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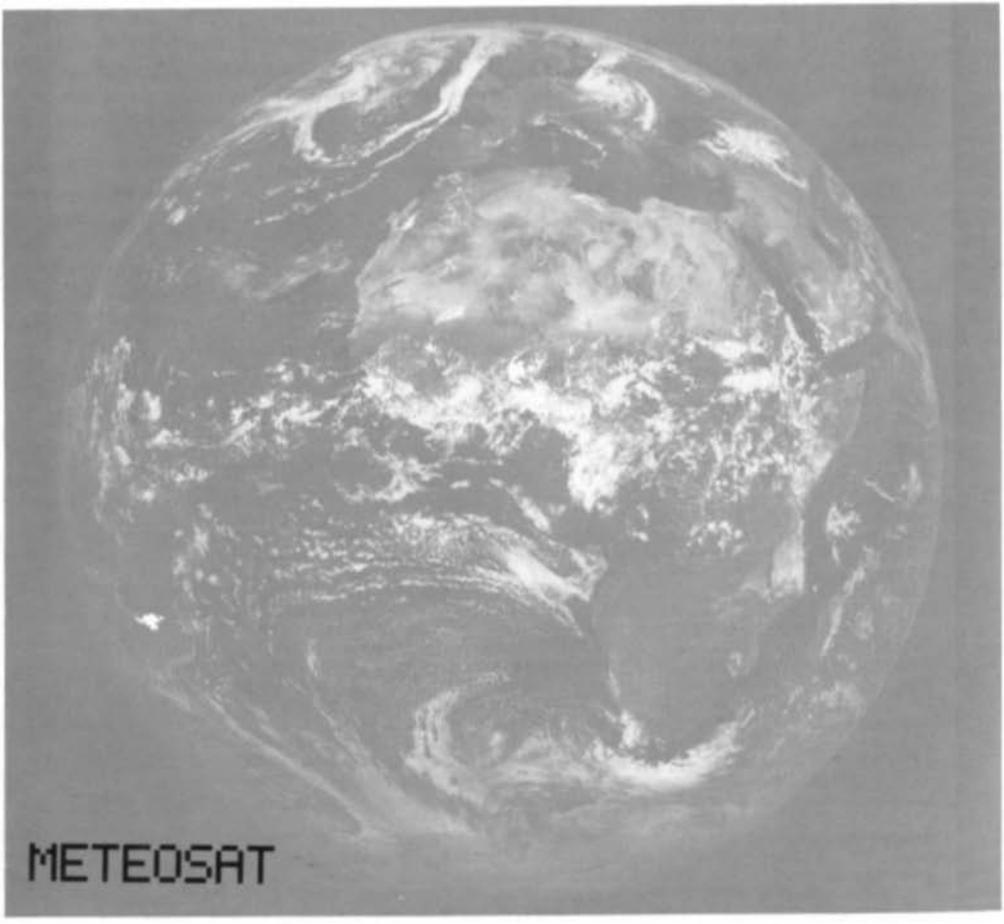

VHF COMMUNICATIONS

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SYNTHESIZER FOR THE 2 m BAND IN C-MOS TECHNOLOGY

by G. Heeke, DC 1 QW

Phase-locked circuits were firstly conceived in 1932 by de Bellescize. In amateur radio circles, phase-locked loop circuits have been used as long as there have been integrated circuits made especially for this application.

At first, PLL oscillators offered very unfavorable technical specifications. Better specifications and high reproducibility can now be achieved by use of modern components and consideration of the special demands of phase control circuits.

A 80 channel PLL synthesizer is to be described that exhibits a high spurious signal rejection, short switching times and very favorable power consumption. Any required frequency shift can be switched, and a fixed-channel programming can be provided. The circuit concept allows various variations so that it is possible to use the synthesizer in equipment having different intermediate frequencies.

1. OPERATION OF THE CIRCUIT

The operation of phase-locked loops and multi-channel oscillators were described in the previous publications (1), (2), (3).

The phase-locked oscillator is now to be described in conjunction with the block diagram given in **Figure 1**.

The actual oscillator is a voltage-controlled oscillator and buffer. The frequency range is switchable. Separate, two-stage amplifiers are provided for the transmitter and the receiver. This means that an RF-switching is not necessary when changing over from transmit to receive.

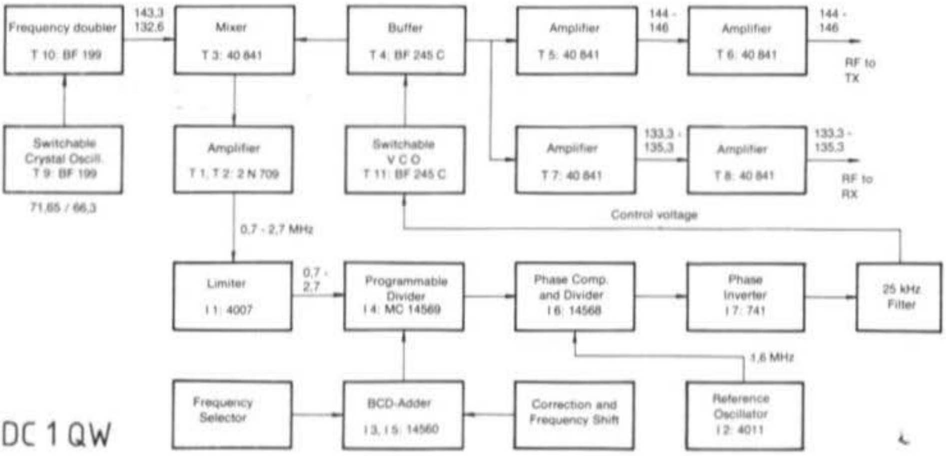


Fig. 1: Block diagram of the phase-locked oscillator

Since the frequency divider used in the control loop is only able to process frequencies up to approximately 2 MHz, a switchable crystal-controlled oscillator with doubler as well as a mixer are provided in order to convert the generated VCO frequency to a lower frequency level. The required VCO frequency range and the associated crystal oscillator are electronically switched so that they are suitable for the transmit and receive mode. If a frequency conversion system is used in the transmitter, it will only be necessary for one crystal oscillator to be provided, since the same oscillator frequencies can be used for transmitter and receiver.

The mixer stage feeds the frequency difference between the VCO and the doubled crystal oscillator frequency to a two-stage transistor amplifier. This is followed by a limiting integrated amplifier equipped with field effect transistors. The output of this stage is at C-MOS-level, which is required for driving the subsequent programmable divider.

The frequency division factor of the programmable divider is dependent on the required output frequency of the VCO, and also on the crystal oscillator frequencies, as well as possibly on a required frequency shift (repeater). An adding circuit ensures that the correct division ratio is maintained.

The phase comparator, in which the reference frequency is compared to the output frequency of the programmable frequency divider and a control voltage is generated, is then followed by a phase reversal stage. This is necessary in order to provide the control circuit with a signal of correct polarity. The control loop is closed via a filter for suppression of any residual phase comparator frequency of 25 kHz, and the output voltage of the filter is fed as control voltage to the VCO.

The reference frequency used to maintain an exact channel spacing is obtained from an integrated crystal oscillator using an 1.6 MHz crystal.

2. CIRCUIT DESCRIPTION

The RF portion of the synthesizer is shown in **Figure 2,** and **Figure 3** shows the circuit of the digital portion. The individual stages of the PLL oscillator are to be described in the following section. Characteristics that lead to the selection of special circuits are to be discussed in detail.

It has already been mentioned that the overall concept can be varied. The following description is to be based on a synthesizer for a receiver having an intermediate frequency of 10.7 MHz, and direct generation of the transmit frequency.

2.1. The Voltage-Controlled Oscillator

In the case of an intermediate frequency of 10.7 MHz, it is necessary for the VCO to operate in the range of 133.3 MHz to 135.5 MHz in the receive mode, as well as in the range of 144.0 MHz to 146.0 MHz in the transmit mode.

The noise-characteristics of oscillators are mainly dependent on the Q of the resonant circuits, and transistor noise (4). Field effect transistors offer considerable advantages here as they provide a more favorable noise behaviour in the RF-range than bipolar transistors. One cause of this is the lack of Shot-effect noise of the minority currents (5). An oscillator circuit using the FET BF 245 C was selected for the VCO. This transistor has proved itself in the oscillator described in (3).

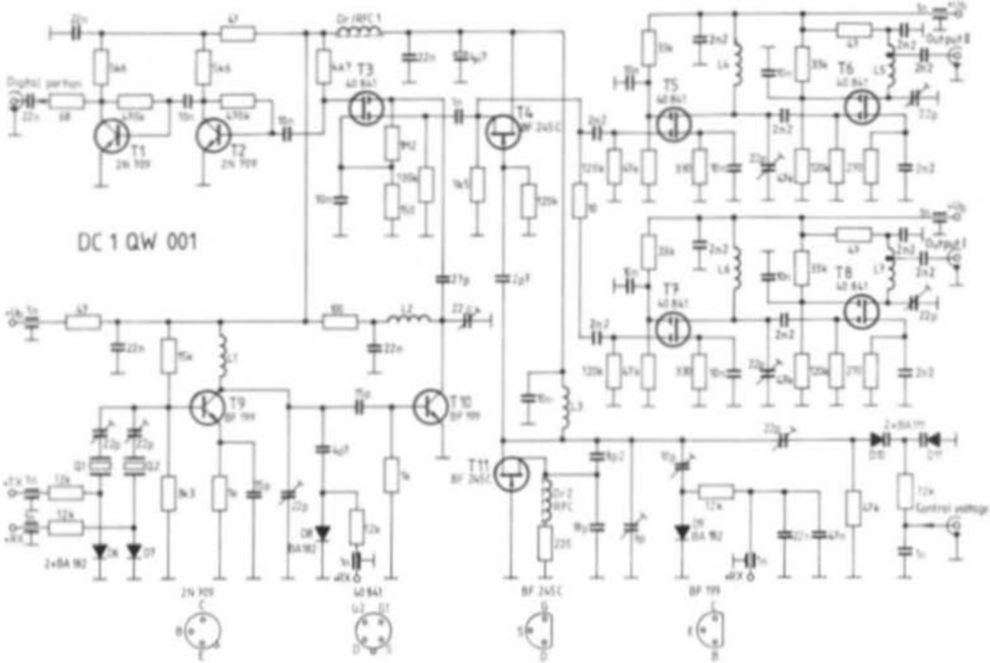


Fig. 2: Circuit diagram of the RF-portion of the synthesizer DC 1 QW 001

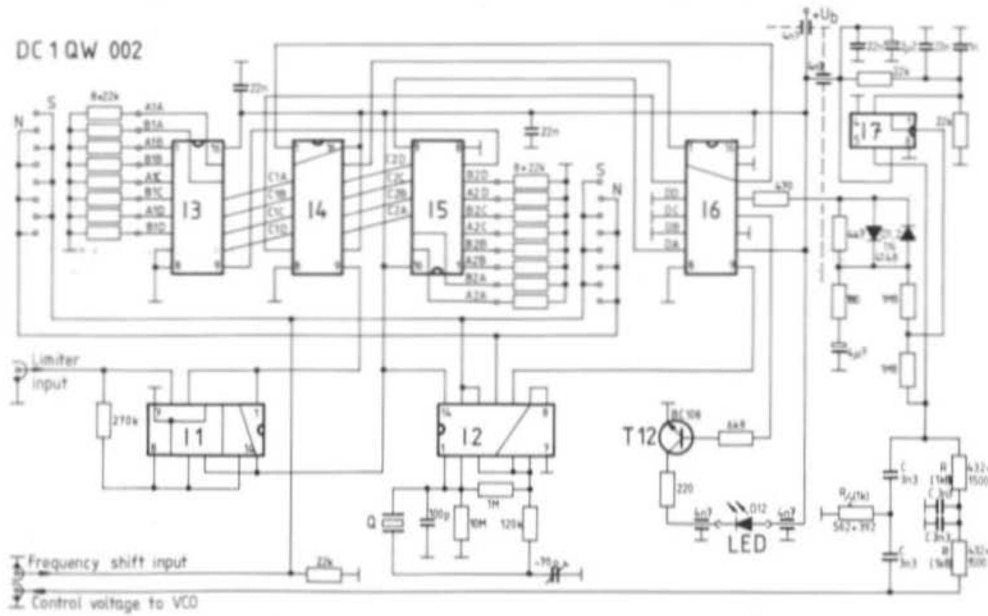


Fig. 3: Circuit diagram of the digital portion of the synthesizer DC 1 QW 002

The frequency variation is made using two varactor diodes in a circuit that avoids unwanted rectification effects. The coupling of the varactor diodes to the resonant circuit is variable using a trimmer capacitor, which means that the required tuning slope can be adjusted easily when any inavoidable spreads of the components must be compensated for.

A diode is used to switch in an additional capacitance in the receive mode so that the oscillator operates at a lower frequency, corresponding to the difference between transmit and intermediate frequency. This preliminary frequency alignment allows a relatively loose coupling of the varactor diodes and thus plays its part in the purity of the oscillator spectrum.

2.2. Buffer und Amplifier

The VCO is followed by a buffer equipped with a BF 245 C field effect transistor. It is true that the voltage gain of such a source follower is less than 1, however, it is possible together with the small coupling capacitor to decouple the oscillator and amplifier stages well, and thus load the oscillator resonant circuit to a minimum.

