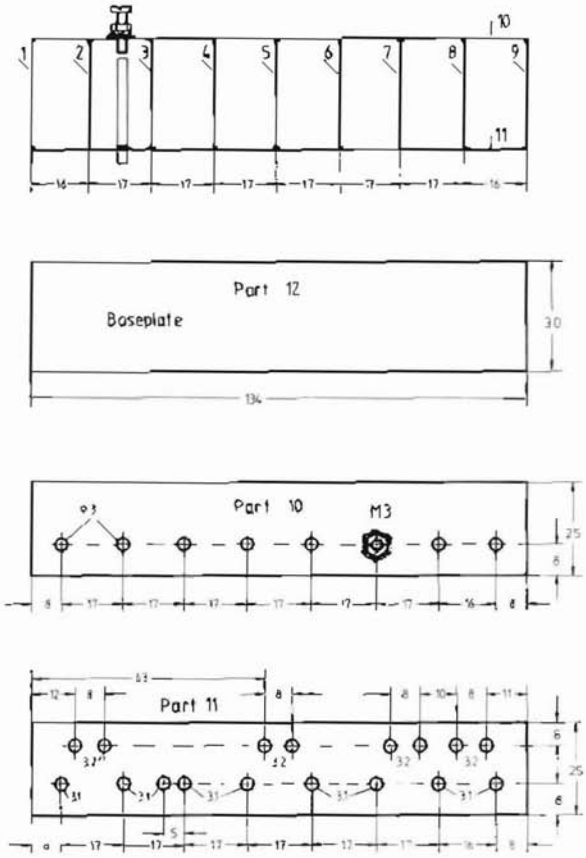


Fig. 5:
Dimensional drawings of the metal-plate parts.
Material: Tin-plate 0.3 to 0.4 mm thick

The three amplifier transistors are now connected to the operating voltage and their quiescent currents are set individually to approximately 5 mA. This is a value at which one is easily able to see small fluctuations of the required signal at 2320 MHz, and is a great help for the preliminary alignment.

The 144 MHz drive signal is now connected to input "B" and should have a level of maximum 300 mW. The author uses an IC-202 with an output power of 3 W as driver. For this reason, it is necessary to provide an attenuator of approximately 10 dB. For experimentation, it is very advisable that the attenuation can be varied slightly. Before switching on the driver, the circuits comprising L 3 and L 4 should be aligned to the oscillator frequency of 2176 MHz, which will be seen as an increase of the collector current of T 2 to approximately 8 to 10 mA. The tuning screws should then only be approximately 0.3 mm from the hot ends of the circuits.

The exciter is now switched on and the tuning screws of L 3 and L 4 are slowly rotated out. One will soon find the positions of the screws where the current of T 2 increases to 7 to 8 mA. Ensure that one is really tuned to 2320 MHz by switching off the 2 m drive signal – the current of T 2 should then drop immediately to 5 mA.

The subsequent stages are then also aligned to 2320 MHz, and a few mW of output power should already be measureable at the output. It is now possible for the quiescent currents of the transistors to be aligned to their final values with the exciter switched off: T 2 and T 3 approx. 20 mA, and T 4 approx. 30 mA. The final alignment is made alternately by carefully bending the coupling tabs, and tuning the alignment screws. The coupling tabs at the base have a spacing of approximately 0.8 to 1 mm from the line circuits, whereas the tabs at the collector can go down to 0.5 mm from the resonant lines. If the coupling tabs are bent towards the resonant lines, this means that the resonant circuit will have a greater capacitive load, and that the resonant frequency will be pushed downwards. The tuning screws must therefore be rotated out. It may be necessary in some cases for the line circuits to be shortened. This can be carried out easily with the aid of a soldering iron and pliers. The M3 nuts are countered after completing the alignment.

The preliminary alignment to the nominal frequency of 2320 MHz can, of course, be carried out more easily, if a separate receive converter is available so that one can monitor the transmit signal. However, the author has always been able to find the correct 2320 MHz resonance even without this. It is recommended that the output frequency is measured with the aid of a simple absorption wavemeter such as described in (3).

4.1. Measured Values

After constructing the first prototypes, the author was able to gain access to professional measuring equipment, which means that the following values can be classed as reliable.

- Output power:
  According to drive power at 2176 MHz and spread of the transistors: 100 – 200 mW
- Overall current drain:
  Without drive approx. 80 mA
  With drive approx. 100 mA
- Suppression of unwanted conversion products:
  Better than 40 dB
- Suppression of the local oscillator frequency
  Better than 40 dB
- Suppression of the image frequency:
  Better than 60 dB

5. EXPERIENCE GAINED DURING ASSEMBLY AND ALIGNMENT

Four prototypes of the described transmit converter have now been built up and all were constructed using tin-plate. Once, silver-plated brass tubing was used for the resonant lines, once copper tube, and twice blank brass tube. No differences could be seen in the suppression of unwanted conversion products, and the oscillator and image frequency suppression were also identical in all cases.

A cover plate can be soldered or screwed to the chambers, but was not found to be necessary in practice when using this construction It was found that the suppression of unwanted frequencies was practically identical with and without cover (of course, after realignment)

The author had carried out extensive experiments with respect to the transistor complements and the selection of the mixer diodes. The most favorable complement was found when using three BFR 34 A, and a BFR 96 in the output stage Attention should be paid, however, that only good-quality semiconductors are used.

Approximately 10 different types were tried as mixer diode. The diode with the least problems, which was also very cheap, was the Schottky diode HP 5082-2800 If the local oscillator power is too low, the HP 5082-2817 can provide a certain improvement. Cheap switching diodes such as the 1N4148 were also suitable, however, were never able to provide the maximum possible output power

Heat-conductive paste can be placed between all transistors and the screening panels, which improves the thermal stability

The described converter is sufficient to drive a 2C39BA driver and output stage combination up to an output power of approximately 25 W

5.1. Another Version for 9 cm

Based on the good experience obtained at 2320 MHz, a further transmit converter was constructed according to the same principle for 3456 MHz. The transistor complement is identical and a HP 5082-2817 is used as mixer diode. Since the inexpensive transistors used are in the vicinity of their limit frequency at 3456 MHz, the gain is, of course, not so high as at 2320 MHz. Output powers of between 20 and 30 mW were achieved in two prototypes.

The mechanical dimensions of the $\lambda/4$-circuits should be changed as follows:

L 1 to L 6, and L 8 are 13 mm long.
L 7 is 10 mm.
This is the length in the chamber.

The dimensions of the intermediate panels change to a width of 20 mm; the height of 25 mm remains

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Carsten Vieland, DJ4GC

A Sensitive Thermal Power Meter

A measuring instrument is to be described that has seven measuring ranges from 100 μW to 300 mW, and whose upper frequency limit is way up in the X-band! Construction should not be difficult for those readers having adequate mechanical skill, and a magnifying glass; only a few special parts are required which are easily available.

1. POWER MEASUREMENT PROBLEMS

For radio amateurs, power measurement is probably one of the most difficult areas in radio frequency measuring technology. The various types of diode voltmeters, **see Fig. 1**, have three distinct disadvantages:

1. The junction capacitance of the test diode (1-4 pF) represents a parallel capacitance to the load resistance. For instance, the amount of the capacitive reactive impedance will be less than the 50 Ω load resistor when using a Schottky diode HP 2800 even at 1.6 GHz. In conjunction with the inavoidable circuit inductivity, this will lead to noticeable resonance effects, which limit this type of power measurement to frequencies up to approximately 1 GHz, if a special scale calibration is not used.

2. The non-linearity of the diode characteristic will be noticeable at low AC-voltages. This leads to a non-consent mathematical scale calibration inspite of subsequent amplification and delogarithming. If one is to avoid the very extensive compensation method, it will be necessary for the scales to be calibrated point by point, for instance by calibrating it against a precision meter.

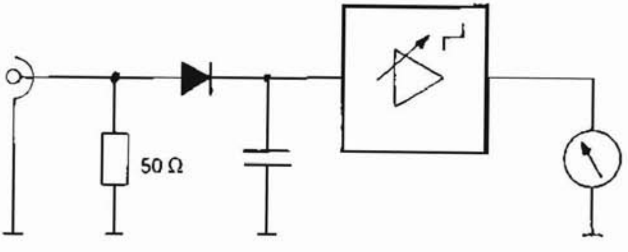

Fig. 1: A diode voltmeter as power meter

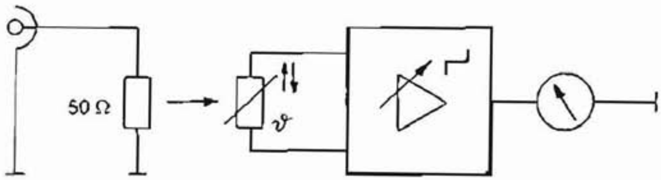

Fig. 2:
Principle of a thermo-power meter operating according to the bolometer principle

3. The calibration of the diode voltmeter is made in RMS-values, however, the measurements are made with peak voltages. In the case of subsequent measurements on amateur equipment, the required sinewave signal will be superimposed with harmonics, subharmonics, conversion products, and unwanted oscillations. When the maximum values of the individual voltages coincide, peak voltages are provided to the diode, which have no relationship to the RMS-value. The output power of oscillating stages can be even higher than the power consumption from the power line.

The described disadvantages of diode voltmeters can be avoided or at least reduced when using the bolometer principle (see **Figure 2**), since the load resistance is only to be found in the RF-circuit. The heating is a linear function of the RMS-value of the RF-power, at least at low temperatures. The temperature increase is measured with the aid of a NTC-resistor, which will lead to a power-linear scale calibration. Calibration and accuracy measurements on such a unit can be made with the aid of DC-voltages.

Fundamental considerations were made in (1), (2), (3) and (4). A suitable construction was described in (1) Higher sensitivities can be achieved with the aid of thermo-elements using thin-film technology (5).

The described meter has seven measuring ranges from 100 μW to 300 mW (FSD). Its upper frequency limit is in the X-band. One disadvantage is the somewhat long transient time of this method (50% of full scale after 1 s), which means that no modulation measurements can be made

2. COMPONENT SELECTION

The 50 Ω load resistor should be as small as possible. A small mass results in a short thermal transient time, as well as a high temperature coefficient (meter sensitivity), and has a positive effect on the upper limit frequency The smallest, inexpensive, and available resistor (51 Ω) uses a flat metal-glazed conductor and is sometimes designated as micro-miniature resistor (62.5 mW). It is in the form of a bead-type microchip resistor that has been dipped in lacquer, and they are usually used in layer circuits. After carefully removing the lacquer, one will obtain a ceramic chip whose dimensions are 2.2 mm × 1.2 mm × 0.8 mm.

The temperature-probe resistor should also have a low mass and thus a short transient time. In addition to this, high-impedance resistors are preferable, since these exhibit the lowest intrinsic heating as result of the connected test voltage. The Siemens Thernewid-NTC resistor type K 19 is very suitable. This component comprises a glass bead of 0.4 mm diameter and possesses virtually invisible connection wires. This component is so sensitive that it will react to the radiation heat of one's hand without delay, even at a spacing of 1 meter. Unfortunately, this resistor, which can also be supplied in pairs, is not inexpensive, but it is also offered by several other manufacturers.

Experiments made with the thermoprobes SAK 1000 and KTY 11 resulted in an inferior limit sensitivity, and the transient time was at least ten times longer.

