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for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

VHF communications

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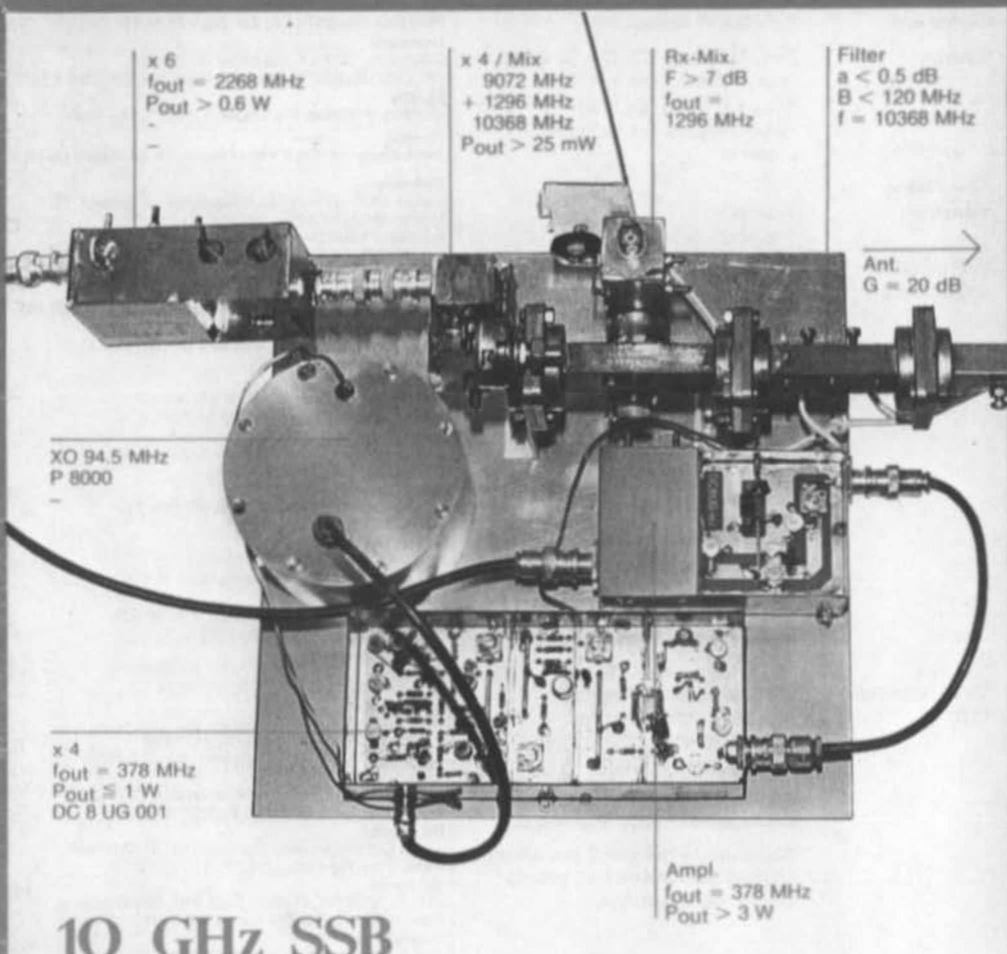

10 GHz SSB

You will have noticed that we have brought several descriptions recently that are of a far higher state-of-the-art than equipment already available on the market. This is not only the case with equipment for UHF and SHF but also such equipment as the DJ 7 VY and DJ 3 VY converters for 2 m and 70 cm. Also of interest is the design SSB equipment for 10 GHz. These are all designs that still make home construction worthwhile inspite of the multitude of equipment that is offered on the amateur market from Japan.

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A Publication for the Radio Amateur
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The long evenings of the autumn and winter months are ideal for construction projects. We hope that you will find something of interest in this, past or future editions of VHF COMMUNICATIONS.

We would like to point out that VHF COMMUNICATIONS is still not paying its way and is still being subsidized by the German language edition. It is therefore very important to increase the number of subscribers if we are to continue publishing the English-language version. We would therefore be grateful if you could support your VHF-UHF magazine by recommending it to your friends and other club members.

Happy building – G 3 JVQ / DJ 0 BQ

SSB on the 10 GHz Band

Part 1: Generation of the Local Oscillator Frequency

by H. Fleckner, DC 8 UG and
G. Börs, DB 1 PM

This three-part article introduces an SSB system for the 3 cm band having a first intermediate frequency in the 23 cm band.

Part 1 of this article is to describe the generation of the local oscillator frequency, Part 2 describes the waveguide modules, such as transceive and receive mixers, waveguide switch and filter. Finally, Part 3 will describe modifications for various different intermediate frequencies, as well as describing a fixed station and discussing the experience gained using the system.

Figures 1 and 2 show two low-power SSB systems, (30 mW) that are mainly designed for portable operation. The system shown in Figure 1 uses a waveguide switch that feeds the oscillator signal to the receive mixer in the receive mode using a shutter. The second design uses a straight-through mixer similar to that used in the Gunn oscillator technology (1). The individual modules can still be used when extended with preamplifier and power amplifier, or even when using a different intermediate frequency. The photograph on the front

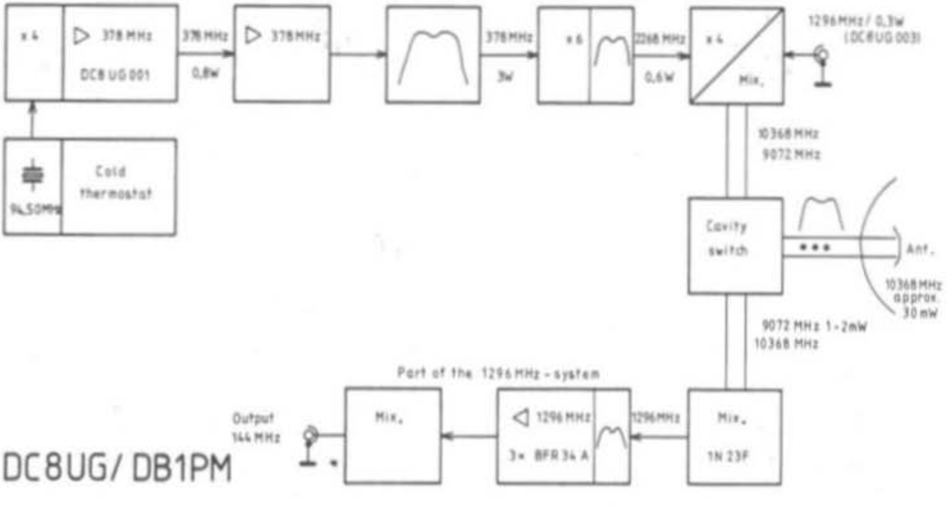


Fig. 1: A 10 GHz SSB system with cavity switch

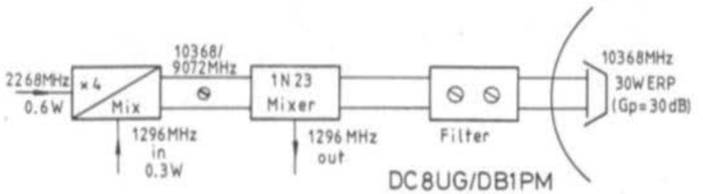


Fig. 2: A 10 GHz SSB transverter with through-line mixer

cover of this magazine shows the complete system.

1. GENERATION OF THE LOCAL OSCILLATOR FREQUENCY

As can be seen in the block diagram given in Figure 1, the subharmonic transmit mixer must be provided with a local oscillator frequency of 2268 MHz. This local oscillator signal is obtained by frequency multiplication of a crystal oscillator frequency of 94.5000 MHz. This is obtained using the following modules: crystal oscillator with oven, 378 MHz multiplier, amplifier, and frequency multiplier (x 6).

1.1. Crystal Oscillator

SSB-operation in the 10 GHz band requires an extremely stable local oscillator frequency for the first mixers, if the advantage of this technology with respect to wideband systems is to be utilized to the full. When using a first IF of 1296 MHz, it is possible to use a crystal-controlled frequency of 94.5 MHz and to multiply this 96 times to 9072 MHz in a frequency multiplier chain. Due to the high multiplication factor, three stability demands are placed on the crystal oscillator:

- Long-term stability
- Short-term stability
- Mechanical stability.

The long-term stability determines the accuracy of the signal frequency and can be improved by using an aged crystal with better temperature characteristics than the crystals usually used for amateur applications.

In order to minimize the effects of external temperatures, the oscillator should at least be insulated thermally (cold-crystal oven).

The short-term stability, as well as sideband and phase noise are mainly determined by selection of the oscillator circuit as was shown in (2). At high frequency multiplication factors, it is necessary for an oscillator to be used that has a very low sideband noise level per Hz of bandwidth, otherwise the receive mixer can be blanketed already by signals adjacent to the receive frequency within the receive bandwidth. Two-stage oscillators operate in class A equipped with high-current FETs that proved to be very satisfactory. Values of up to – 165 dB/Hz have been reached at a carrier frequency spacing of 100 kHz. Even a single-stage P 8000 overtone crystal oscillator as was used in (3) is able to achieve a sideband noise of < – 160 dB/Hz at > 100 kHz carrier spacing. This means that it is also suitable for feeding a frequency multiplier chain, and is to be used in the author's prototypes.

The circuit diagram, and a small PC-board (DC 8 UG 004) for constructing this oscillator, are given in **Figures 3 and 4.**

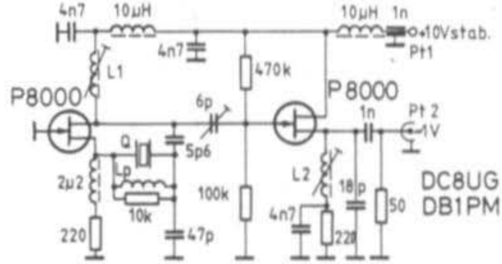


Fig. 3: Circuit diagram of a low-noise crystal oscillator

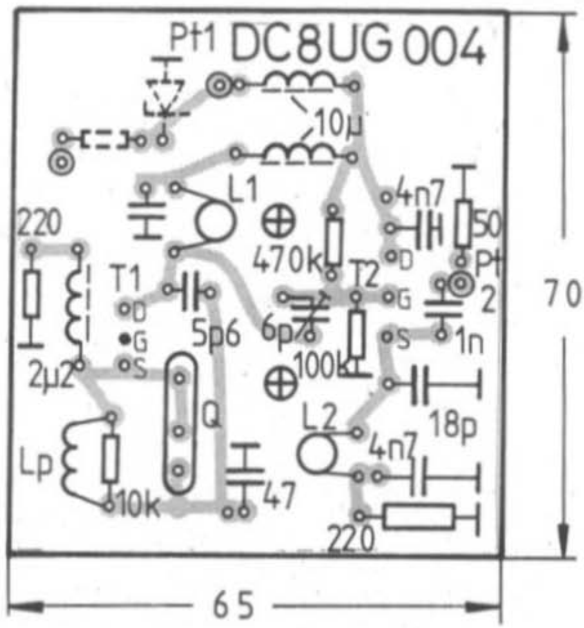


Fig. 4:
PC-board DC 8 UG 004 for the crystal oscillator given in Figure 3

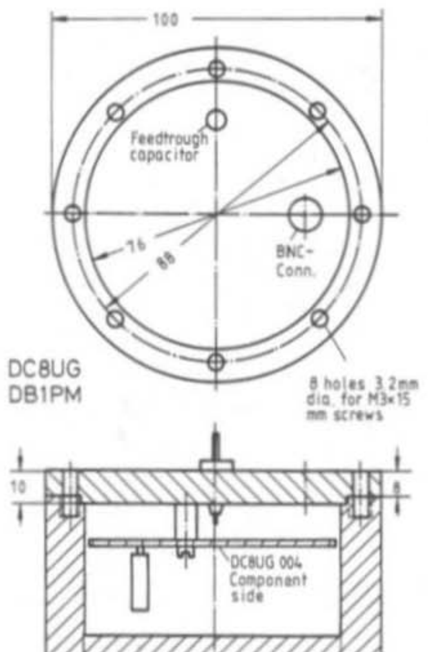


Fig. 5:
A cold thermostat for the crystal oscillator using a thick metal case

1.1.1. Components for the Crystal Oscillator

2 power-FET P 8000 or subsequent type
Miniature chokes for 15 mm spacing:
2 x 10 µH (value uncritical), 1 x 2.2 µH

L 1:	5 turns of 1 mm dia. silver-plated copper wire wound on a 5 mm coil former with VHF-core
L 2:	4 turns, otherwise as L 1

$L_p = 1/\omega^2 C_0$ provides approx. 0.5 µH together with C_0 = 6 pF: approx. 18 turns of 0.3 mm dia. enamelled copper wire wound on a 10 kΩ resistor of 3 mm dia.

Crystal: 94.50000 MHz with C_0 = 6 pF; HC-6/U or HC-18/U

Alignment accuracy:
$\leq \pm 5 \times 10^{-6}$ at 20°C

Temperature response:
$\leq \pm 10 \times 10^{-6}$ between – 20 and + 70°C.

In order to avoid microphonics, (mechanical stability), it is advisable to mount the oscillator board in rubber in a very strong case. This case also serves as a cold crystal oven. **Figure 5** gives the dimensions of the case used by the author and is 100 mm in

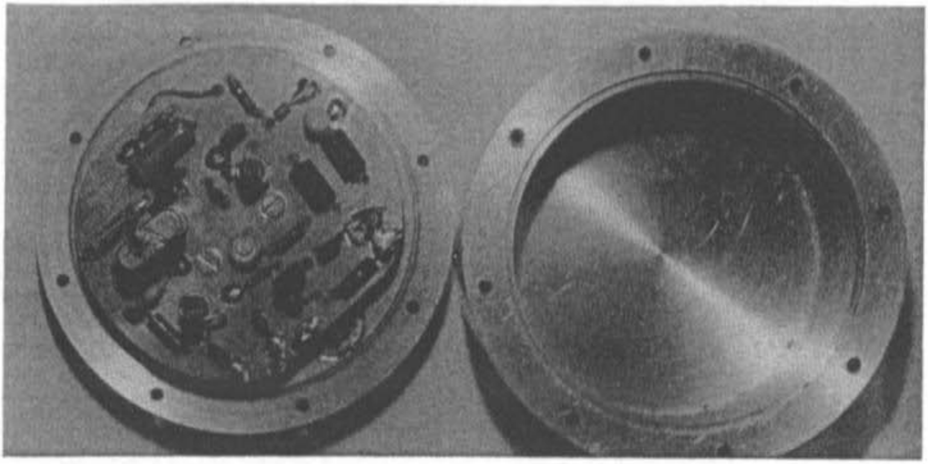


Fig. 6: Photograph of the oscillator board mounted to the lid of the cold thermostat

diameter and has been lathed from Dural. However, this is only one of many possibilities, and is thus not to be described in detail here. The shape of the oscillator board can also be made to suit the case to be used. Components are mounted on the ground side of the board, and the board itself is then fixed to the cover which should be provided with feed-throughs for voltage and RF (see **Figure 6**).

Attention should be paid that no noise-generating zener diodes in the stabilizing circuit of the power supply will deteriorate the sideband noise value of the oscillator.

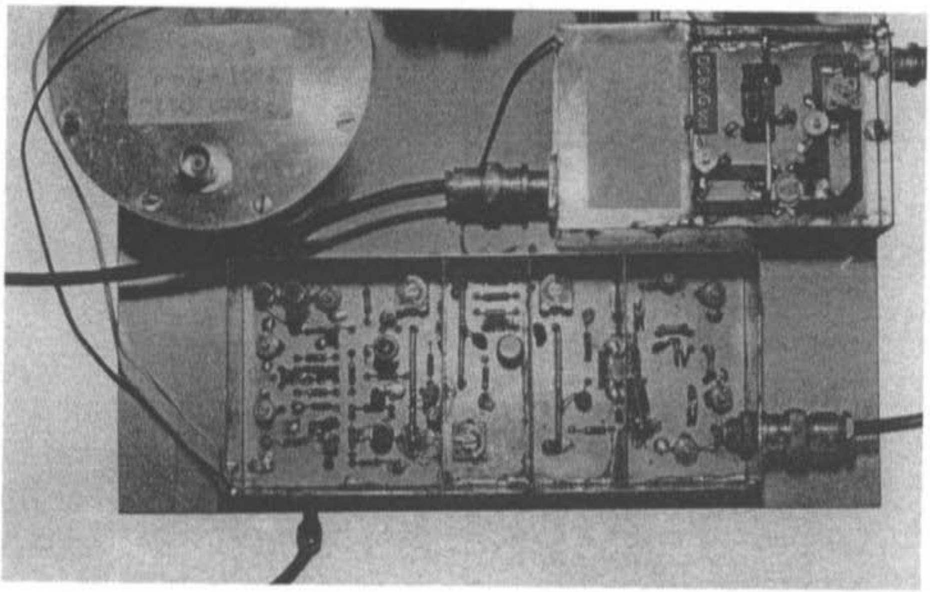


Fig. 7: Photograph of the 378 MHz multiplier constructed on PC-board DC 8 UG 001

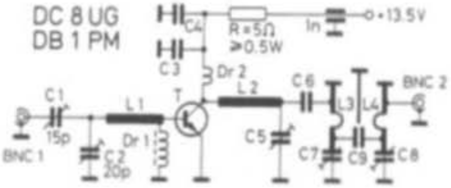


Fig. 8: A power amplifier for 378 MHz (6 dB/3 W)

The author (DC 8 UG) uses a buffered NiCd battery and operates the oscillator continuously so that no transcient time is required, which can amount to as much as 10 kHz per hour at 10368 MHz.

1.2. 378 MHz Frequency Multiplier DC 8 UG 001

The 378 MHz signal is generated in module DC 8 UG 001. The following modifications are required:

1. Instead of using the crystal, the oscillator signal is coupled to the base of T 1, using a 10 pF trimmer.
2. Trimmer C 1 is replaced by a 1 nF-capacitor.
3. A 270 Ω resistor is soldered in parallel to RFC 2 in order to avoid parametric oscillation.

With the exception of this, the board can be constructed and aligned as described in (4) **Figure 7** shows a photograph of this module.

1.3. 378 MHz Amplifier

Figure 8 shows the circuit diagram of this amplifier which is equipped with the CTC transistor C 3-12. This module is constructed on PC-board DC 8 UG 002 (4), and the required modifications are given in **Fig. 9**. The prototype is able to provide an output power of more than 3 W at a gain of 6 dB.

1.3.1. Components

T : C 3-12 (CTC)

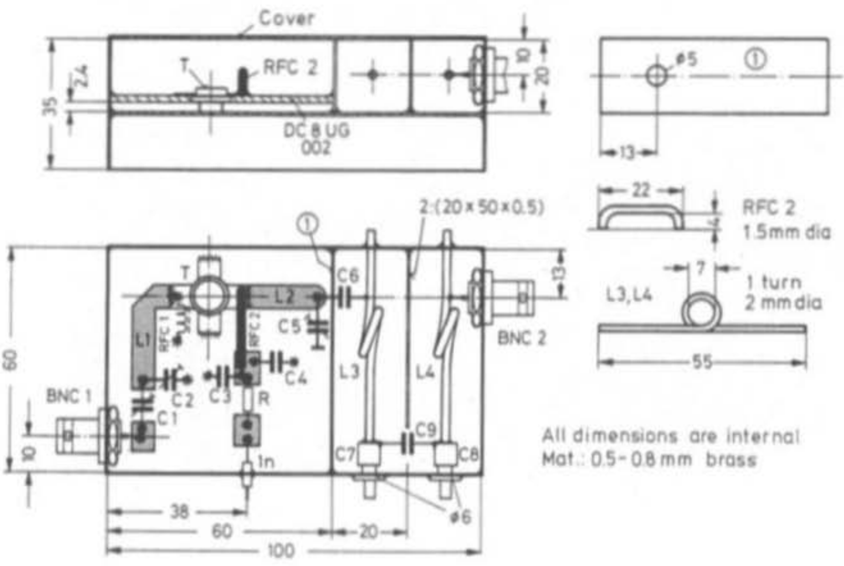


Fig. 9: Arrangement of the parts in the power amplifier

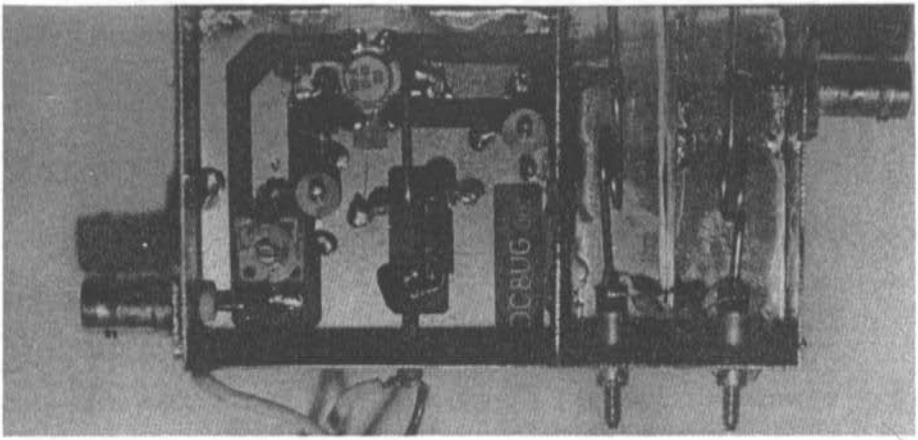


Fig. 10: Photograph of the 378 MHz amplifier without cover

L 1, L 2:	Etched on DC 8 UG 002
L 3, L 4:	Silver-plated copper wire of 2 mm dia., bent as shown in Figure 9
RFC 1:	1.5 turns of 0.4 mm dia. enamelled copper wire fed through a 5 mm ferrite bead
RFC 2:	Silver-plated copper wire of 1.5 mm dia., bent as shown in Figure 9 **It is also advisable to use this choke in the tripler described in (4) !**
C 1:	Air-spaced trimmer approx. 15 pF
C 2, C 5:	Plastic foil trimmers, green (approx. 20 pF)
C 3, C 4:	Ceramic disk 330 pF
C 6:	Tubular capacitor 100-560 pF
C 7, C 8:	Tubular trimmer, 6 pF (Philips)
C 9:	2 ceramic disk capacitors of 1.5 pF in series

Figure 10 shows a photograph of the 378 MHz-amplifier before soldering the cover over the bandpass filter.

1.4. Frequency Multiplier (x 6)

A frequency multiplier is used for multiplying the input frequency of 378 MHz by 6 to 2268 MHz. The circuit diagram of this is given in **Figure 11**. The reasons for multiplying by 6 were discussed in (5). The author's prototypes operate with a varactor diode DH 110 manufactured by Thomson-CSF. The frequency multiplier possesses one idler circuit each for the second and fourth harmonic, as well as a bandpass filter at the output which suppresses the subharmonics by more than 25 dB. This value is satisfactory for the transmit mixer.

The capacitive coupling is variable using the thread of a BNC-socket for single-hole mounting. This simplifies the matching to the subsequent reactance stage.

The efficiency was found to be 20 to 24 % in several prototypes using comparable diodes (5). The prototypes provide an output power of 0.7 W into 50 Ω when measured selectively for an input drive of 3 W. The multiplier is also suitable, without modification, to provide the 2160 MHz signal for 13 cm transverters.

Figure 12 provides the dimensions for construction of the frequency multiplier and shows all required parts; **Figure 13** shows the photograph of an author's prototype constructed using simple means.