The amplifier stages equipped with protected DG-MOSFETs do not have any special features and are built up nearly identically. Besides providing an additional decoupling, they are provided to amplify the oscillator signal so that sufficient power is available for driving ring mixers with their relatively high power requirements. The high output power level is also favorable for direct drive of a transmit amplifier. If required, it is possible for the gain per stage of the DG-MOSFET and thus the output power to be reduced by decreasing the voltage divider resistors between gate 2 and ground.

The output signal of the oscillator amplifier chain can be coupled out at high or low impedance according to whether a short connection to a high-impedance consumer (e.g. gate of a DG-MOSFET) or to whether a low-impedance coaxial cable is to be connected, e.g. for feeding a ring mixer.

2.3. Crystal Oscillator and Frequency Doubler

The generated crystal oscillator frequencies serve to transpose the frequency of the VCO together with the mixer stage into a frequency range that can be processed by the programmable frequency divider, which operates in C-MOS-technology.

The selection of the crystal frequencies is dependent on the frequency divider used. Experiments have shown that the frequency divider used can process frequencies satisfactory up to 3.4 MHz. If the 2.0 MHz required to cover the 2 m amateur band is subtracted from this 3.4 MHz, a safety range of 1.4 MHz remains for frequency shifts. This safety range is split up so that the same amount of reserve is provided above and below the VCO frequency range, e.g. 0.7 MHz. This means that the oscillator frequencies will be 0.7 MHz under the lowest VCO nominal frequency of 133.3 MHz and 144.0 MHz. If crystal oscillator frequencies of 66.3 MHz and 71.65 MHz are selected, output frequencies of 132.6 MHz are provided after doubling in the receive mode, and 143.3 MHz for transmit. This guarantees that the VCO is able to securely phase-lock when its frequency is up to 0.7 MHz above or below the nominal frequency due either to temperature, or operating voltage fluctuations. Normally, such large oscillator frequency drifts are not present in practice.

Conventional bipolar transistors are used in the crystal oscillator and doubler stages. The oscillator operates in a common base circuit with internal capacitive feedback. Class A-operation was selected to obtain low distortion and good frequency stability. A frequency shift is possible by providing a capacitance across the resonant circuit and switching the crystal. The selected silicon planar diodes exhibit a very good switching behaviour, even at higher frequencies. The required crystal is switched on by connecting the operating voltage to the dropper resistor of the diode. It is not necessary for a negative voltage to be used to block the other diode.

The frequency doubler stage operates in a common emitter circuit. The resonant circuit at the output provides sufficient bandwidth, which means that no frequency switching is required.

2.4. Mixer

The mixer is equipped with a dual-gate MOSFET, type 40 841, which allows a well decoupled mixing at high conversion gain. The bias voltage for gate 2 is generated as the voltage drop across the source resistor. The source and gate 2 are decoupled using a high-impedance resistor.

2.5. Amplifier and Limiter

Switching transistors with storage times in the order of nano-seconds are used in the transistor amplifier. A voltage feedback coupling was used for temperature stabilization. The low-impedance resistor through which the output signal is taken from the collector of the second transistor, suppresses any interference due to reflections, even when a long coaxial cable is used for interconnection to the subsequent limiter.

The task of the limiter amplifier is to process the output signal from the two-stage transistor amplifier so that it is suitable for processing in the frequency divider. Two complementary field effect transistor pairs and an inverter stage are integrated in I 1. They amplify the input signal that can be anywhere in the frequency range, to a value of approximately 10 V (peak-to-peak), which corresponds to the required C-MOS level. At low input frequencies, e.g. 1 MHz, the output voltage is a clean square wave, however, the slope obtained is sufficiently great even at the highest possible input frequencies.

2.6. The Programmable Divider

The limiter provides frequencies of 0.7 MHz to 2.7 MHz both in the transmit and receive mode if the VCO is varied over its nominal frequency range, and when the recommended crystal oscillator frequencies are used for mixing. A frequency of 0.7 MHz corresponds to the lower band limit of 144.0 MHz and 2.7 MHz correspond to the upper band limit of 146.0 MHz. In the case of a 25 kHz spacing, the given frequencies must be divided by the value of the phase comparator frequency of 25 kHz.

This results in:

$$\frac{0.7\ \text{MHz}}{0.025\ \text{MHz}} = 28 = n_{00}$$

This represents the lowest division factor that can be selected.

The largest division factor is:

$$\frac{2.7\ \text{MHz}}{0.025\ \text{MHz}} = 108 = n_{80}$$

Since the programmable divider has three positions, it possesses three decades. Two of these are provided in the integrated circuit MC 14569 (I 4), whereas the third decade is accommodated in the phase comparator MC 14568 (I 6). Both integrated circuits are matched to another and were especially designed for use in frequency synthesizers.

The programmable divider itself has several special features. It is in the form of a cascaded, selectable downwards counter. In order to achieve the highest possible input frequency, a special circuit was used for the zero-state, and a Johnson counter used in the unit decade. Although this would be possible, the divider is not used as binary counter.

The programmable divider is provided with its coded input information and the division factor via a total of 12 data inputs, since a 4-Bit code is necessary for each decimal position and the divider comprises a total of three decades. The data lines are designated as follows in the circuit diagram given in Figure 3:

First decade:	(unit, least significant digit):	C 1 A to C 1 D
Second decade:	(tens):	C 2 A to C 2 D
Third decade:	(hundreds, most significant digit):	DA to DD

It will be seen in the circuit diagram that three of the four data inputs (DB, DC, DD) of the third decade (accommodated in I 6) are grounded, and are thus continuously provided with »low« level. This is necessary since only the digits 0 (for division factors up to 99) or 1 (for division factors from 100 upwards) can be adjusted, using this decade.

2.7. The Adding Circuit

BCD-adding circuits (I 3, I 5) are provided in front of the programmable divider. They offer three advantages over other well-known circuits:

1. Only two coding switches are required, although the highest division factor of 108 possesses three digits.
2. It is not necessary for the actual division factors of 28 to 108 to be selected on the coding switches, but the required channel number. This means that the channel numbers from 00 to 80 (corresponding to 144.0 MHz to 146.0 MHz) can be directly selected.
3. A frequency shift can be realized without providing further crystals. For instance, a programmed frequency shift of 600 kHz is suitable for repeater operation. In addition to this, inverted operation of transmitter and receiver allow the input frequency of the repeater to be checked, or communication to be made with other stations when no repeater is present in the area.

The task of the two BCD adders is to add a digit to the channel number selected on the coding switch (between 00 and 80) so that the required division factor results as sum. In normal operation, it is necessary for a digit of 28 to be added.

If a lower digit is added, the synthesizer will provide a correspondingly lower output frequency. This means that a frequency shift can be realized in a very simple manner. It is only necessary for the division factor to be decreased by 600 kHz ÷ 25 kHz = 24 in the case of a 600 kHz shift, which means that a digit of 4 should be added to the selected channel number when operating with a frequency shift instead of a digit of 28.

Figure 4 shows how the digits to be added are fed to the BCD adders.

The inputs of I 3 and I 5 are provided with »pulldown«-resistors. These ensure that a »low« level is present, as long as no digits are to be added.

As has already been shown, the adders must be provided with a digit of 28 or 4, according to whether simplex or frequency shift operation is required. If the frequency shift mode is required, the line designated with »S« will be at »high« level. An inverter stage in I 2 ensures that the line designated »N« is at »low« level. Input B 1 C will receive »high« level via diode D 3, which means that the BCD-coded digit 4 will be fed to the unit adder.

For normal operation without frequency shift, the logic levels on lines »N« and »S« are reversed. Since the »N« line is now at »high« level, a digit of 28 will be realized in the BCD coding circuit with the aid of diode D 4 and D 5.

If different frequencies than those recommended for the crystal oscillators are to be used, or different frequency shifts are required, a different coding will be required. The PC-board to be described allows such modifications. The BCD codes are given in **table 1**.

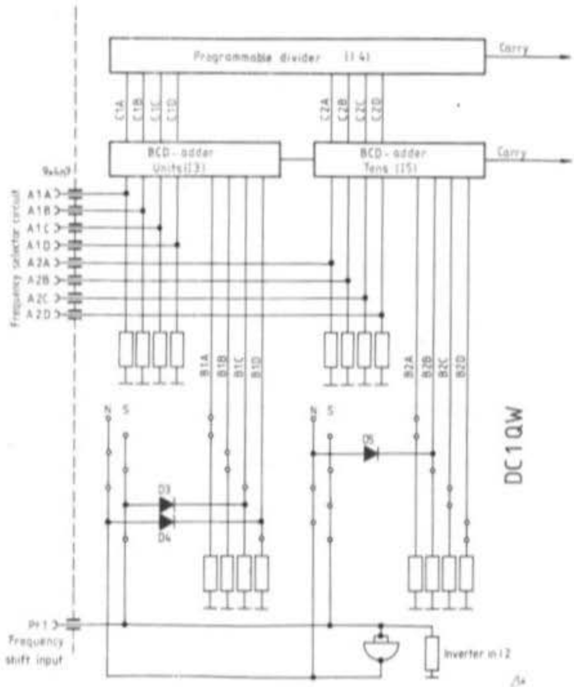


Fig. 4: Part of the circuit of the BCD-adder and programmable divider

Decimal	Dual A	B	C	D
O	O	O	O	O
1	L	O	O	O
2	O	L	O	O
3	L	L	O	O
4	O	O	L	O
5	L	O	L	O
6	O	L	L	O
7	L	L	L	O
8	O	O	O	L
9	L	O	O	L

Table 1

The required channel number (00 to 80) is fed to the adder circuit as second term of the sum. This is also made in BCD-code. Inputs A 1 A to A 2 D are provided for this purpose.

After the digits to be added are fed to the adding circuits, the sum will be present at the outputs designated with C 1 A to C 1 D and C 2 A to C 2 D of the integrated circuits MC 14560 (I 3, I 5), which are directly connected to the programmable divider.

2.8. Reference Oscillator

The phase-comparator frequency of 25 kHz is derived from a 1.6 MHz oscillator. A crystal oscillator of 1 MHz was not used since its ninth harmonic would interfere if a receiver using a 9 MHz intermediate frequency were used. The selected frequency is also favorable since no further dividers are required in addition to that contained in I 6 (64 : 1).

In contrast to oscillators using TTL circuits, the oscillator circuit used here with MOS inverters provides a low loading of the crystal, which in turn provides a high long-term stability. The fine alignment of the frequency is made with the aid of a trimmer.

2.9. The Phase Comparator Circuit

The integrated circuit MC 14568 (I 6) comprises, in addition to the previously mentioned dividers, an arrangement of slope-triggered flipflops and a pair of MOS transistor switches. These combine to form a phase comparator. An external capacitor is switched using the switching transistors and resistors either to the operating voltage or to ground, according to the phase difference. When the phase difference is zero (zero frequency difference), the two input signals switch off both transistors so that the momentarily present capacitor voltage will remain as long as no unwanted discharge takes place.

The capacitor forms a lag-lead filter together with the current limiting resistors, which suppresses any interfering AC-voltage components of the control voltage. This filter configuration was chosen in order to obtain a favorable phase response.

As soon as the voltage drop across the 4.7 kΩ resistor due to the capacitor charge current is greater than 0.6 V, one of the two anti-phase diodes will conduct. It is now possible for a capacitor charge or discharge to be carried out rapidly. The transient time of the control circuit is drastically shortened in this manner, which is very favorable when switching between two frequencies that are far apart.

When the input signals are of equal phase, the comparator will provide a control signal. This is then fed to a LED via a transistor for amplification of the output current. This indicator diode is very useful during the alignment, since it indicates whether a frequency is synchronized, or not.

2.10. The Phase Reversal Stage

The control voltage from the phase comparator, which is stored in a capacitor, is then fed as input signal to the subsequent phase inverting stage. An operational amplifier is used here that is switched as a subtractor. The capacitor voltage is subtracted from the operating voltage. This means that an increasing control voltage will be converted into a decreasing voltage and vice versa.

Due to the high-impedance input of the operational amplifier and the 1,8 MΩ coupling resistor, the previously mentioned charge capacitor will be only slightly loaded.

Only a positive supply voltage is used for the operational amplifier instead of the usual positive and negative supply voltages. This means that its output voltage could not fall below approximately 1.8 V. The Q of varactor diodes that are driven from the generated control voltage is far higher in the range above 1.8 V than at lower voltages. The previous measures assure that the varactor diodes of the VCO always operate at high Q, which has a very positive effect on the overall Q of the oscillator resonant circuit.

2.11. The Output Filter

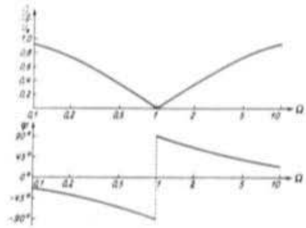
A disadvantage of many of the present synthesizer circuits is the insufficient suppression of output frequencies that are spaced to the value of the phase comparative frequency above and below the nominal frequency. It is hardly possible for these interference frequencies to be suppressed in the transmitter or receiver. It is therefore necessary to ensure that such spurious signals are not generated. They are caused by a residual phase comparator frequency on the control voltage line, which causes a frequency modulation of the oscillator.