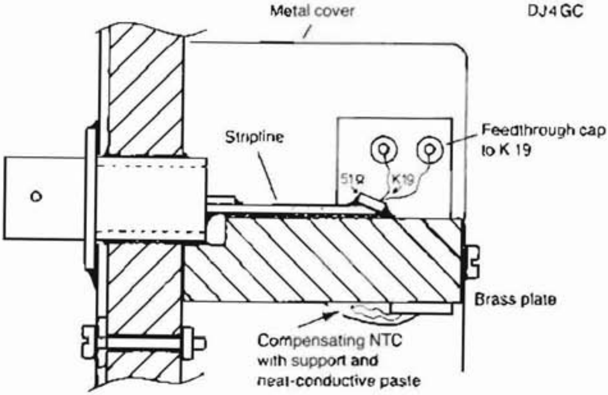

Fig. 3: Suitable construction of the bolometer

3. CONSTUCTION OF THE RF-CIRCUIT

Of course, the NTC-resistor must be directly glued to the load resistor (with very little two-component adhesive). However, due to its high sensitivity, it should be thermally decoupled from the input connector. For this reason, it is not recommended for the load resistor to be directly soldered to the RF-connector, since the mechanical tension passed via the inner conductor could destroy the chip resistor. A favorable solution was found by using a 50 Ω stripline in conjunction with a heat sink (brass plate) for interconnecting the load resistor to the input connector. This type of construction is shown in **Figure 3.**

In order to ensure a high cutoff frequency, the stripline should be ideally on a PTFE-material (double-coated). A stripline width of 2.3 mm will result when using 0.8 mm thick RT/duroid 5870 material. In the author's prototype, the stripline is 12 mm in length. Of course, epoxy PC-boards can be used up to several GHz without problems due to the non-resonant conductor lane. When using 1.5 mm thick epoxy PC-board material, the stripline width is 3.1 mm.

Special care must be taken in the transition between the coaxial connector and the stripline. Although N-connectors have more favorable RF-characteristics than BNC-connectors, the former will exhibit are more noticeable incontinuity at the transition. Professional users specify SMA-connectors up to 18 GHz.

In order to achieve the shortest possible transient time of the bolometer, a good heat dissipation is obtained at the cost of maximum sensitivity. Heat-conductive paste should be provided between the stripline board and the brass heat sink, which is also placed around the chip resistor. Temperature fluctuations coming from the input connector are compensated for with the aid of a second brass plate **(Figure 4).** Since the thermal probe still reacts to the radiation heat reaching the case, the whole bolometer is surrounded in a metal case.

Any excessive solder on the stripline should be removed with a file in order to ensure a low heat delay. The NTC-resistor should be glued into position only after this has been carried out.

The fragile connection wires of the K 19 are supported on the RF-side with the aid of feedthrough capacitors, and on the lower side with the aid of a small board that has been glued into place.

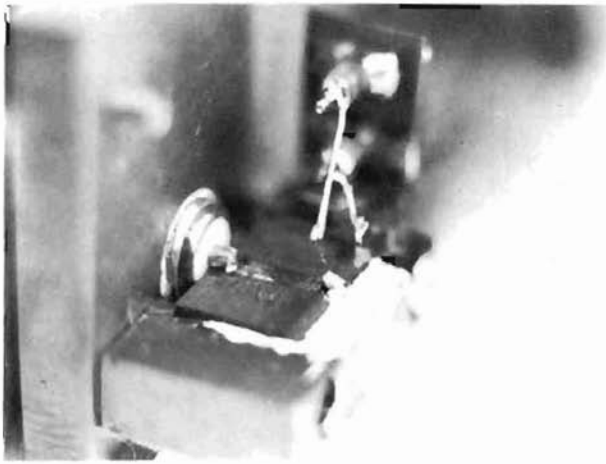

Fig. 4:
RF-circuit with bolometer and heat sink

The author's prototype is mounted in a compact, standard metal box whose dimensions are 111 mm × 73 mm × 50 mm (see **Figure 5**).

4. MEASURING AMPLIFIER

In order to maintain the zero-point stability, and the calibrated meter sensitivity, it is recommended that a brigde circuit be used together with a second K 19 (paired to have the same temperature coefficient), in order to

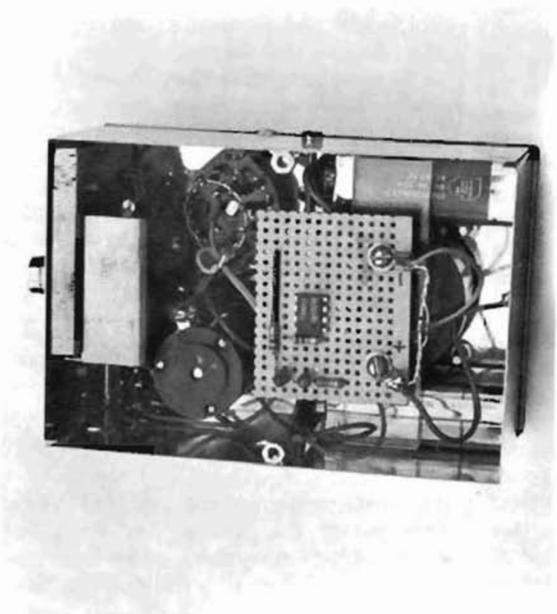

Fig. 5:
Photograph of the author's prototype; RF-portion under the metal cover.

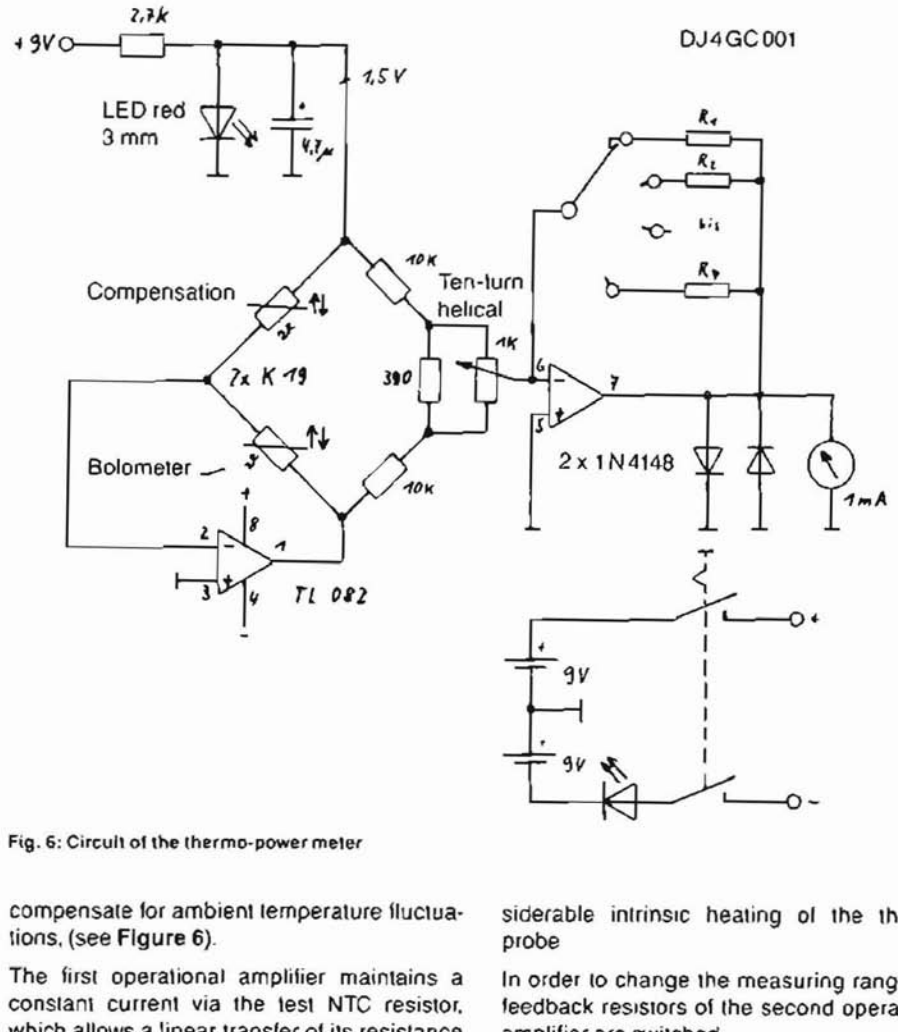

Fig. 6: Circuit of the thermo-power meter

compensate for ambient temperature fluctuations, (see **Figure 6**).

The first operational amplifier maintains a constant current via the test NTC resistor, which allows a linear transfer of its resistance value to the actual test amplifier. The zero-point should be adjusted before commencing the measurement with the aid of a low-impedance, shunted, ten-turn helical potentiometer. If a larger case is used, it is also possible for less-expensive coarse and fine controls to be used.

The reference voltage is provided by a LED, which is connected as zener diode. Higher voltages than 1.5 V will, however, improve the sensitivity of the reading, however, will lead to considerable intrinsic heating of the thermal probe

In order to change the measuring range, the feedback resistors of the second operational amplifier are switched.

The resistance values and the measuring ranges (full-scale deflection) are:

R_1 = 2.2 MΩ (0.1 mW)
R_2 = 680 kΩ (0.3 mW)
R_3 = 220 kΩ (1 mW)
R_4 = 68 kΩ (3 mW)
R_5 = 22 kΩ (10 mW)
R_6 = 6.8 kΩ (30 mW)
R_7 = 1.4 kΩ (300 mW)

The meter cannot be overloaded since the operational amplifier possesses an internal current limiting. Since an offset alignment is not required, it is advisable to use a low-drift dual-operational amplifier in an eight-pin case, such as the TL 082. In the most sensitive range, the flicker noise of the operational amplifier will cause a certain fluctuation of the meter reading.

The operating current is only in the order of 5 mA, which means that two 9 V-batteries can be used as power supply. The meter will also operate perfectly at ± 5 V.

5. ALIGNMENT

The calibration of the meter is made with direct current. It is advisable to adjust the current to the full-scale deflection of the appropriate range with the aid of a digital multimeter

The feedback resistors R_1 to R_6 of the test amplifier are selected to have the highest accuracy from a large selection of resistors. Since the sensivity of the bolometer is greatly dependent on the mechanical construction, the given values are only for orientation.

Due to the non-linear relationship between the temperature and the resistance value of the NTC-resistor, it is necessary for the highest range of 300 mW to be calibrated independently on the scale, (see Figure 7). In the 30 mW-range, the deviation is still only a maximum of 4%, and should thus be acceptable

Fig. 7:
This photograph shows the scale calibration with the 300 mW scale at the bottom

6. MEASURED VALUES

It was possible before manufacturing the power meter to measure the input return loss of the RF-circuit with the aid of a network analyzer It was found that a return loss of 20 dB (corresponding to approx. 1.2 VSWR) can be achieved up to a frequency of 2 1 GHz. The return loss of 10 dB (approx. 2 VSWR) is only exceeded at 11 GHz.

A 3 GHz oscillator having an output power of 25 mW with an accuracy of ± 0 1 dB was now connected to the meter and this power was indicated with an accuracy of the meter-needle. The Gunnplexer manufactured by Microwave Associates (15 mW at 10 36 GHz) provided 12 mW of heat after being adapted from waveguide to BNC.

The meter reaches 50% of the full-scale value after approximately 1 second; 90% of the final value is passed after 3.4 s. The transient time τ (63% of the final value) is in the order of 1.5 s.

7. PRACTICAL EXPERIENCE

Due to the short length of the stripline used, a certain temperature sensitivity exists via the inner conductor of the input connector, since both NTC-resistors are not heated simultaneously. In the very low power range, it is advisable to work together with an intermediate cable which remains connected to the meter. Otherwise, the zero-point stability is so high that it is possible to carry out measurements directly after switching on. In the case of the two most sensitive ranges, it is advisable to leave a warm-up time of appoximately three minutes

The calibration was made at 20°C. A further test in a refrigerator at 5°C did not show large deviations.

The speed of the reading is approximately as high as that of a dampened laboratory meter. Taking all advantages of this measuring system into consideration, it will not be found that alignment work is made more difficult due too slow an indication

The dynamic range of the power meter can be increased by adding wideband amplifiers (e g as described by DJ7VY), attenuators, or directional couplers. However, the frequency range will be limited by this. It is possible, for instance, to use a directional coupler with an attenuation of 40 dB to increase the measuring range up to 3 kW. The exact value of the loss can be measured previously using this meter.