1.4.1. Special Components

D : DH 110 (Thomson-CSF)

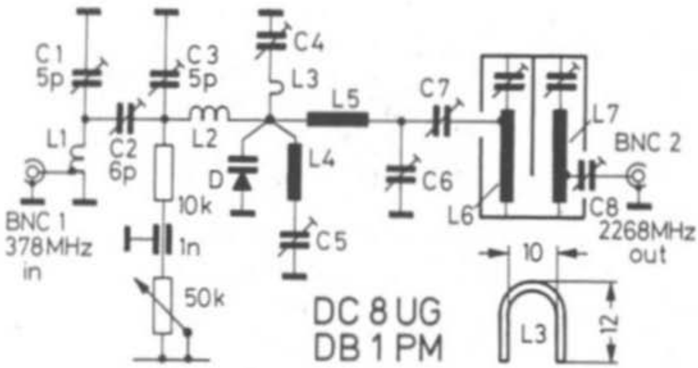


Fig. 11:
Circuit diagram of the frequency multiplier 378/2268 MHz

L 1, L 2:	2.5 turns of 1 - 1.5 mm dia. silver-plated copper wire wound on a 6 mm former, bent as required; L 1 tap: 0.75 turns from the cold end, L 2 as close-wound as possible
L 3:	2 mm dia. silver-plated copper wire, 12 mm wide, and 12 mm high and as shown in Fig. 11
L 4:	Brass or copper strip of approx. 0.5 mm thick, 22 x 5 mm, soldered at right angles to L 5
L 5:	as L 4, but only 12 x 5 mm
L 6, L 7:	Brass tube of 10 mm dia., 24 mm long, with a M 4 nut soldered to one end
C 1, C 3 - C 6:	Ceramic tubular trimmer 6 pF (Philips)
C 2:	Plastic-foil trimmer 6 pF, gray (Philips)
C 7:	Copper strip (see Part 11 in Fig. 12)
C 8:	Brass drawing pin of 9 mm dia. soldered to the inner conductor of the BNC-connector (shorten the pin as much as possible !)
Input:	BNC-connector with flange
Output:	BNC-connector for single-hole mounting

2. ALIGNMENT INSTRUCTIONS AND MEASURED VALUES

Measuring equipment:

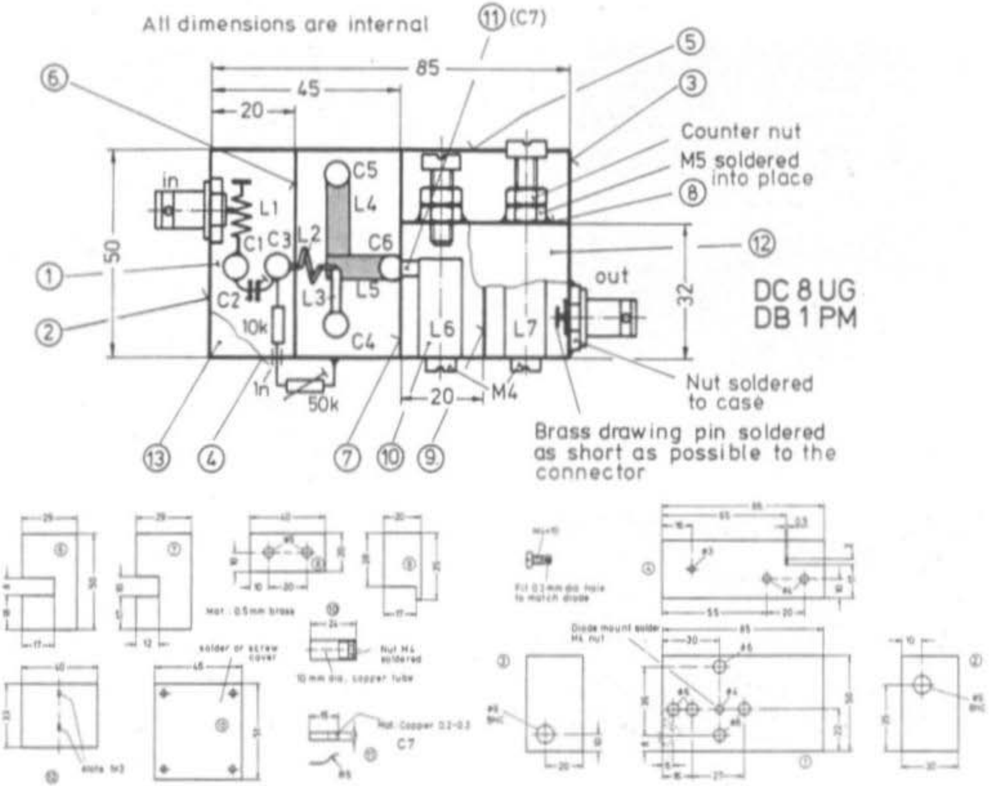
Power:	HP 435 and Bird wattmeter Thruline 43
Spectrum:	TS 1916 VPM 84 A, Polarad

2.1. Oscillator Module DC 8 UG 004

The frequency of the oscillator should be checked with a frequency counter and aligned to the required frequency by adjusting L 1. The oscillator should be checked for correct synchronization. Normally, this point is to be found just before maximum oscillation, and approximately 1 kHz below the nominal frequency ! L 2 is aligned for maximum. The 6 pF trimmer is now aligned to obtain a HF-voltage of 1 V (≙ 20 mW) on the external 50 Ω terminating resistor, which is followed by correcting the detuned circuit L 1. As can be seen in **Figure 14**, the oscillator spectrum is free from unwanted resonancies down to 70 dB. It was not possible to measure the noise sidebands.

2.2. 378 MHz Multiplier and Amplifier

The frequency multiplier will provide 1 W into 50 Ω at a spurious rejection of more than 40 dB [see (4)]. The C 3-12 amplifier will provide an output power of 3 to 3.5 W at 13 V at this drive level, and will also provide a spurious and harmonic rejection of 40 dB. Since the signals are not radiated, but are processed further, their spectrums are sufficiently clean **(Figure 15)**.



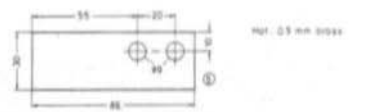


Fig. 12:
Important dimensions for the construction of the frequency multiplier

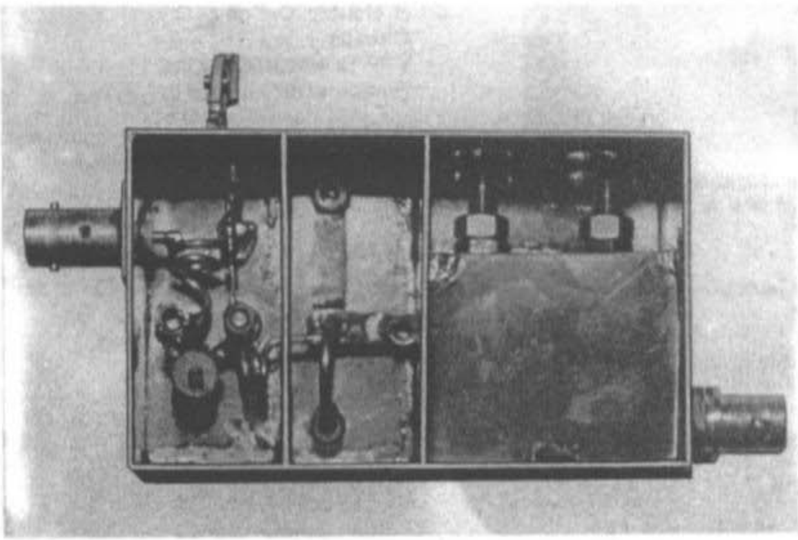


Fig. 13: Photograph of the prototype frequency multiplier

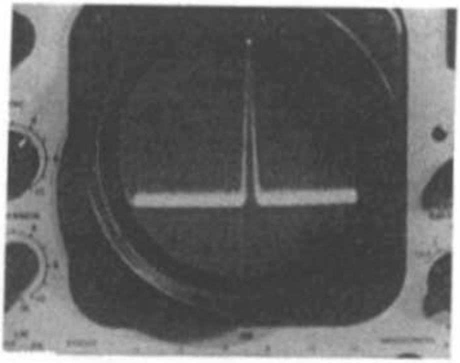


Fig. 14: Spectrum of the 94.5 MHz crystal oscillator
V: 10 dB/cm; H: ± 25 MHz

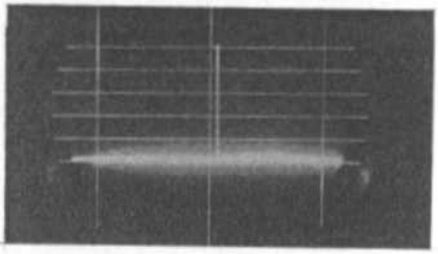


Fig. 15: Spectrum at 378 MHz
V: 10 dB/cm; H: ± 50 MHz

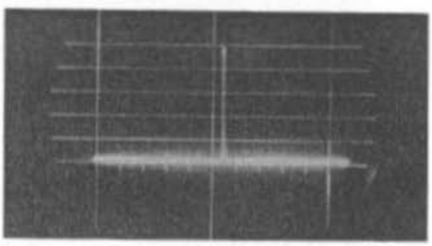


Fig. 16: Spectrum at 2268 MHz
V: 10 dB/cm; H: ± 50 MHz

2.3. Frequency Multiplier (x 6)

Attention should be paid during the alignment of this frequency multiplier that it is not aligned to the fifth harmonic instead of the sixth. This should be checked most favorably using a wavemeter or spectrum analyzer.

Firstly, the frequency multiplier is connected and driven with the 378 MHz signal. Trimmers C 1, C 2 and C 3 should be aligned alternately for maximum voltage drop across potentiometer R 2. This is followed by resonating the output circuit and indicating the power output using a suitable measuring device. If the idler circuits are in resonance, this will increase the voltage drop across R 2 considerably; the output power will be also increased at the same time.

The fine alignment is achieved by alignment of all variable capacitors and resistor R 2, and attention should be paid to the interaction of the idler circuits. A mismatch of the varactor stage will be seen immediately as a wide noise spectrum that will be audible in a 13 cm receiver. The cause of this is usually a mismatched input circuit. For this reason, all stages should be aligned to 50 Ω before interconnecting. The cover of the output circuit filter is soldered into place after aligning C 7. As can be seen in **Figure 16**, the final output spectrum possesses a spurious and harmonic rejection of more than 40 dB.

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VHF COMMUNICATIONS 10, Edition 3/1978, pages 145-153

Home-Made Parabolic Dishes for Microwave Applications

by S. Reithofer, DL 6 MH

The interest of radio amateurs in microwave technology has increased considerably in recent times. For this reason, the author is to describe a simple manner of constructing efficient parabolic dish antennas for the 10 GHz and 24 GHz bands.

It is true that horn radiators are simpler to construct, however, they become too large and unhandy when constructed for high antenna gains. This can be seen by comparison: A horn radiator for 10 GHz with approximately 25 dB gain has a length of approximately 60 cm and an aperture of 25 cm x 19 cm. In comparison, a parabolic dish of 30 cm diameter already exhibits a gain of 27 dB at this frequency and is considerably more favorable both in its dimensions, weight, and transportability than a horn radiator.

Manufacture of Parabolic Antennas

Of the many possibilities for manufacturing parabolic dishes, the author has found the following to be most favorable:

The main surface of the dish is made from a wire mesh grid which should be as fine as possible. It should not be more than 0.1 λ, and 0.05 λ would be better. This corresponds to a maximum of 3 mm for the 10 GHz band, and a maximum of 1.25 mm for the 24 GHz band. In the case of the author's prototype, a grid of 2.5 mm was used for 10 GHz, and 0.8 mm for 24 GHz. Such metal grids are readily available from hardware stores. The grid for 10 GHz should be welded at the crossover points; for 24 GHz, the author used a fine brass grid.

Construction of the Parabolic Form

The following equation is valid for construction of a parabolic:

$$y^2 = 4 \times f \times x$$

where y is the spacing of a point of the parabolic in the vertical plane from a central line, x is the spacing of each point from a vertical tangent in the horizontal plane and f the focal depth, or spacing of the focal point from the vertical tangent on the center line.

These magnitudes are given in the form of a drawing in **Figure 1**.

After transposing the parabolic equation, the following results:

$$x = \frac{y^2}{4 \times f}$$

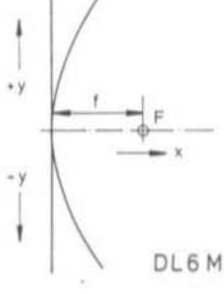

Fig. 1: The parabolic shape

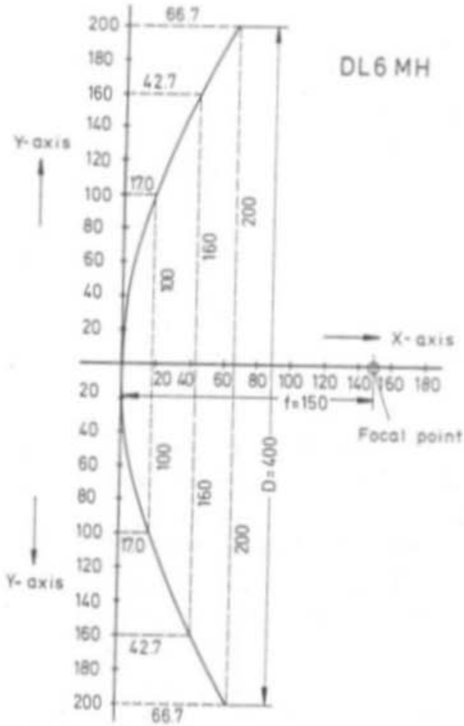

Fig. 2: Construction diagram of a parabolic dish of 40 cm diameter and a focal depth of 15 cm (f/D = 0.38)

For the given values of y, the following x-values result with f = 150 mm and D = 400 mm:

y mm	x mm
20	0.7
40	2.7
60	6.0
80	10.7
100	17.0
120	24.0
140	32.7
160	42.7
180	54.0
200	66.7

Several of these values are given in **Fig. 2.** For construction of the antenna, the above values are drawn to scale. This is followed by making a template. The required values for smaller or larger parabolic dishes can be calculated as required according to the above equation.

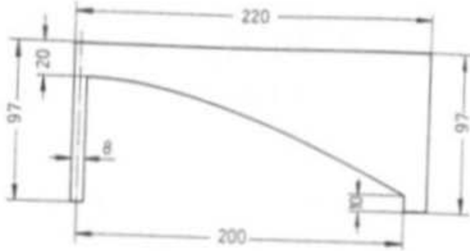

Fig. 3: Template for the plaster form of a dish as given in Figure 2

Construction of the Parabolic Template

The practical construction of the antenna is commenced by cutting a template in the shape of half of the parabolic form from metal plate. A metal plate of approximately 2 mm thickness is sufficient for constructing the 40 cm dish. In order to ensure that the template can be rotated around the central point, it is provided with a central pivot **(Figure 3)**. The dish is constructed on a large wooden board of approximately 50 cm x 50 cm. A tube of approximately 10 mm inner diameter is fixed in the center of this board with the aid of a wooden block or similar so that its top is mounted 77 mm from the surface of the board. The photograph given in **Figure 4** shows this phase of construction.

Fig. 4: Preparations for making the plaster form

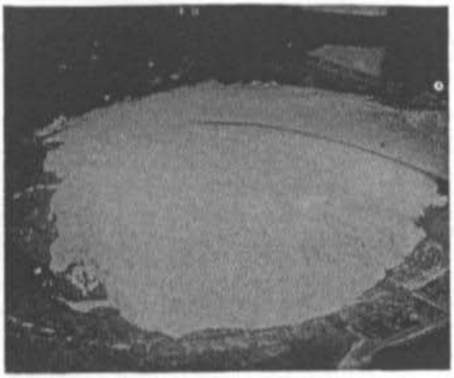

Fig. 5: Placing the plaster around the central hole

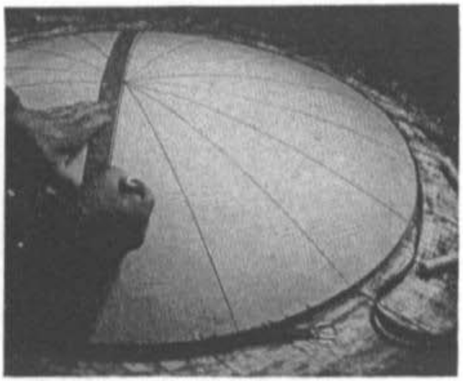

Fig. 7: The segments are now marked on the form

Manufacturing the Plaster Form

The plaster is now mixed with water and placed around the central tube. This can be mixed with sand for the central core to save the amount of plaster required. The plaster is now placed in roughly the required parabolic shape after which the template is placed in the central tube and scraped around the edge to remove excess amounts of the soft plaster.

In the case of the last layer, the plaster surface should be kept somewhat softer by adding water. In this manner it is possible to obtain a flat surface with the aid of the template. This stage of construction is shown in **Figure 6**. The form can be completely smoothed using a thick mixture of plaster and water and using a wide brush.

Fig. 6: Scraping off the excessive plaster using the template

The form is now allowed to harden for one or two days, after which it is divided to 10 or 12 segments with the aid of a fiber pen or pencil **(Figure 7)**.

Wire Grid Segments

The corresponding segments are now cut from a wire grid using shears, allowing approximately 10 mm extra for the outer edge. These segments are firstly formed to a rough parabolic shape on a hard surface using a light hammer.

After having formed the segments to the required shape, they are laid one after another on the plaster form and soldered firstly at three to five points using a large (approximately 150 W) soldering iron. Attention should be paid that the segments correspond to the shape of the form and are in direct contact with it. After completing the 360°, narrow, thin metal strips of approximately 8 to 10 mm in width (tin plate or brass) are placed into position in front of the soldered joints and cut so that they stop approximately 30 mm before the center of the dish. These metal strips are soldered continuously to the seams of the segments.

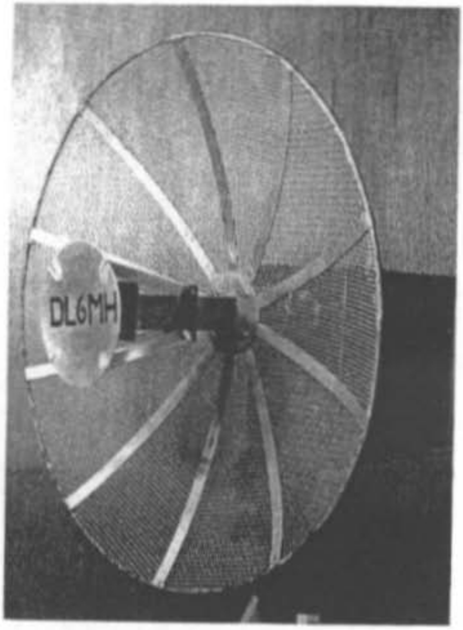

Fig. 8: The round plate for mounting the waveguide can be seen in the center of the completed 10 GHz antenna

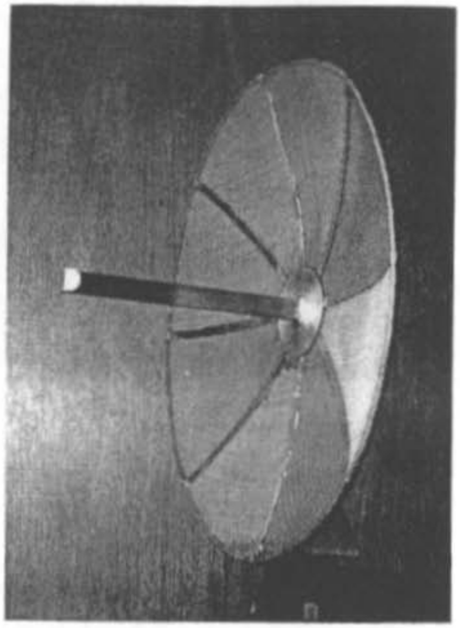

Fig. 10: An antenna for the 24 GHz band with radiator (D = 30 cm, f = 15 cm)

A ring of 4 mm diameter brass or copper wire is now placed around the edge of the plaster form and the excess grid at the edge of the dish is bent around this and soldered into place.

This will provide sufficient stability in the case of a parabolic dish of 40 cm in diameter.

In the case of larger dishes, vertical metal strips should be provided vertically on the rear of the dish for stability.

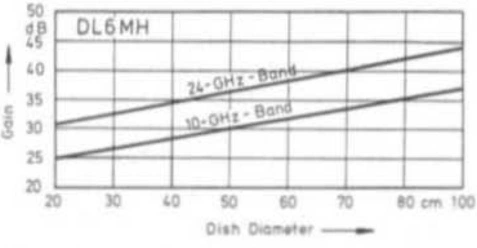

Fig. 9: Average antenna gain as a function of dish diameter

Waveguide Mount

The ends of the segments are cut at the center of the dish so that a hole of approximately 60 mm diameter is provided. An approximately 2 mm thick brass plate with a rectangular cutout for the waveguide is soldered into place above this hole. It is possible to provide such a metal plate at the front and back of the dish to provide improved stability, if required. The central plate of a 40 cm dish for mounting the waveguide can be seen in **Figure 8**.

Gain of Parabolic Antennas

The larger the diameter of the parabolic dish, the higher will be the gain and thus the narrower the beamwidth. The diagram given in **Figure 9** gives the approximate gain values that can be obtained in the 10 GHz and 24 GHz band.

Parabolic dish diameters of 30 to 40 cm have been found satisfactory for practical operation, especially during contests.

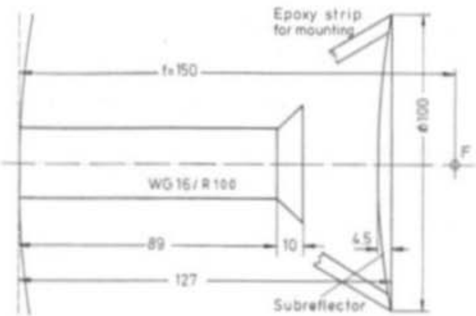

Fig. 11: A 10 GHz Cassegrain radiator for 40 cm parabolic dishes

Larger dishes have a very narrow beamwidth which means that it is very often difficult to set up the antenna to a required direction especially in the case of poor visibility. If the exact location of the station is known, it is possible with some experience for the antenna direction to be determined using a map and compass.

Radiators for Parabolic Dishes

The illumination and thus the overall efficiency of the antenna is mainly dependent on the radiator (sometimes called primary radiator or exciter). A very simple radiator was described in (1). The required cutout on the waveguide and the shape of the cap can be constructed without problems. The author has also used this type of radiator with success at 24 GHz. **Figure 10** shows a parabolic dish of 30 cm diameter which is equipped with this radiator.

A somewhat more extensive radiator is described in the VHF/UHF Manual published by the RSGB using a dipole and reflector for the 10 GHz band. It was described extensively (2) so that it is not necessary to go into it in detail here. The author has also tested such radiators successfully on the 10 GHz band.

If the parabolic dish is only to be used for transmitting, it is possible for the Gunn oscillator module to be mounted together with a short horn radiator of approximately 3 to 5 cm in length at the focal point of the dish. Suitable horn radiators were described in (3) and (4) together with their dimensions.

When using transceivers for the 10 GHz band such as the Microwave-Associates Gunn-Plexer that uses a varactor diode and a DC voltage for remote tuning, it is also possible for it to be installed in the focal point of the dish together with a small horn radiator. The mounting is usually made with three or four pieces of insulating material (strong strips of etched epoxy board material).

A somewhat more extensive method that can be easily achieved by radio amateurs, is the use of a Cassegrain system. In this

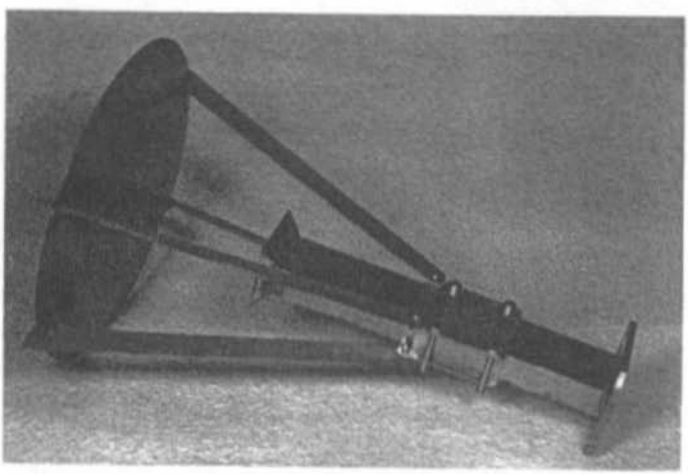

Fig. 12: Prototype of a Cassegrain radiator

case, the waveguide ends with a small horn which illuminates a round subreflector which reflects the energy efficiently to the main dish.

The construction of this Cassegrain system is described in detail in the VHF/UHF Manual published by the RSGB. A drawing with dimensions for the described 70 cm dish is given in **Figure 11**. The design given in **Figure 12** illuminates the dish well and possesses a very short depth.