In the case of the circuit used here, a special phase comparator is used. In comparison to other comparators that provide squarewave alternating voltage which must be integrated, it provides a pure DC-voltage under locked conditions. In addition to this, a filter for the phase comparator frequency is additionally used in the control voltage line. In practice, it is possible to obtain additional spurious voltage rejections of up to 60 dB. The passive double T-network used has a frequency response that is favorable for the control circuit. Its phase and amplitude response is given in **Figure 5**. The equations are given so that the components can be calculated for other applications of the filter:

$$R = \frac{1}{2 \times \pi \times f \times C}$$

$$C = \frac{1}{2 \times \pi \times f \times R}$$

Fig. 5: Phase and frequency response of the filter for suppression of the phase comparator frequency



f = frequency to be attenuated

In practice, one usually inserts a suitable capacitance value, and the resistors are calculated. The attempt is then made to realize the calculated values by combination of various resistors.

Such a double-T-filter should be fed from a low-impedance source (e.g. operational amplifier) and should not be loaded at the output side.

2.12. The Frequency Selector Circuit

In order to ensure that the operation of the synthesizer is simple, a printed circuit board was developed for accommodation of a diode matrix and several additional components. The matrix is programmed by soldering diodes to the correct position so that up to 15 frequencies could be selected using a rotary switch.

If a repeater frequency is to be programmed, it will be necessary to solder in another diode in addition to that for the channel selection of the receive frequency, which is then able to determine the required frequency shift in the transmit mode.

Several NAND-gates are used for logical processes and allow, if required, inverted operation. It is only necessary for the corresponding connection to be grounded. A LED is provided to check whether a frequency shift occurs, which will then be indicated.

A coding switch, with which all 80 channel frequencies on the 2 m band can be selected, is also provided.

Figure 6 shows the frequency selector circuit. A few examples are given to indicate how programming is carried out. In order to achieve a logic »high« level, a diode is connected to the corresponding position of the matrix. A possible programming plan is given in **table 2**.

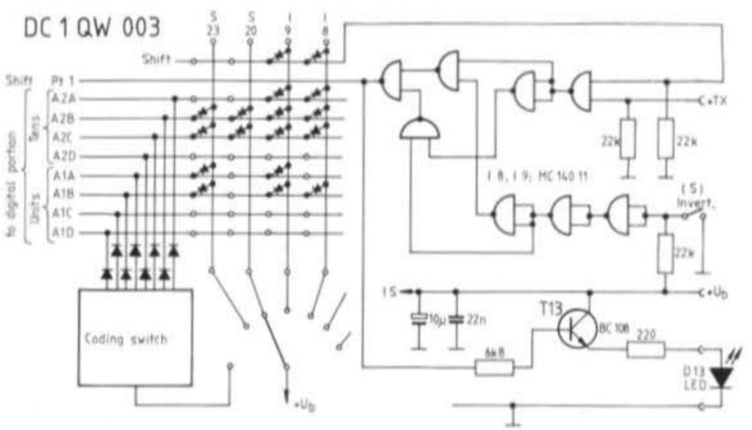


Fig. 6: Frequency selector circuit

Shift	TENS A	B	C	D	UNITS A	B	C	D	Channel No.	Designation	Frequency MHz
L	O	L	L	O	O	O	L	O	64	I 0	145,600
L	O	L	L	O	L	O	L	O	65	I 1	145,625
L	O	L	L	O	O	L	L	O	66	I 2	145,650
L	O	L	L	O	L	L	L	O	67	I 3	145,675
L	O	L	L	O	O	O	O	L	68	I 4	145,700
L	O	L	L	O	L	O	O	L	69	I 5	145,725
L	L	L	L	O	O	O	O	O	70	I 6	145,750
L	L	L	L	O	L	O	O	O	71	I 7	145,775
L	L	L	L	O	O	L	O	O	72	I 8	145,800
L	L	L	L	O	L	L	O	O	73	I 9	145,825
O	O	L	L	O	O	O	O	O	60	S 20	145,500
O	O	L	L	O	L	O	O	O	61	S 21	145,525
O	O	L	L	O	O	L	O	O	62	S 22	145,550
O	O	L	L	O	L	L	O	O	63	S 23	145,575

Table 2

3. CONSTRUCTION

Three PC-boards have been developed for construction of the synthesizer: DC 1 QW 001 for the analog portion **(Figure 7)**, DC 1 QW 002 for the digital portion **(Figure 8)**. The boards are double-coated in order to ensure a good decoupling between the individual components. Since the frequency selector circuit DC 1 QW 003 only possesses DC-voltages, this circuit is built up on a single-coated PC-board as shown in **Figure 9**.

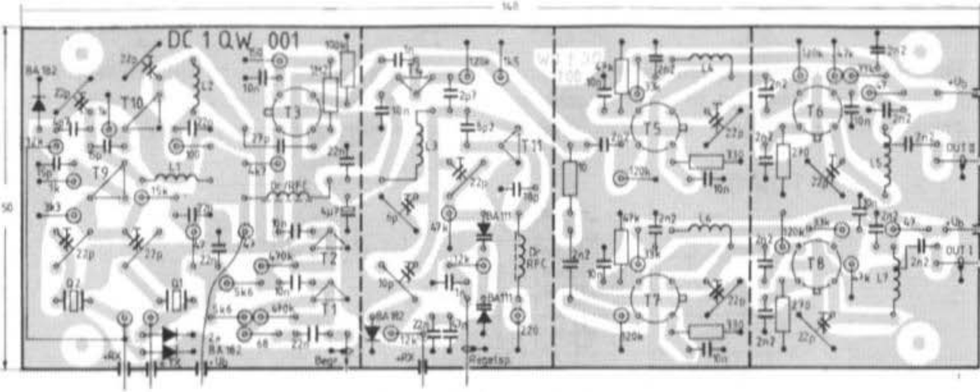


Fig. 7: PC-board DC 1 QW 001 (RF portion)

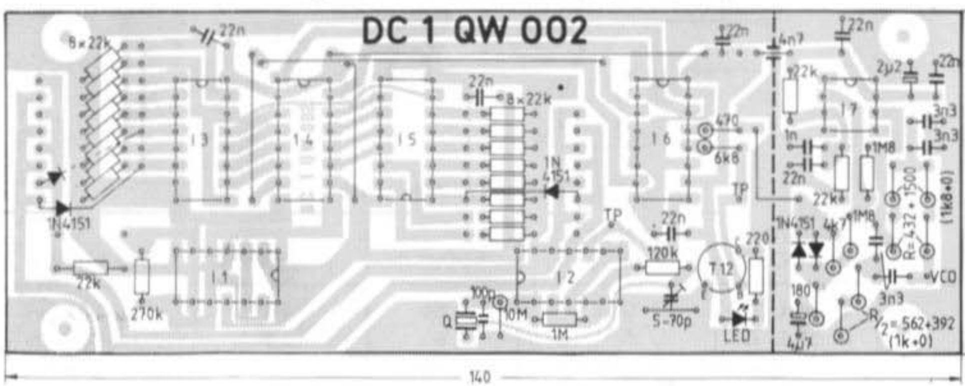


Fig. 8: PC-board DC 1 QW 002 (digital portion)

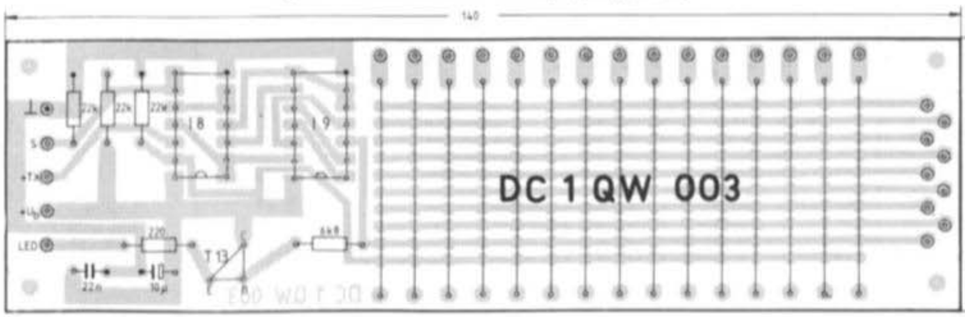
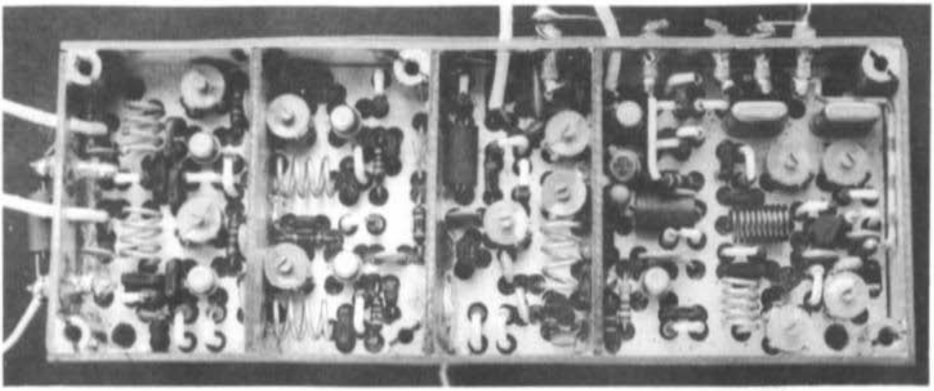
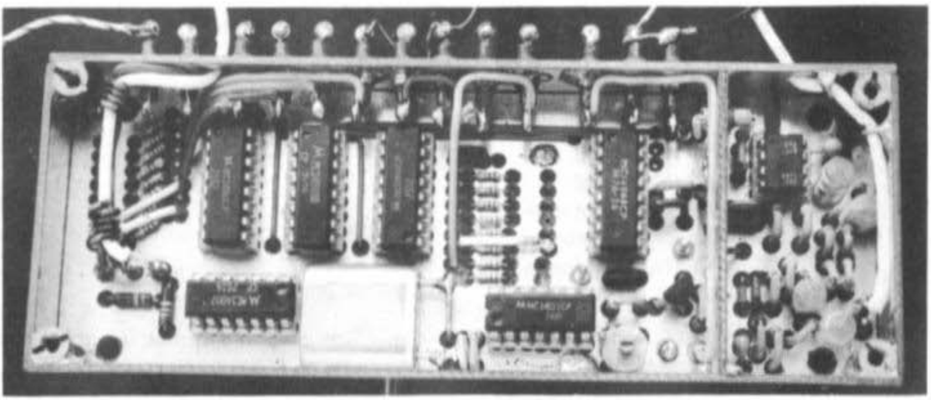


Fig. 9: PC-board DC 1 QW 003 (frequency selector)

All electrical components should be soldered into place with the shortest possible leads, however, attention should be paid that no shorts are made to the ground surface. It is advisable for sockets to be used for the integrated circuits. General instructions regarding working with C-MOS circuits were given in (6). **Figures 10 to 12** show photographs of the author's prototypes.





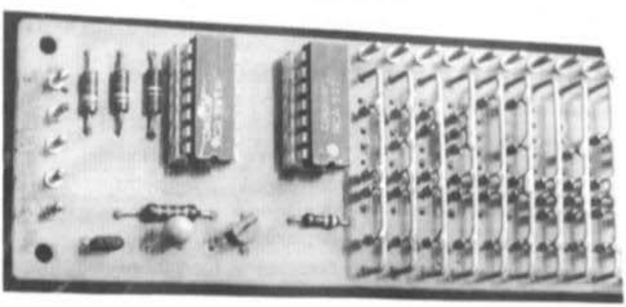


Fig. 10 ÷ 12: Photographs of the author's prototype of the RF-portion DC 1 QW 001 (above), digital portion DC 1 QW 002 (center), and frequency selector DC 1 QW 003

The double-coated intermediate panels from PC-board material are soldered into place before the components are mounted onto the PC-boards. In the author's prototypes these panels have a height of 15 mm. This is followed by soldering the components into place, which is followed by soldering approximately 22 mm wide strips of single-coated PC-board material around the edge of the boards. This results in a very stable construction both mechanically and electrically. Any required feedthrough capacitors are soldered into place afterwards. Threaded bolts can be soldered to the corners of the boxes which then allow covers to be screwed into place.

It is necessary for the control voltage to be fed to the VCO using a screened cable. It is also necessary for a screened cable to be used between the limiter and amplifier (see block diagram). The author used thin 50 Ω coaxial cable such as RG-174 U. It is not necessary for any DC-lines to be screened.