The high sensitivity of this meter (the resolution is in the order of 1 μW) allows one to also measure the frequency, or attenuation characteristics of filters, bandpass filters, directional couplers, frequency multipliers, mixers, low-signal amplifiers, etc., in addition to purely power measurements, and its high dynamic range can be used right up to X-bands.

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H. Wessels, PA 2 HWG

A 1296 MHz/144 MHz Converter equipped with the GaAs-FET 3 SK 97

The autor became interested in experimenting with these transistors after the price of dual-gate GaAs-FETs dropped to a level that was affordable for regular amateurs. Similar articles have already been published for applications in the 144 MHz and 432 MHz bands (1), (2). For this reason, the author decided to experiment with them on the 1296 MHz band.

The circuit shown in **Figure 1** is result of a few weeks of experimentation. The main specifications of this circuit are:

Power gain of T 1:	approx 15 dB
Conversion gain of T 2:	approx. 10 dB
Overall gain:	approx. 25 dB
Total noise figure:	approx. 4 dB

Construction

The case of the converter is built up from 0.8 mm thick brass plate which provides it with a sufficient mechanical stability A piece of Vero-board

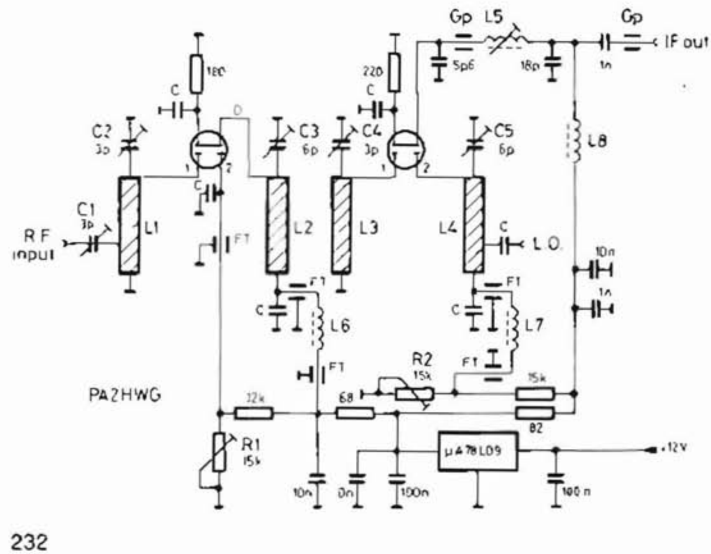

Fig. 1: A 1296 MHz converter with the DG-GaAs-FET 3 SK 97 in preamplifier and mixer stage

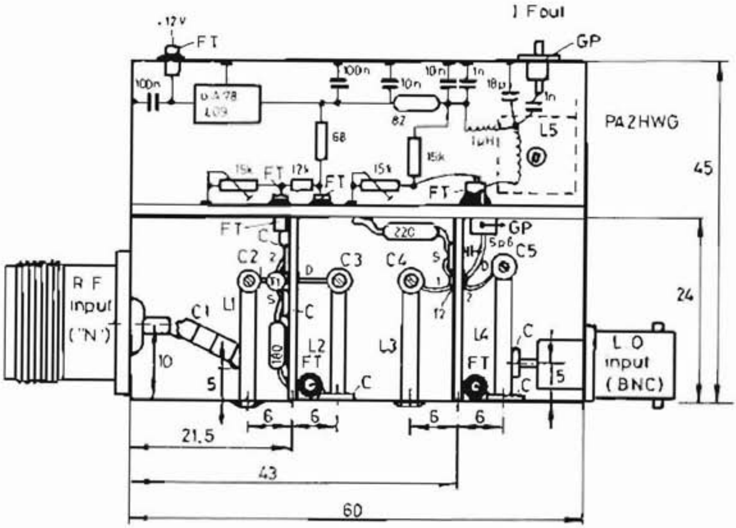

Fig. 2: Dimensions of the metal box as seen from above

is used for the DC-portions of the circuit and for the intermediate frequency output. It should be possible with the aid of **Figure 2 and 3** to construct the converter without difficulties.

Components

T 1, T 2: 3 SK 97, S 3030, MRF 966
C. Chip capacitor approx. 200 pF
FT Feedthrough capacitor, approx. 1 nF
GF: Low-capacitance feedthrough (glass)
L 1 ... L 3: 13 mm length of 2 mm dia silver-plated copper wire
L 4 as L 1, but 15 mm long
L 5: 5 turns of 0.8 mm dia. enamelled copper wire wound on a 5 mm coil former with VHF core
L 6, L 7: 0.56 µH miniature choke
L 8: 1 µH miniature choke

Alignment

For aligning the converter a local oscillator is required that is able to provide a local oscillator signal of approximately 3 mW at 1152 MHz, furthermore a 144 MHz receiver, and a test signal from a local station or beacon. The alignment is commenced by placing the trimmer potentiometers R 1 and R2 to their center positions.

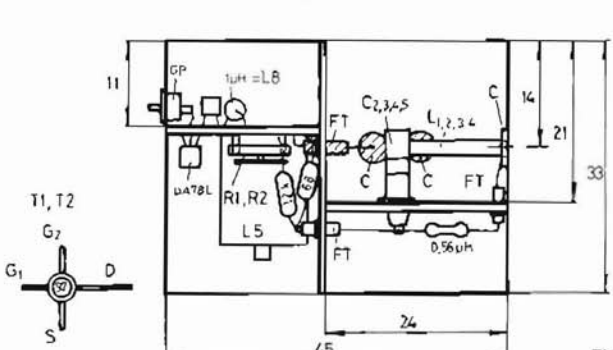

Fig. 3: Side-view from the left-hand side

Trimmer capacitors C 1 to C 5, and the IF-inductance L 5 should now be aligned for maximum S-meter reading on the 144 MHz receiver. After this, the adjustment of R 1 and R 2 is varied so that the maximum reading results on the S-meter. Trimmer capacitors C 1 to C 5 should now be corrected. Finally, C 1 and C 2 should be aligned for best signal-to-noise ratio, which is made in conjunction with a weak, stable signal, or with a noise source. The converter is then ready for operation.

Local Oscillator Circuit

A large number of suitable oscillator circuits has been described. The author uses module DF 8 QK 002 (3). This is just as suitable for operation when the described converter is to be used for amateur TV-application in the 24 cm band. In this case, it is only necessary for the crystal in the local oscillator module, and the IF-filter in the converter (5p6 – L5 – 18p) to be changed.

Finally, the author would like to thank PE 1 HVX who carried out most of the mechanical work on the converter.

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Guenther Borchert, DF 5 FC

A 2 m/70 cm Transmitter with High Spurious Rejection Concluding Part II

4. CONSTRUCTION

All modules are accommodated in standard metal boxes of 30 x 74 x 111 (mm) or 30 x 74 x 35 (mm). PC-boards have been designed whose dimensions are 72 x 109, or 72 x 33 (mm). Both the components and the filters are mounted on these boards. The interconnection of the individual modules between one another is made exclusively using SMC-connectors and RG-174 or similar PTFE-cable. These connectors can be used up to frequencies in the order of 10 GHz. **Figure 10** shows the five, interconnected modules, the covers over the helical filters of module DF5FC 001 have been removed.

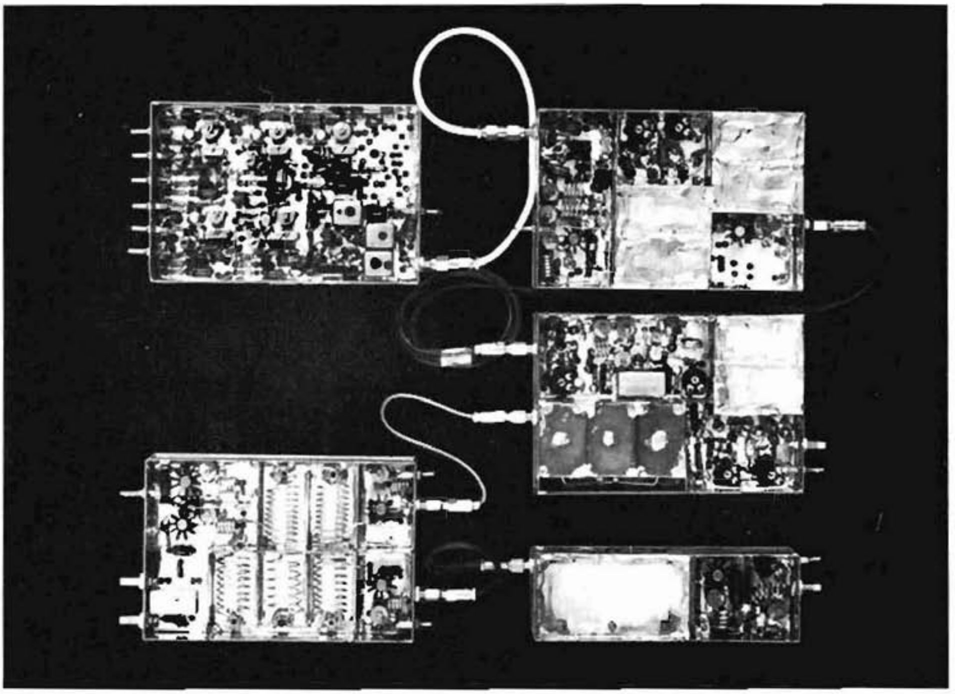

Fig. 10. The five modules of the 2 m/70 cm SSB transmitter with low spurious rejection are shown connected together

It is advisable to firstly solder the prepared and drilled PC-board into the case before commencing construction After this, the helical filters are prepared and installed. The tubular trimmers should be soldered into place before inserting the intermediate panels. This sequence is recommended since the heat loading of the other components especially the semiconductors, would be considerably high if they were mounted into place before soldering in the intermediate panels and soldering the PC-board into the case. The only components that are soldered into place before total assembly are the disk capacitors as used in the DF7VY amplifier, since these are very difficult to mount afterwards.

The holes in the case for the feedthrough capacitors and SMC connectors should be made before assembly according to the given dimensions. Holes of 5 mm diameter are required for the SMC connectors

The tuning trimmers of the helical filters protrude into the chamber from the base surface. and for this reason, it is possible for the filters

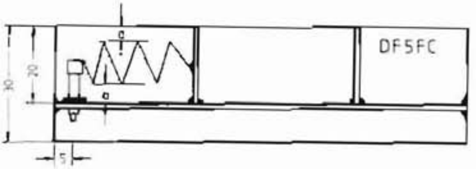

Fig. 11: Installation of the board in the metal box (dimension a = b)

to be arranged, as required, in a module. The side panels of the case therefore remain free for the connectors and for mounting. The modules are designed so that all connections are made on the narrow sides of the case, which simplifies installation in a cabinet later.

Silver-plated copper wire of 1.5 mm diameter is used for the filters at 304 MHz and 432 MHz. Similar wire with 1.0 mm diameter is used for the 127 MHz and 145 MHz filters. The exact data for the inductances is given in the associated circuits. The inductances should be mounted as near as possible to the center of the chamber and be soldered well at both ends (if possible without mechanical tension).