The subreflector is made from soft brass or aluminium plate which has been hammered into shape on a metal plate. The required shape is given in the drawing. The subreflector is held into place using four strips of etched epoxy board material which are in turn mounted to the waveguide with the aid of a sliding bracket. The spacings such as the distance of the small horn from the center of the dish, and the spacing from the horn to the subreflector are not very critical and can be set using either an available measuring line for most favorable SWR, or according to the S-meter during operation.

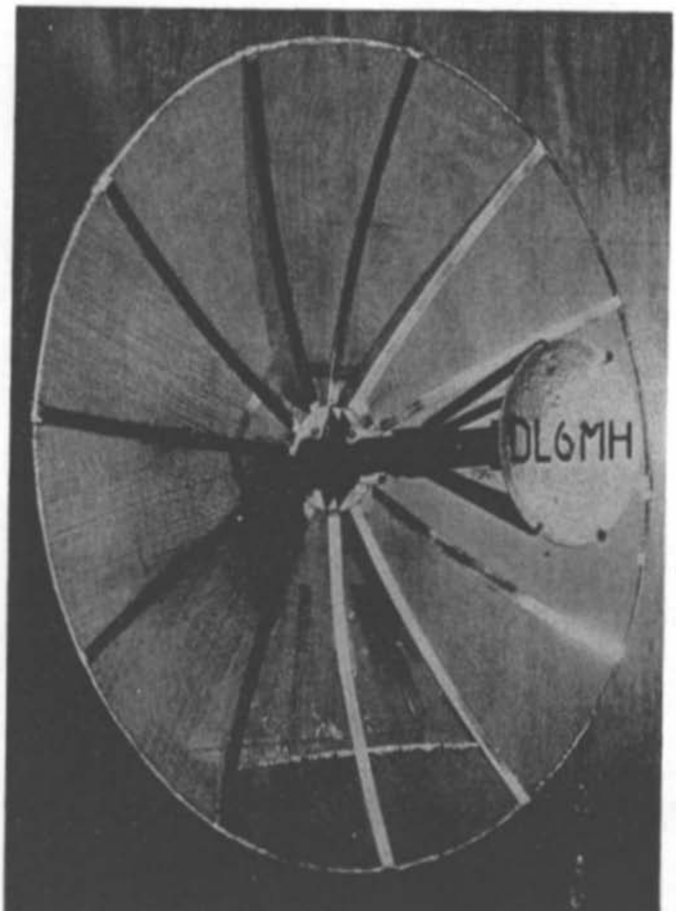

Fig. 13: Combined antenna for the 10 and 24 GHz band with Cassegrain radiator

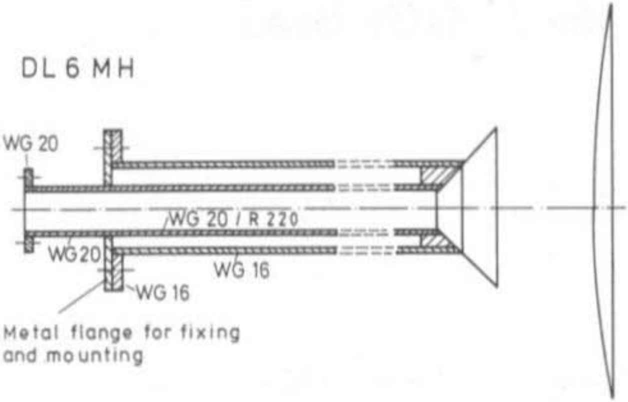

Fig. 14:
Combined radiator
for either 10 or 24 GHz

Combined Radiator for 10 GHz and 24 GHz

In order to save using two completely different antennas, the author has developed an antenna that can be used on both bands. In order to ensure that the antenna is also efficient at the higher frequency of 24 GHz, a wire grid of 0.8 mm spacing was used. The diameter of the antenna shown in **Figure 13** amount to 60 cm. A Cassegrain radiator for 10 GHz is fixed in the center of the dish. For 24 GHz, a second waveguide for 24 GHz (WG 20) is inserted into the 10 GHz waveguide (WG 16).

A brass block is soldered into place at the end of the WG 20 waveguide whose external dimensions are such that it just slides into the WG 16 waveguide. The brass block has a rectangular hole with the outer dimensions of the WG 20 waveguide. This aperture is filed out in the form of a horn.

It will be seen in **Figure 14** that the small horn of the 10 GHz radiator forms the extension for the 24 GHz radiator. The rear end of the WG 20 waveguide is screwed to the WG 16 flange.

If the dish is to be used in the 10 GHz band, the 24 GHz waveguide is pulled out and the antenna connected to the 10 GHz station. Since 10 GHz signals are stronger over greater distances than at 24 GHz, and since the dish will possess larger beamwidths at the lower frequency, it is easier to find the required station by setting the antenna up at 10 GHz, after which the dish is fixed into position and the WG 20 waveguide for 24 GHz placed into the WG 16 waveguide. This means that it should be possible to make communications on 24 GHz without directional problems.

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Waveguide for the 24 GHz Band

by E. Schaefer, DL 3 ER

Three types of waveguides are suitable for use on the 24 GHz band (amateur band from 24.000 to 24.250 GHz):

R 220 = WR 42 = WG 20
R 260 = WR 34 = WG 21
R 320 = WR 28 = WG 22

As can be seen in the data sheets of these waveguides, or in the table given in (1), the last mentioned type can only be used with some limitations at 24 GHz; however, it possesses a number of advantages, which are to be discussed.

Designation Systems

- IEC-standard: R 220

  The number following the R (R = rectangular waveguide) indicates the central operating frequency (x 10 GHz); in our example: 22 GHz

- US-standard: WR 42

  The number following the designation WR indicates the waveguide dimension »a« in decimal inches; in our example: a = 0.42 inches

- British standard: WG 20

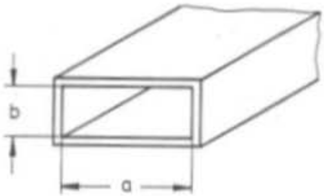
  The number following the designation WG (Wave Guide) is only a serial number that increases with frequency.

Dimensions

The waveguide dimension »a« (see drawing) of the three previously mentioned waveguides is:

R 220: a = 10.668 mm = 0.42"
R 260: a = 8.640 mm = 0.34"
R 320: a = 7.110 mm = 0.28"

The dimension »b« is only important for the breakdown voltage (radar-transmitters !). Of course, it does have some effect on the installation of semiconductors, and is also standardized. This will be discussed later.



Operating Frequency Range

The operating frequency range f_{op} given in the data sheets is as follows:

R 220: 17.6 – 26.7 GHz
R 260: 22.0 – 33.0 GHz
R 320: 26.4 – 40.1 GHz

f_{op} is defined as:

$$f_{op} = (1.25 \text{ to } 1.9)\, f_{co} \qquad (1)$$

and is only valid for the H_{10}-wave. This is the most important waveguide-wave type; it is usually exclusively used in the amateur radio technology.

The cut-off frequency f_{co} mentioned in equation (1) results from the limit wavelength λ_l:

$$f_{co} = \frac{c}{\lambda_l} \qquad (2)$$

where c is the propagation speed of the electromagnetic waves:

$$c = 300 \times 10^8 \text{ cm/s}$$

and $\lambda_l = 2a$

where a is the inner width of the waveguide (see drawing).

Example: Let us calculate the operating frequency range of waveguide type R 320:

$$\lambda_l = 2a = 1.422 \text{ cm}$$

$$f_{co} = \frac{300 \times 10^8 \text{ cm}}{1.422 \text{ cm} \times \text{s}} =$$

$$f_{co} = 210.911 \times 10^8 \text{ 1/s} = 21.091 \text{ GHz}$$

According to equation (1), the operating frequency range is then:

$$f_{op} = (1.25 \text{ to } 1.9)\ 21.091 \text{ GHz}$$
$$f_{op} \approx 26.4 \text{ to } 40.1 \text{ GHz}$$

Wavelength in the Waveguide

The wavelength in the waveguide λ_H is important for the design of equipment. It differs from the free-space wavelength λ_o by the ratio of λ_o/λ_l. The relationship is as follows:

$$\lambda_H = \frac{\lambda_o}{\sqrt{1-(\lambda_o/\lambda_l)^2}} \qquad (3)$$

Where $\lambda_o = c/f_o$

c = propagation speed
f_o = operating frequency in Hz
λ_l = 2a

It will be seen in equation (3) that the wavelength in the waveguide will approach infinite values when λ_o approaches the value λ_l, in other words, when λ_o/λ_l approaches zero. This means that a waveguide can no longer be used in the vicinity of its cut-off frequency f_{co}.

Finding One's Own Waveguide »Standard« ?

You may have noted that the waveguide dimension »b« is not mentioned in the equations. Practically speaking, it only determines the breakdown voltage of the waveguide at high power levels. For this reason, radio amateurs need not necessarily be bound to the sometimes very expensive, and difficult to obtain standard waveguides. There is nothing to stop us using metal profiles available in metalwork and model-building shops. Suitable brass center profiles are available that are very suitable for installing components.

The decision not to use standard waveguide parts means that it is also necessary to construct such things as the flahges. Also, one will not be able to connect standard measuring equipment to a home-made waveguide system unless suitable, calculated transitions from one system to another were provided.

Calculated Values for the Waveguide Wavelength

According to equation (3) the following values result for the wavelength λ_H for the corner frequencies of the 24 GHz amateur band:

Frequency in GHz	R 220/WR 42 λ_H	R 320/WR 28 λ_H
24.000	15.42 mm	26.19 mm
24.250	15.18 mm	25.07 mm

Waveguide Type R 320

This type of waveguide (WR 28) is used outside of its operating frequency range in the vicinity of its cut-off frequency. As will be seen in the above table, the waveguide wavelengths are approximately twice as long as in the larger waveguide type R 220. This is a small advantage: The electromechanical system will be increased by factor 2, which means that the mechanical tolerances can be greater and thus construction less critical.

This advantage must be compared with a considerable disadvantage: All discontinuances in the waveguide will lead to an increase of the standing wave ratio (SWR) and will excite higher wave types. These have an adverse effect on the fundamental wave (H_{10}), and extract energy from it, thus increasing the power loss. For this reason, it should be noted that waveguide type R 320 (WR 28) can only just be used for the 24 GHz amateur band when bends and other complicated configurations such as waveguide switches are not used and when restricted to short waveguide lengths. Waveguide type R 220 (WR 42), on the other hand, can be used without limitation.

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Modern Receive Converter for 70 cm Receivers Using DJ 7 VY 002 on the 70 cm Band

by M. Lass, DJ 3 VY

After constructing the large signal, low-noise 2 m converter DJ 7 VY 002 (1) without difficulties, the author decided to construct a 70 cm version using the same board. DJ 7 VY provided a number of tips regarding the construction of the doubler stages which must be very clean in the selected intermediate frequency range of 28 to 30 MHz. Of course, the oscillator signal should also possess a very low noise spectrum.

Since five crystals are required to cover the whole 70 cm band, the crystal oscillator is to be accommodated on a separate PC-board. The original board DJ 7 VY 002 is therefore suitable for use at the higher frequencies. Stripline circuits made from silver-plated copper wire contained in screened chambers are used instead of the coil bandpass filters used at 2 m. In order to use the available conductor lanes as much as possible, the coupling points were selected to suit the available construction.

The circuit is virtually the same as that for 2 m, which means that all details given in (1) are also valid for the 70 cm version. The input preamplifier was redesigned according to information given in (2).

1. CIRCUIT DESCRIPTION

The circuit diagram of the 70 cm converter without crystal oscillators is given in **Fig. 1**. The dimensions of the line circuits together with their coupling points are given in the circuit diagram; all dimensions are given in mm.

The ring mixer type TAK-1-WH is a high-level mixer of + 17 dBm for the frequency range from 5 to 750 MHz. When using this mixer, the 1 dB compression point will be reached at a signal level of + 14 dBm, whereas this will occur at + 1 dBm when using the standard-level mixer SRA-1 (+ 7 dBm).

Intermodulation details are also given in the data sheet for the TAK-1 WH:

The third-order intermodulation is at least 55 dB, typically 60 dB, down on the IF-signal with two input signals of 0 dBm each (200 and 202 MHz) and when using an oscillator power of + 17 dBm at 180 MHz.

The mixer is terminated at the IF-port in a wideband manner with 50 Ω using the diplexer described in (2).

Since the author's receiver already possesses a noise limiter in the 9 MHz-IF, this stage was deleted in the converter. Inspite of this, two amplifier stages equipped with the power FET P 8000 were used in order to achieve the same overall gain as was provided by other converters: If all converters exhibit a gain of 20 dB, this means that the calibration of the S-meter will be valid for all bands. The adjustment of the gain is made by selecting a suitable resistor value between the drain of the first P 8000 (T 3) and the operating voltage.

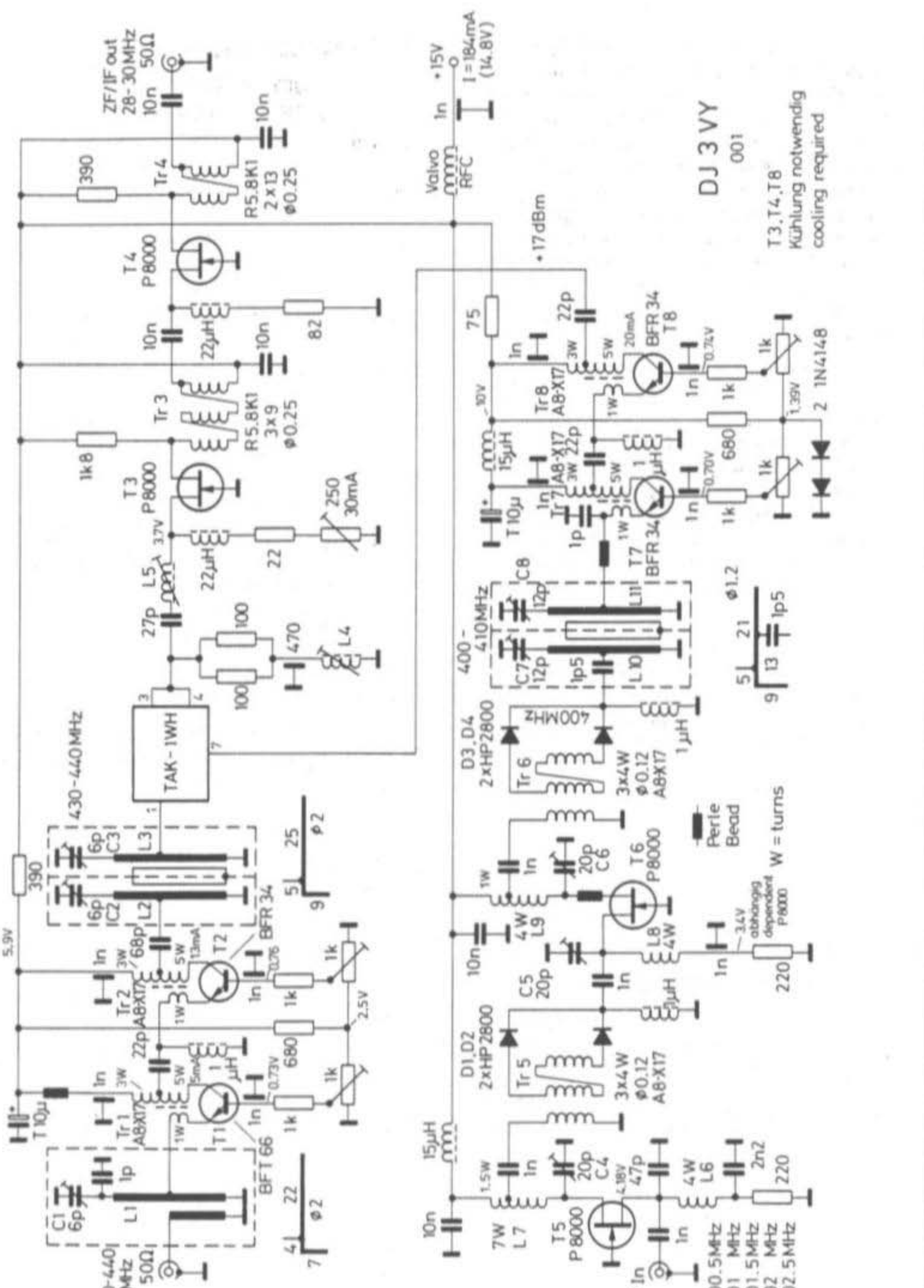

Fig. 1: The 70 cm version of the DJ 7 VY converter (without oscillators)

The circuit of the crystal oscillators was taken from (3) and somewhat modified according to (4) to make it suitable for 100 MHz, fifth overtone crystals. The final values are given in the circuit diagram given in **Figure 2**. An output power of approximately 200 mV into 50 Ω is provided after the single-stage buffer and the transformer. The individual oscillators (normally 5) are interconnected at the primary side of the toroid transformers; the selection of the required crystal oscillator is made by connecting it to ground.

The 100 MHz output power of the oscillator module DJ 3 VY 002 is fed via a coaxial cable to the local oscillator input of the converter module DJ 3 VY 001. The frequency is then multiplied in two frequency doublers equipped with Schottky diodes to 402 to 410 MHz. The required amplification is made in two FETs type P 8000. The circuit of this can easily be accommodated in the first chamber of the original 2 m converter.

A two-circuit stripline bandfilter is now used instead of F 4. A wideband amplifier equipped with two BFR 34 now amplifies the oscillator power in the same manner as in the 2 m version to a level of + 17 dBm (50 mW).

1.1. Special Components

T 1:	BFT 66 (Siemens)
T 2, T 7, T 8:	BFR 34 A (Siemens)
T 3 - T 6:	P 8000 (Texas Instruments) or P 8002 (different case !)
T 9a - e:	2 N 2222 (different manufacturers)
T 10a - e:	BF 246 C (TI)
Mixer:	TAK-1 WH (Mini Circuits)
D 1 - D 5:	HP 2800 (Hewlett Packard)

Stabilizing diodes: 1 N 4148 or similar

C 1 - C 3:	6 pF air-spaced trimmer with two pins (Tronser)
C 4 - C 6:	20 pF ceramic or foil trimmer (Philips), 5 mm diameter
C 7, C 8:	12 pF air-spaced trimmer with two pins (Tronser)
C 9a - e:	12 pF ceramic or foil trimmer

8 ceramic chip capacitors of approx. 1 nF without connection wires

1 ceramic chip capacitor 470 pF

2 ceramic feed-through capacitors for solder mounting, approx. 1 nF

2 tantalum electrolytics, 10 µF

All other capacitors: ceramic chip capacitors for 2.5 mm or 5 mm spacing.

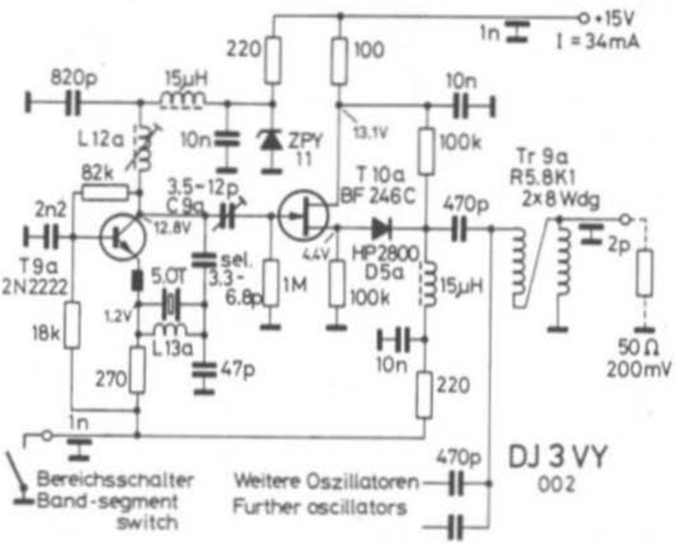

Fig. 2: Circuit diagram of one of the oscillator circuits

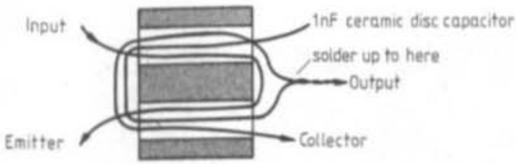

Fig. 3: Winding diagram for the two-hole transformers

Tr 1, Tr 2, Tr 7, Tr 8: Two-hole core Siemens B 62152-A 8-X 17, wound according to Fig. 3 with approx. 0.12 mm dia.enamelled copper wire
R = 1; m = 3; n = 5

Tr 5, Tr 6: Two-hole core as for Tr 1, trifilar wound with 4 turns of appr. 0.12 mm dia. enamelled copper wire (wire dia. uncritical)

Tr 3: Toroid core R 5.8 (mat. K 1) Siemens B 64290-A 0056-X 001 Trifilar wound with 9 turns of approx. 0.25 enamelled copper wire. Make full use of the core.

Tr 4: Toroid core and wire as Tr 3, bifilar wound with 13 turns.

Tr 9a - e: Toroid core and wire as Tr 3, bifilar wound with 8 turns.

L 1: made from 2 mm dia. silver-plated copper wire, bent slightly in the shape of a »W« (see photograph), dimensions as given in Fig. 1. Antenna coupling: 1 mm dia. silver-plated copper wire spaced 2 to 2.5 mm in parallel to L 1.

L 2, L 3: made from 2 mm dia. silver-plated copper wire as shown in Fig. 1 and photograph; coupling link: 1 mm dia. wire (see photograph) insulated in the intermediate panel.

L 4: 2 turns of 0.3 mm enamelled copper wire in special coil set (64 nH) (Vogt D 41-2165) with core Fi 05 f 7, but without core caps and w/out screening can.

L 5: 1.2 μH. 10 turns of 0.3 mm dia. enamelled copper wire wound in a single layer in special coil set Vogt D 41-2165, with core and cap, but without screening can. See details in (1) for other frequencies.

L 6: 4 turns of 0.5 mm dia. silver-plated copper wire wound on a 5 mm former, self-supporting. Aligned to 100 MHz by spreading one of the windings.

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

L 8, L 9: 4 turns, otherwise as L 6; coil tap on L 9: 1 turn from the cold end.

L 10, L 11: Constructed from 1.2 mm dia. silver-plated copper wire according to Fig. 1 and photograph; coupling link from 1 mm dia.(see photo), insulated in the intermediate panel.

L 12a - e: 5 turns of 0.6 mm diameter enamelled copper wire wound on a coil former of 5 mm dia. with VHF-core (Vogt Sp 3.5-2348 C with Gw 3.5/8 x 0.5 Fi 05 f 7 - D)

L 13a - e: 12 turns of 0.22 mm diameter enamelled copper wire wound on a 2.5 mm former, self-supporting.

Fixed inductances such as Delevan or Amphenol of 2.4 mm dia. with the following values:

5 x 1 μH, 2 x 15 μH, 2 x 22 μH as well as 1 x 1 μH and 2 x 15 μH per crystal in the crystal oscillator circuit.

2 ferrite beads

1 six-hole core choke (Philips)

Crystals: approx. 100 MHz, series resonance at fifth overtone, HC-43/U case (wire connections). KVG type XS 6306 are recommended.

2. CONSTRUCTION

As was mentioned previously, the described converter is accommodated on PC-board DJ 7 VY 002 which is double-coated and

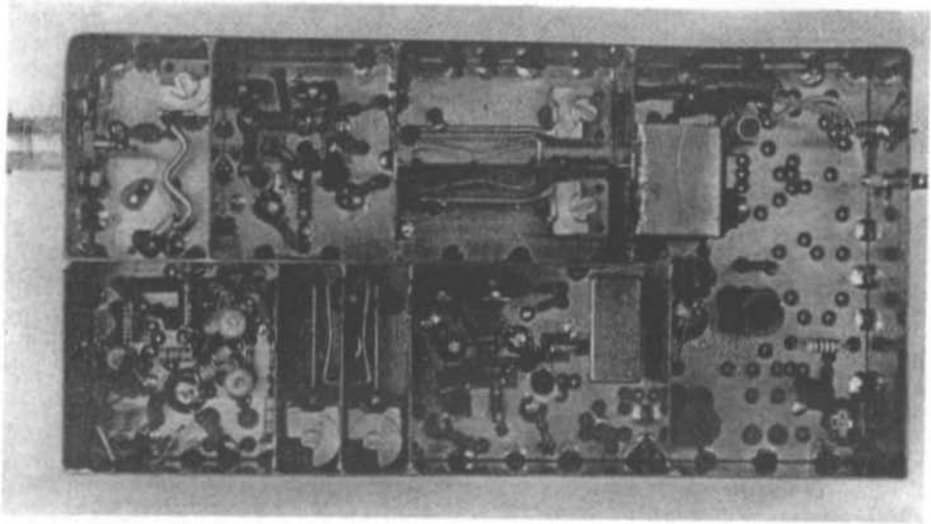

Fig. 4: Photograph of the 70 cm converter DJ 3 VY 001 accommodated on board DJ 7 VY 002

possesses through-contacts. All outer and intermediate panels are arranged in the same manner as for the 2 m converter. Two additional panels are required for the two stripline filters that can be seen in the photograph of the author's prototype given in **Figure 4**.