3.1. Special Components

T 1, T 2:	2 N 709
T 3, T 5 - T 8:	40 841 (40 673) or similar DG-MOSFET
T 4, T 11:	BF 245 C
T 9, T 10:	BF 199 or similar
T 12, T 13:	BC 108 or similar

D 1 - D 5, as well as all matrix diodes: 1 N 4148 or similar silicon switching diode

D 6 - D 9:	BA 182 (special switching diode, Siemens)
D 10, D 11:	BA 111, BA 124, BA 125/45
D 12, D 13:	LED (50 mA)
I 1:	MC 14007 (Motorola)
I 2, I 8, I 9:	MC 14011 (Motorola)
I 3, I 5:	MC 14560 (BCP) Motorola
I 4:	MV 14569 (BCP) Motorola
I 6:	MC 14568 (BCP) Motorola
I 7:	741, TBA 221 B (Siemens)
L 1:	10 turns of 0.7 mm dia. enamelled copper wire on 5 mm former, length 10 mm
L 2:	5 turns of 1 mm dia. silver-plated copper wire on 5 mm former, length 8 mm
L 3:	3.5 turns of 1 mm dia. silver-plated copper wire on 5 mm former, length 11 mm
L 4:	4 turns of 1 mm dia. silver-plated copper wire on 5 mm former, length 10 mm
L 5:	4 turns of 1 mm dia. silver-plated copper wire on 5 mm former, length 10 mm

RFC 1, RFC 2: Wideband chokes such as 6-hole ferrite cores

Capacitors in the filter:
2.5 % tolerance (e.g. styroflex)
Resistors in the filter: 1 %

Q 1:	71.65 MHz, HC-25/U, series resonance
Q 2:	66.30 MHz, HC-25/U, series resonance
Q 3:	1.60 MHz, HC- 6/U, parallel resonance, load 30 pF

All trimmers: plastic foil trimmers (Philips)
Coding switch: BCD-code, positive logic

4. ALIGNMENT

A frequency counter with a frequency range up to at least 150 MHz should be available for the alignment. A voltmeter with RF-probe or a simple RF-indicator are also required.

After connecting the individual modules together, the operating voltage of + 10 V to + 12 V (it need not be stabilized) is connected. The 1.6 MHz crystal oscillator in the digital portion is now aligned to its nominal frequency with the aid of the trimmer capacitor. This means that this module is completely aligned.

In the RF-portion, the transmit crystal oscillator (+ TX) is switched into operation. The collector resonant circuit is now aligned to the frequency of the crystal Q 1 (71.65 MHz). Finally, the operating voltage is switched from + TX to RX. This switches on crystal Q 2 (66.3 MHz), as well as an additional capacitor in the resonant circuit. A value of 4.7 pF is only for orientation, but is usually suitable.

After confirming that the crystal oscillators are working correctly and can be switched, the output resonant circuit of the frequency doubler stage is aligned to a frequency that is approximately between the two doubled frequencies (appr. 138 MHz). This alignment is uncritical.

For alignment of the VCO, the varactor diodes are fed with a variable DC-voltage of between approximately 2.5 V and 8 V from a potentiometer instead of the control voltage line. Firstly, connection + RX is not connected and the trimmer (1.4 - 10 pF) switched by the diode is aligned to its minimum capacitance value. The frequency counter is connected to the output of one of the oscillator amplifiers. The VCO frequency is adjusted by varying the band trimmer (1.2 - 6 pF) and the trimmer capacitor (2 - 22 pF) via which the varactor diodes are connected so that a frequency variation of approximately 144 MHz to 146 MHz is obtained when the varactor diodes are provided with a voltage of 2.5 V to 8 V.

Finally, the connection + RX is connected to the operating voltage. The trimmer (1.4 - 10 pF) that is now switched into circuit, is now aligned so that the frequency of the VCO amounts to 133.3 MHz with a varactor diode voltage of 2.5 V.

After completing this coarse alignment, the varactor diodes are connected to the control voltage line by connecting them to the filter. The oscillator should now synchronize securely at all transmit and receive frequencies when the coding channel switch is switched to any digit between 00 and 80. Functional checks can be made by monitoring the transmit signal in a receiver, and is also indicated on the LED D 12. If this LED does not light up to full brightness, a fine alignment should be made of the VCO. Attention should be paid that the control voltage does not exceed the limit values of approximately 2.5 V and 8 V.

Finally, the oscillator amplifiers are aligned for maximum output power. The center frequencies are 145 MHz for the transmitter and 134.3 MHz for the receiver.

If the output signals are not to be fed to consumers having an impedance of 50 Ω, and when a higher output voltage is required, it is possible for the output tap on L 5 or L 7 to be placed further towards the hot end of the output resonant circuit. It would also be possible for an inductive coupling to be used. Sufficient output power will be provided in all cases.

The synthesizer can be modulated by feeding an audio signal to the control voltage connection via a coupling capacitor. In order to achieve a frequency deviation of ± 3 kHz, a mean value of 15 mV voltage variation (peak-to-peak) is required at the varactor diodes. The transmission characteristics are dependent on the parameters of the control circuit and, of course, on the microphone and AF amplifier. However, this is not to be discussed in detail here.

5. MEASURED RESULTS

The output spectrum of the synthesizer was examined in a frequency range of 100 kHz to 1.5 GHz. When using a good screening of the modules, and good, low-impedance ground connections, as well as carefully blocked supply voltage lines, it was possible to achieve a spurious signal rejection of 75 dB throughout this frequency range. Even when construction is made less carefully, a spurious signal rejection of 65 dB can be expected. The following harmonic suppression values were measured: First harmonic: 32 dB, second harmonic: 50 dB, third harmonic: 40 dB, all higher-order harmonics were at least 50 dB down. It is especially favorable that the phase-comparator frequency does not generate any side lines, that are not suppressed by at least 75 dB with respect to the required signal.

On channel selection, the oscillator must firstly jump to the new frequency. This transient process is extremely short when using the described circuit. A transient time of only 0.5 ms was measured in practice.

The output power of the synthesizer amounts to 14 mW. This means that an output voltage of 836 mV is available at an impedance of 50 Ω. Correspondingly more voltage can be provided at higher impedances.

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DIODE APPLICATIONS IN FREQUENCY MULTIPLIERS FOR THE MICROWAVE RANGE

by H. Fleckner, DC 8 UG

Frequency multipliers are used at higher frequencies when the required frequency cannot be generated at the final frequency with the required power and frequency stability. This is usually possible at one of its subharmonics. If a crystal-controlled signal is required in the X-band (amateur radio band 10.0 - 10.5 GHz), this can be obtained by frequency multiplication and, if necessary, also amplification, of the RF signal from a crystal oscillator.

In principle, any two-pole with non-linear characteristic will generate harmonics of the original oscillation. However, in the case of non-linear resistors, this energy will be dissipated so that only a small portion of the fundamental wave power will be converted into harmonics. Under favorable conditions, the multiplication of f to n x f, the efficiency is $\eta = 1/n^2$ (1).

If non-linear energy storage is used for frequency multiplication, it is possible, in the ideal case, to do this at no loss. If the power P at the frequency f is fed to such an energy storage, and only harmonic power P_n is taken at n x f, $P_n = P$ must be valid due to the energy conservation, since the power distribution law according to Manley-Row simplifies the energy conservation when only one frequency is used for excitation. PN varactor diodes are especially suitable for frequency multiplication because their construction is simple, and the losses are very low right up to the higher frequencies.

The following can be used as non-linear energy storage:

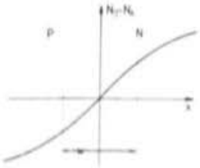
1. The drive-dependent junction capacitance (junction varactor)
2. The charge storage in flow direction (storage varactors), or
3. Both effects together (bi-mode diode).

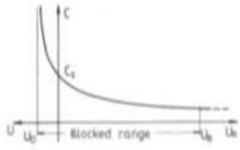
Furthermore, it is possible to use PIN structures as storage diodes, and Schottky diodes as junction varactors for frequency multiplication applications. All these types of energy storage with their special applications are to be discussed together with examples in the following sections.

1. FREQUENCY MULTIPLICATION USING JUNCTION VARACTORS

The junction varactor can be represented in an equivalent circuit diagram as a series circuit of non-linear capacitance and conductor resistance R. The capacitance-voltage characteristic and the characteristic curves of virtually any symmetrical PN-junction is shown in **Figure 1**. The conductor resistance has a direct effect on the cut-off frequency f_c of the diode and should be as low as possible for low-loss applications. The concentration curve of the junction determines the number and amplitude of the generated harmonics. The characteristic exponent γ is important for this together with the relationship between the standardized charge and voltage drive. Principally speaking, one differentiates between two types of junction varactors:

1. The abrupt PN junction with $\gamma = 1/2$ and
2. The linear PN junction with $\gamma = 1/3$





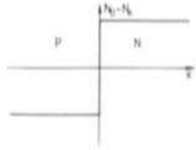


Fig. 1a:
Symmetrical concentration curve at the PN-junction, junction width W, donator concentration N_D and acceptor concentration N_A

Fig. 1b:
Capacitance-voltage characteristic of a junction varactor having

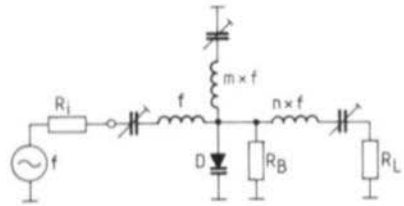
$$C = \frac{C_0}{\left(1 - \frac{U}{U_D}\right)^{\gamma}}$$

Fig. 2:
Abrupt PN-junction (γ = 1/2; u = q^2)

1.1. Abrupt PN-Junction

Figure 2 shows the curve of this junction. The donator and acceptor concentration is homogenous up to the dose limit. The result of this is a **square** relationship between the voltage and the charge drive in the blocked range. This means that it is only possible to double using these varactors ! A basic circuit is shown in **Figure 3**.

Fig. 3:
Principle circuit diagram of a frequency multiplier having an idler circuit for any harmonic higher than n > 2



If higher harmonics are to be generated, it is necessary to provide reactive circuits. In the case of a tripler, the reactive current is generated in the varactor at the frequency of the second harmonic, which then mixes with the fundamental wave to produce 3 f = 2 f + f. Two reactive circuits are required for a quadrupler, so that the fourth harmonic is obtained by mixing the third harmonic to the fundamental wave.

1.2. Linear PN-Junction

In the case of a linear junction, there is a proportional concentration curve at the dose limit (Figure 4). The characteristic exponent γ is 1/3 which results in $q^{3/2}$-relationship between voltage and charge.

The Fourier analysis shows that power can be produced at higher-order harmonics using such diodes directly and without reactive circuits. However, when using reactive circuits, it is possible for the efficiency to be increased considerably due to the improved energy utilization. This is valid generally for all characteristic exponents γ. The efficiency of frequency triplers with linear and abrupt junction as a function of the ratio of fundamental wave frequency to cut-off frequency is given in **Figure 5**. The abrupt varactor is approximately 5 % better than a linear varactor with the same cut-off frequency, (2).

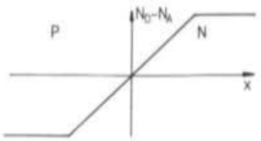


Fig. 4:
Linear PN-junction
(γ = 1/3; u = $q^{3/2}$)

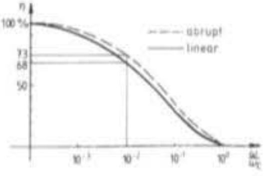


Fig. 5:
Efficiency of a frequency multiplier with abrupt or linear PN-junction as a function of ω/ω_C.
ω_C = Cut-off frequency of the diode

1.3. Summary

In the desire to obtain the highest possible efficiency, the following design and selection critera result:

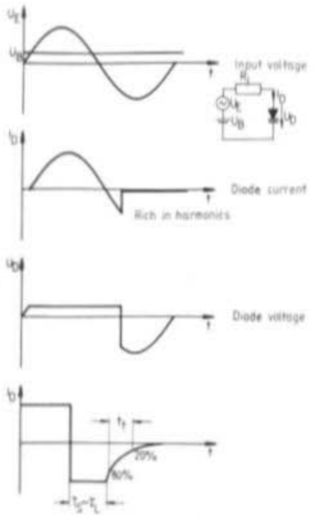
1. Cut-off frequency should be large in comparison to the fundamental wave ($f_C \geqq 100$ f)
2. Always provide reactive circuits (at $n \geqq 3$)
3. Junction varactors are advantageous at low frequency multiplication ratios ($n < 4$) and at high output frequencies ($n \times f \geqq 10$ GHz)
4. The input power should not be much greater than 1 W, since the efficiency will fall when overdriven.
5. The abrupt junction is advisable since the junction capacitance can be kept low. The characteristic exponents can be found in the detailed data books.

If higher powers at high efficiencies are to be multiplied up to a frequency of approximately 12 GHz, it is possible to use a diode that is driven into the flow range: storage, step-recovery, or snap-off diode.

2. STORAGE DIODES IN FREQUENCY MULTIPLIER CIRCUITS

When one drives a PN junction from the blocked into the flow range, a further storage mechanism is exhibited in addition to the junction capacitance.

The diffusion capacitance caused by the increase in concentration of the minority carriers in the conductive area of the N-semiconductor allows a capacitive behaviour of the diode. Whereas this is an undesirable effect in the case of switching diodes, this characteristic is desirable in the case of storage diodes and is increased technologically using an overlapping concentration that produces a drifting field having an effect on the minority carriers. The stored charge carriers must be removed on switching into the blocked range in order to ensure that the diode blocks. If the charge carriers are removed, the diode current will jump abruptly to its static, blocked current I_r. This results in current pulses having a high harmonic content such as can be seen in a Fourier series expansion. **Figure 6** shows the switching behaviour of these diodes. One can differentiate between two characteristic times which determine the frequency range. The first is the storage time, which approximates the life of the minority carriers and the second is the switching time or snap-off time. The dose profile of PSN storage diodes is given in **Figure 7**, S designates a small dose.