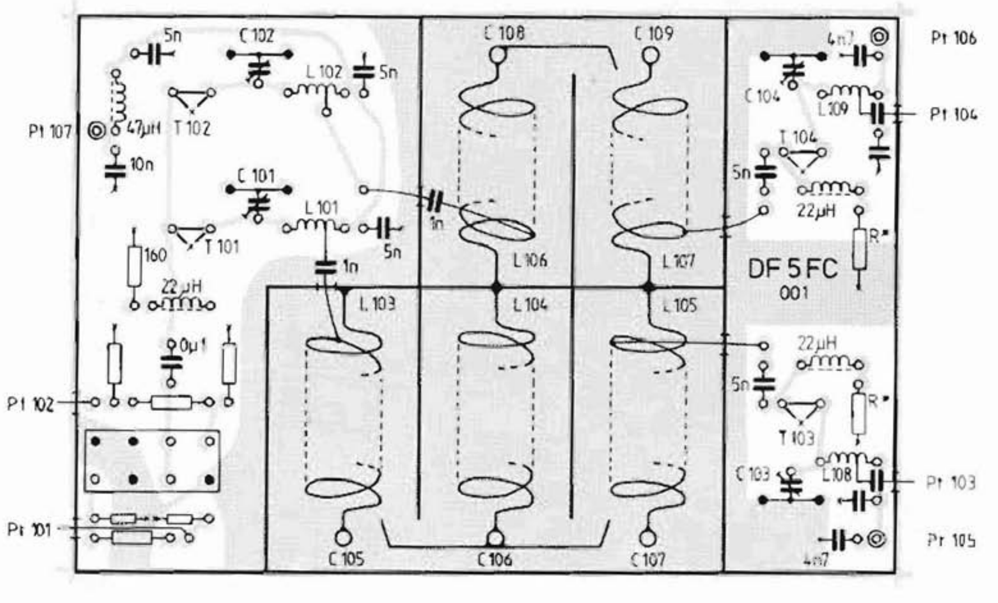

Fig. 12: Component locations on the 144 MHz/126 MHz mixer board DF5FC 001

The exact coil taps are partially to be found experimentally, and the given positions are only values of orientation. It is important that especially the 432 MHz and 304 MHz filter chambers are soldered all around hermetically sealed, otherwise, the insertion loss of the filter would increase considerably.

It was found that the size of the coupling slot was sufficient; readers who have access to suitable measuring equipment will be able to optimize the selectivity curves. Exact data are not possible, since the individual filters are somewhat different from another mechanically.

The PC-boards are mounted exactly 20 mm deep from the top edge of the case, which results in the required depth of the filter. This arrangement is shown in **Figure 11** in the form of a drawing.

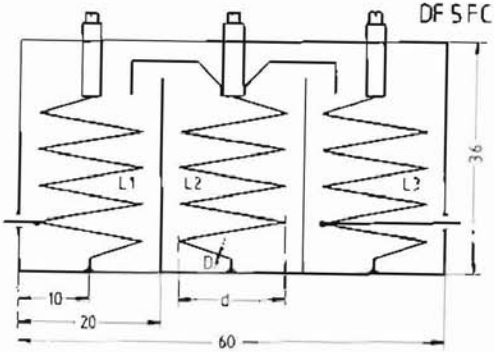

Fig. 13: Construction of the three-stage filter for 145 or 126 MHz

4.1. MIXER 145 MHz/126 MHz

The component location plan of the double-coated PC-board DF5FC 001 (without through contacts) is shown in **Figure 12**. The filters are built up according to **Figure 13**. In

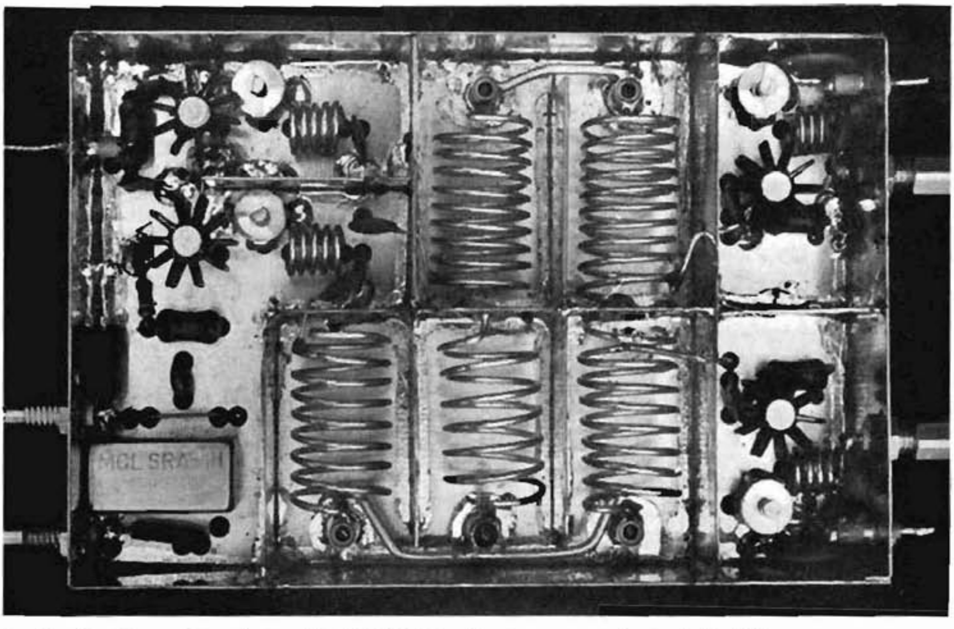

Fig. 14: Prototype of the mixer module DF 5 FC 001 without cover over the two helical filters; the coupling wires can be seen

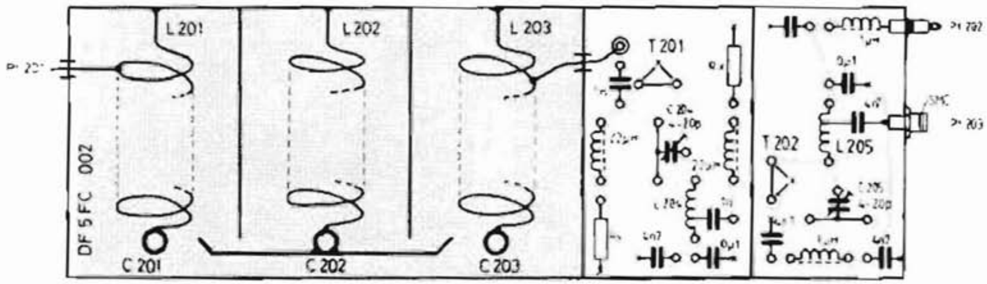

Fig. 15: Component locations on the filter/amplifier module board DF 5 FC 002

the case of the inductances, attention should be paid that the adjacent circuits are wound in the opposite direction. This means that the center circuit of the three-stage filter and the second circuit of the two-stage filter must be wound in the opposite direction.

The two attenuators at the mixer must be suitable for the actual power levels. The common source resistor of the two FETs coupled onto the mixer is set so that a total current drain of approximately 30 mA results. The source resistors of the output amplifier should be set to the corresponding values. Data regarding the semiconductors, inductances, and trimmers are given in Section 4.5. **Figure 14** shows a completed module without cover over the helical circuits.

The modules are individually aligned before installation in a cabinet and the cover is attached with a few solder points so that the module can be opened later without difficulty. A subsequent total alignment should not be required if a careful alignment has been made, since all modules are built up for 50 Ω impedance.

The operating voltage supply is made as shown in the block diagram for the active modules. It is sufficient to switch over the voltage on changing from 2 m to 70 cm. In the case of the first mixer, attention should be paid that only the required output amplifier is in operation. If this is not the case, the image suppression will decrease.

4.2. FILTER AMPLIFIER

This board is also double-coated and not provided with through-contacts. The component location plan of this board is given in **Figure 15**. The same specifications are valid for the filter as for the three-stage filter on the mixer board. The two coils of the FET amplifier have 5 turns, each, and have a coil tap at approximately 0.5 to 1 turn from the cold end. The first FET can be a 2N 4856A, which is operated with approximately 25 to 30 mA; the second FET should be a P 8000 or P 8002, since it is easier to cool. The current drain should be set to 40 to 50 mA.

The input connector protrudes into the filter chamber and is directly soldered without isolating capacitor; the output connector is provided with a coupling capacitor of 4.7 nF, which is self-supporting within the chamber. The coupling capacitor between the two stages is mounted directly between the inductance and the board. If a copper plate is used as screening and heat sink for the second FET, it is possible for the stage to be operated with a higher voltage of between 18 and 24 V, which leads to a higher linearity and a higher output power. In the case of a voltage of > 18 V, it is recommended to use also a P 8000 or P 8002 for the first FET. **Figure 16** shows a completed prototype

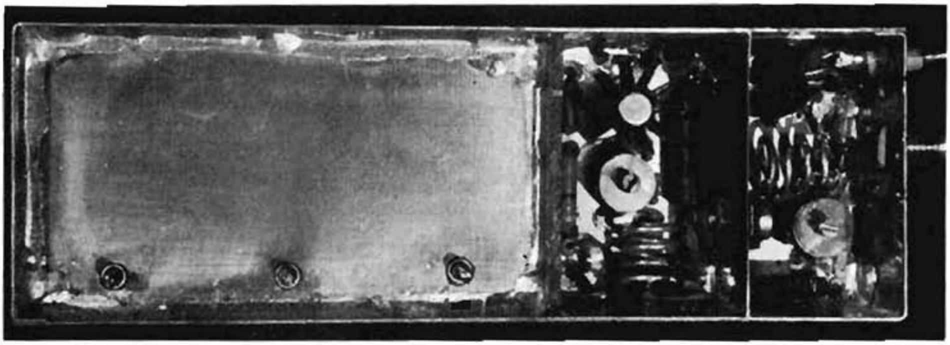

Fig. 16: Photograph of the prototype module DF 5 FC 002

4.3. PROCESSING OF THE SECOND OSCILLATOR

The crystal oscillator was either a module designed by DJ3VY, or a module of standard size that was provided with a doubler. The construction of the oscillator using the double-coated PC-board DF5FC 003 (without through-contacts) is uncritical and the information given by DJ3VY is valid. The frequency doubler stages are equipped with Schottky diodes. Attempts were made to save the wide-band transformers using the two-hole cores, however, this led to alignment problems and the coupling of the diodes was different depending on the depth of the tuning core. For this reason, the transformers were once again provided in the final version of the module. Figure 18 shows the construction together with the five crystals.

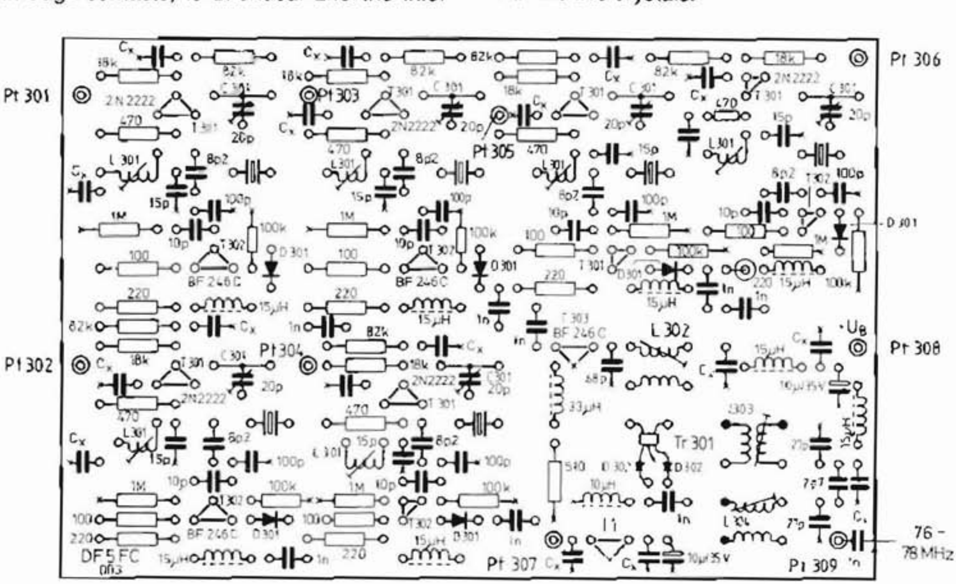

Fig. 17: PC-board DF5FC 003 can accommodate a maximum of 6 crystal oscillators, as well as an amplifier, doubler, and filter.