The input BNC connector protrudes into the chamber of the input circuit and provides a support for the coupling link. The first IF-transistor (T 3) is provided with a heat sink as in the 2 m version, whereas the two doubler transistors (P 8000) do not require such a heat sink. Transistor T 8 that provides the 50 mW for the mixer, should be placed on the ground surface of the PC-board and provided with heat-conductive paste for cooling.

When high demands are to be placed on the screening of the converter, one should provide a base plate and cover that is able to provide perfect contact to each point of the intermediate panels. For applications where these demands are not so great, it is possible to delete both covers of the intermediate bandpass filter. The author has found this to be suitable both for the 2 m version and the 70 cm version of the converter.

A double-coated PC-board with through-contacts of 155 mm x 41 mm was developed for accommodating the crystal oscillators. This board, which has been designated DJ 3 VY 002, is shown in **Fig. 5**. The dimensions are suitable for mounting it in a TEKO tin plate case. This oscillator board provides sufficient room for mounting 8 crystal oscillators so that an additional 3 can be provided for special frequencies (e.g. for OSCAR) in addition to the 5 crystals required to cover the whole of the 70 cm band. **Figure 6** shows a photograph of part of the author's prototype.

The original circuit was taken from (3) and modified according to (4). The circuit now operates in a more stable manner. The transcient drift has been reduced to approximately 1/10 of the original value using the professional bipolar transistor type 2 N 2222 as oscillator transistor instead of a FET.

3. ALIGNMENT

The ring mixer should not be installed at this point.

It is not possible to align the input impe-

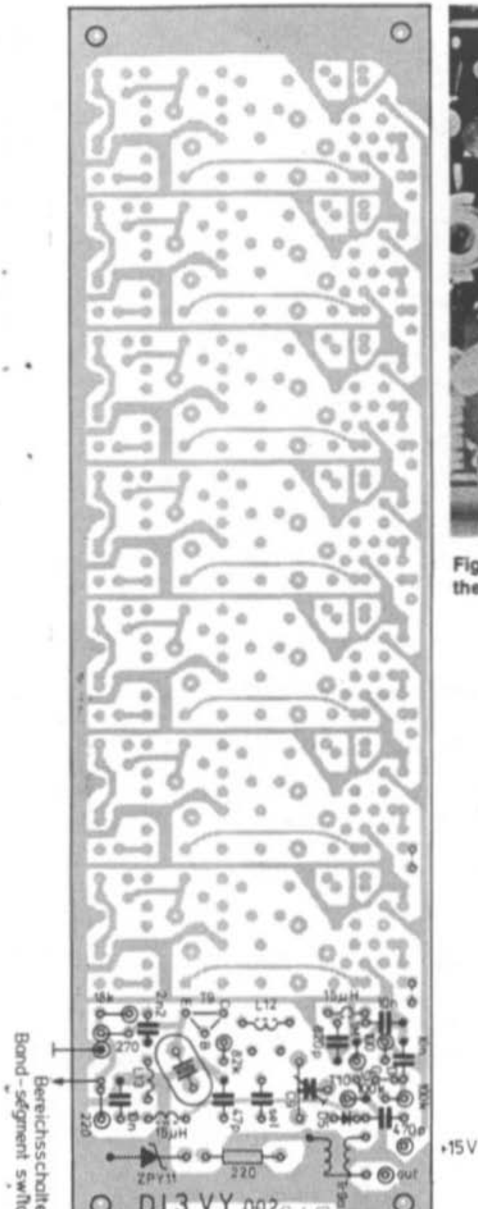

**Fig. 5:
PC-board DJ 3 VY 002 with component locations**

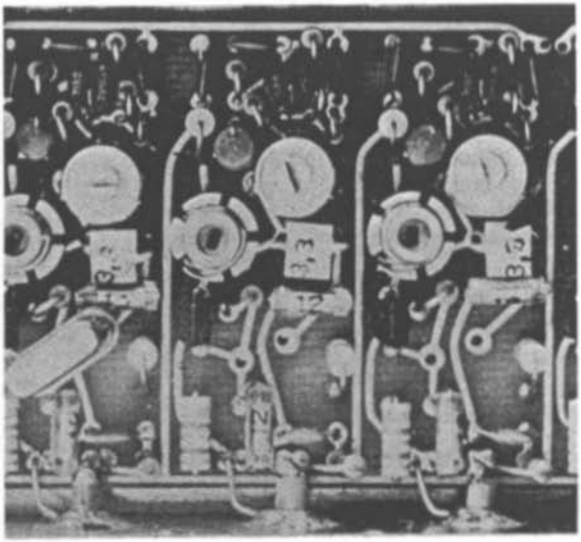

Fig. 6: Part of the author's prototype oscillator; the original version is shown.

dance and the bandwidth of the stripline filter exactly if a swept frequency generator is not available. A demodulator with 50 Ω input impedance is connected into place instead of mixer port 1 and the passband curve adjusted for a bandwidth of 10 MHz by adjusting the air-spaced trimmers and bending the coupling link. The very wide input circuit is tuned to the center of the band so that it exhibits a ripple of less than 0.5 dB over the whole band.

The bandpass filter after the frequency doublers can be aligned by connecting the swept frequency generator via a 1.5 pF capacitor. The filter is then aligned to 400 to 410 MHz if the whole 70 cm band is to be covered. If only one 2 MHz range is to be used, it is possible for the filter to be aligned for maximum output power. Both amplifier stages are so wide that it is sufficient to align for maximum using one of the crystals near the center of the band.

Since the diplexer subsequent to the mixer is exactly the same as for the 2 m version, all details regarding this were already given in (1). If this large overall gain is not required, it is possible for the second P 8000 (T 4) to be deleted. This then means that the large-signal handling demands on the subsequent receiver are reduced.

4. MEASURED VALUES

Both the 2 m converter and the described 70 cm version are used in conjunction with a HF-receiver having an input intercept point of 28 dBm. They feed into a two-stage bandpass filter with 50 Ω input and output impedance and a bandwidth of 2 MHz to a SRA-1 H. The following measured values were determined for this setup:

Sensitivity at 432 MHz (SSB):
SINAD* 10 dB: 0.1 μV
SINAD 20 dB: 0.22 μV

IP at antenna input: + 2.5 dBM
(Δf = 20 kHz; 433 MHz or 439 MHz)

* For definitions see (5)

The author would like to thank DJ 7 VY for his advice and tips, as well as DL 1 BU for the measurements.

A noise figure of 1.8 dB was measured on the author's prototype in the publisher's laboratory.

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(2) M. Martin: A New Type of Preamplifier for 145 MHz and 435 MHz Receivers
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(3) M. Martin: Rauscharmer Oszillator für ein Empfänger-Eingangsteil mit großem Dynamikbereich
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(4) B. Neubig: Design of Crystal Oscillator Circuits
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An Automatic SWR-Meter

by J. Kestler, DK 1 OF

VSWR measurements on antennas, amplifiers and matching systems are complicated and time-consuming when one has to switch the measuring instrument to »forward« power, adjust it for full scale, and then set it back to »reflected« power. The following is to describe a measuring system that allows a direct read-off of the SWR, or reflection factor with the aid of an analog multiplier, without needing to carry out any adjustment on the meter. Such a unit is very useful during alignment, since the tendency (improvement or deterioration) can be seen immediately during the adjustment process.

1. DIRECTIONAL COUPLER

The various forms of directional couplers (reflectometers) and their characteristics are not to be dealt with in detail here. In order to understand the following article, it is only necessary to know that a directional coupler (SWR-bridge) is connected between the signal source and the test object and that two voltages are provided at the two outputs whose amplitude is proportional to the forward and reflected wave. These RF-voltages are detected with the aid of diodes and fed to the indicator unit. Due to the non-linear diode characteristics, considerable errors occur during the rectification process (1).

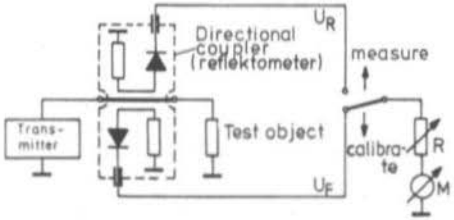

Fig. 1: Principle of a standing wave meter

Figure 1 shows the circuit diagram of a conventional SWR-meter. In the »calibrate«-position, the dropper resistor R is adjusted so that meter M indicates full scale. After switching to »measure«, the ratio of U_R/U_F can be read off. This corresponds to the reflection factor r:

$$r = \frac{U_R}{U_F}$$

For amateur radio application it is usual to give the standing wave ratio S (SWR or VSWR) instead of the reflection factor. This can be calculated as follows:

$$S = \frac{1 + r}{1 - r}$$

If the voltages U_R and U_F are fed to a divider circuit, to obtain the ratio U_R/U_F, it will not be necessary to carry out the previously mentioned calibrating process, and the measuring result can be indicated directly. The following section is to show how such a circuit can be realized with the aid of an analog multiplier.

2. THE ANALOG-MULTIPLIER AD 532

Whereas it was necessary for multiplying circuits to be constructed in hybrid technology or completely discretely several years ago, integrated circuits are now available that operate with sufficient accuracy but are relatively inexpensive. **Figure 2** shows the block diagram of such an IC. The inputs (X_1, X_2 or Y_1, Y_2) are followed by differential amplifiers which feed the actual multiplier »X«. The output voltage U_{out} is coupled out via a further amplifier (G = – 1). The operation of the circuit can be described with the aid of the following equation:

$$U_{out} = \frac{(X_1 - X_2) \times (Y_1 - Y_2)}{10\ G}$$

The IC is connected as divider in **Figure 3**. The transfer function is as follows:

$$U_{out} = -10\,G \times \frac{U_Z}{U_X}$$

It should be noted that U_X is negative with respect to ground and must be greater than approximately 0.5 V.

3. CIRCUIT OF THE SWR-METER

The complete circuit diagram of the indicator unit is given in **Figure 4**. The voltages from the directional coupler U_R and U_F (both positive) are fed via Pt 1 or Pt 2 to the two preamplifiers I 1 and I 2 whose gain is automatically switched according to the amplitude of the input signal (x 1 or x 10).

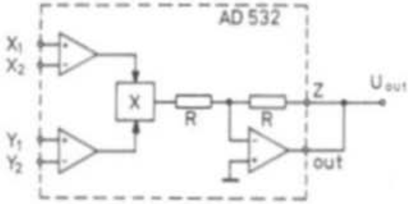

Fig. 2: IC AD 532 as multiplier

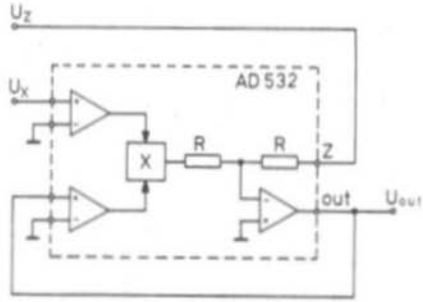

Fig. 3: IC AD 532 as divider

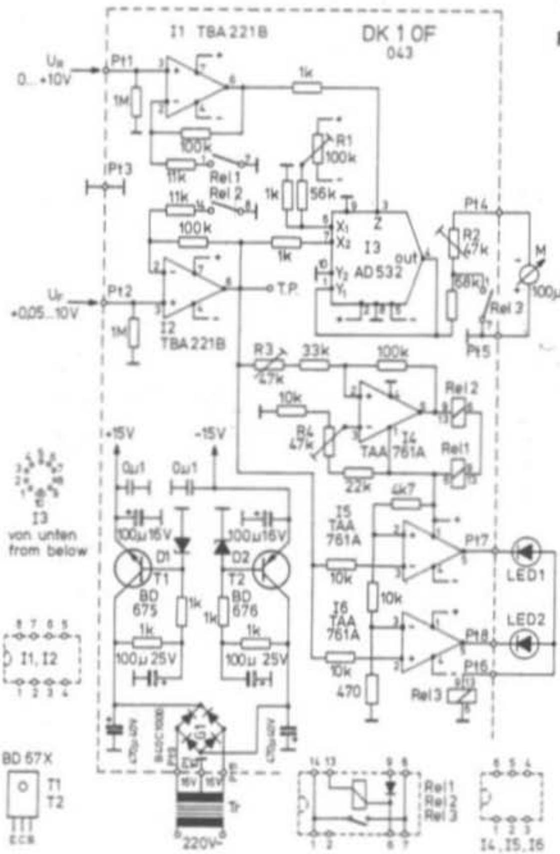

Fig. 4:
The automatic SWR-meter

The switching is made with the aid of the comparator I 4 and relay Rel 1/Rel 2 and has the task of keeping the input level of divider I 3 within a favorable range. The switching points of the comparator are adjusted with the aid of R 3 and R 4; R 1 allows the offset-alignment, whereas R 2 determines the full-scale deflection of the meter.

Integrated circuit I 5 and I 6 are two further comparators; too low an input voltage (U_F < 50 mV) is indicated by LED 2. If U_F is too high (> 10 V), LED 1 will light up. Relay Rel 3 is provided to switch off meter M if the voltage is too low or too high, since the output voltage of I 3 would then be non-defined.

The operating voltages of the amplifiers (± 15 V) are provided by the power supply shown in the lower left-hand corner of Figure 4. This circuit is very simple and need not be discussed here.

4. CONSTRUCTION

The described circuit can be accommodated on PC-board DK 1 OF 043. This single-coated board is 90 mm x 77.5 mm. The component locations are given in **Figure 5**, and **Figure 6** shows a photograph of the author's prototype.

Special Components

I 1, I 2:	TBA 221 B, 741 CN (8-pins) DIL case)
I 3:	AD 532 JH or AD 532 KH (TO-100 case)
I 4, I 5, I 6:	TAA 761 A, TAA 765 A (6-pin DIL case)
T 1:	BD 675 or similar NPN Darlington transistor (SOT-32)
T 2:	BD 676 or similar PNP Darlington transistor (SOT-32)

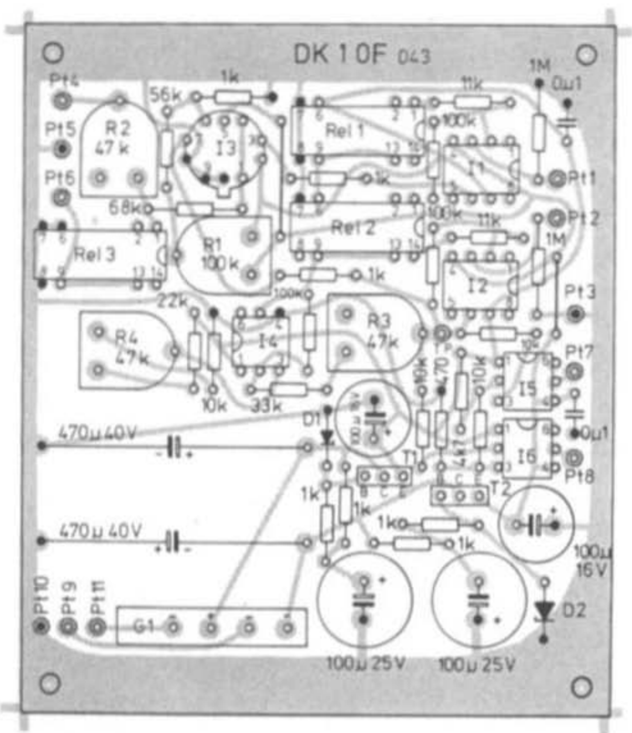

Fig. 5:
PC-board DK 1 OF 043

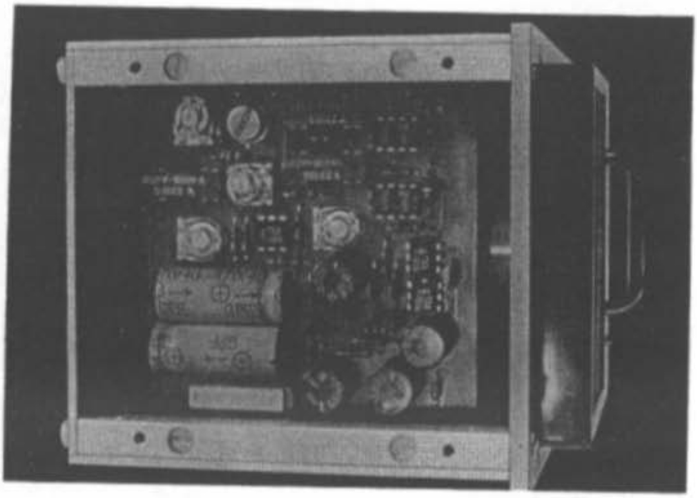

Fig. 6: Photograph of the author's prototype

D 1, D 2:	BZX 97 / C 16 or other 16 V zener diodes
G 1:	B 40 C 1000/1500 Bridge rectifier
Rel 1 - Rel 3:	Reed-relay in DIL-case, e.g. Siemens V23100-V4005-A010
M :	Moving coil meter 100 μA, fsd

Trimmer resistors: horizontal mounting, spacing 10/5 mm

5. ALIGNMENT

After checking the operating voltages, a voltmeter is connected between the test point (T.P.) and ground (range approx. 10 V). Pt 2 is then provided with an adjustable bias voltage (potentiometer approx. 10 to 100 kΩ between + 15 V and ground). R 3 and R 4 are now aligned so that relay Rel 1 and Rel 2 are deenergized at a voltage at TP of + 10 V and are energized at a voltage of + 0.9 V at TP (at this moment, U_{TP} will jump to + 9 V). In this case R 4 will determine the value and R 3 the spacing (hysteris) of the switching points.

This is followed by connecting Pt 1 to ground (Pt 3) and connecting a voltage to Pt 2 that results in a voltage of approximately 1 V at TP. R 1 is now adjusted so that »0« is indicated on the meter. This must be the case for all voltages at TP between approximately 0.5 V and 10 V; any deviations should be less than 1 to 2 % of full scale.

Finally, Pt 1 and Pt 2 are connected to the variable bias voltage which is selected so that + 1 V is present at TP. It is now possible for R 2 to be aligned for full scale on the meter. The indication should also be independent of the absolute value of the voltage at TP ($0.5\ V \leq U_{TP} \leq 10\ V$).

Figure 7 gives a nomogram for calculating the reflection factor (as indicated), and standing wave ratio.

6. REFERENCES

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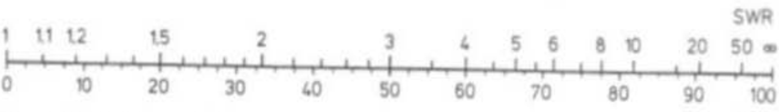


Fig. 7: Monogram of calibration

A Measuring System for Determining the Temperature Response of Crystals

by M. Arnoldt

It is very necessary to know the temperature response of the resonant frequency of crystal oscillators used for precision applications such as clock generators in clocks, time bases in frequency counters, etc. These can be in the form of ambient temperature oscillators (ATXO), temperature compensated oscillators (TCXO), or crystal-oven oscillators (OCXO).

However, the temperature curve of the crystals available to radio amateurs is often unknown. This article is to describe a measuring system for determining the temperature range up to approximately 100°C.

CRYSTAL CUT

The manufacturer can select several different crystal cuts according to the frequency range in order to obtain optimum behaviour. By fine variations within a crystal cut class, it is also possible to match th crystal to the required temperature range of operation.

Table 1 contains the preferred crystal cuts for various frequency ranges. The manufacturing problems are not to be discussed here, but **Figure 1** shows the influence the crystal cut has on the temperature response.

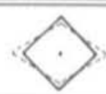
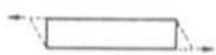
Mode of Oscillation	Frequency Range	Cut	
Flexure vibrator	8 kHz to 85 kHz 20 kHz to 100 kHz	XY' NT	
Extensional vibrator	55 kHz to 200 kHz	X	
Face-shear vibrator	180 kHz to 350 kHz 300 kHz to 1000 kHz 400 kHz to 800 kHz	DT CT SL	
Thickness-shear vibrator Fundamental Third to ninth overtone	 1 MHz to 35 MHz 2.5 MHz to 250 MHz	 AT AT	

Table 1

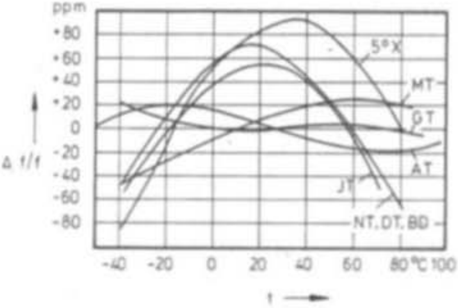


Fig. 1: Effect of the crystal cut on the temperature behaviour

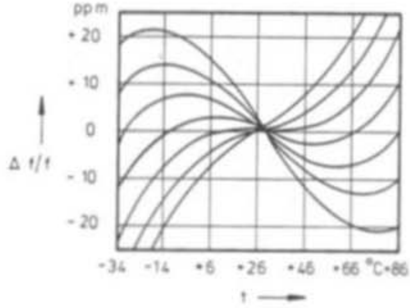


Fig. 2: Temperature response of AT-crystals

The most popular crystal cut in the MHz-range is type AT. **Figure 2** shows which variations result even with small differences of the cutting angle. The parameter-cutting angle changes from line to line only by approximately two angular minutes.

Especially two consequences can be directly read off from the family of curves given in Figure 2:

1. A relatively narrow and virtually linear temperature run can be achieved when the so-called zero angle is selected. This angle amounts to approximately 35° 13' at a crystal disk diamater of 10 mm. For larger ambient temperature ranges, one will usually select an angle that leads to inversion points that are further apart.
2. Crystals that are to be used in a crystal oven should be cut so that the inversion point coincides with the temperature of the crystal oven. In this case, the effect of the residual temperature fluctuations, which cannot be completely avoided even in the crystal oven, remain at a minimum. It should be noted that the crystal oven temperature need not be higher than the maximum ambient temperature to be expected.

PRINCIPLE OF THE MEASURING CIRCUIT

The following considerations are valid for the construction of the measuring system:

1. The crystal whose temperature response is to be measured must oscillate in the oscillator circuit. This circuit should be variable and be stable enough with respect to fluctuations of the ambient temperature, operating voltage and load impedance.
2. The crystal should be mounted in a crystal oven that allows the temperature to be varied slowly in the required range.
3. The circuit should have a temperature measuring and indicating system that is able to indicate the actual temperature of the crystal as accurately as possible.

The third demand represents the greatest problem, since it is not usually possible to measure the temperature of the crystal directly on the disk.

For this reason, a heating system is used that allows two identical crystal cases (e.g. HC-6/U) to be accommodated and be heated identically. One of the two cases contains the crystal, and the other a temperature probe, a so-called phantom crystal.