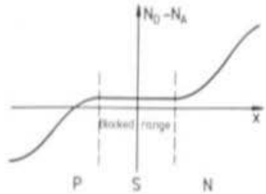


Fig. 7: Concentration curve of a PSN-structure

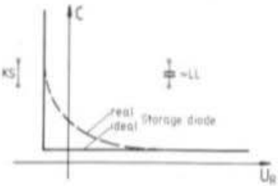


Fig. 8: Capacitance-voltage characteristic of a real (γ = 0.5 - 0.1) and an ideal (γ = 0) storage diode

Fig. 6: Switching behaviour with sine-wave or square-wave drive of a storage diode
T_S = Storage time ~ life t_L
t_t = Snap-off time

The best storage diodes can, however, be constructed using PIN-structures since the junction of these is virtually conductive. All diodes manufactured today are PIN-diodes, although one has traditionally kept this designation for diodes that are used exclusively as switching diodes or RF-resistors. The capacitance-voltage characteristic of the storage diode is given in **Fig. 8**. An ideal storage diode has a voltage-independent capacitance curve, whereas the real storage diode exhibits characteristic exponents of $\gamma = 0.5 - 0.1$.

2.1. Frequency Range, Power and Efficiency

Storage diodes provide all harmonics of the fundamental wave whereby the Fourier coefficients mean that the higher-order harmonics are especially present. This means they are especially suitable for circuits having a high frequency multiplication ratio ($n > 4$). Whereas only losses in the conductor are present with the ideal diodes, and no limitation in the frequency range is present, a real diode can only operate at high efficiency over a certain frequency range. The recombination and switching losses limit the applicational range to 50 MHz to 25 GHz.

The lower frequency limit is determined by the storage time which is not infinite. The upper frequency limit of the individual diode is determined by the switching or snap-off time. The range is usually listed by the manufacturer. If this is not the case, it can be estimated as follows:

Lower frequency limit: $10 \leqq 2\pi f_u \times t_L < 100$ (with $\eta = \eta$ max) ($t_L \sim T_S$)

Upper frequency limit: $T_{RF} > t_t \rightarrow f_o \sim \frac{1}{t_t}$

The following table gives the efficiency levels of several frequency multipliers of factor eight and factor three having different reactive circuit frequencies [see (2)]. The construction of storage diode multipliers is the same as that of junction varactors, but the operating point should be adjustable.

n = 8	m = 0	m = 2	m = 6	(m = idler circuit harmonic)
-	17.3 %	23.3 %	25 %	storage diode
n = 3	75 %	84 %	-	storage diode
-	0 %	73 %	-	abrupt PN varactor

with $f/fc = 10^{-2}$

2.2. Summary

Storage diodes are especially suitable for circuits having a high frequency multiplication ratio, and high input power level ($n > 4$, $P > 1$ W). The efficiency can be improved considerably when using a reactive circuit. A large number of reactive circuits should, however, be avoided at high output frequencies since the impedance at the diode could become too low, and cause matching losses.

2.3. Practical Realization of a Frequency Multiplier from 1296 to 10368 MHz (x 8) using a Storage Diode

Gunn oscillators exhibit poor short-term and long-term stability. The moderate short-term stability is observed as a high FM-noise level which will lead to low signal-to-noise ratios of receive and transmit systems used in the amateur range of the X-band. This means that narrow-band stations operating with crystal control are 40 to 50 dB better than the unstabilized Gunn oscillator systems.

Narrow-band frequency multiplier circuits comprising several frequency multipliers one after another are, in principle, possible, but are difficult to construct for radio amateurs. The reasons for this are: Varactor diodes are two-poles that require a low-reflection termination for correct operation. If isolators are not used between circuits, it will not be possible to satisfy this demand without using a spectrum analyzer, even when only two varactor frequency multiplier stages are connected together. This will become worse and worse with the number of frequency multiplier stages that are connected. All impedances of a varactor are greatly dependent on the drive level. Any incorrect matching will cause reflected power which will radically alter the impedance relationships of the previous stage, which usually causes completely unstable operation. This leads to a non-permissible noise level at the output, or a thermal destruction of the power amplifier driven by the varactor chain.

This means that such chains of frequency multipliers can only be constructed using expensive isolators, and when they can be aligned using a swept frequency generator and spectrum analyzer.

The same result can be achieved far easier when the characteristics of storage diodes are used where higher-order harmonics are accentuated in the frequency multiplication process. Such a circuit is shown in **Figure 9** using printed lines on a PTFE board. The low-frequency portion uses discrete circuit elements (Giga trimmer), whereas the high-frequency circuits are in the form of striplines that provide selectivity and transformation. The reactive circuit is in resonance for the sixth harmonic. The varactor impedance is matched using a transforming line onto which capacitances are provided at suitable positions. The theoretical efficiency has been calculated to be 24 %, which means that efficiencies of 15 to 20 % can be obtained in practice. If the same diode, e.g. DH 256 or DH 292 (Thomson-CSF) is used, hardly any alignment difficulties will be presented. The subsequent X-band waveguide represents a high-pass filter that will attenuate the lower frequency harmonics.

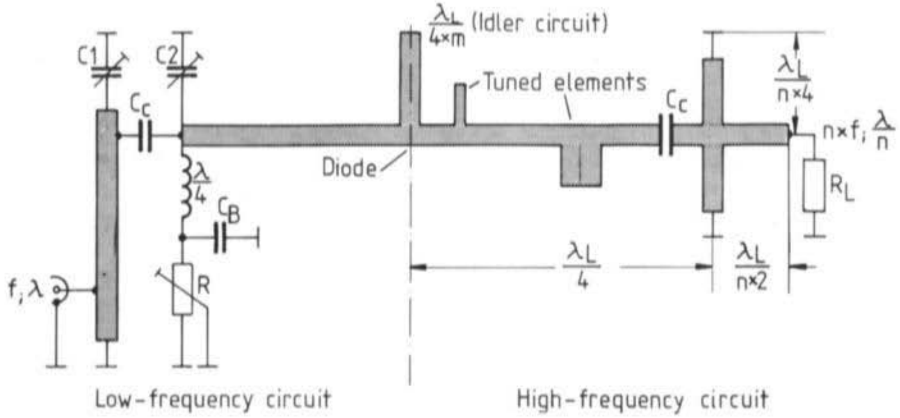


Fig. 9: Circuit diagram of a frequency multiplier (x 8) with storage diode and idler circuit at m = 6

3. TRAVELLING WAVE MULTIPLIER USING STEP-RECOVERY DIODES

The travelling wave multiplier (3) is a relatively unknown frequency multiplier circuit that is especially suitable for high frequency multiplication ratios ($n > 10$). The circuit diagram in **Figure 10** shows the principle of operation.

The PIN diode represents a switch which switches off the current flowing via inductance L so that a high self-induction voltage is generated at the diode due to the inductance and the low series impedances. The slope of these can be influenced by the selection of L and C. A mis-matched $\lambda/4$ line is now fed with this impulse. A damped sine-wave oscillation will be generated across the load resistor having a frequency of $f = n \times f_0$, which will reappear after a delay of $4 \times \lambda/4$, **(Figure 11)**. If a bandpass filter of frequency f is coupled to this line, one will obtain a selective carrier signal. The advantage of this circuit is that no idler circuits are required, and it is thus suitable for frequency multiplication at high-order harmonics.

Figure 12 shows the circuit of a frequency multiplier from 144 MHz to 2304 MHz (x 16) which provides a selectively measured efficiency of 5 %, and a sideband suppression of 22 dB.

Fig. 10:
Principle circuit diagram of a travelling wave frequency multiplier.
C compensates for the inductive input impedance

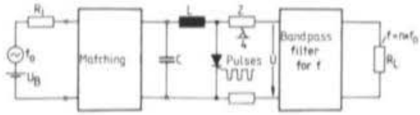


Fig. 11:
Voltage characteristic at the end of the travelling wave line

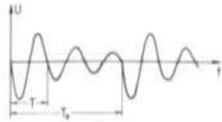
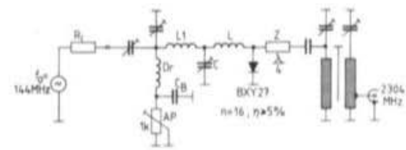


Fig. 12:
Circuit diagram of a frequency multiplier from 144 MHz to 2304 MHz



4. SCHOTTKY-DIODE MULTIPLIERS

A Schottky diode operates similar to a point contact diode and has a metal semiconductor junction that possesses the following advantages and disadvantages with respect to a PN-junction:

Advantages:

Only a N-conductive area : high cut-off frequency

Exhibits considerable capacitive-voltage characteristics with low capacitance values $\gamma = 1/2$ (abrupt)

Ga-As Schottky diodes operate up to the mm wave range

Disadvantages:

Low breakdown voltage: low power

No storage effect, only operation in the blocked range possible

Schottky-diode multipliers are built up in the same manner as PN-junction varactor circuits.

5. OVERALL SUMMARY

The criteria for selection of the diodes in the most favorable frequency multiplier circuit are: power, frequency multiplication ratio, frequency, but also the price. PN-junction varactors should be used at low frequency multiplication ratios ($n < 4$), medium power ($P < 1$ W), and high frequency ($f > 10$ GHz).

Storage diodes or PIN diodes should be used for high frequency multiplication ratios ($n > 4$) and high power ($P > 1$ W) in the frequency range of 50 MHz to 25 GHz.

Schottky diodes are suitable for low power levels ($P < 0.5$ W), high frequencies and low frequency multiplication ratios ($n < 4$).

5.1. Some suitable Diode Types

5.1.1. Bi-mode Diodes

These can be used as storage and junction varactors.

Type	Dissipation Power	Transit Frequency	Made by
BAY 96	10 W	> 34 GHz	Valvo
1 N 4885	10 W	> 34 GHz	Valvo
1 N 5152	5 W	> 55 GHz	Valvo
1 N 5153	5 W	> 55 GHz	Valvo
1 N 5155	3 W	> 100 GHz	Valvo
1 N 5157	2,5 W	> 180 GHz	Valvo
BXY 27	4 W	100 GHz	Valvo
BXY 28	3,5 W	100 GHz	Valvo
BXY 29	1 W	120 GHz	Valvo
VAB 811 EC	10 W	100 GHz	Varian

5.1.2. Storage diodes
Manufacturer: Philips

Type	Dissipation Power (Heat resistance)	Transit Frequency f_c	Switching time t_t	Storage time t_s
BXY 92	1 W	150 GHz	$\leqq$ 150 ps	50 ns
BXY 35	$R_{th\,G} \leqq$ 10 grd/W	$\geqq$ 25 GHz		
BXY 36	$R_{th\,G}$ = 20 grd/W	75 GHz	$\leqq$ 500 ps	150 ns
BXY 37	$R_{th\,G}$ = 20 grd/W	100 GHz	$\leqq$ 350 ps	100 ns
BXY 38	$R_{th\,G}$ = 30 grd/W	120 GHz	$\leqq$ 300 ps	75 ns
BXY 39	$R_{th\,G}$ = 40 grd/W	150 GHz	$\leqq$ 200 ps	50 ns
BXY 40	$R_{th\,G}$ = 50 grd/W	180 GHz	$\leqq$ 150 ps	50 ns
BXY 41	$R_{th\,G}$ = 50 grd/W	200 GHz	$\leqq$ 100 ps	25 ns

5.1.3. PIN Storage Diodes
Manufacturer: Thomson-CSF

Type	P_{out}	Frequ.range	t_t	t_s
DH 294	0,5 W	0,2 - 2 GHz	400 ps	125 ns
DH 200	20 W	0,5 - 2 GHz	1000 ps	250 ns
DH 270	15 W	2 - 3 GHz	700 ps	160 ns
DH 110	9 W	2 - 4 GHz	400 ps	100 ns
DH 293	6 W	3 - 6 GHz	250 ps	60 ns
DH 252	3 W	2 - 8 GHz	200 ps	95 ns
DH 256	2 W	5 - 12 GHz	120 ps	20 ns
DH 292	0,6 W	8 - 16 GHz	75 ps	10 ns
DH 267	0,2 W	10 - 25 GHz	60 ps	6 ns

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INTERDIGITAL CONVERTERS for the GHz AMATEUR BANDS

Interdigital Filters Extended to Form Receive Converters

by J. Dahms, DC 0 DA

It is just as important on the GHz amateur bands as on the lower bands that the converters do not only have sensitivity and gain but also selectivity. Filters suppress interference caused by unwanted signals, and increase the sensitivity by suppressing the noise component of the image frequency (1). Of course, selectivity cannot be obtained in the microwave range without mechanical construction. Interdigital filters are simple to construct, are of high quality electronically and are extremely reproducable. Such interdigital filters have been described several times in VHF COMMUNICATIONS (2), (3). It is possible for them to be connected in front of an available wideband converter constructed in PC-board technology.