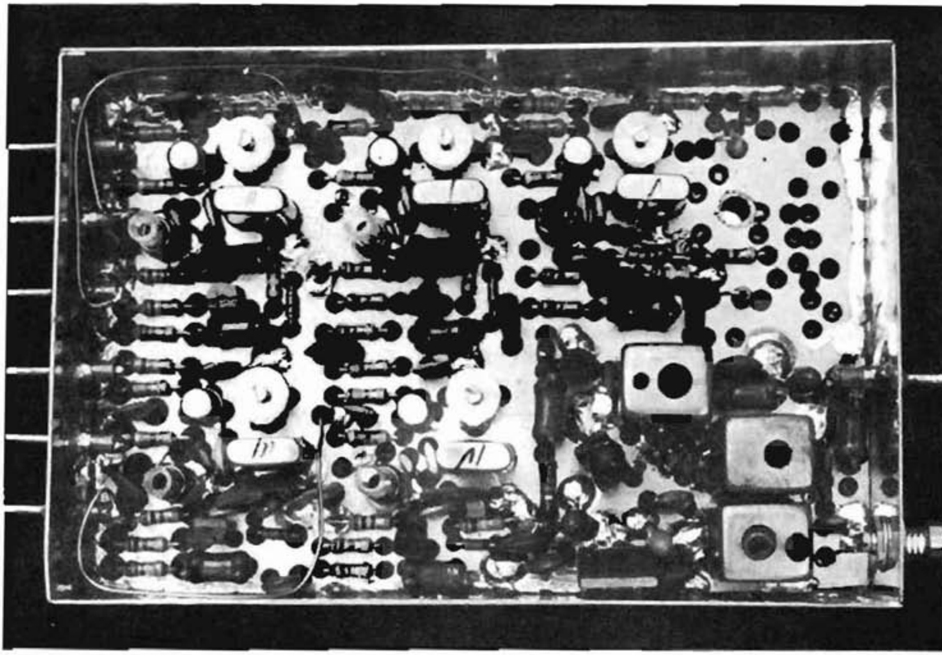

Fig. 18: The prototype DF 5 FC 003 can be equipped with 5 crystals in the 38/39 MHz range

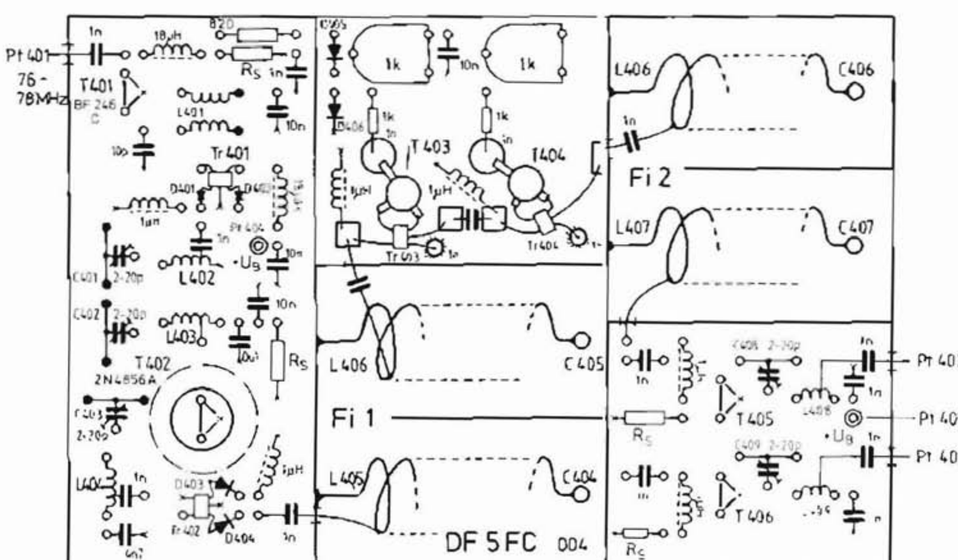

Fig. 19: Component locations on the frequency multiplier board DF 5 FC 004

The frequency multiplier (×4) DF5FC 004 is also constructed similar to that of DJ3VY. Attempts were also made here to delete the two-hole cores, however, this was not successful for the same reasons as given before. The 76 MHz circuits on both boards were wound in the special coil kits D41-2438.1 (see Section 4.5.).

The double-coated PC-board DF5FC 004 is also not provided with through-contacts. Attention should be paid during the construction of the doubler stage shown in **Figure 19** that the cathode sides of the diodes (side with the 1 μH-choke) should have as low a capacitance as possible, otherwise, it is possible that storage effects would reduce the required AC-voltage components (similar effect to that of a filter electrolytic subsequent to a rectifier). In order to reduce further parasitic capacitances, the diodes are only soldered to the board on the cathode side, and the anode sides are directly connected to the secondary windings of the transformer.

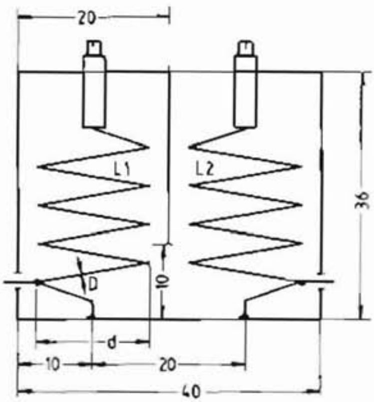

Fig. 20: Construction of the two-stage helical filter for 304/432 MHz with inductive coupling

The helical filter is built up as two-stage circuit and provided with radiation coupling. The mechanical specifications are given in **Figure 20.**

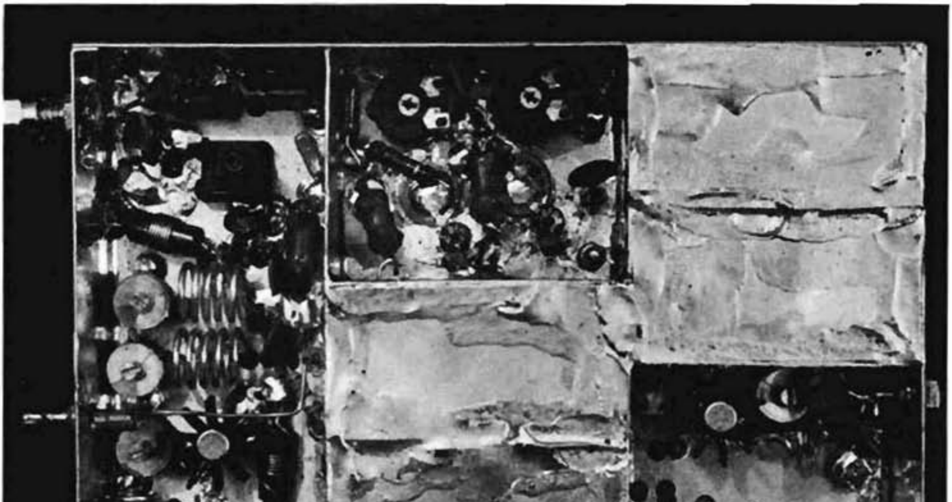

Fig. 21: In this prototype of the frequency multiplier, the second output for a receive mixer was not provided.

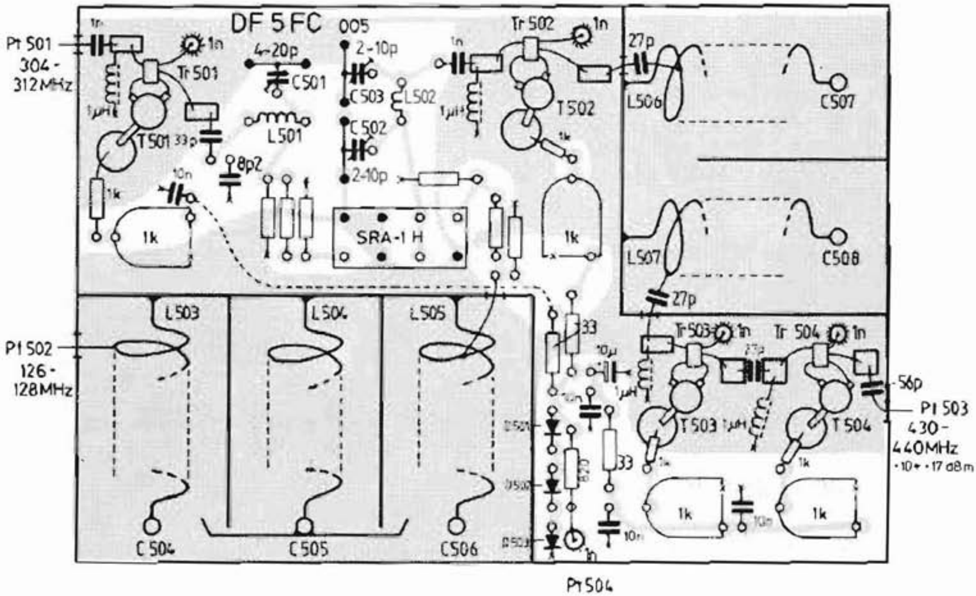

Fig. 22: Component locations on the 70 cm mixer board DF5FC 005

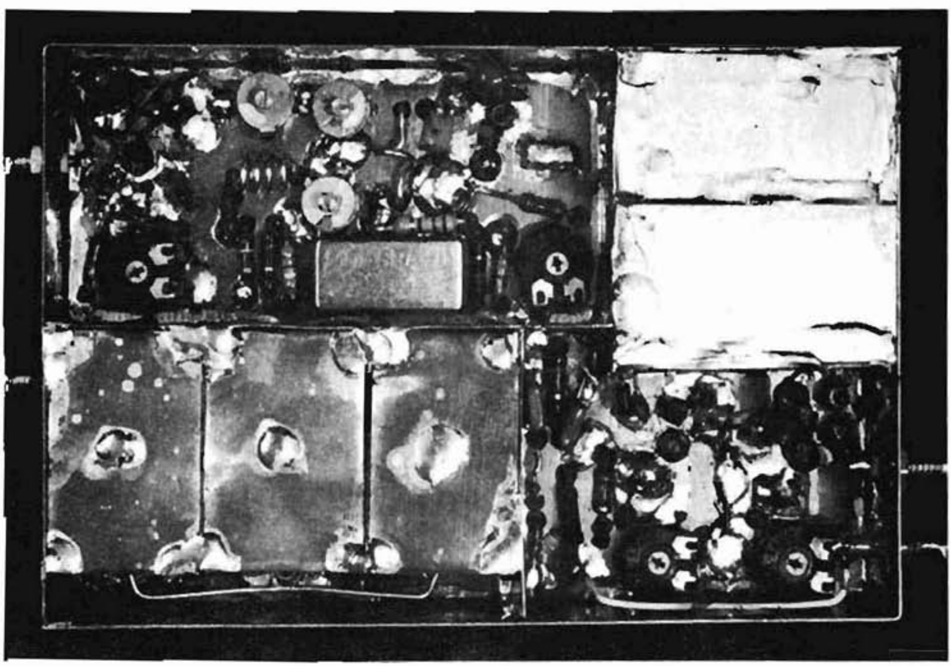

Fig. 23: Prototype of the 70 cm mixer DF5FC 005

The intermediate amplifier conforms to DJ7VY and is built up according to the original description (4). The disk capacitors for the base connections of the transistors should be soldered into place before installing the board into the case. The positive operating voltage is not bypassed using disk capacitors, but using 1 nF feedthrough capacitors. This simplifies the conductor lanes considerably, since the operating voltage can then be directly fed via the resistors to the capacitors on the lower side.

The same is valid for the FET output stages as was mentioned with respect to the output power, operating voltage, and cooling of other amplifiers. If the stages are to be operated with a higher voltage than the system voltage, an extra feedthrough capacitor should be provided in the last chamber. A completed module 004 is shown in **Figure 21.**

The output power of the last (output) stage can be varied with the aid of the operating voltage and current. It is also possible for the oscillator amplifier on the 70 cm mixer board to be deleted, and to feed into the low-pass filter directly, or to match the power level using an attenuator.

4.4. 70 CM MIXER

The required bypass capacitors should be built into this module before the double-coated PC-board (without through-contacts) is installed in the case.