CRYSTAL OVEN AS HEATING SYSTEM

The heating system used is a crystal oven type XT-2 whose temperature is controlled using a bi-metal strip and a heating winding (two-point control). The heater winding goes around both crystals, which are pressed against each other with the aid of

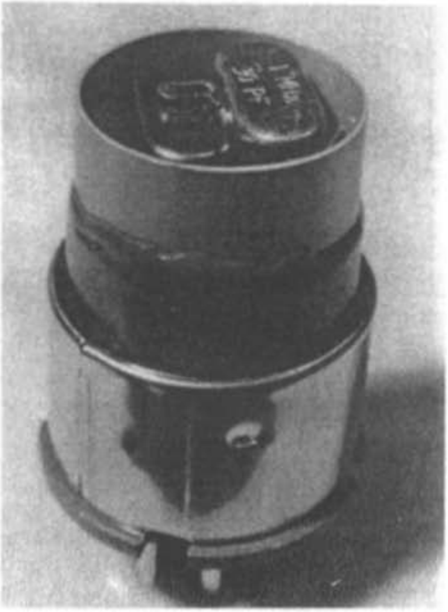


Fig. 3: Photograph of the crystal oven XT-2

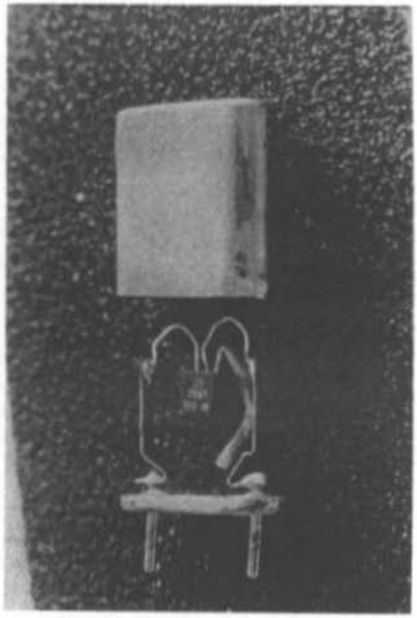


Fig. 4: Photograph of the phantom crystal using temperature probe LM 335 Z

springs. By bridging the 75°C bi-metal switch (see **Figure 14**), it is possible for the measuring range to be increased in excess of + 75°C. **Figure 3** shows a photograph of the crystal oven with the crystal and phantom crystal mounted into place.

The great advantage of this arrangement over temperature measurements on the case is evident: Temperature fluctuations take longer to reach the actual crystal disk than they do the case. For instance, if a temperature probe indicated a stabilization of the case temperature, this would still not be valid for the actual crystal disk itself.

CONSTRUCTION OF THE PHANTOM CRYSTAL

The case should be as similar as possible to that of the crystal to be measured. The temperature probe itself should be mounted in the same position within the case as the crystal disk. The connection wires of the temperature probe are satisfactory for mounting.

The author's prototype used a temperature probe type LM 335. Both the plastic and the metal version have been found successful in practice. These probes provide a temperature-linear voltage with a variation of 10 mV/K (10 mV/°C). It is not absolutely necessary for a calibration to be made.

There are 3 possibilities for construction:

1. A defective crystal is opened by unsoldering the case, removing the crystal and mounting the temperature probe in the crystal holder, whereby one holder is connected to the output of the temperature probe, and the other to the ground contact. The photograph given in **Fig. 4** shows the construction of the phantom crystal using a plastic temperature probe (TO-92).
2. If no defective or obsolete crystal is available in the right type of case, such a case comprising the socket and case itself is usually available from crystal manufacturers.

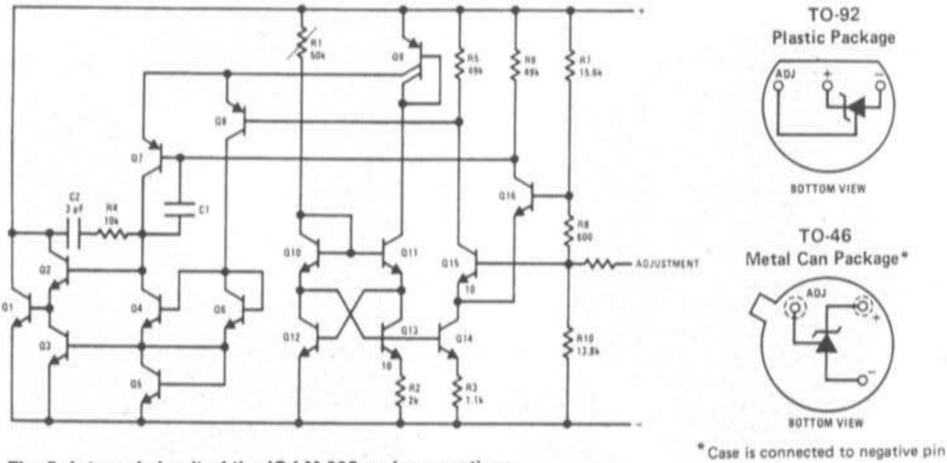


Fig. 5: Internal circuit of the IC LM 335 and connections

3. In addition to this, a complete phantom crystal is available for approximately DM 10,— under the designation QT-T 1025. This phantom crystal has the advantage over the previously mentioned version that it is enclosed in the same nitrogen gas as are most modern crystals. This gas atmosphere exhibits a low dew point of – 55°C, which ensures that no humidity results on the crystal resonator or temperature probe down to this temperature.

THE TEMPERATURE MEASURING CIRCUIT

Since the temperature probe LM 335 provides a temperature-linear output voltage, no linearization is required. The typical linearity error amounts to only 0.3 K. The typical indication error without alignment is given as 4 K for the inexpensive version LM 335.

The internal circuit and the external connections of the three-pole temperature probe are given in **Figure 5**. The function of the temperature probe corresponds to that of a zener diode. It operates in a current range of 0.4 to 5 mA at a maximum indication error of 0.3 K.

The actual measuring circuit is given in **Figure 6**. The temperature probe is fed with

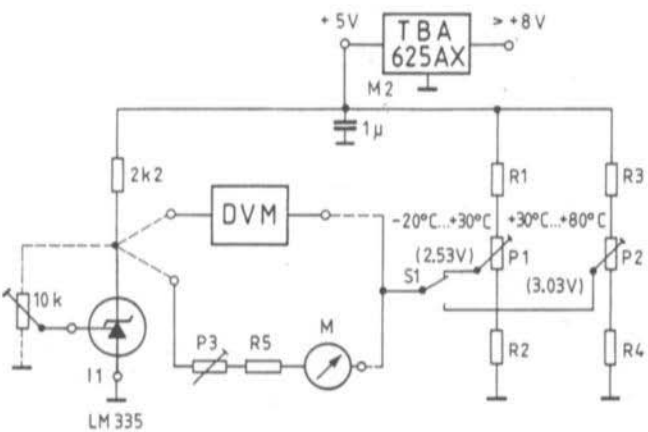


Fig. 6: Temperature measuring circuit

Indicator	U_{P1}/V	U_{P2}/V	R_1/kΩ	R_2/kΩ	P_1/kΩ	R_3/kΩ	R_4/kΩ	P_2/kΩ
DVM	2.73	–	2.3	2.7	0.1	–	–	–
Voltmeter								
– 20°C to + 30°C	2.53	–	2.5	2.5	0.1	–	–	–
+ 30°C to + 80°C	–	3.03	–	–	–	2.0	3.0	0.1

Table 2

a current of approx. 1 mA via a resistor from a 5 V voltage stabilizer. The output voltage amounts to 273 x 10 mV at 0°C and increases per degree of temperature by 10 mV. The most favorable means of indicating the temperature is the use of a digital voltmeter (DVM) having a resolution of 1 mV ≙ 0.1 K. In order to avoid that the zero value of the indicator (theoretically) is at the absolute zero point (– 273.15°C), and a value of 293.0 will be indicated for + 20°C, the minus input of the DVM should be connected to a potentiometer that is aligned to 2.73 V (≙ 0°C). It is then possible to read off the temperature in °C with a resolution of 0.1 K.

In order to ensure that the circuit is not only limited to use with a DVM, the PC-board is designed so that a meter can also be connected. Since the resolution of normal meters on the market is not usually more than 50 scale markings, whereas the temperature measuring range of interest can amount to 100 K (e.g. – 20°C to +80° C), the measuring range can be switched with the aid of S 1 to a lower and upper range.

If the zero point of the lower measuring range is set to – 20°C, the associated voltage divider P 1 should be adjusted to 2.53 V. The zero point of the upper measuring range then corresponds to + 30°C ≙ 3.03 V.

Table 2 contains the required resistance values. It is easily possible to change the measuring ranges. Resistors R 5 and P 3 are in series with the meter M and can be aligned so that the full scale deflection amounts to 500 mV (≙ 50 K).

Since the maximum measuring error of up to 4 K is no longer permissible during measurements on crystals, it is recommended that this error be compensated for. The actual response of the output voltage with respect to the ideal characteristic of exactly 10 mV/K can have a linearity error of typically 0.3 K, and additionally a constant shift of typically 4 K ≙ 40 mV. This shift can be compensated for by connecting the adjustment connection of the temperature probe LM 335 to a potentiometer of e.g. 10 kΩ, as shown as a dashed line in Fig. 6, or by varying the opposite voltage that is

					Temperature error/K, uncalibrated			
	Temperature range		Linear.error/K		at 25°C		Total temp. range	
Type	T_{min}/°C	T_{max}/°C	typ.	max.	typ.	max.	typ.	max.
LM 135 A	} – 55	+ 150	0,3	0,5	0,5	1	1,3	2,7
LM 135			0,3	1	1	3	2	5
LM 235 A	} – 25	+ 100	0,3	0,5	0,5	1	1,3	2,7
LM 235			0,3	1	1	3	2	5
LM 335 A	} 0	+ 100	0,3	1,5	1	3	2	5
LM 335			0,3	1,5	2	6	4	9

Table 3

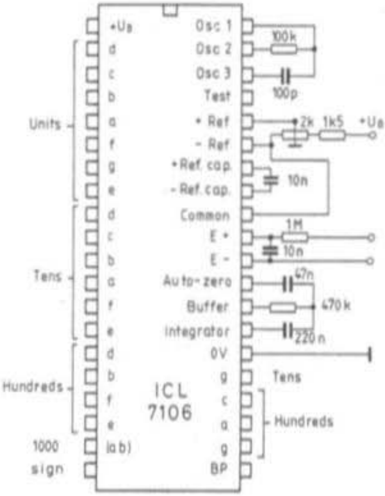


Fig. 7: Digital voltmeter: Connections of the IC ICL 7106

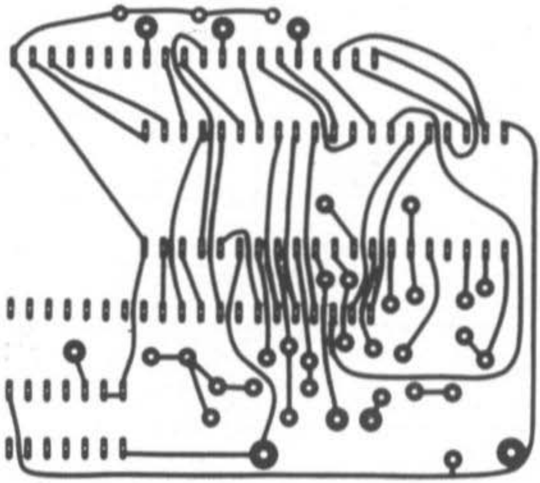


Fig. 8: Conductor lanes of the DVM

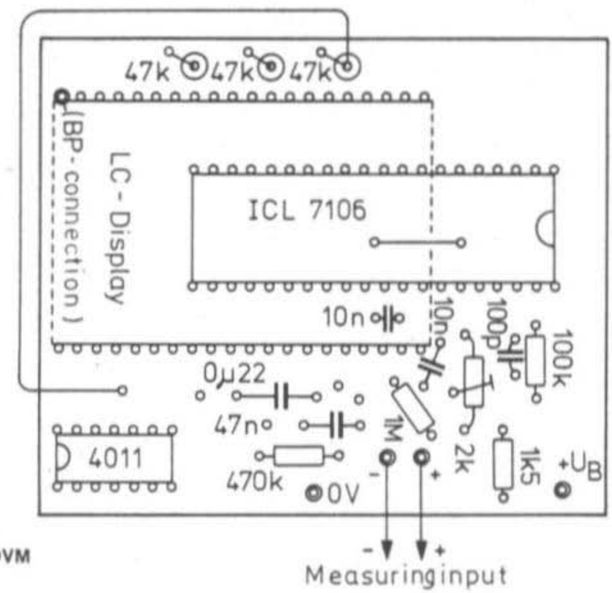


Fig. 9: Component locations of the DVM

present on P 1 (P 2). Both methods are equally good. Since the crystal case only possesses two connections, in contrast to the three connections of the temperature probe, the adjustment connection is not connected, and for this reason the alignment is made with the aid of P 1 (P 2).

If, on the other hand, the crystal case itself is used as ground connection, it should be ensured that the voltage drop caused by the heating current on the ground connection to the heating winding, which is electrically connected to the intermediate panel between the cases, will not cause a measuring error.

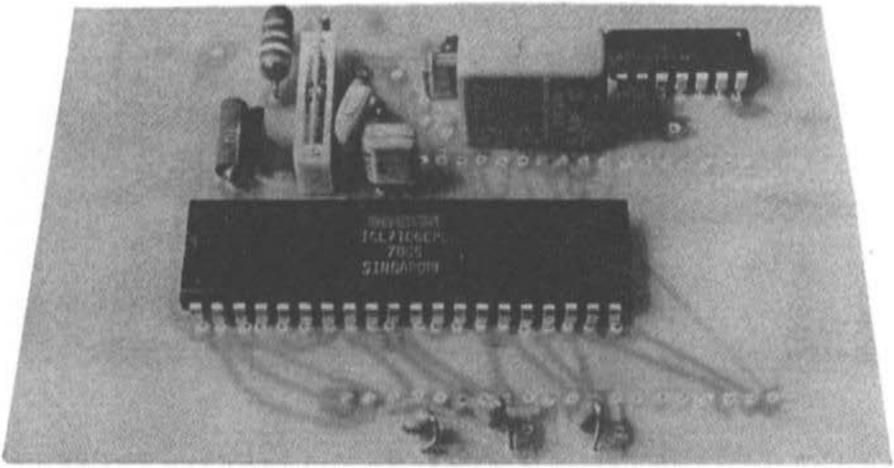


Fig. 10: Photograph of the DVM

For applications where higher measuring accuracy is required, it is possible for temperature probes to be used that possess a lower tolerance. **Table 3** lists some characteristics of these types.

INDICATION USING A DVM

Basically, any DVM is suitable for indicating the temperature having a minimum resolution of 10 mV (better will be 1 mV). The author's prototype uses a DVM equipped with the integrated circuit ICL 7106 and a LCD readout.

The operation of this digital voltmeter is not to be described in detail here, but only the most important details. The external circuitry has been selected so that a measuring range of ± 2000 mV results. A spindle trimmer is used for alignment. The zero input of the measuring system is not identical to the ground connection of the DVM.

Figure 7 shows the external circuit of the IC. The LCD-readout is mounted on the conductor side of the board. This means that the module is flat and allows suitable dimensions of the board. **Figure 8** shows the conductor side of the board, **Figure 9** the component locations, and **Figure 10** a photograph of the completed module.

The indication of the zero point requires a zero point segment voltage that runs inversely proportional to the backplane voltage. It is generated in an inverter 4011.

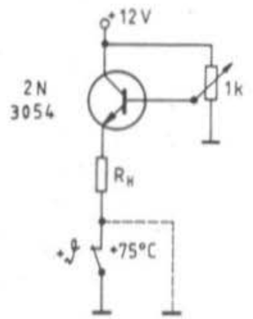


Fig. 11: Heater circuit

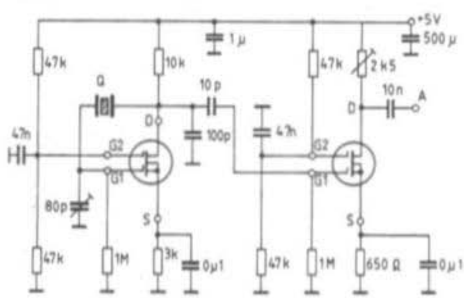


Fig. 12: Low-reactive oscillator circuit

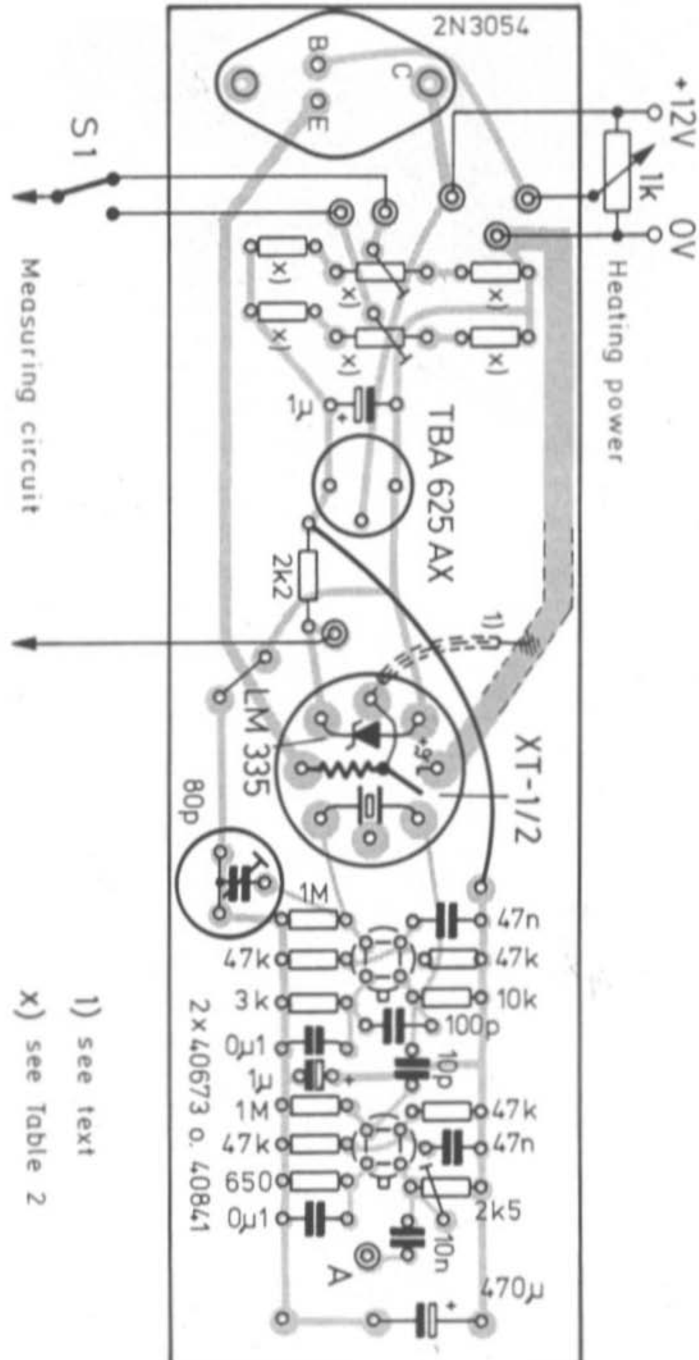


Fig. 13: Component locations on the main board

HEATER CIRCUIT (Figure 11)

The heating system of the two-point crystal oven XT-2 exhibits a resistance of approx. 6 Ω in the case of the author's prototype. If the temperature switch is not to be used, it is possible for heating power and temperature to be selected freely, varying the heating current up to approximately 1 A. A 1 kΩ potentiometer is used for adjusting the heating current. A transistor 2 N 3054 is used as current amplifier. The heating winding is connected to the emitter. The heating current flow follows the potentiometer voltage relatively linearly. The heater voltage amounts to approximately 12 V. It is necessary for the power transformer to be suitably selected. If – according to the selected heating power – large variations of the operating voltage take place, it is recommended that the operating voltage for the DVM is stabilized, for instance to 8 V using an IC 7808, or similar.

THE OSCILLATOR CIRCUIT

The measurement of the temperature characteristic is best made in conjunction with an oscillator circuit. Besides providing an alignment possibility, the crystal oscillator should be able to provide a sufficiently high voltage to a frequency counter at low reactance, and that the circuit should allow the crystals to commence and maintain oscillation over a wide frequency range. The circuit given in **Figure 12** fulfills this conditions. The RMS value of the output voltage amounts to 1.2 V at 1 MHz, and 0.15 V at 10 MHz. The reaction is decreased sufficiently by using a dual gate MOSFET 40673 or 40841 (C_r = 0.02 pF) as a buffer.

COMPONENTS

Crystal oven: Type XT-2 available from the publishers

LM 335: National Semiconductor

Phantom crystal: Quarztechnik

2 MOSFETs 40673 or 40841: RCA

ICL 7106: Intersil

4011: RCA

TBA 625 A (X), L 129, 7805, or similar

2 N 3054: RCA

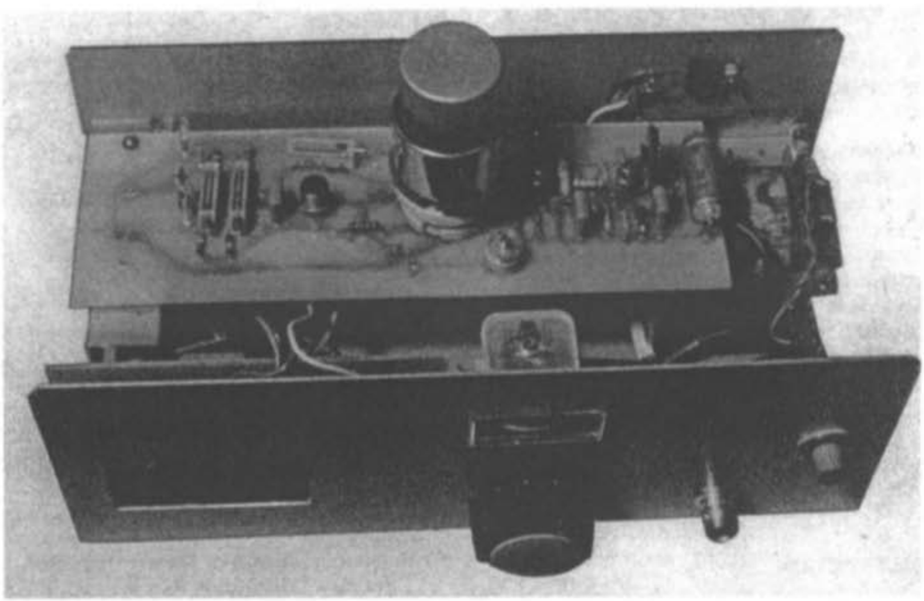


Fig. 14: Photograph of the main board of the completed module

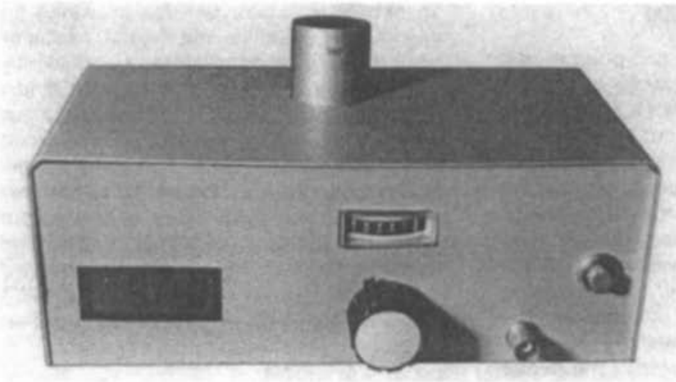


Fig. 15:
Photograph of the author's prototype

ALIGNMENT

The alignment process is mainly limited to the temperature measuring system. It should be commenced at the DVM or meter. This is followed by selecting the required zero point of the measuring range with the aid of P 1 (P 2). When using the DVM, P 2 will not be required. P 1 is then adjusted to 2.73 V. The fine alignment is then made by adjusting P 1 (P 2) at a certain temperature value that has been set with another, sufficiently accurate means (e.g. clinical thermometer between 36 and 42°C).

The measuring system will not provide satisfactory results if the reference frequency of the frequency counter used is inaccurate and/or unstable. It is therefore recommended to either use a commercially available frequency counter whose error is always less than 10^{-7}, or for the crystal oscillator of the time base used to be locked to a standard frequency transmitter such as Droitwich 200 kHz.

Of course, the described measuring unit can form the basis for constructing RCXOs, TCXOs or OCXOs, e.g. for frequency counters.

CONSTRUCTION

The main board (see **Figure 13**) accommodates the oscillator circuit, the eight-pin octal socket of the crystal oven XT-2 and part of the measuring circuit, whereas the DVM is constructed and mounted in a different position. **Figure 14** shows a photograph of the completed module without cover, and the completed unit with cover is shown in **Figure 15**.