The overall construction is very simple when using the interdigital converter described here: This is virtually an interdigital filter which has been extended to form a receive converter by installing a mixing diode. This part of the article is based on a publication given in (4). When equipped with a suitable local oscillator module, and a very low-noise IF preamplifier, complete converters result, which are to be described for the GHz amateur bands at 23 cm, 13 cm, and 9 cm.

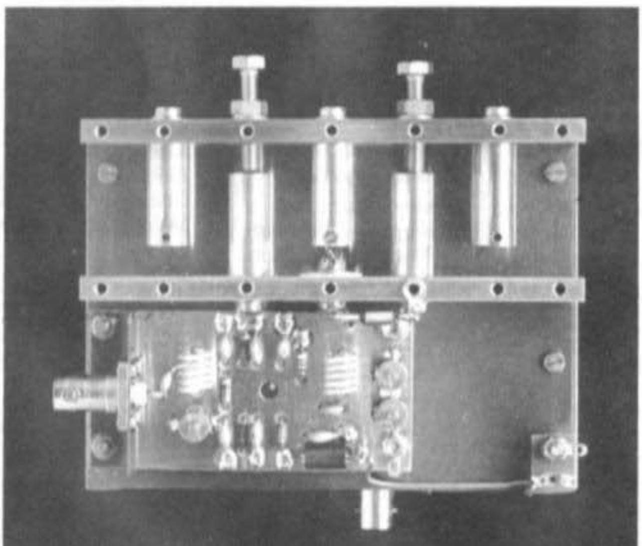


Fig. 1: Interdigital 13 cm converter, without cover

below: IF-preamplifier

The photograph given in **Figure 1** shows the principle of operation in conjunction with the 13 cm converter: The antenna is connected to the hot end of the right-hand resonator – this finger does not contribute to the selectivity. The second resonator from the right is tuned to the input frequency. The mixer diode is connected to the third (center) finger – which means that this finger will not possess any noticeable resonance. The left-hand finger is connected to the local oscillator – also at the hot end, which means that no resonance is present either. Finally, the fourth resonator from the right is aligned to the oscillator frequency. Although virtually only one resonator is effective for each of the two frequencies, it will be seen from the specifications that a sufficient image frequency rejection results.

The converters were, however, designed for an intermediate frequency of 144 to 146 MHz.

It may be of interest that the converters were not developed according to increasing frequency, e.g. firstly for 23 cm, then 13 cm etc. The first of the converters to be made was the 9 cm converter. The good specifications of this converter led to the construction of the other two converters using the experience gained on 9 cm. This resulted in a system of three virtually identical, selective converters for the three microwave bands of 23 cm, 13 cm and 9 cm.

1. SPECIFICATIONS

Table 1	23 cm	13 cm	9 cm
Input frequency	1296-1298	2304-2306	3456-3458
Intermediate frequency (MHz)	144-146	144-146	144-146
Oscillator frequency (MHz)	1152	2160	3312 (1104)
Noise figure (SSB)	8 dB	8.5 dB	11 dB
Overall gain	18 dB	18 dB	20 dB
3 dB bandwidth	3.5 MHz	7.5 MHz	21.8 MHz
Image frequency suppression	26 dB	24 dB	24 dB
Mixer diode	HP 5082-2817	HP 5082-2579	HP 5082-2579
Mixer diode current	1 mA	1 mA	0.8 mA
Multiplier diode	–	–	HP 5082-2835
Oscillator power	1 mW	1 mW	30 mW (1104 MHz)
IF transistor	BF 900	BF 900	BF 900

2. CONSTRUCTION OF THE CONVERTER

The construction of the interdigital converter is similar to that described in (4) where converters for 23 cm and 13 cm with intermediate frequencies of 144 MHz and 28 MHz were described. The components used, however, were varied considerably from the original so that a converter resulted which could be realized easily in practice. **Figure 2** shows the principle of the circuit in the form of a diagram. It shows how the local oscillator frequency is tripled in the converter using diode D 2, however, this is only valid for the 9 cm converter. In the case of the other two converters the local oscillator frequency is generated externally up to the final frequency where it is fed to the first resonator from the right via a connector (13 cm converter), or via a low-capacitance feedthrough (23 cm). The construction shown in Figure 2 is less favorable, since only one resonator provides selectivity, and the mixer diode will be provided with spurious signals at too high a level. These could cause a noticeable deterioration of the noise figure.

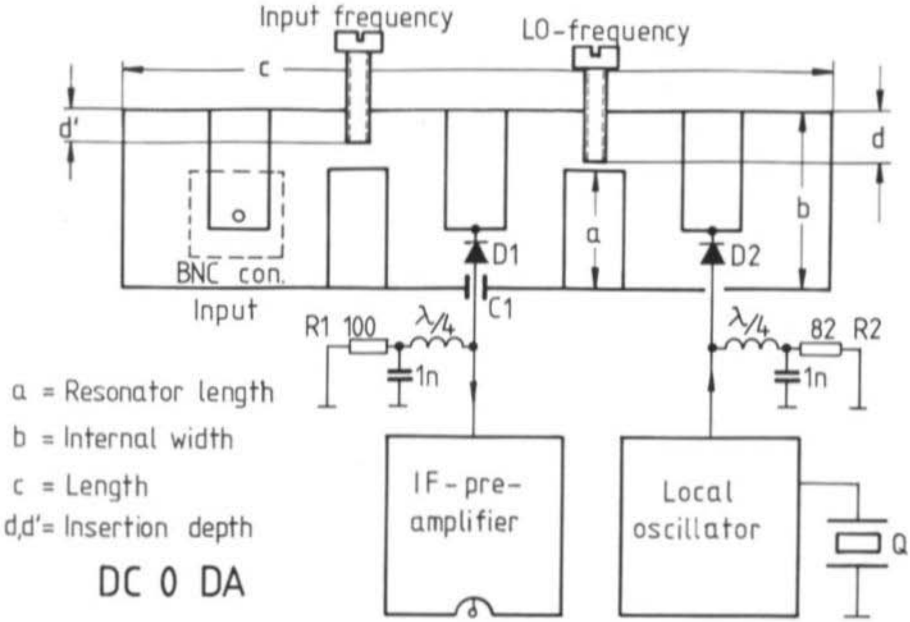


Fig. 2: Principle of the converter; variable dimensions given in Table 2

The dimensions to Figure 2, that vary from band to band, are given in **Table 2**.

Wavelength	Dimensions (mm)					Recommended Mixer Diode
	a	b	c	d	d'	
23 cm	51	57	120	12.5	8.5	HP 5082-2565
13 cm	25	32	120	8.0	6.5	HP 5082-2565
9 cm	16	24	120	6.5	4.5	HP 5082-2565

The spacings between the individual resonators, their diameters, as well as the height of the filters are identical for all three versions. These dimensions are given in **Figure 3**. The side panels are made from brass plate of 20 mm in width and 6 mm thick, which is readily available in hobby shops. Copper tubing of 10 mm outer diameter and 1 mm thick, is used for the resonators.

The two side panels are marked according to Figure 3, drilled, and provided with the required threads. After this, they should be polished, and if possible, silver-plated. The copper tubes are now cut to the required length with the aid of a saw, and both ends are filed until they are straight. This can be achieved by placing them in a bench drill (without denting or scratching them) and polishing the ends with fine Emery cloth at high speed. Hextagonal brass nuts with 4 mm thread are soldered to one end. It has already been explained in previous articles that the nuts should not be placed directly at the end of the tube but slightly inside so that a good contact is made on the outer edge on screwing into place. A completed resonator for the 9 cm converter is shown in the lower part of **Figure 4**. It is also advisable for these parts to be polished, and subsequently silver-plated, if possible.

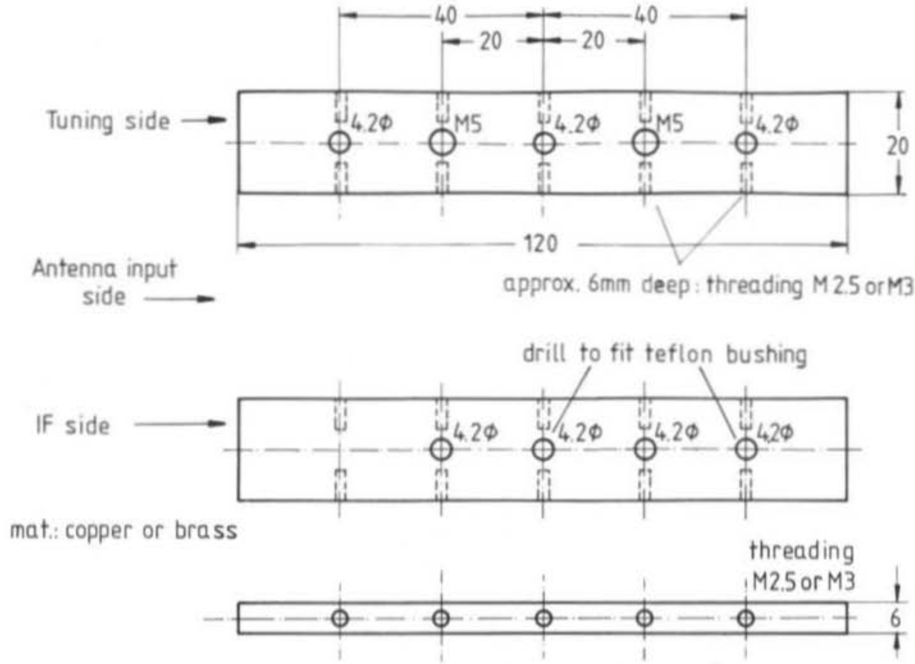


Fig. 3: Dimensional drawings of the side panels; dimensions the same for all versions

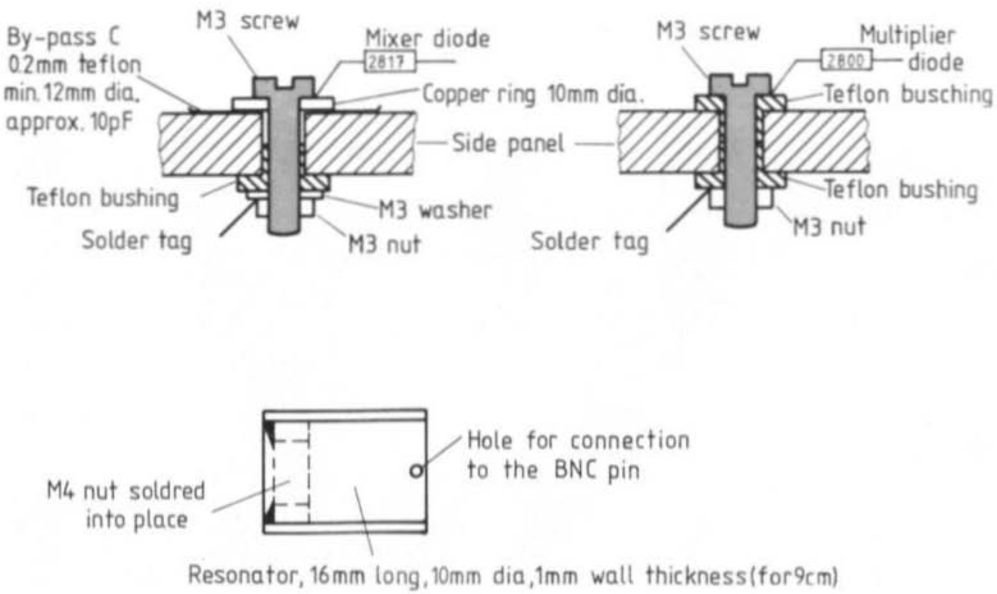


Fig. 4: The feedthroughs for IF and oscillator frequency, as well as the resonators

The base plate of the converter can be made from copper-coated PC-board material. The dimensions should be selected so that sufficient room is available for the IF preamplifier besides the converter. The local oscillator module is, on the other hand, accommodated on the rear panel, which can be seen in the following photographs. This is followed by drilling 2 x 5 holes for the 2 mm or 3 mm screws in the base plate having a spacing of b + 6 mm.

The converter cover has dimensions of (b + 12 mm) x c and is made from brass or copper plate of 0.5 to 1 mm thickness. It is provided with holes for one BNC connector or two such connectors for the 13 cm version.

It is not necessary for the narrow sides of the converter to be sealed since no fields are present here because the resonators are virtually short-circuited with 50 Ω. It is possible to solder the pin of the BNC connector to the input coupling through these apertures.

It should now be described how the mixer diode and, in the case of the 9 cm converter, the multiplier diode are to be installed, since these are located in critical positions. These two positions are shown in Figure 4 where the diodes, and especially their connections are only sketched roughly. The insulating piece designated »teflon bushing« in Figure 4 may also be made from Nylon or Trovidur. The bypass capacitor for the mixer diode (approx. 10 pF) is constructed from a brass or copper disk of approximately 10 mm diameter and a disk of approximately 0.2 mm thick PTFE-foil (Teflon). The screw on the inside is cold with respect to UHF, and only the IF is tapped off. The diode connections should be kept as short as possible to the bypass capacitor (where it is clamped or soldered carefully) and to the edge of the center resonator. A small screw of 2 mm or 2.6 mm diameter can be used at this position.

The mounting of the frequency multiplier diode in the 9 cm converter can be made in the same manner, however, there is no bypass capacitor. In this case, a Teflon bushing is also used on the inside for insulation.