The collector voltage of the amplifier is also fed in using feedthrough capacitors. The voltage supply is made below the board via the collector resistors which are supported by a common disk capacitor.

The three-stage filter for 127 MHz is constructed in exactly the same way as the three-stage filter in the 144 MHz path. The specifications of the inductances are given in Section 4.5. The coil taps are approximately 0.5 turns from the cold end. The transformers for the amplifiers are all wound in the same manner with m = 3, n = 5, r = 1.

The amplifiers in front of the mixer and in front of the output are aligned for currents of 20 mA, the others to 13 mA at a collector voltage of 10 V.

The 432 MHz filter is built up in the same manner as the oscillator filter, however, the inductance specifications are somewhat different (see Section 4.5.). The coil taps are approximately 0.3 to 0.5 turns from the cold end, and the exact position should be found experimentally. A photograph of the completed module is given in **Figure 23.**

4.5. SPECIAL COMPONENTS

Semiconductors

T 101 - T 104:	2N 4856A or P 8000/P 8002
T 201:	2N 4856A or P 8000/P 8002, see text
T 202:	P 8000/P 8002
T 301:	2N 2222, 6 pcs. required for full complement
T 302:	BF 246C, 6 pcs. required for full complement
T 303:	BF 246C
D 301:	1N 4148 or similar, 6 pcs. required
D 302, D 303:	HP 2800, only with 38 to 39 MHz version (see text)
I 301:	Voltage stabilizer 7809
T 401:	BF 246C
T 402:	2N 4856A or P 8000/P 8002
T 403, 404:	2N 4856A or P 8000/P 8002
T 405, 406:	BFR 34A
D 401 - D 404:	HP 2800
D 405 - 406:	1N 4148 or similar
T 501 - 504:	BFR 34A
D 501 - D 503:	1N 4148 or similar

Trimmer Capacitors

C 101 - C 104:	2-20 pF plastic foil trimmer colour code: green
C 105 - C 109:	0.5-6 pF tubular trimmer, 3 mm dia.
C 201 - C 203:	0.5-6 pF tubular trimmer

C 204, 205:	2-20 pF plastic foil trimmer, green
C 301:	2-20 pF plastic foil trimmer, green
C 401 - C 403;	
C 408, 409:	2-20 pF plastic foil trimmer, green
C 404 - C 407:	0.5-6 pF tubular trimmer
C 501:	2-20 pF plastic foil trimmer, green
C 502, 503:	2-10 pF plastic foil trimmer, yellow
C 504 - C 508:	0.5-6 pF tubular trimmer

Two-Hole Core Transformers

All transformers built up on two-hole cores type A 8 X 17, manufactured by Siemens. In the case of the frequency multiplier stages, the cores are built up with 3 × 4 turns of enamelled copper wire of 0.12 mm dia. using three wires, twisted together, which are connected as shown in the circuit diagram.

The transformers for the BFR 34A amplifiers are wound according to the circuit diagram with $r = 1$, $m = 3$, $n = 5$; the exact description is given in the article of DJ7VY in (4).

Inductances

L 101:	4 turns of 1 mm dia. silver-plated copper wire wound on a 5 mm former
L 102:	5 turns, otherwise as L 101
L 103 - L 107:	Helical filter, see text
L 108:	as L 101
L 109:	as L 102
L 201 - L 203:	Helical filter, see text
L 204 - L 205:	as L 101

Crystals for 38 to 39 MHz:

L 301:	9 turns of 0.4 mm dia. enamelled copper wire wound on a 4 mm former
L 302:	8 turns of 0.4 mm dia. couple, winding 1-1.5 turns in special coil set D 41-2165

Crystals for 76-78 MHz:

L 301:	6 turns of 0 4 mm dia., wound on a 4 mm former
L 302:	6 turns of 0.4 mm dia., couple winding 1-1.5 turns in special coil set D 41-2438.1
L 303, 304:	6 turns of 0.4 mm dia. enamelled copper wire, couple winding 1-1.5 turns onto the cold end in special coil set D 41-2438.1 (10 × 12 - 5140400004)
L 401:	6 turns of 0.4 mm dia. enamelled copper wire, couple winding 1-1.5 turns onto the cold end in special coil set D 41-2438.1
L 402 - L 404:	5 turns of 1 mm dia. silver-plated wire wound on a 6 mm-former; tap 1 turn from the cold end
L 405 - L 408:	Helical filter, see text
L 409, 410:	2 turns of silver-plated wire, 1 mm dia., wound on a 5 mm former, tap approximately 0.5 turns from the cold end
L 501:	4 turns of silver-plated wire, 1 mm dia., wound on a 4 mm former
L 502:	1 turn, otherwise as L 501
L 503 - L 507:	Helical filter, see text

5. ALIGNMENT OF THE MODULES

The following individual oscillator signals must be provided for the alignment. In the case of the 2 m mixer, these are:

− 10 dBm to − 5 dBm at 9 MHz
+ 17 dBm from 135 to 137 MHz

A frequency of 304 to 312 MHz is required for alignment of the 70 cm mixer, which can be provided from modules DF5FC 003 and 004. Either +8 dBm will be required for the version with oscillator amplifier, or +17 dBm in other cases. Furthermore, one will require a system voltage of 15 V, as well as higher voltages for the output amplifiers, if necessary.

The following measuring equipment is required: A mA-meter, a frequency counter, and a power meter that can indicate up to +20 dBm, or a VSWR-meter and a terminating resistor, as well as a RF-probe for lower power measurements.

The first alignment of all modules is that of the DC-voltage values. The transistors operating with 20 mA should be provided with a little heat-conductive paste on the copper surface so that they are cooled sufficiently. It is recommended that these transistors are aligned for a preliminary current of approximately 17 mA until the current increase during the warmer period is completed, after which the final value can be selected. The voltage at the collectors should amount to approximately 10 V after the alignment.

5.1. 145/126 MHz MIXER AND FILTER

After completing the DC-alignment, the oscillators of the correct output power level are connected to the module. The 126 MHz output amplifier is provided with its operating voltage. It is now aligned in the center of the band for maximum output power. This preliminary alignment in the center of the band is necessary, since it would be possible that no alignment can be found if the individual circuits are misaligned. This is followed by connecting the operating voltage to the 145 MHz amplifier, which is then also aligned in the center of the band for maximum output power. The three-stage filter should now be aligned carefully for the required bandwidth by alternate careful alignment of the threestages. This alignment is also valid for the filter amplifier.

The 126 MHz filter is also aligned for a wider passband.

If a swept-frequency generator is available, this will simplify the alignment considerably. The mixer is still not soldered into place, and the filter amplifier is switched directly behind the 2m-circuit. In conjunction with a signal at the center of the band, all circuits are aligned for maximum at both frequencies, after which they are aligned in conjunction with the swept-frequency generator for the required filter characteristic. Attention should be paid that the flatest top of the characteristic is obtained simultaneously with steep slopes of the filter.

Under these conditions, approximately a level of 13 to 17 dBm is available at the output of the amplifier, and approximately +3 dBm at 126 MHz. Spurious waves are suppressed by more than 75 dB at 145 MHz.

5.2. SECOND LOCAL OSCILLATOR

The alignment of the crystal oscillator DF5FC 003 is limited to an alignment of the inductances for maximum output power after the doubler.

If an oscillator should not commence oscillation, the two feedback capacitors should be altered slightly. Further details are given in articles of DK1AG (5) and DJ3VY (6).

In the case of the frequency multiplier DF5FC 004, the circuits for 76 MHz and 152 MHz are aligned so that approximately the same voltage is present at the input of the first filter (measured with the aid of an RF-probe) when switching between the individual oscillators. After this, the first filter is placed in the circuit and the output signal is measured subsequent to the amplifier. This is then aligned for a most constant output voltage. Finally, the powermeter or reflectometer with terminating resistor is connected to the output amplifier and the final circuits are aligned for constant output power, as previously described, by switching the individual oscillators. In the case of a 16 V operating voltage and a current of 30 mA to the final transistor, an output power of approximately +10 dBm should be available. In the case of 24 V and 50 mA it is possible to achieve approximately 13 to 17 dBm according to the construction.

5.3. 70 CM MIXER

The ring mixer SRA-1H should not be installed during the alignment of module DF5FC 005. After aligning the DC-values of all amplifier stages, the 126 MHz-filter is aligned, and this process has already been described in conjunction with the 2m-mixer. After completing the alignment, approximately −5 dBm to −10 dBm should be present at the output of the mixer. In the case of a higher power level, the attenuator that should be provided is designed so that the required output power is present. One should not be tempted to use a higher input level; this would bring, of course, more output power, however, would increase the intermodulation products rapidly.

The output of the filter should be terminated with 50 Ω for alignment of the 126 MHz-filter, since the filter would be detuned otherwise after installation of the mixer.

The mixer is installed after aligning the filter. After connecting the two oscillators with the previously described output power levels, the helical filter is firstly aligned for maximum at the center of the required band. If one is only to operate in a 2 MHz range, this will complete the alignment. An output power of approximately 50 mW should be measured at this time.

When using several ranges, it is necessary for the filter to be aligned to a wider bandwidth. This is made by switching the crystals of the second oscillator.

Even in the case of a wideband alignment, a power of approximately 50 mW ≙ + 17 dBm is available at the output. Although this circuit only uses until now a two-stage filter at the output frequency, the spurious wave rejection is already approximately 60 dB, which is the result of the large spacing of all frequencies used from the required frequency. Inspite of this, bandpass filters should still be provided in a subsequent amplifier.

6. USE OF THIS FREQUENCY CONCEPT IN A RECEIVER

If the described concept is to be used in a transceiver, an output can be provided at the second oscillator for this purpose. The converter would then convert a 2 MHz wide portion of the 70 cm band to 126-128 MHz and from there to 9 MHz. When using the modern converters, as already described in VHF COMMUNICATIONS, this is best made with the aid of a relay which is connected directly in front of the mixer. This then switches the appropriate input filter. Attention should be paid here that a careful blocking of 2 m frequencies is made, since these could otherwise simulate signals on the 70 cm band.

7. REFERENCES

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Harald Braubach, DL 1 GBH

Measuring Aid and a Harmonic Filter for the V-MOS Transistor 100 W-Power Amplifier for 144 MHz

As mentioned at the end of the article describing V-MOS transistor power amplifiers in the last edition of VHF COMMUNICATIONS, this article is to describe a 30 dB coupler, a test detector, and a harmonic filter suitable for handling 100 W. They can be constructed from inexpensive components, and allow a clean alignment and operation of this modern power amplifier.

6. MEASURING AIDS - HOME-MADE

Those readers that have access to measuring equipment and are able to carry this out, can delete this section of the article and commence reading with the harmonic filter.

6.1. 30 dB Coupler

In contrast to a directional coupler that requires a considerable amount of construction and alignment, the described simple coupler provides 1/1000 of the power independent of direction, and is virtually frequency-independent.

The simple circuit diagram is shown in **Figure 20**. The four resistors have the following values:

1 × 1.5 kΩ, 1/2 W (0309)
2 × 4.7 kΩ, 1/4 W (0207)
1 × 56 Ω, 1/4 W (0207)

One can use normal composite carbon resistors with a tolerance of 5% (gold ring). At a power rating of 100 W, they will be loaded together with approximately 4-5 W, which is permissible during swept-frequency operation over a longer period.