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(5) B. Neubig: Design of Crystal Oscillator Circuits VHF COMMUNICATIONS 11, Edition 3/79, pages 174-190

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A System for Reception and Display of METEOSAT Images

Part 5

by R. Tellert, DC 3 NT

The last part of this article described the power supply and frequency generator module. This is now to be followed by the description of the video board DC 3 NT 005.

Before going into details, it should be mentioned that the equipment described in this series of articles has been improved as they go along. For instance, a control bus and a system board have been added. This not only simplifies the wiring, but also simplifies operation of the system considerably. These boards will be described in part 6 of the series, together with the start-stop evaluation (007) and a circuit for the drum motor (DC 3 NT 010).

We would also like to point out that the system is also suitable for recording the APT-transmissions from the three GOES satellites located over North America, as well as from the Japanese GMS satellite. However, the latter uses a far higher frequency deviation which requires a modified IF-module in the VHF-receiver.

All important functions can be controlled via the 12 Bit bus. The switching is made using C-MOS analog switches which means, for instance, that the expensive 4-wafer switches will not be required for the seven speeds. It is now possible using a simple single-wafer switch and a diode matrix to »program« the unit for all facsimile standards from VLF to HF, and for the various weather satellites. The extension to include the other weather satellite and weather map sources was made necessary due to the failure of METEOSAT 1, and after it was found that too many possibilities of incorrect operation were possible when switching the system to one or another »standard« when not using the control bus.

The control bits are as follows:

3 bit for the 7 speeds
2 bit for the AF-source
Tape/VLF/HF/VHF

2 bit for video switching
AM/FM (± 150 Hz)/FM (± 400 Hz)

2 bit for the normal
TIROS-NOAA-CH 1/CH 2

1 bit for the index of cooperation
(monitor 288/576)

1 bit for normal/weather map

1 bit for the METEOR burst 250/1200 Hz

The system board provides the interconnection between the following modules:

Power supply DC 3 NT 006

Start-stop logic DC 3 NT 007

Frequency generator DC 3 NT 008

Monitor interface X-Y-Z DC 3 NT 009

New MOS-switching module DC 3 NT 011

In addition to this, several switches of the data bus (2 IC's) are also accommodated.

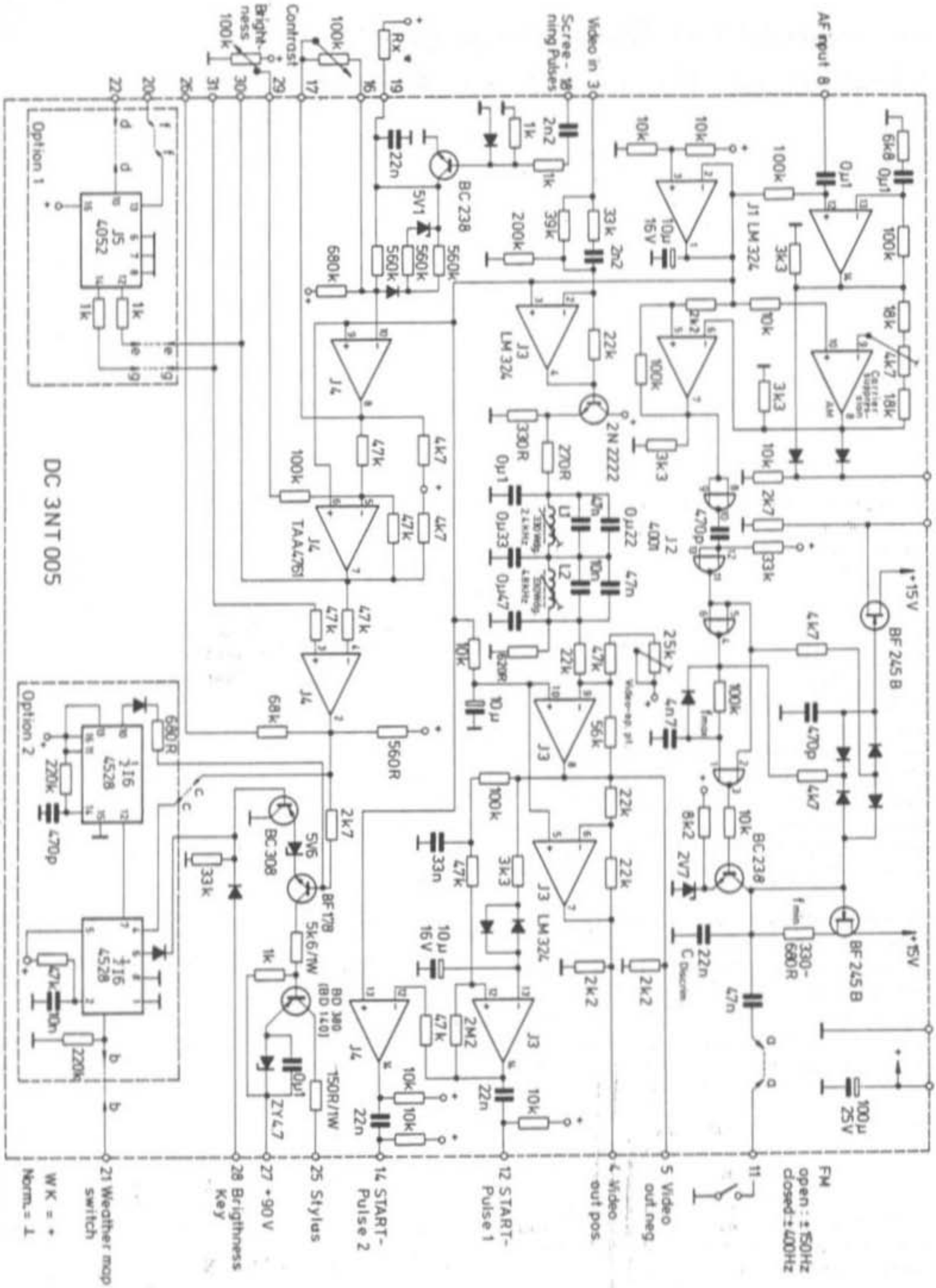

Fig. 37: Circuit diagram of the video board DC 3 NT 005

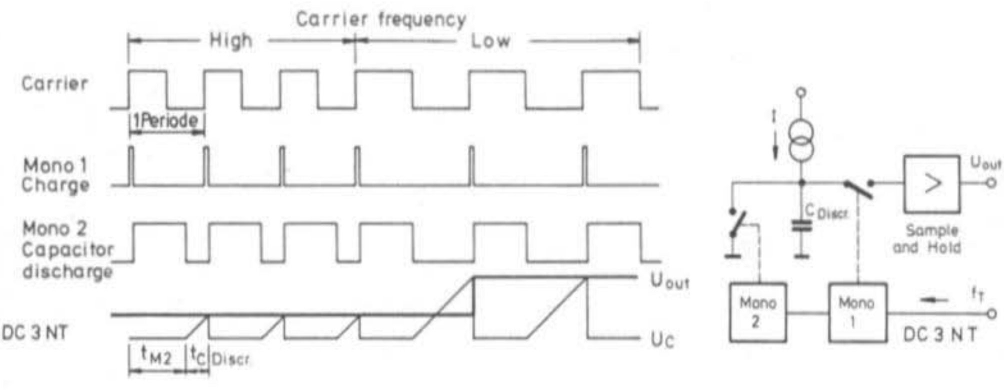

Fig. 38: Basic circuit and pulse diagram of a special FM-demodulator

5.4. DC 3 NT 005

As was mentioned in the block diagram (Figure 26 of part 3), module DC 3 NT 005 accommodates the following circuits:

a) for CRT and FAX-machine:
   2400 Hz preamplifier
   AM-demodulator
   FM-demodulator
   Video filter
   Video amplifier (pos. and neg.)
   Start pulse processing

b) Only required for driving the stylus of the FAX-machine:
   Pulse-width modulator
   Brightness control
   Contrast control
   Two-channel switch for monitoring with an oscilloscope
   Pulse-width correction for weather maps
   Stylus current drive

The special features of the individual stages are now to be discussed with the aid of the circuit diagram of module DC 3 NT 005 which is given in **Figure 37**.

5.4.1. Demodulators

The low frequency spacing between the subcarrier of 2400 Hz and the video frequency of max. 1700 Hz place high demands on the demodulator. Only such circuits should be used that suppress the carrier. A full-wave detection is used for the AM-demodulations (The two upper amplifiers of I 1 together with the two subsequent diodes). The carrier will be suppressed in the case of an exact alignment of the anti-phase amplitudes using the trimmer potentiometer. The strong component of 2 f_{carr} (4800 Hz) generated in this manner can be easily suppressed in a subsequent filter.

The selection of a circuit for the FM-demodulator is more difficult since the carrier frequency varies. Known demodulators will not operate successfully in this application, which meant that a new circuit had to be developed that had a very high carrier suppression and a high video cut-off frequency. In addition to this, it is even possible to increase the slope of the frequency-to-voltage characteristic.

Figure 38 shows the basic circuit together with the corresponding pulse diagram. A Schmitt-trigger (not-shown) shapes the sinewave carrier frequency voltage to a square-wave voltage. The time spacing of the rise slopes corresponds to the period duration of the carrier frequency (1/Hz). This time is measured. For this. a capacitor C is charged with a constant current and periodically discharged with the previously mentioned rise-slopes of the square-wave voltage. The achieved voltage U_C is stored just before its discharge. The stored voltage is a measure of the previous period duration and corresponds to the inverse value of the carrier frequency.

In the case of a constant carrier frequency, a DC-voltage will be present at the output. In the case of a continuous frequency variation, the output voltage U_{out} of this demodulator will be provided in steps; for instance, at a carrier frequency of 3 kHz, 3000 steps per second will result. This low residual AC will only occur during changes of the modulation and will only have odd harmonics according to the Fourier analysis

$$(f_{carr} + \frac{1}{3} 3 f_{carr} + \frac{1}{5} 5 f_{carr} + ...)$$

These can be easily suppressed in a subsequent filter.

The basic discriminator slope is shown in **Figure 39**. In this case, f_{max} is determined by the pulse duration of Monoflop 2 (R = 100 kΩ/C = 4.7 nF between the two gates of I 2), whereas f_{min} is not only determined by this pulse duration, but also by the maximum charge time of the discriminator capacitor.

Fig. 39: Discriminator characteristics of the circuit given in Fig. 38

The greatest possible frequency deviation $f_{max} - f_{min}$ can therefore be influenced by the selection of the capacitance value and the amount of charge current. This is utilized in order to obtain two different characteristics of optimum slope for the two frequency deviation values required. When connecting a subcarrier of 2400 Hz to Pt 8 of DC 3 NT 005, the output voltage at connection Pt 10 should amount ± 1.5 V at a frequency deviation of ± 150 Hz. If this value is not obtained, it is necessary to change the source resistor of the constant-current FET BF 245 B; a greater resistor value results in a lower output voltage. This is the most favorable operating range for VLF facsimile transmissions.

For short-wave transmissions with ± 400 Hz shift, connection Pt 14 is grounded thus increasing the value of the discriminator capacitor by 47 nF. As can be seen in Figure 39, the center frequency will be shifted slightly down, which has no importance for our application.

5.4.2. Video Module

The video signal is fed from the outputs of the demodulators (Pt 9 and Pt 10), via switches, and via Pt 3 to the video amplifier. Any residual carrier or harmonics thereof are now suppressed and the DC-voltage levels are brought to the following uniform values:

Signal	Video in Pt 3	Video out Pt 4	Video out Pt 5
Black	7.0 V	9.5 V	5.5 V
Grey	8.8 V	7.5 V	7.5 V
White	10.5 V	5.5 V	9.5 V

The frequency response should be as linear as possible from 0 to 1.7 kHz. To obtain this, a video filter was optimized that possesses two poles: at 2.4 kHz and at 4.8 kHz. This greatly suppresses the main components of the residual AM-carrier.

The video amplifier has one output each for a positive and a negative video signal (Pt 4 and Pt 5). In addition to this, a circuit is provided for determining a black/white jump + an inverter provides an impulse in the case of a white/black jump. These impulses are used after receiving the starting tone to start the machine, or the electron beam, at the correct phase position. These outputs are Pt 12 and Pt 14.

5.4.3. Stylus Drive

The following circuits are only required for the FAX-machine.

The metalized paper used (BOSCH) can basically only provide white (silver) or black. If one attempts to obtain a grey tone by varying the stylus current, this will lead to a non-uniform coverage, which virtually only shows the coarseness of the paper. For this reason, it was decided to screen the image in a similar manner to that used in newspapers by obtaining a grey tone using a series of small white and black image points. Since our images are written in lines, it is possible for the grey tone to be obtained by varying the length-ratio of black to white image points. The best results were obtained by screening in synchronous with the drum speed. **Fig. 40** shows the basic circuit of a (synchronous) pulse-width modulation.

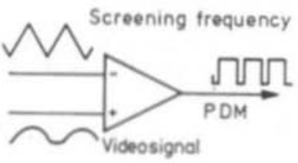

Fig. 40: Basic circuit of a synchronous pulse-width modulation

It is possible using a comparator and a sawtooth or triangular voltage to chop the video input voltage in time with the screening frequency and to vary it according to the keying ratio. The following possibilities are provided:

1. A shift of the DC-voltage level generates a brightness variation
2. A variation of the sawtooth amplitude results in a contrast variation.
3. The transmission characteristic can be influenced by the curve shape of the voltage at the minus input.

Point 3 is realized firstly in the associated circuit (Fig. 37 from Pt 18, 16, 17, 29 to 31 via I 4) which is given in the right-hand part of the circuit diagram:

In order to correct the inherent degradation of the dark grey steps of this method, a linear sawtooth voltage is not used, but an expotential waveform. This is obtained by charging a capacitor (22 nF) via a resistor. Needle pulses are obtained from the screening frequency present at Pt 18 by differentiating, and these pulses discharge the capacitor 720 times per drum revolution with the aid of a transistor (BC 238). The time constant of R and C should correspond to 3 τ. In order to obtain the same waveform at all speeds, it is necessary for the time constant to be switched together with the drum speed. For this reason, the charge resistor is not mounted on board DC 3 NT 005, but is combined with the speed switching.

The sawtooth voltage is now corrected in a network of diodes and resistors and controlled to obtain a variable amplitude, (Pt 16). This is the contrast control (point 2 above). At zero position of the sawtooth voltage, the contrast will be infinite, which means that no grey tones will be provided. This position is used for weather **maps** (in contrast to weather or better cloud **images**).

Finally, the DC-voltage level can be altered in the subsequent stage (Pt 29), and this is the brightness control (point 1 above).

The integrated circuit I 4 can, of course, not directly drive the stylus, since a voltage of 80 V and a current of approximately 300 to 500 mA are required to burn off the aluminium coating. This is a power of 25 to 40 W; however, this is only required for a period of µs even in the case of complete black. This drive is provided by the power transistor BD 380, which switches the stylus voltage from Pt 27 to Pt 25; a resistor of 150 Ω and an external 250 Ω potentiometer are provided for limiting the stylus current.

The stylus current can be electronically blanked, using a further transistor (BC 308). This is used to suppress current to the stylus when the FAX-machine is stationary, and can also be used for blanking out the unwanted channel in the case of TIROS and NOAA transmissions.

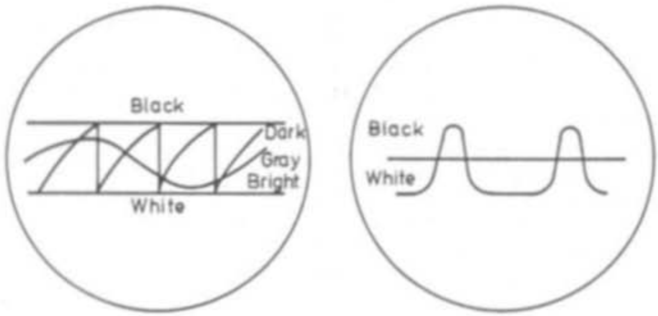

Fig. 41: Oscilloscope traces of an image with and without gray tones

5.4.4. Two Accessory Circuits

To complete the description of module DC 3 NT 005, two useful circuits are to be described that are not absolutely necessary. These are to be designated as »Option 1« and »Option 2«. Both of them are also used only for the FAX-machine.

Option 1: Two-Channel Switch

An electronic switch switches the two input signals of the comparators used for the pulse-width modulation alternately to a common output (Pt 22). This allows the adjustment of the brightness and contrast of the video signal to be carried out in conjunction with a simple single-trace oscilloscope. Switching is made using the VCO-frequency of 86.4 kHz obtained from module DC 3 NT 008. The chopper does not only save a dual trace oscilloscope, but also the exact adjustment of the two channels with respect to gain and zero line.

Figure 41 shows an oscilloscope trace for frequency of 86.4 kHz obtained from module DC 3 NT 008. The chopper does not only save a dual trace oscilloscope, but also the exact adjustment of the two channels with respect to gain and zero line.

Option 2: Weather Map Switching

Most symbols and signs are at the resolution limits when writing weather maps. The image contents consist mainly of a white surface with short black points, which results in the highest transmission rate. Since the transmission bandwidth is limited, the individual image point will be cut off. As is indicated in **Figure 42**, a transcient process is caused from the originally sharp white/black transition that has a characteristic rise and fall time dependent on the bandwidth.

If the comparator threshold is in the center, the smallest image points will not be recorded; if the threshold is lower, this will cause a lengthening of the black image points.

In addition to this, the stylus will increase the width of the black point to the value of its own width **(Figure 43)**.

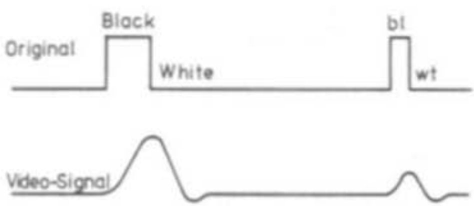

Fig. 42: A transcient process will result from a sharp white/black/white transition

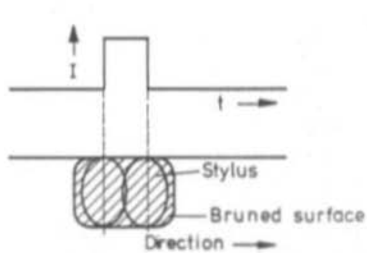

Fig. 43: A black point will be lengthened by the actual width of the stylus

This error is decreased using two Monoflops (I 6):

The commencement of the stylus current is delayed to the value of time corresponding to the stylus width on the paper. This means that an individual black image point is brought back to approximately the original length. If the image current pulse is, however, already completed before the delay time has elapsed, this would mean that this image point would be missing completely. For this reason, Monoflop 2 produces a short impulse after the switching of Monoflop 1 in order to ensure that an image point is generated in all cases. This point has practically the size of the stylus, and is thus the smallest possible image point.

Since the described correction method is only operative in the case of true black-and-white images, the contrast should be set to an infinite value in this mode (contrast control bridged). This additional circuit must be switched off when producing grey tones, so that the screened image points are not »corrected«. Connection point Pt 21 is provided for switching option 2 on and off: It is connected to + 15 V for weather maps, and grounded for cloud images.

5.4.5. Construction of DC 3 NT 005

The circuit given in Figure 37 is accommodated on a single-coated PC-board which has been designated DC 3 NT 005. This board is shown in **Figure 44**. This board is also of standard Europa-card size and is provided with a 31-pin connector that is plugged into the system board to be described later.

Since the board is single-coated, a few bridges are required that are given in the component location plan, and can be seen in the photograph of the author's prototype given in Figure 45.

Please check: There is a total of 6 bridges on DC 3 NT 005. The two »options« and their connection points 'b' to 'g' are especially marked in the component location plan. These must be connected with the aid of connection wires as can be seen in the photograph.

5.4.5.1. Components of DC 3 NT 005

14 diodes 1 N 4148 or similar

1 zener diode C 2 V 7

1 zener diode C 5 V 1

1 zener diode C 5 V 6

1 power zener diode ZY 4,7

1 switching transistor 2 N 2222

2 AF-transistors BC 238, BC 413 or similar
2 FETs BF 245 B

1 AF-PNP transistor BC 308, BC 415
or similar

1 HF-transistor BF 178 or BF 258 or similar

1 PNP power transistor BD 380, BD 140

Integrated C-MOS circuits:

1 x 4001, 1 x 4052, 1 x 4528

1 quadruple Op-Amp TAA 4761 (Siemens)

2 quadruple Op-Amp (LM) 324

2 Siemens coil sets 11 x 7
mat. N 22, with alignment core,
coil former and holder

Plastic foil capacitors for 7.5 mm spacing:

3 x 470 p, 1 x 2n2, 1 x 4n7, 2 x 10n,
4 x 22n, 1 x 33n, 3 x 47n, 4 x 0µ1
1 x 0µ22, 1 x 0µ33, 1 x 0µ47

1 ceramic or foil cap. 820 pF,
spacing 5 mm

1 ceramic cap. 2n2, spacing 5 mm

3 tantalum electrolytics (drop type):
10 µF / 25 V

1 alu-electrolytic, round, spacing 5 mm:
100 µF / 25 V

1 trimmer potentiometer, each, 4k7 and
25 kΩ, spacing 10/5 mm

1 resistor 5k6/1 W

1 resistor 150 Ω 1 W

All other resistors for 10 mm spacing

1 31-pin connector (DIN 41617)

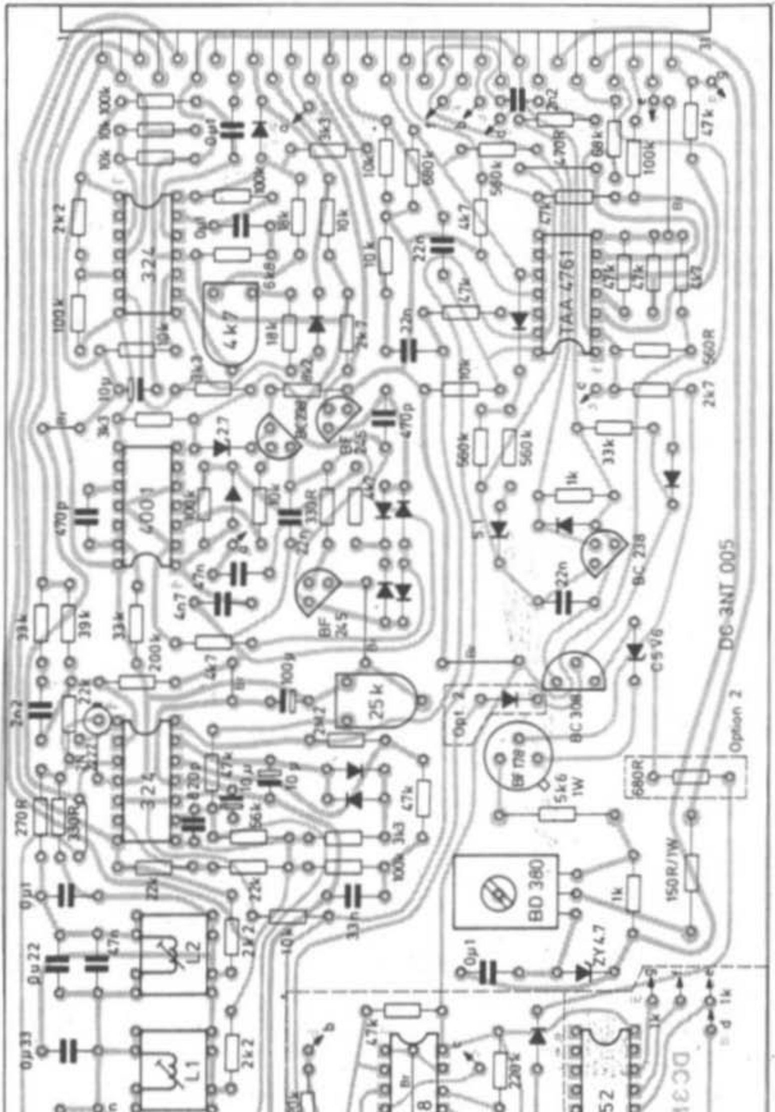

Fig. 44: Component location plan on PC-board DC 3 NT 005 (video board)