Figure 5 shows a photograph of the three converter versions in their completed, ready-to-operate condition. **Figure 6** shows converters for 23 cm and 9 cm with the cover of the inter-digital filter removed.

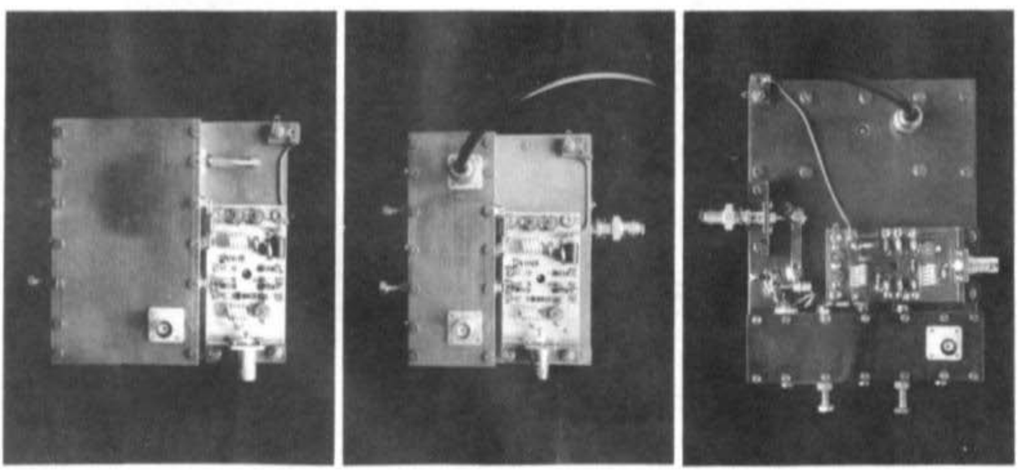
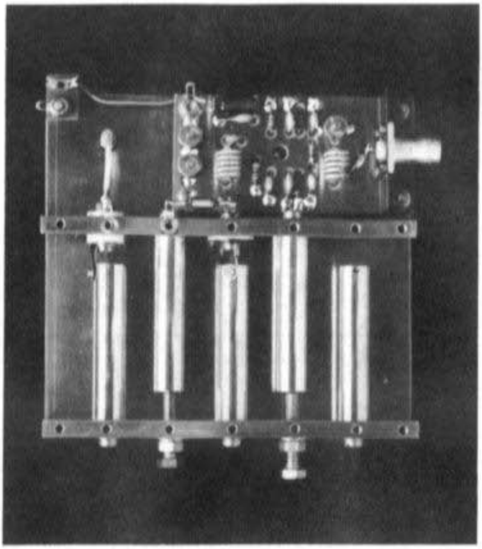


Fig. 5: Prototype converters for 23 cm, 13 cm, and 9 cm



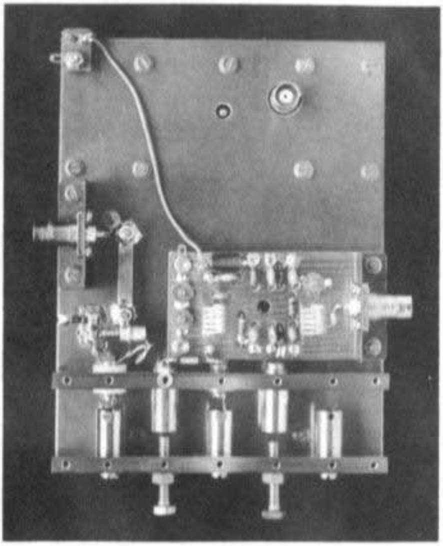


Fig. 6: Converters for 23 cm and 9 cm with the covers removed

Before commencing the alignment, further details are to be given regarding the IF-preamplifier and the local oscillator module.

3. IF-PREAMPLIFIER

The circuit diagram of a well-proved IF-preamplifier is shown in **Figure 7**. This circuit is equipped with a dual-gate MOSFET, type BF 900. This transistor possesses far better specifications with respect to noise figure and gain than conventional dual-gate MOSFETs due to its modern technology of ion implantation, nitrid passivation, and the use of a stripline case. It is possible to achieve a noise figure of 1.3 dB at 145 MHz and a gain of approximately 20 dB (in front of a converter having F = 2.6 dB; measured by DL 3 WR). The DC-operating point of the transistor is stabilized according to recommendations by DJ 4 BG.

In the application in question, the matching between the mixer diode and the input of the BF 900 is the most critical point. A Pi-filter is used to obtain the required impedance transformation. The mixer diode current can be measured as voltage drop across resistor R 1. The RF-voltage is isolated from this resistor using a low-loss $\lambda/4$ air-spaced choke. C 1 is the home-made bypass capacitor in the converter.

The following is valid for the two inductances in the IF-preamplifier:

5 turns of 1 mm diameter silver-plated copper wire wound on a 6 mm former, self-supporting. The $\lambda/4$ choke is made from 50 cm enamelled copper wire of approximately 0.5 mm diameter, using a former of approximately 3 mm diameter. The choke in the supply line is a 6-hole ferrite choke manufactured by Philips. The PC-board for this preamplifier, which can also be used for other applications, is given in **Figure 8**.

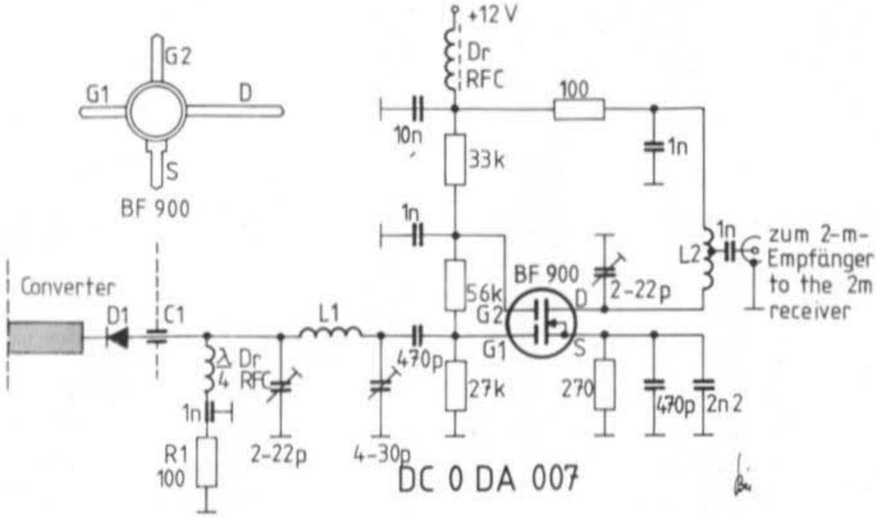


Fig. 7: Circuit diagram of the low-noise IF-preamplifier

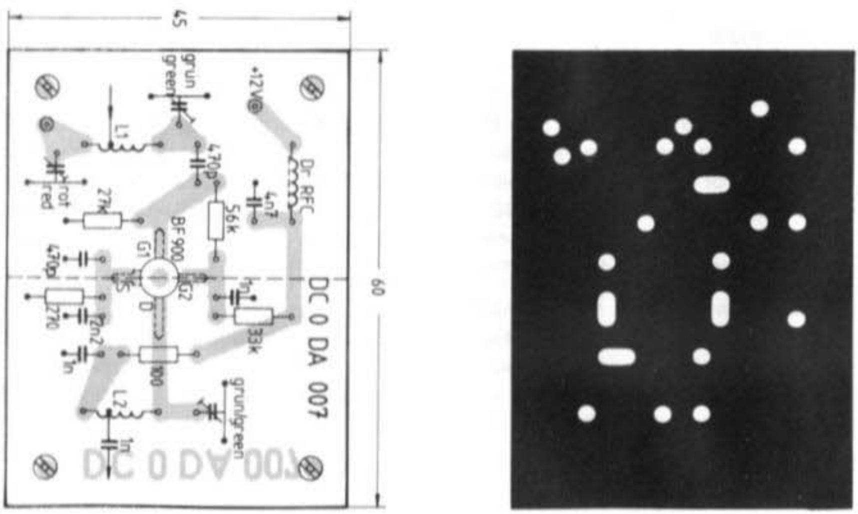


Fig. 8: PC-board DC 0 DA 007 for the 144 MHz preamplifier shown in Figure 7

4. LOCAL OSCILLATOR MODULE

It is not the intention of this article to add another local oscillator module to the numerous descriptions, or to add anything to them. **Table 3** is only to give a list of the frequencies, frequency multiplication steps and the semiconductors that have proved themselves in the author's experiments.

Oscillator for the 23 cm band

Crystal	96 MHz	BF 173
x 3	288 MHz	2 N 5179
x 2	576 MHz	BFW 92
x 2	1152 MHz	BFW 92

Oscillator for the 13 cm band

Crystal	90 MHz	BF 173
x 3	270 MHz	2 N 5179
x 2	540 MHz	2 N 5179
x 1	540 MHz	BFW 92
x 4	2160 MHz	1 N 914

Oscillator for the 9 cm band

Crystal	92 MHz	BF 173
x 3	276 MHz	2 N 5179
x 2	552 MHz	2 N 5179
x 1	552 MHz	BFW 92
x 1	552 MHz	BFW 16 A
x 2	1104 MHz	1 N 914
x 3	3312 MHz	HP 2835

The frequency plan for the 23 cm band is shown in **Figure 9**. It will be seen that air-spaced striplines are used at the base and collector of the last frequency doubler. The output coupling is made approximately 1 cm from one end of the 1152 MHz λ/2 circuit; a thin coaxial cable is used to feed the output power through the base plate to the interdigital filter.

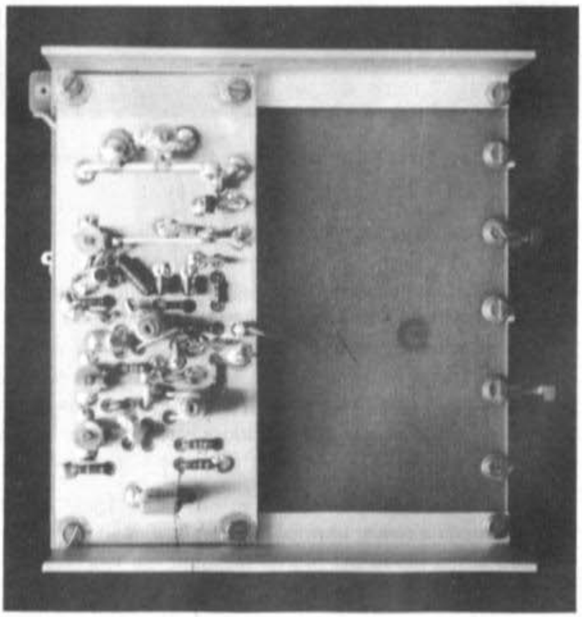


Fig. 9: Photograph of the local oscillator module for the 23 cm converter

A local oscillator module for the 13 cm band was already described in (5) under the designation DC 0 DA 003. As can be seen in the above table, a lower power version is used for our application: the 540 MHz power amplifier equipped with a transistor BFW 16 A has been deleted. This power level is sufficient when using a receive converter with only one mixer diode as long as a good diode has been found for the frequency multiplier, and when the matching has been optimized. Further details as to the construction are given in **Figure 10**, and the technology has been described in detail on a number of occasions elsewhere.

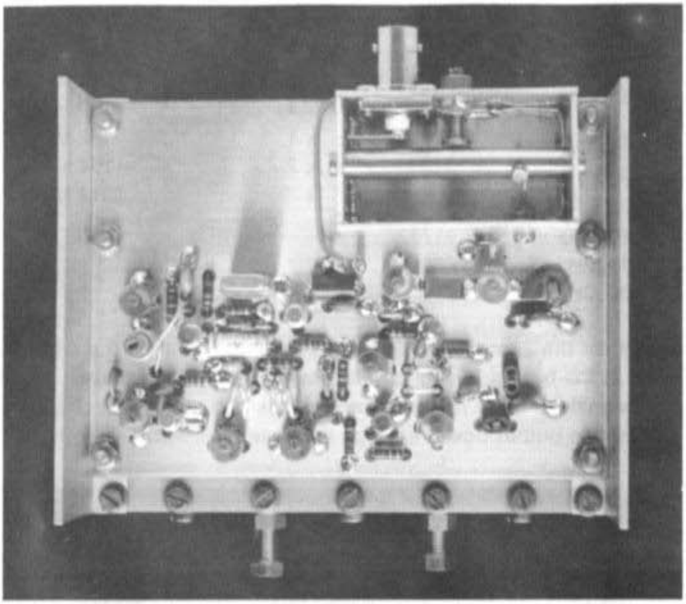


Fig. 10: Local oscillator module for the 13 cm converter

4.1. Local Oscillator Module for the 9 cm Converter

The last frequency multiplication is carried out within the interdigital converter. Since the selectivity is not very high, it is necessary for the 1104 MHz signal to be as clean as possible. The author was able to examine several different local oscillator modules on a spectrum analyzer. It was found that when using the simple, open construction, it was only possible to obtain a clean 1104 MHz signal when doubling from 552 MHz. Due to the large frequency spacing from 552 MHz, the spurious signals no longer caused any interference during the subsequent tripling process.