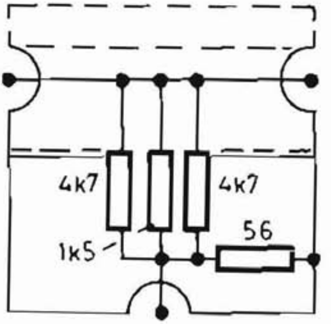

Fig. 20:
Circuit diagram and constuction of a 30 dB coupler

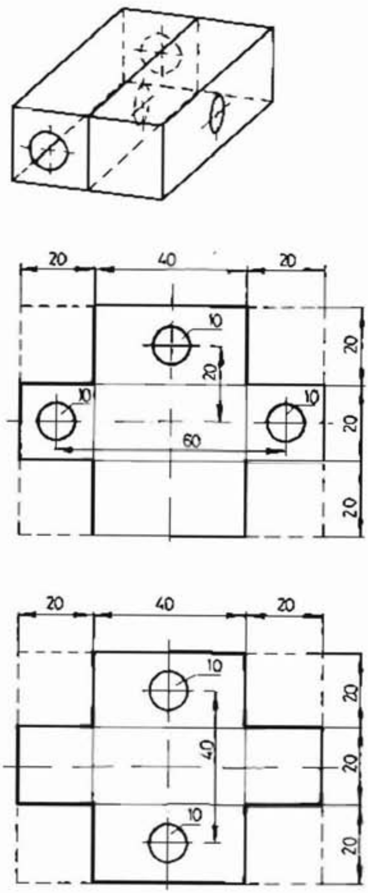

Fig. 21:
Perspective diagram and dimensions of the two metal-plate sections for the case of the 30 dB coupler. After completion, a cover is provided to ensure that the coupler is RF-tight.

Figure 21 gives a recommendation for the mechanical construction of the coupler from metal plate. It comprises two chambers with a hole in the intermediate panel through which the three high-impedance resistances protrude into the second chamber from the continuous 50 Ω line.

Figure 22 shows the author's prototype of the 30 dB coupler, and the measured attenuation values are given in **Table 3.**

6.2. Test Detector

As can be seen in **Figure 23,** the 50 Ω termination is followed by a voltage doubler rectifier and a wideband RF-bypass and a choke. The measured curves given in **Figure 24** result when feeding the circuit from a signal generator. The circuit is built up in a self-supporting manner in a small, screened case. **Figure 25** shows the author's prototype. The two 100 Ω resistors are connected directly from the center pin of the input connector to the right and left and are soldered to the case. A disk or chip capacitor is used for coupling to the Schottky diodes. Two ceramic supports are used to mount the subsequent component connections. These are just as uncritical as the component values themselves, since a calibration must be made if one is to use the detector for exact measurements.

No calibration is required for the swept-frequency alignment of the power amplifier described in Part 1. **Figure 26** shows how the equipment is to be connected together. It is important that the detector is sufficiently RF-tight so that it is only driven via the 30 dB coupler.

f MHz	Attenuation I/II	Attenuation I/III	Return loss at I. II and III terminated with 50 Ω
100	0.2 dB	31 dB	25 dB
200	0.2 dB	30.5 dB	19 dB
300	0.3 dB	29.8 dB	21 dB
400	0.3 dB	29 dB	17 dB
500	0.2 dB	28.5 dB	20 dB

Table 3: Measured values of the 30 dB coupler from 100 to 500 MHz.

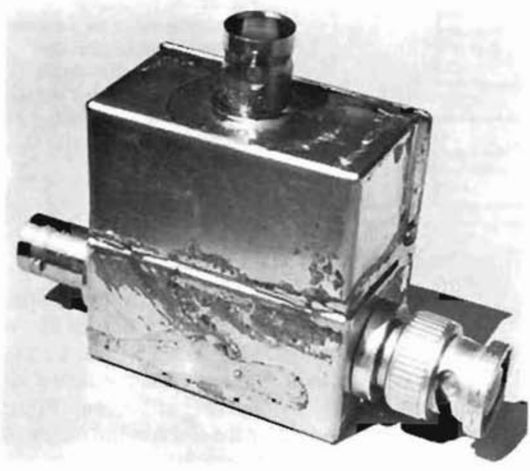

Fig. 22:
Photograph of the author's prototype 30 dB coupler - Simple but effective!

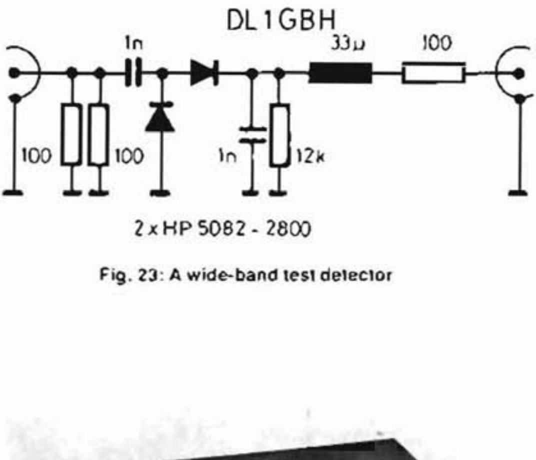

Fig. 23: A wide-band test detector

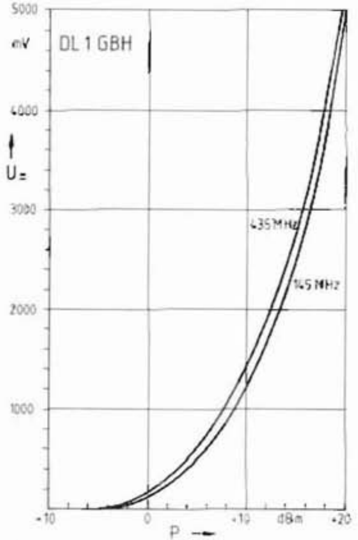

Fig. 24:
The two measured curves indicate that the detector could be used over a considerably greater frequency range.

Fig. 25:
The detector was installed in an available case. The two holes have nothing to do with the described application.

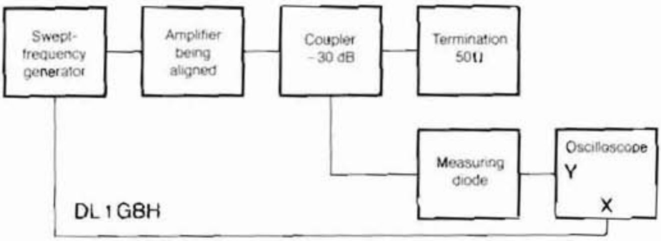

Fig. 26:
Measuring setup for a swept-frequency alignment: The 50 Ω termination must be able to handle the full output power of the amplifier.

7. HARMONIC FILTER FOR THE POWER AMPLIFIER

Transistor power amplifiers usually have a harmonic suppression of only approximately 30 dB to a maximum of 45 dB. This is usually not sufficient. For this reason, a filter is to be described that allows the second and third harmonic to be further suppressed usually by at least 60 dB.

The most simple filter for attenuating the harmonics is a low-pass filter. However, simple low-pass filters do not provide sufficient attenuation of the unwanted harmonics and must be made steeper.

If one constructs a steeper low-pass filter in conventional technology, one will require components having a higher breakdown voltage, since voltages of several kV can be present under resonant conditions. In the case of inductances, the breakdown voltage capabilities can be achieved easily, however, one must usually use expensive mica capacitors. Furthermore, one requires relatively complicated measuring equipment for aligning such a filter.

These problems can be solved when using the following filter where coaxial lines are used for increasing the steepness of the low-pass filter. These coaxial lines are λ/4 for the harmonics. In the pass-band range, the capacitance of the lines is taken into consideration in the parallel capacitance of the low-pass filter. One can obtain a high reduction of the harmonic content in this manner together with a low insertion loss in the passband range, and obtain a good reproducibility.

7.1. Calculation of the Filter

Firstly, a 5-pole Tschebyscheff low-pass filter was calculated according to the Book of Tables given in (1). After this, λ/4-lines for the second and third harmonic were inserted instead of the parallel capacitances. Since these lines represent too low a capacitance at the required frequency range of 145 MHz, additional capacitances are added, also in the form of pieces of coaxial cable.

7.1.1. The Basic Filter

Low-pass filter type TS/21 dB was selected, which is a 5-pole Tschebyscheff low-pass filter with a return loss of 21 dB (≙ VSWR = 1.2). An (upper) cutoff frequency of 150 MHz was inserted, and a reference impedance of 50 Ω. The following low-pass coefficient can now be taken from the Book of Tables:

$a_1 = a_5 = 0.932714$
$a_2 = a_4 = 1.366557$
$a_3 = 1.762498$

These allow the reference inductances and capacitances to be calculated and one will obtain:

$L_B = 53.05$ nH
$C_B = 21.22$ pF

The component values of the filter **(Figure 27)** can be calculated from the coefficients and the reference values:

$L_1 = L_B \times a_1 = 53.05$ nH $\times 0.932714 =$ **49.5 nH**
$L_3 = L_B \times a_5 = L_1 =$ **49.5 nH**
$L_2 = L_B \times a_3 = 53.05$ nH $\times 1.762498 =$ **93.5 nH**
$C_1 = C_2 = C_B \times a_2$ or $a_4 =$ **29 pF**

The calculated filter attenuates the second harmonic at 290 MHz by 30 dB, and the third harmonic at 435 MHz by 48 dB.

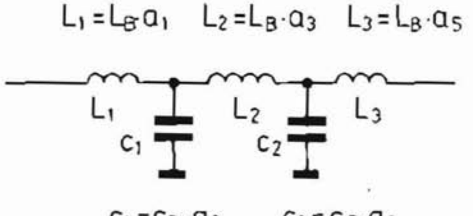

Fig. 27: Calculating the low-pass filter

7.1.2. Steepening the Filter Slopes

The $\lambda/4$ lines at 290 MHz and 435 MHz, respectively, are used instead of the parallel capacitances C_1 und C_2. They represent short-circuits for both these harmonics, which considerably increases the harmonic attenuation

These lines represent capacitances at the fundamental frequency of 145 MHz, which are less than the 29 pF calculated in section 7.1.1. The missing „additional capacitances" are also obtained using pieces of cable (Figure 28).

C_1 is replaced by a $\lambda/4$ line at 290 MHz This cable is $\lambda/8$ at 145 MHz and has an impedance of $-j50\ \Omega \triangleq 21.95$ pF This means that 7.05 pF are required in order to obtain the required 29 pF. The piece of cable used for this must have an input impedance of $-j155.76\ \Omega$, which is obtained with an electrical length of $0.04944\ \lambda$.

C_2 is obtained using two open lines of $\lambda/4$ length for 435 MHz. These lines will have an electrical length of $\lambda/12$ at 145 MHz which corresponds to an input impedance of $-j86.6\ \Omega \triangleq 12.67$ pF. Since two lines are used here, they result in a total capacitance of 25.34 pF, and a further 3.66 pF are required to complete the required 29 pF. The required piece of cable must have an input impedance of $-j300\ \Omega$, which means that it must have an electrical length of $0.02625\ \lambda$ at 145 MHz.