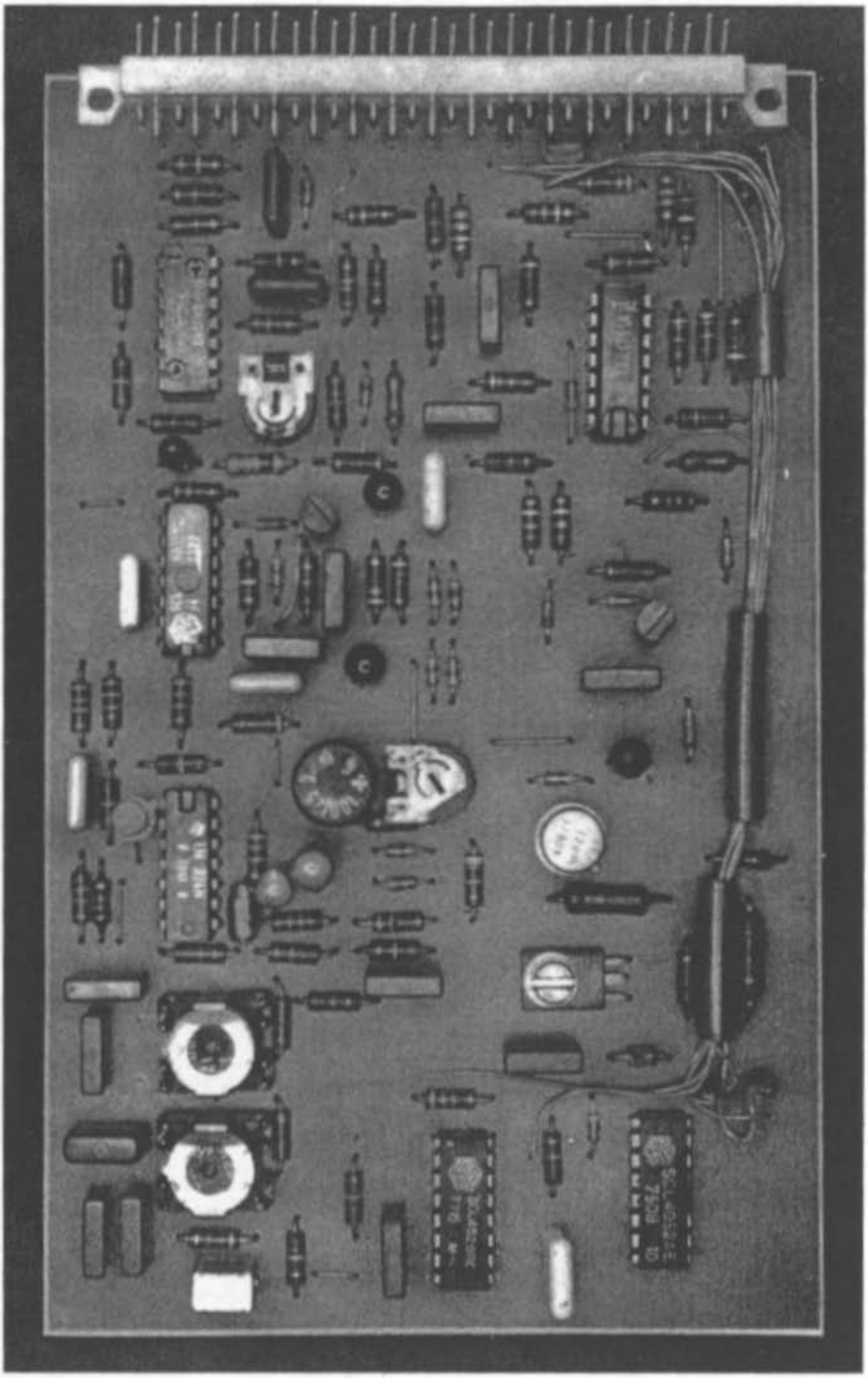

Fig. 45: Photograph of the author's prototype of video board DC 3 NT 005

5.4.6. Alignment of DC 3 NT 005

The alignment controls on board DC 3 NT 005 are:

1 potentiometer for carrier suppression in the AM-mode,

1 potentiometer for the video operating point,

1 inductance for AM-carrier suppression (2400 Hz),

and 1 inductance for suppression of twice the subcarrier frequency (4800 Hz). The alignment is made in the same order.

Preparation:

Connect an operating voltage of + 15 V to Pt 1/2, and ground to Pt 6. Set an AF-generator to 2400 Hz/sinewave (if possible, check with a frequency counter !) and connect this to Pt 8 and ground. Set the output voltage to 235 mV (665 mV peak-to-peak value). Connect an oscilloscope to Pt 9, and adjust the carrier suppression potentiometer so that both halfwaves are of the same amplitude, (white = 5 V peak-to-peak). Reduce the output voltage from the AF-generator by 50 % so that the white-value amplitude is 2.5 V peak-to-peak.

Connect Pt 3 to Pt 9, and connect a multimeter to Pt 4 or Pt 5. Adjust the voltage at these two points with the aid of the video operating point potentiometer to the same amplitude of approximately 7.5 V DC.

The potted cores of the LC-video filter only have a small alignment range. This means that the tolerances of the parallel capacitors sometimes make an alignment impossible. For this reason, two capacitor positions are provided so that the lower value can be varied if an alignment is not possible with the original complements.

The oscilloscope should now be switched to AC-input and is then connected to Pt 5. The residual carrier will then be visible and can be adjusted to a minimum by aligning those inductances. The AF-generator should now be tuned away from the frequency in order to establish whether the minimum is really at 2400 Hz.

The stylus current circuit does not require alignment !

A Microcomputer for Amateur Radio Applications

Part 3: Memory and System Bus

by W. Kurz, DK 2 RY

Part 2 described the construction and operation of the central processing unit (CPU). Since there was no room to include the circuit diagram of the CPU, this is now to be given here in Figure 14.

As is mentioned in Part 1, the microcomputer also requires a memory. A fixed program memory (8 K) and a random access memory (8 K) are now to be described, as is the bus-board.

No decision has been made as to whether PC-boards and kits are to be offered for construction of this microcomputer, and we would like to hear whether our readers are in favour of this or not.

THE MEMORIES

Memories are components that store the inputted information. Fundamentally speaking, there are two different types:

- Fixed-program memories (PROM, ROM)
- Random access memories (RAM)

In contrast to random access memories, fixed program memories will maintain their program even after the operating voltage has been switched off.

There are four different types of fixed-program memories:

- ROM (Read-Only Memory)
- Bipolar PROM (Programmable ROM)
- EPROM (Erasable programmable ROM)
- EEPROM (Electrically erasable PROM)

ROMs are fixed memories that have been programmed with the user program by the manufacturer. The manufacture of the required masks is only worthwhile for large quantities; on the other hand, these mask-ROMs are the cheapest memory components when required in large quantities. ROMs can only be programmed once.

The bipolar PROMs can also only be programmed once. The memory element is either a diode path, or a type of fuse that is burnt through during the programming process. Bipolar PROMs are cheaper than EPROMs, but usually do not have such a high memory capacity.

EPROMs are the most popular memories used as present in microcomputer technology. With these components, the program can be erased by eradiation with ultraviolet light (wavelength approx. 300 to 400 nm). This erases the content of all memory cells.

In contrast to EPROMs, it is possible with EEPROMs (sometimes also called EAROMs) for individual memory cells to be localized and erased. At present, this is the most expensive form of fixed-program memories.

Random access memories (RAMs) maintain their program content as long as the operating voltage is present. As is indicated by the designation, it is possible to enter any required information, and it is usually data that is stored in such memories. There are four types of RAMs:

- Bipolar RAMs
- N-MOS-RAMs
- C-MOS-RAMs
- Dynamic RAMs

Bipolar RAMs have, at present, a relatively low memory capacity and also require a high operating current. The advantage of these components is the short access time of typically 23 ns. This is the time required to obtain certain data.

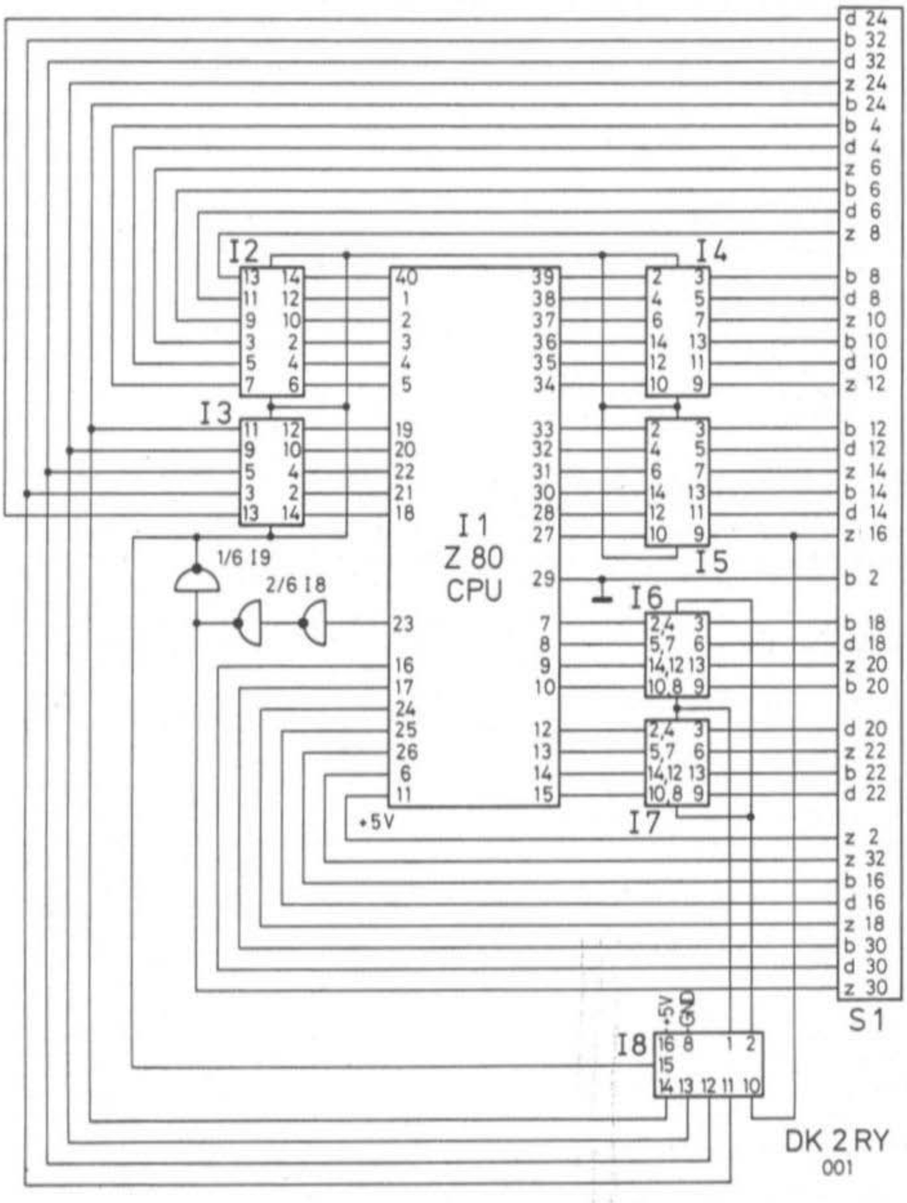

Fig. 14: Circuit diagram of the CPU

N-MOS-RAMs such as type 2114 have a considerably longer access time (typically 0.5 to 1 µs). However, they exhibit a higher memory density (at present up to 8 kBit/component) and only require a low operating current of approx. 30 to 150 mA in comparison to their memory capacity. At the present time, these are the most popular RAMs, especially since they are available inexpensively.

C-MOS-RAMs exhibit approximately the same access time and memory density as the N-MOS-RAMs. However, their main advantage is the low operating current. For example, only approximately 1 to 10 µA flow in the »Stand-by« mode in the case of type 5101. The disadvantage is that C-MOS-RAMs cost up to four times the price of a similar N-MOS-RAM.

Dynamic RAMs have a very high memory capacity (at present up to 32 kBit). These components use the junction capacitance as memory element. However, since the connected MOSFET transistors possess a finite input impedance, these storage cells will discharge. For this reason, it is necessary for the stored information to be continuously refreshed, otherwise it would be lost. This requires a considerable refresh-logic. Dynamic RAMs have a shorter access time than N-MOS-RAMs (typically 200 ns) and exhibit a relatively low »Stand-by« operating current (for example 5 mA in the case of the 16 k RAM MK 4027).

After discussing the memories in general, the types used in the described computer system are now to be described in more detail.

2.1. The Fixed-Program Memory

This memory has a capacity of 8 kByte and stores the programs required for operation. These can be requested by the command keyboard of the periphery unit. The required order sequences are to be described in a later article.

The well-known PROM type 2708 is used, which is lighter and less expensive than the 2 kByte-memory 2716 or even the 4 kByte-memory 2732. This component is a EPROM having a memory of 1 k x 8 memories that can be erased with ultraviolet light. 1 k means that the memory possesses 1024 cells with a word width of 8 Bit. The memory can be erased with ultraviolet light through a small window on the upper part of the case. However, this process can only be carried out approximately 30 to 100 times, and it has been found in a few cases that it is no longer possible to program the EPROM after the first erasure. After the PROM has been programmed, it will hold its program as long as required (up to the next erasure), even when the operating voltage is switched off, or the EPROM taken out of its socket.

Figure 15 shows the connections of PROM type 2708. Three different voltages are required for operation: + 12 V, + 5 V, and − 5 V. The individual memory cells can be addressed via the address lines A 0 to A 9. If a binary value is fed to the address lines, the stored binary value will appear on outputs 01 to 08.

In addition to this, the EPROM 2708 possesses an enable input (CE = Chip Enable). If high-logic level is present at this input, the tri-state outputs will be at high impedance which means that other memories can be active. If, on the other hand, low-logic is connected, this will mean that this memory is active. It is possible in this way for any number of memories to be connected in parallel to increase the capacity.

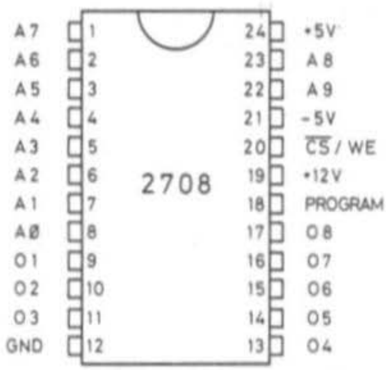

Fig. 15: Connections of the PROM 2708

2.1.1. Programming the 2708

If one wishes to develop one's own program for the microcomputer, it is necessary for this to also be stored. A programming unit is to be described in a later article that allows a programming with the aid of the CPU and an input. The required program will be contained in the operating system. It is also used for comparison, in which the CPU checks whether the program stored in the 2708 memory corresponds to the main program.

2.2. 3 Bit Demultiplexer

This circuit (SN 74 LS 138 or 8205) is used for selection of the various modules such as memory, inputs etc. Its outputs 00 to 07 are selected via its three address inputs A 0 to A 2. If a binary value is present on the address input (for example A0 = 1, A1 = 1, A 2 = 0, corresponding to 011 in binary code), the associated output will be set to L (02 in our example), and all other outputs will be H. However, this is only the case when the circuit is activated. Three enable-inputs E 1, E 2 and E 3 are used for activation. The 3-bit demultiplexer is only active when L is present at the first two, and H at E 3.

Figure 16 shows the connections of the 16-pin case, and **Table 1** shows the logic relationships.

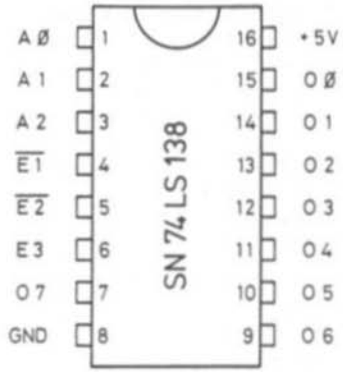

Fig. 16: Connections of the 3 bit/demultiplexer 74 LS 138

2.3. Construction of the Fixed-Program Memory

The 8 k memory is accommodated on a double-coated PC-board of 162 mm x 101 mm with through-contacts. This board is designated DK 2 RY 004 and is in the form of the standard Europe card. The component locations are given in **Figure 17**, and a photograph of the author's prototype is given in **Figure 18**.

The previously mentioned 3-bit demultiplexer 74 LS 138 is used for address coding of the eight integrated circuits. In order to

E 1	E 2	E 3	A 0	A 1	A 2	00	01	02	03	04	05	06	07
L	L	H	L	L	L	L	H	H	H	H	H	H	H
L	L	H	H	L	L	H	L	H	H	H	H	H	H
L	L	H	L	H	L	H	H	L	H	H	H	H	H
L	L	H	H	H	L	H	H	H	L	H	H	H	H
L	L	H	L	L	H	H	H	H	H	L	H	H	H
L	L	H	H	L	H	H	H	H	H	H	L	H	H
L	L	H	L	H	H	H	H	H	H	H	H	L	H
L	L	H	H	H	H	H	H	H	H	H	H	H	L
H	x	x	x	x	x	H	H	H	H	H	H	H	H
x	H	x	x	x	x	H	H	H	H	H	H	H	H
x	x	L	x	x	x	H	H	H	H	H	H	H	H

Table 1: Logic table of the 74 LS 138
x = as required L or H

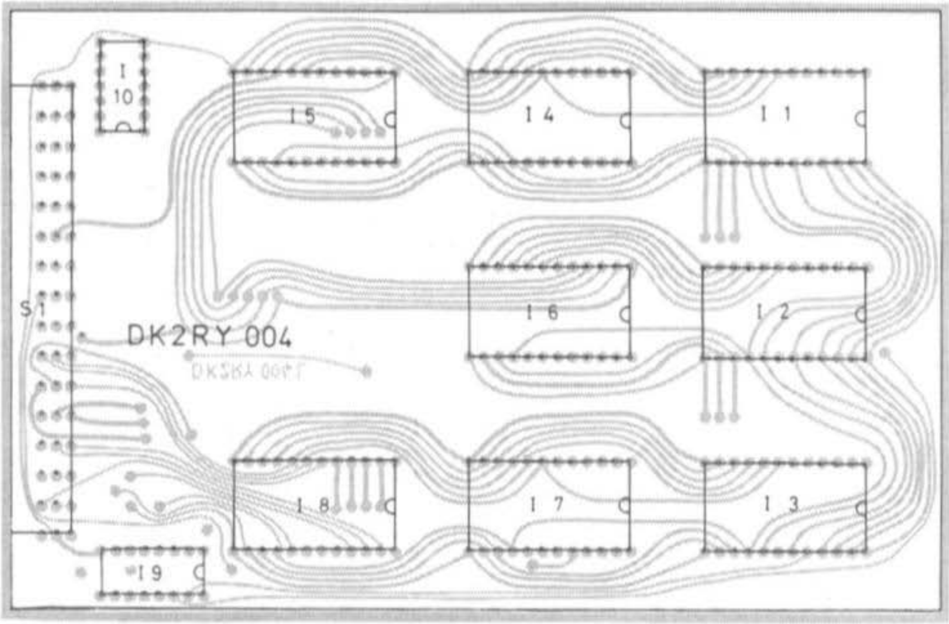

Fig. 17: Component locations on the PROM-board DK 2 RY 004

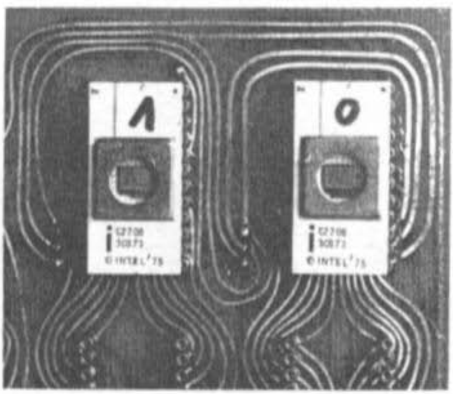

Fig. 18: Photograph of parts of the PROM-board

ensure that the memory is only active when a memory-request is present from the CPU, E 3 of this decoder is connected via the bus-board to a NOR-gate, the MREQ, and A 15. For selection of the individual IC, it is necessary for the address inputs A 0 to A 2 of the decoder to be connected with the address outputs A 10 to A 12 of the CPU. In addition to this, the following interconnections exist: E 1 of the 74 LS 138 to A 13 of the CPU, and E 2 of the 74 LS 138 to A 14 of the CPU **(Figure 19)**.

This allows a correct addressing of the PROM: The PROM-range is addressed by the address range 0000H to 1FFFH. It should be noted that each individual PROM-IC may only be placed in the appropriate socket. If the ICs are exchanged, the program will not function. For this reason, the actual address range of the individual PROM-ICs is given in the following **Table 2**:

IC-No.	Address Range
1	0000H to 03FFH
2	0400H to 07FFH
3	0800H to 0BFFH
4	0C00H to 0FFFH
5	1000H to 1BFFH
6	1400H to 17FFH
7	1800H to 18FFH
8	1C00H to 1FFFH

Table 2: Address range of the PROMs

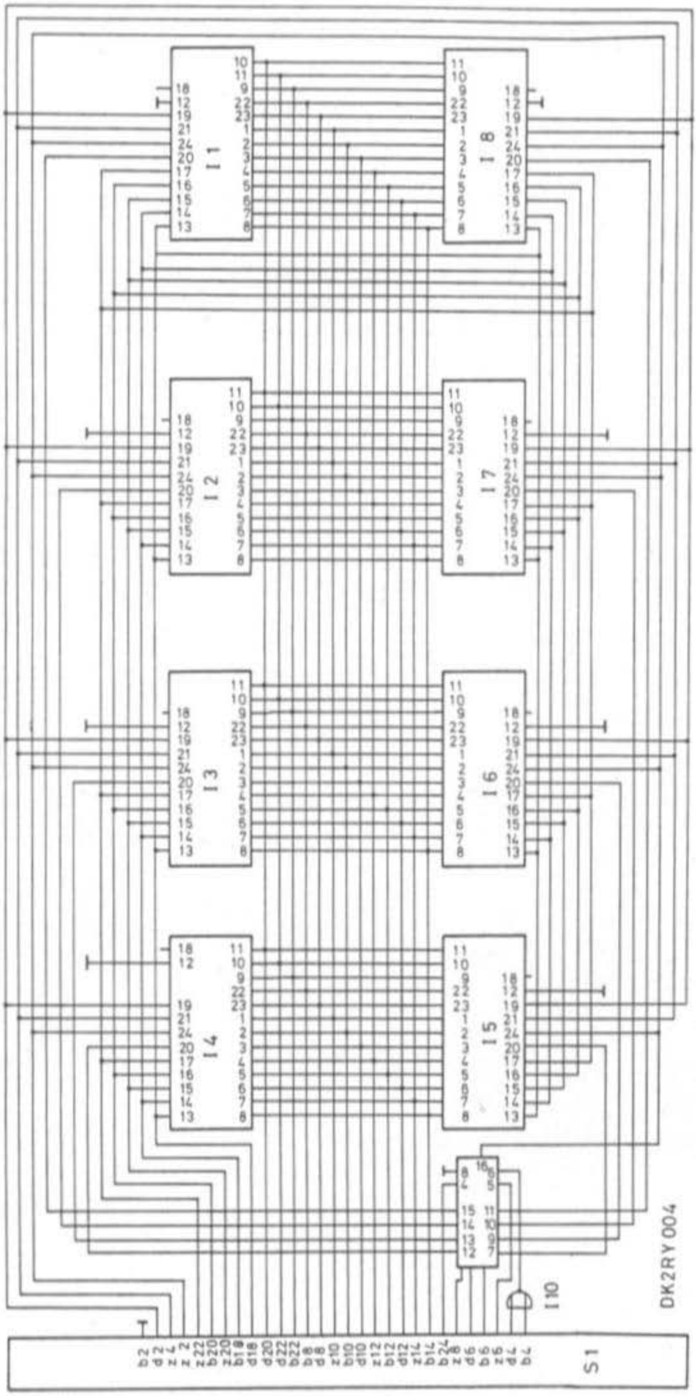

Fig. 19: Circuit diagram of the fixed-program memory

Module DK 2 RY 004 is plugged into the connector strip on the bus-board. This ensures that all required connections are made and no further wiring is required.