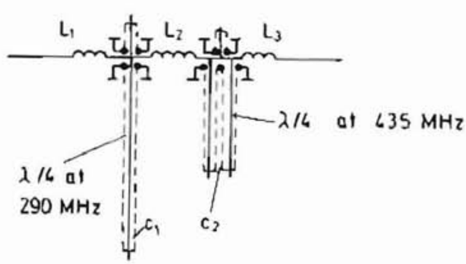

Fig. 28:
Realization of the steepened low-pass filter

One requires the mechanical lengths of the cables for construction. If RG-188/U is used, having a velocity factor (VF) of 0.71, the following lengths will result:

$\lambda/4$ at 290 MHz:

$$l_1 = \frac{c \times VF}{4f}$$

$$= \frac{3 \times 10^8 \times 0.71}{290 \times 10^6 \times 4}$$

$$= 0.184\ m = 184\ mm$$

$\lambda/4$ at 435 MHz:

$$l_2 = 0.122\ m = 122\ mm$$

Additional capacitances to l_1

$$l_{ad.} = 0.04944 \times \lambda_{144} \times VF$$
$$= 0.0726\ m = 72.6\ mm$$

Additional capacitance to l_2:

$$l_{ad.} = 0.02625 \times \lambda_{144} \times VF$$
$$= 0.03855\ m \approx 38.6\ mm$$

7.1.3. Inductivities

The inductances are calculated according to the following equation:

$$L = \frac{D^2 \times N^2 \times 10^3}{45D + 100l}$$

L in nH
D in cm
l in cm

$$N = \sqrt{\frac{L \times (45D + 100l)}{D^2 \times 10^3}}$$

In the case of $l = 4$ cm and $D = 1.6$ cm, the following winding measurements for the air-spaced coils will result:

$L_1 = L_3 = 49.5$ nH $\quad N = 3.02$ (turns)
$L_2 = 93.5$ nH $\quad N = 4.15$ (turns)

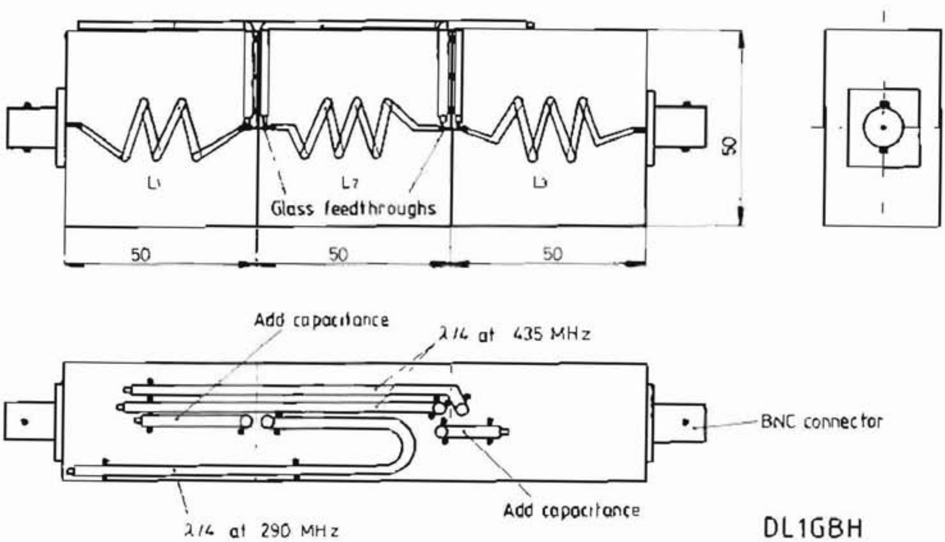

Fig. 29: Construction of the steepened low-pass filter

7.2. Construction

Figure 29 shows a drawing of the whole harmonic filter, a photograph of the author's prototype is given in **Figure 30.** A metal box can be provided with two intermediate panels and constructed as shown in **Figure 31**, and the three inductances can be made according to **Figure 32**. The five pieces of cable are prepared as shown in **Figure 33**. The additional capacitances should be cut somewhat longer and shortened later during the alignment to a minimum insertion loss of the filter at 144 MHz.

The coaxial cables are fed into the chamber through holes on the outside, and passed along the inside of the intermediate panels. They are finally soldered to the glass feedthroughs. As indicated by the black dots in

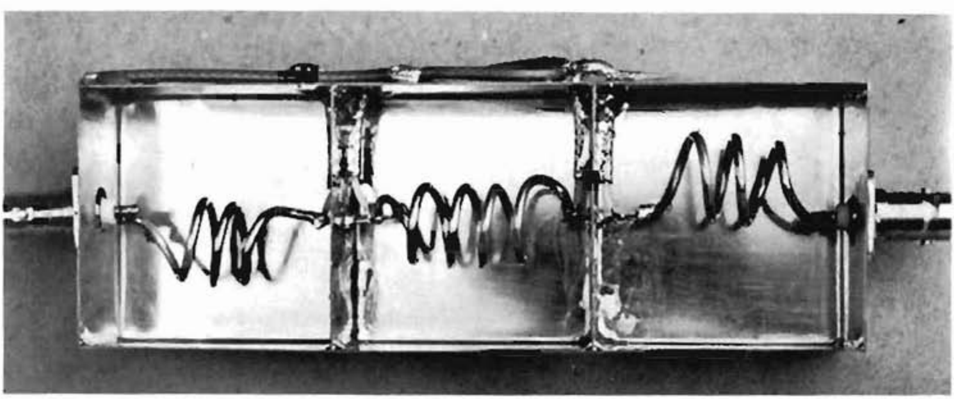

Fig. 30: The final filter is provided with a cover during operation in order to ensure that it is RF-tight.

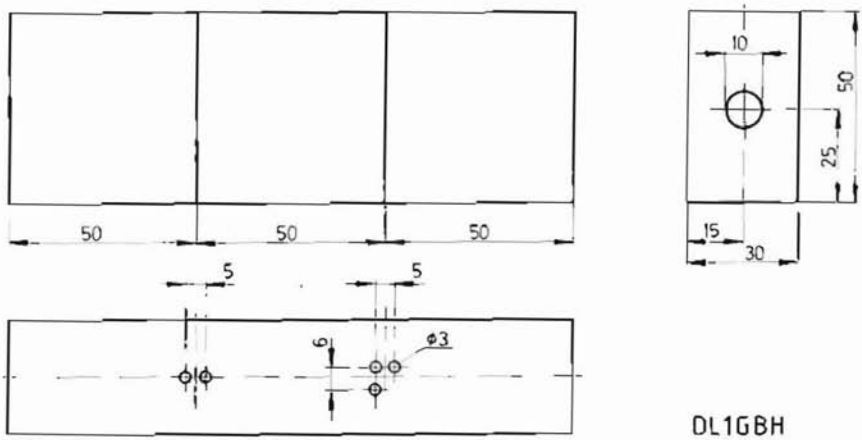

Fig. 31: Since the dimensions of the case are uncritical, not all individual dimensions are given. 0.7 mm thick tin plate is suitable for construction.

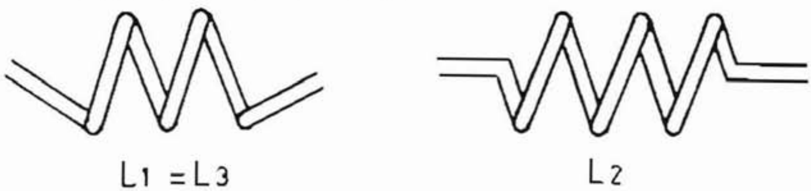

Fig. 32: L 1 = L 3 = 3 turns
L 2 = 4 turns of 2mm dia. silver-plated copper wire wound on a 14 mm former, self-supporting.

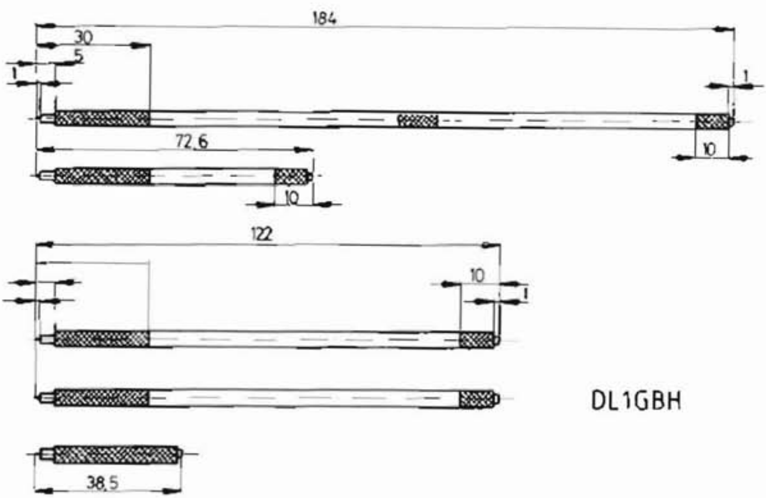

Fig. 33: All cables are from RG-188/U (50 Ω). Solder the two upper lines between L 1 and L 2, and the lower three cables between L 2 and L 3.
Shaded pieces: Remove the outer insulation and solder the sheathing into place after installation.

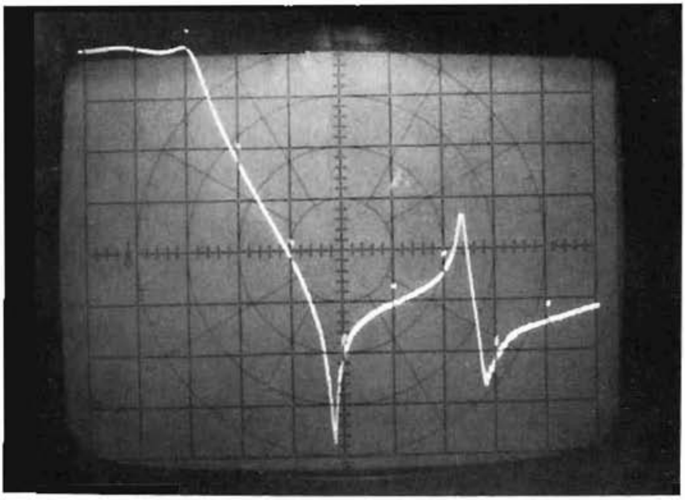

Fig. 34: Frequency response with markers every 50 MHz (at the cutoff frequency: 150 MHz). Vert.: 10 dB/line (at the limit frequency: approx. 0 dB).

Figure 29, the outer conductor of the cable is soldered at several positions to the metal case. Since PTFE cable is to be used, this does not represent any problem

7.3. Results

The filter was tested in conjunction with a tube power amplifier stage with a continuous output power of 500 W. No breakdown voltages were observed, and the inductances became only handwarm. The frequency response between 50 and 550 MHz is shown in **Figure 34.** One can clearly see the additional suppression caused by the λ/4 cable at 288 MHz and 435 MHz, as well as a over – resonance effect at approximately 420 MHz The latter has no effect since an amateur transmitter for the 2 m band does not have any harmonics at this frequency.

The return loss in the 2 m amateur band is greater than 20 dB, and the insertion loss is less than 0.3 dB

7.5. References

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2) G. Rose
Große Elektronik-Formelsammlung
Franzis-Verlag Munchen

3) J. Kammerloher.
Hochfrequenztechnik I
C. F. Winter'sche Verlagshandlung Prien

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described in edition 4/1983 of VHF COMMUNICATIONS

DJ4GC	A sensitive Thermal Power Meter		Art. No.	Ed. 4/1983
PC-board	DJ4GC 001	50 mm long piece of RT/duroid 5870 with etched 50 Ω stripline		
Parts	DJ4GC 001	1 pair of NTC resistors K 19, 12 kΩ/10%, 3 micro chip-resistors 51 Ω (2 of these as "spare parts"), 1 dual OpAmp TL082CP, 1 red and 1 green LED, 2 switching diodes		
Kit of special parts, DJ4GC 001, with the above mentioned parts			6796	DM 135,—

DC0RZ	A 30 MHz FM-Receiver for SHF receive systems			Ed. 4/1983
PC-board	DC0RZ 001	single-coated, without component location plan, drilled		
Parts	DC0RZ 001	3 transistors, 3 ICs, 3 diodes, 6 ready wound coils, 1 ceramic filter, 22 resistors, 18 ceramic and 5 tantalum electrol capacitors, 1 tinned-metal case		
Kit	DC0RZ 001	complete with above parts	6797	DM 99,—
Module	DC0RZ 001	ready to operate	6798	DM 165,—

Special parts:

Dual-gate GaAs-FET S3030 (TI) or MRF 966 (Motorola)	S3030	9073	DM 35,—
	MRF 966	9698	DM 34,50

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12 segments to be screwed or riveted together,		Riveting machine + rivets	0105	93.00
3 plastic supports for radiator,		1.7 GHz Cavity radiator kit	0091	90.00
mast-mounting parts with elevation mechanism		3 radiator supports	0106	29.00
		Mast-mounting parts	0107	85.00
Low-noise amplifier for 1.7 GHz (Originally described for use at 2.4 GHz, this unit is tuned to 1.7 GHz)	1/1980	DJ6PI 010	6565	225.00
METEOSAT Converter, consisting of	4/1981	DJ1JZ 003	6705	189.00
two modules – Output first IF = 137.5 MHz)	1/1982	DJ1JZ 004	6714	185.00
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