Components

I 1 - I 8: 2708 (Intel) or 8708 (Siemens)
I 9: SN 74 LS 138 (TI)
I 10: SN 74 LS 02 (TI)

1 connector strip
(Siemens C 74334-A 40-A 60)

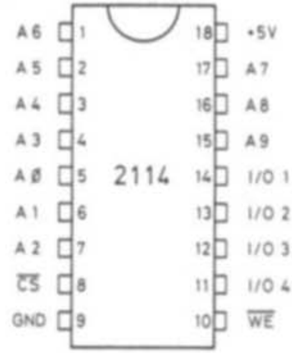

Fig. 20: Connections of the RAM 2114

2.4. Random Access Memory

The RAM has a storage capacity of 8 kByte. This means that a capacity of 16 kByte is available together with the fixed-program memory (16384 memory cells of 8 Bit word-length). If further memory capacity is required, a 16 kByte RAM using dynamic RAMs will be described later which allows a total addressable memory capacity of 64 kByte to be equipped.

Memories type 2114 are used. These are a 1 k x 4 RAM that possess 1024 memory cells with a wordlength of 4 Bit, this is the reason why two integrated circuits are used. The 2114 is a static RAM; the individual memory cells are in the form of Flip-flops (4096 Flipflops are integrated into the 2114).

The RAM type 2114 only requires an operating voltage of 5 V and is activated in the same manner as the 2708 via the CE-input. The $\overline{WE}$-input (Write enable) is used for inputting into the memory; if L is present at this input, the information present at the inputs-outputs (I/01 to I/04) will be inputted into the memory whose address is fed to the address inputs A 0 to A 9. If, on the other hand, H is present at the $\overline{WE}$-input, the opposite will be true and the information from the actual memory cell will be read out.

Figure 20 shows the pin connections of the 18-pin case of the RAM 2114.

2.4.1. Bi-Directional Bus Driver

It is possible to transport data in both directions through a bi-directional bus driver. This IC increases the fan-out of the memory circuits 2114 (1 TTL-load) to 50 low-power TTL-loads, so that larger systems can be connected to the bus. In our case, a bi-directional 8-Bit bus driver type 74 LS 245 is used that is accommodated in a 20-pin case. The connections are given in **Figure 21**.

Fig. 21: Connections of the bi-directional bus driver 74 LS 245

Connections A 0 to A 7 are the inputs-outputs on the data side, and connections B 0 to B 7 are the inputs-outputs on the bus side. Connection DIR is used for directional control according the method shown in **Table 3**. The IC is activated via connection CE.

$\overline{CE}$	DIR	Operation
L	L	Data transport B to A
L	H	Data transport A to B
H	X	Tri-state, high impedance

X = either H or L

Table 3: Data transport of the 74 LS 245

2.5. Organisation of the Random Access Memory

A total of 16 ICs type 2114 are used whose address lines are connected in parallel. As was mentioned previously, 2 ICs form a 1 k x 8 Bit data wordlength. The individual memory cells are addressed via the address lines A 0 to A 9. They are interconnected to the CPU via the bus lines **(Figure 22)**.

Two ICs are activated via the 3-bit demultiplexer (74 LS 138) via the CE-inputs. To achieve this, the address inputs A 0 to A 2 are interconnected to the address outputs A 10 to A 12 of the CPU via the bus in the same manner as used on the PROM board. Input E 1 is connected to MREQ of the CPU in order to only activate the memory in the case of a memory instruction.

A further IC of this type is used for driving the 74 LS 138. The address inputs A 0 to A 2 are interconnected to the address outputs A 13 to A 15. This second 74 LS 138 is also activated by MREQ, that is connected to E 1 and E 2.

The memory is set to the address range of 2000H to 3FFFH. If it is to be set to another range in order to extend the memory, the interconnection between Pt 1 and 0 1 of I 19 should be disconnected and a wire bridge made according to **Table 4** on the conductor side of the board to the appropriate outputs of the 74 LS 138.

The storage can be equipped up to 64 kByte and addressed. A total of 7 RAM boards and 1 PROM board are required for the maximum memory capacity.

Since memories type 2114 can only drive one TTL-load, it is possible that the fan-out is not sufficient for driving larger systems with many memories and inputs-outputs. The »low«-voltage would increase to a non-permissible value, and a false data word would be fed to the bus. For this reason, a double buffering is used, in other words, the memory is connected to the data bus via a buffer stage. The previously described bi-directional 8 Bit bus driver 74 LS 245 is used for this. Its Chip Enable is connected to Pt 1, which means that it will only be active when a memory instruction is given from the 8 k RAM. In all other cases, the tri-state outputs are at high impedance.

Memory range	Address range	Connect Pt 1 to:
0 to 8 k	0000H to 1FFFH	do not connect !
8 to 16 k	2000H to 3FFFH	already connected
16 to 24 k	4000H to 5FFFH	O2
24 to 32 k	6000H to 7FFFH	O3
32 to 40 k	8000H to 9FFFH	O4
40 to 48 k	A000H to BFFFH	O5
48 to 56 k	C000H to EFFFH	O6
56 to 64 k	D000H to FFFFH	O7

Table 4: Extension and addressing of the memory

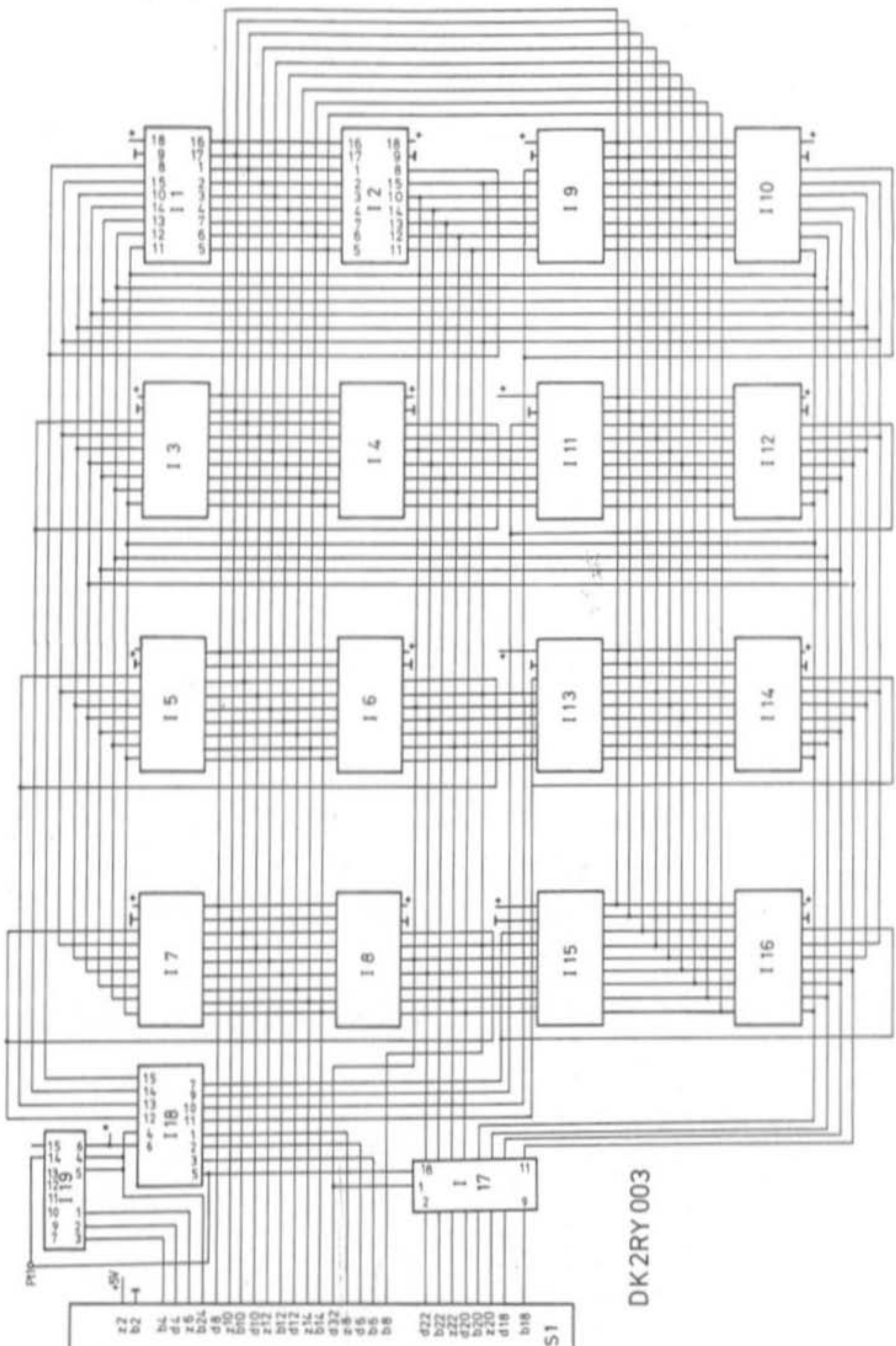

Fig. 22: Circuit diagram of the random access memory

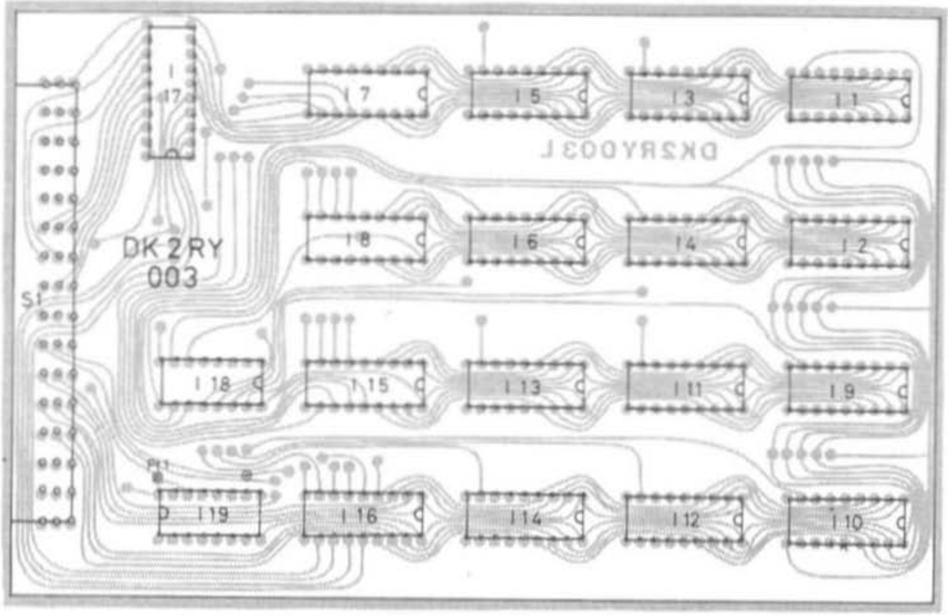

Fig. 23: Component locations on the RAM-board DK 2 RY 003

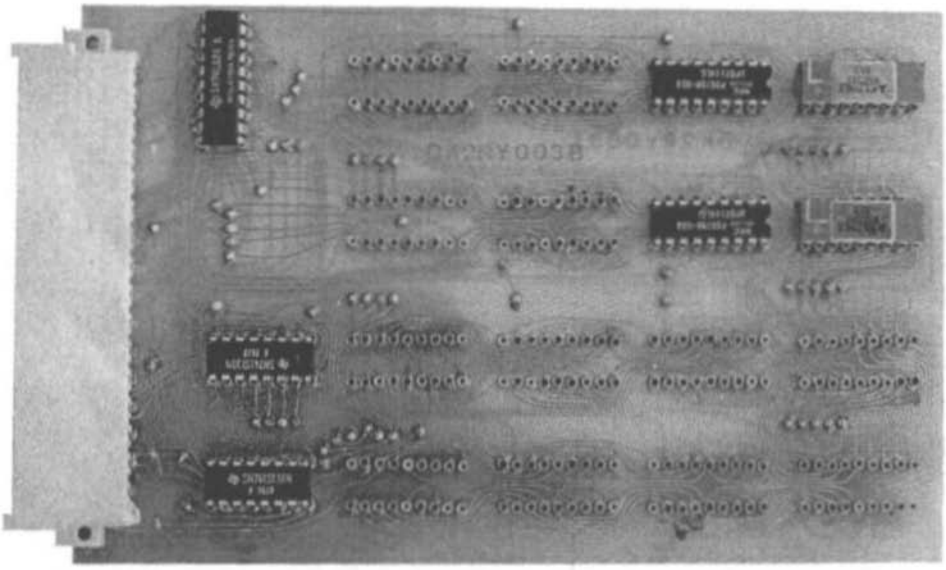
Fig. 24: Photograph of the RAM-board

If the microcomputer is not to be so expensive (for example only a I/0-unit, 1 RAM with 8 k and 1 PROM), it will not be necessary to use a 74 LS 245. However, it is then necessary for the connections of the A-row to be connected to the associated connections of the B-row (A 0 to B 0, A 1 to B 1 etc.).

2.6. Construction of the Random Access Memory

The 8 k RAM is accommodated on a double-coated PC-board with through-contacts having the dimensions of 162 mm x 101 mm. This board is designated DK 2 RY 003 (see **Figure 23**). The sockets are inserted on the »B« side of the board. This module is ready for operation as soon the connector strip is soldered into place and the integrated circuits are plugged in **(Figure 24)**.

Components

I 1 to I 6:	2114 (Intel, NEC)
I 18, I 19:	SN 74 LS 138 (TI)
L 17:	SN 74 LS 245 (TI)

1 connector strip
(Siemens C 74334-A 40-A 60)

2.6.1. Memory Capactiy

The memory capacity requirement naturally depends on the program. The highest memory capacity is probably the contest program which may require up to 64 k. Maybe, only approximately 1 k is required for amateur satellite operation (OSCAR) or for RTTY or CW decoding. The exact memory capacity requirement – only taking the RAMs into consideration – will be given after completing the programs.

It is absolutely necessary to equip the first kByte, since 1 k are required for the stack and the data memory. The remaining 7 k can be equipped as required in 1 kByte steps. It is only necessary for the associated RAM-IC to be inserted into its socket. **Table 5** gives the addresses required by the random access memory:

IC-No.	Address Range	Notes
1 + 2	2000H to 23FFH	Stack + Data memory
3 + 4	2400H to 27FFH	free
5 + 6	2800H to 2BFFH	free
7 + 8	2C00H to 2FFFH	free
9 + 10	3000H to 33FFH	free
11 + 12	3400H to 37FFH	free
13 + 14	3800H to 3BFFH	free
15 + 16	3C00H to 3FFFH	free

Table 5: Address range of the random access memory

Since the random access memory does not have a fixed program, free range can be used as required in contrast to that of a fixed program memory. It should only be noted that it is only possible to write in, or read out of the equipped part. As was the case with the fixed-program memory, it is only necessary for the random access memory board to be plugged into the system board; no further wiring is required.

3. THE SYSTEM BUS

As has already been mentioned, the system bus interconnects all individual modules, which includes all interface units and the power supply. The interconnection is made using the previously mentioned connector strips and parallel lines on the board.

3.1. Construction of the System Bus

The bus-board is constructed using a 185 mm x 102 mm large double-coated PC-board with through-contacts. The PC-board is designated DK 1 RY 002, and is shown in **Figures 25 and 26.**

The bus-board is provided with nine positions for the individual modules. Attention should be paid during construction that all connector strips are mounted in the same direction. One end of the connector strip is somewhat longer than the other, and this longer side should always be mounted on the same side of the bus-board (see **Fig.27**). If attention is not paid to this, the wrong connections will be made between the individual modules, which could lead to the destruction of individual integrated circuits.

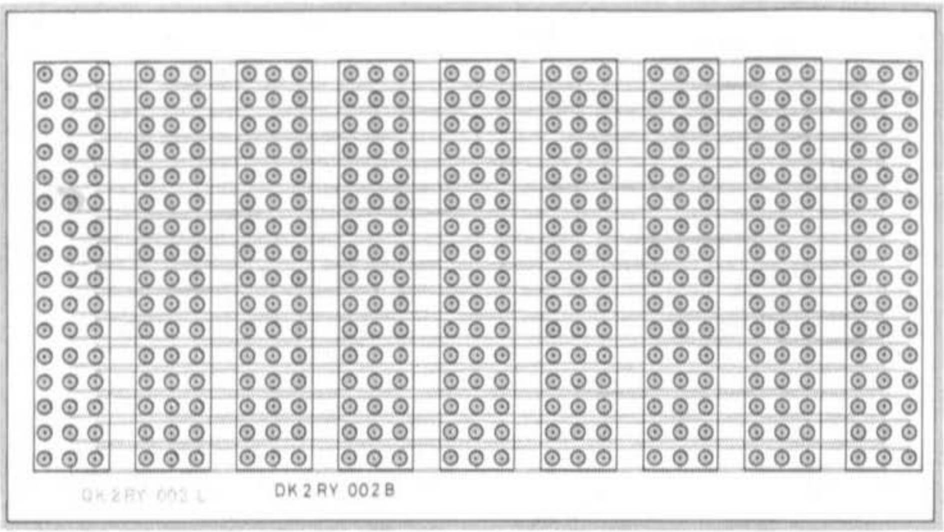

Fig. 25: The bus-board DK 2 RY 002

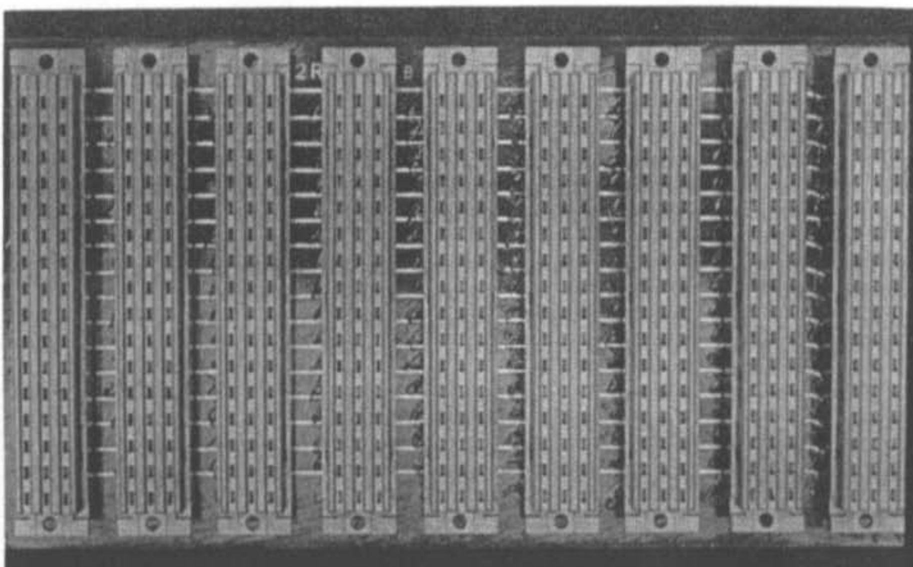

Fig. 26: Photograph of the bus-board

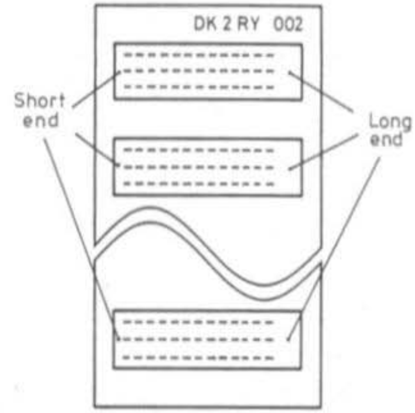

Fig. 27: It is important to mount the connector strips in the same direction

If several bus-boards are to used in order to extend the memory capacity, these should be connected using wire bridges on the lower side of the board.

The bus-board can be equipped with the modules in any order. It is advisable to mount the bus-boards on the rear panel of a suitable case.

Components

9 connector strips
(Siemens C 74334-A 80-A 40)

REFERENCES for Part 3

Data sheet of the 2114 (NEC)
Data sheet of the 2708 (Intel)
Data sheet of the SN 74 LS 138 (TI)
Data sheet of the SN 74 LS 245 (TI)

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 3/1980 of VHF COMMUNICATIONS

DJ 3 VY	**Low-noise 70 cm Converter**		**Ed. 3/1980**
PC-board	DJ 7 VY 002	double-coated with thru-contacts	DM 27.—
PC-board	DJ 3 VY 002	(oscillator) double-coated with thru-contacts	DM 29.—
Semiconductors	DJ 3 VY 001/2	for 5 bandsegments: 18 transistors, 9 Schottky, 5 zener, and 4 diodes	DM 175.—
Ring mixer	TAK-1 WH	high-level mixer upto 600 MHz	DM 120.—
Minikit 1	DJ 3 VY 001/2	for 5 bandsegments: 3 ceramic, 5 air-spaced and 5 plastic-foil trimmers, 9 chip caps., 2 feedthru caps., 2 tantalium caps.	DM 42.—
Minikit 2	DJ 3 VY 001/2	for 5 bandsegments: 6 two-hole ferrite cores, 7 toroids, 2 diff. dia. enamelled copper wire, 2 different dia. silver-plated wire, 2 coilsets, 5 coilformers with core, 24 miniature chokes, 2 ferrite beads, 1 six-hole ferrite core	DM 97.—
Crystals	HC-18/U	100.5 / 101.0 / 101.5 / 102.0 / 102.5 MHz at DM 26.— each	DM 120.—
Kit	**DJ 3 VY 001/2**	**complete with above parts**	**DM 610.—**
DK 1 OF 043	**Automatic SWR-Indicator**		**Ed. 3/1980**
PC-board	DK 1 OF 043	single-coated with plan	DM 15.—
Semiconductors	DK 1 OF 043	1 analog multiplier AD 532 kh, 2 op.ampl. 741, 3 op.ampl. TAA 761 A, 2 Darlingtons, 2 zener diodes, 2 LED, 1 bridge rectifier	DM 183.—
Minikit	DK 1 OF 043	3 relays (DIL), 6 alu.-electrolytics, 2 ceramic caps., 4 trimmer pot., 24 carbon resistors	DM 58.—
Kit	**DK 1 OF 043**	**complete with above parts**	**DM 250.—**
DC 3 NT 005	**Weather Satellite Reception System, Video Processor Module**		**Ed. 3/1980**
PC-board	DC 3 NT 005	single-coated, with plan	DM 17.—
Semiconductors	DC 3 NT 005	8 transistors, 18 diodes, 3 C-MOS IC's, 3 linear IC's	DM 54.—
Minikit	DC 3 NT 005	2 potted cores, 67 carbon resistors, 2 resistors 1 W, 2 pots., 23 pl.foil, 1 ceramic, 3 tantalum and 3 electrolytics, 1 31-pin connector	DM 51.—
Kit	**DC 3 NT 005**	**complete with above parts**	**DM 120.—**

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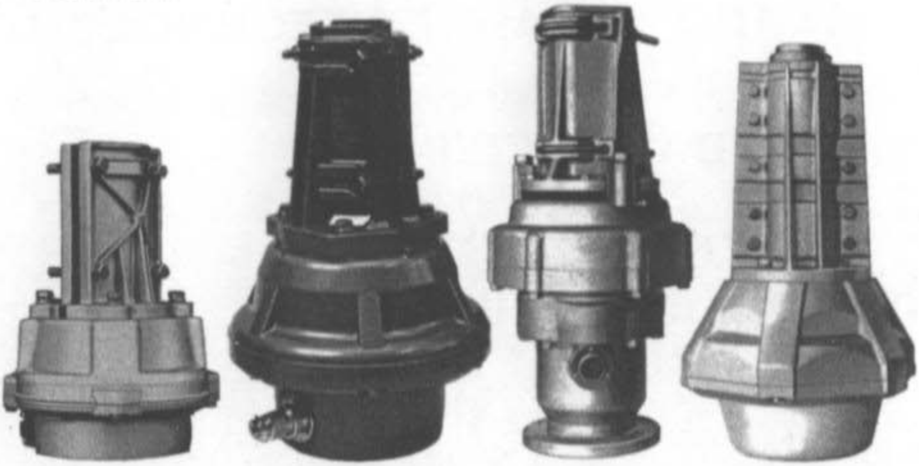

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Ripple		< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 0.5 dB
Insertion loss		< 3 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 6.5 dB
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	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
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