Figure 11 shows the circuit diagram of such a 1104 MHz local oscillator module. The principle of operation is already known; bandpass filters are used at the lower frequencies for selectivity, and air-spaced striplines at the higher frequencies. The matching between transistors T 5 and T 6 is made in a Pi-filter, however, the plastic foil trimmer at the collector of T 5 is not present on the prototype that is shown in **Figure 12**. Transistor T 5 is soldered to the conductor side of the board, in order to achieve the shortest possible connections.

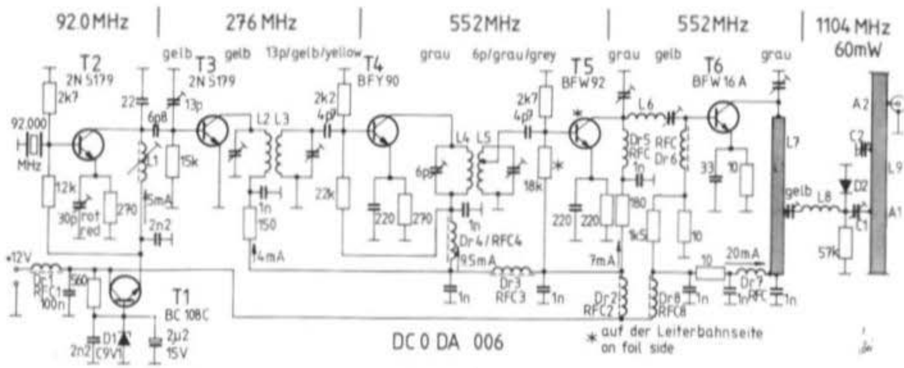


Fig. 11: Circuit diagram of the local oscillator module for the 9 cm converter

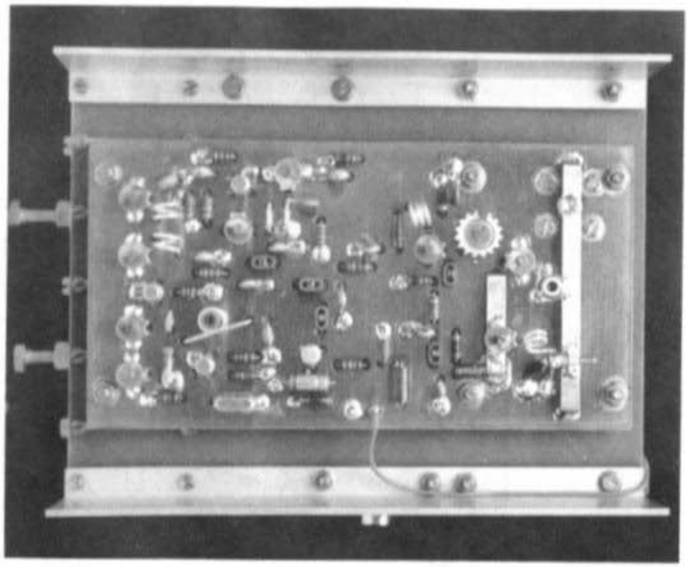


Fig. 12: Photograph of the local oscillator module for the 9 cm converter

The resonant circuit at the collector of T 6 is a capacitively-shortened λ/4 circuit having a good resonance peak. The frequency multiplier diode works into a λ/2 circuit, which is also in the form of an air-spaced stripline. This 1104-MHz stripline is tuned at the center using a ceramic tubular trimmer which is placed through the PC-board. The flange of this trimmer is mounted on the board and is soldered around its circumference to the copper coating. The output signal is coupled out via a BNC connector from the other end of the line to the diode. When using the underlined semiconductor types in the following components list, and when the current values given in the circuit diagram are present, an output power of 60 mW at 1104 MHz will result after careful alignment. However, the tripler in the converter usually only requires about 30 mW, in order to produce a mixer diode current of 1 mA. This means that a certain amount of reserve local oscillator power is available for experimenting with other types. This will probably be necessary in the case of the varactor diode, since the BA 149 is no longer manufactured.

Figure 13 finally shows a drawing of the 150 mm x 80 mm PC-board of the described local oscillator module. The lower side of the board possesses a continuous copper coating. This board has been designated DC 0 DA 006.

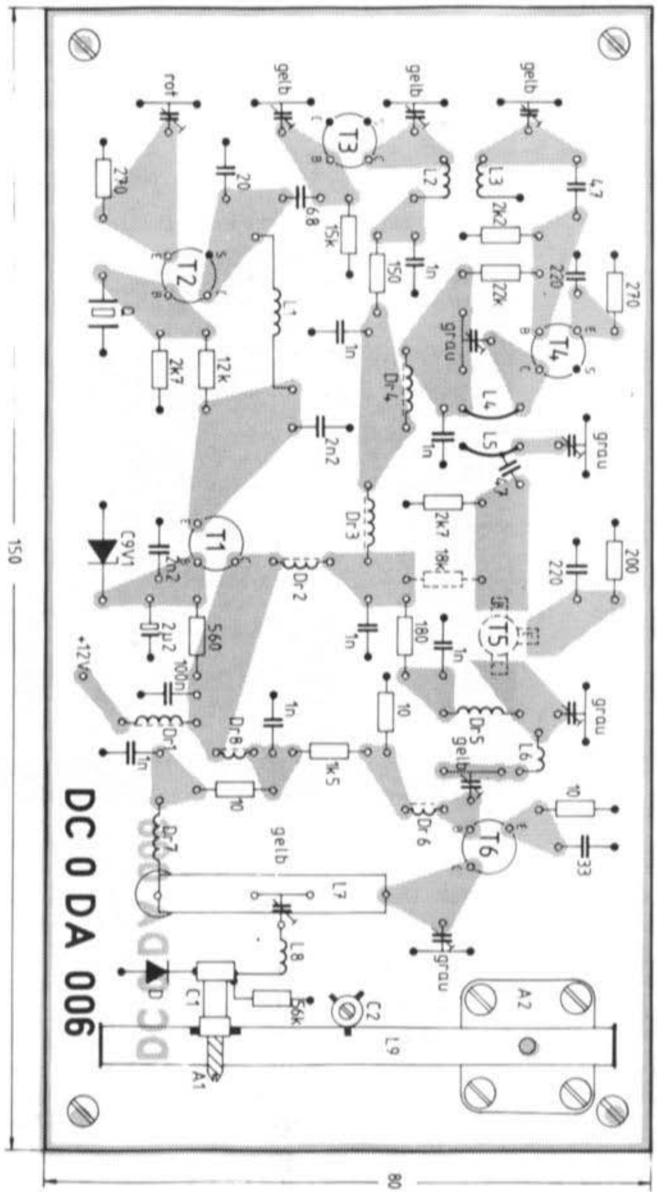


Fig. 13: PC-board DC 0 DA 006

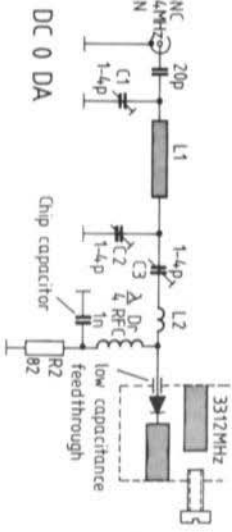


Fig. 14: Matching network for a frequency multiplier diode

4.1.1. Component Details for DC 0 DA 006

T 1:	BC 108, BC 109 or other NPN-AF transistor
T 2, T 3:	2 N 5179
T 4:	BFY 90, BFX 89, 2 N 5159
T 5:	BFW 92, BFR 34 A, BFR 90, BFR 91
T 6:	BFW 16 A, 2 N 3866 (with cooling fins)
D :	BA 149, BA 102, BB 105, BB 141, 1 N 914

Inductances L 1, L 6, and L 8 are made from silver-plated copper wire of 1 mm dia.

L 1:	5 turns wound on a 5 mm dia. coil former with VHF-core
L 2, L 3:	1.5 turns wound on a 5 mm former, self-supporting
L 4, L 5:	0.5 turns, wound on a 6.5 mm former, mounted 10 mm above the ground surface Coil tap on L 5: 7 mm from the hot end
L 6:	2 turns wound on a 4 mm former, self-supporting
L 7:	Metalstrip (silver-plated copper or brass), 5 mm wide, 0.5 mm thick, straight length 27 mm mounted approx. 4 mm above the ground surface, output coupling trimmer approx. 12 mm from the cold end
L 8:	2 turns wound on a 5 mm former, self-supporting
L 9:	Metal strip as L 7; straight length 65 mm, mounted approx. 4 mm above the ground surface; tap for the multiplier: 15 mm from the cold end; tap for the output: 10 mm from the cold end
RFC 1:	6-hole wideband ferrite choke (Philips)
RFC 2,3,4,8:	3 turns of 0.5 mm dia. enamelled copper wire wound through a ferrite bead
RFC 6:	2 turns, otherwise as RFC 2
RFC 5, 7:	$\lambda/4$ choke of 0.5 mm dia. enamelled copper wire wound on a 3 mm former, self-supporting
C 1:	Ceramic tubular trimmer 0.2 - 1.3 pF (miniature version)
C 2:	Ceramic tubular trimmer 1 - 4 pF (low profile)

4.1.2. Matching Network

A network for the matching between module DC 0 DA 006 and the multiplier diode in the 9 cm interdigital converter is shown in **Figure 14**. The inductances and capacitances are aligned for maximum voltage drop across resistor R 2. The module, which can be used for this application also at other frequencies is given in Figure 5 on the right. The following values are valid for the required frequencies:

L 1:	silver-plated copper or brass plate, 5 mm wide, 25 mm long, spaced approximately 6 mm from the ground surface
L 2:	1 turn of 1 mm dia. silver-plated copper wire, wound on a 5 mm former pulled slightly from another

$\lambda/4$ choke for 1104 MHz from approx. 7 cm 0.5 mm dia. wire, wound on a 3 mm former, self-supporting, pulled slightly from another

C 1 - C 3:	approx. 1 - 4 pF ceramic spindle trimmer

5. ALIGNMENT

An incorrect alignment is practically impossible if the given dimensions are kept. It is advisable for the local oscillator module used for driving the mixer, or multiplier to be carefully checked using an absorption wavemeter.

Before commencing alignment, the two tuning screws should be rotated fully anti-clockwise. The resonator for the oscillator frequency is now aligned for maximum DC-current via the mixer diode by slowsly inserting the tuning screw. The maximum diode current should be in the order of 0.8 and 1.5 mA.

The resonance peak is very sharp, which indicates a high Q of the interdigital resonators. The insertion depth of the M 5 - tuning screws given in table 2 can be used as approximate value. The screw for the input frequency resonator is always inserted at least two turns less than the oscillator frequency screw at an intermediate frequency of 145 MHz. However, an exact alignment is only possible when a signal is received via the antenna.

The IF-preamplifier is now aligned for maximum overall gain. Subsequently, it can be aligned for maximum signal-to-noise ratio when receiving a weak signal.

6. NOTES

6.1. Diode Mixer

As is the case with any converter using a passive mixer (diode mixer), the operating conditions of this mixer diode are decisive for the sensitivity of the overall system. A few important points should be remembered regarding this, and two diagrams are to be given that are not readily accessible to radio amateurs. The popular Schottky diode type 2817 (6) is to be used as an example. **Figure 15** shows the obtainable SSB noise figure as a function of the signal frequency. However, these values are only valid when the conditions given in the diagram are fulfilled ! This means the most favorable impedance transformation to the IF-preamplifier, which may not have an additional noise figure in excess of 1.5 dB, as well as an oscillator power of 1 mW. Furthermore, the image frequency rejection should be at least 20 dB. If no image rejection is provided, 3 dB must be added immediately. Also, if the noise figure of the IF-preamplifier is worse than 1.5 dB, this must also be added.

It is also necessary for the DC-path of the mixer diode current to be carefully observed:
As can be seen in **Figure 16**, it is not immaterial whether the diode current is measured using a meter having an impedance of 1 kΩ, or whether the voltage drop is measured on a 100 Ω resistor. The lowest noise figure is achieved using a DC resistance of R_L = 100 Ω, however, one requires at least 3 mW of oscillator power in order to be able to measure an improvement with respect to other load resistance values R_L.

Another important magnitude is the IF load impedance Z_{IF} to which the IF-preamplifier must be matched in order to provide the most favorable IF-impedance to the mixer diode and at the same time the most favorable source impedance for the IF transistor. The IF-impedance is in the order of 250 and 400 Ω in the case of diode type 2817, whereas higher values result at low oscillator power, and lower values at higher oscillator power levels.

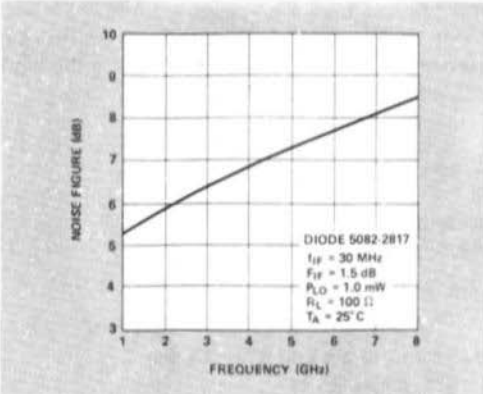


Fig. 15:
Noise figure as a function of frequency

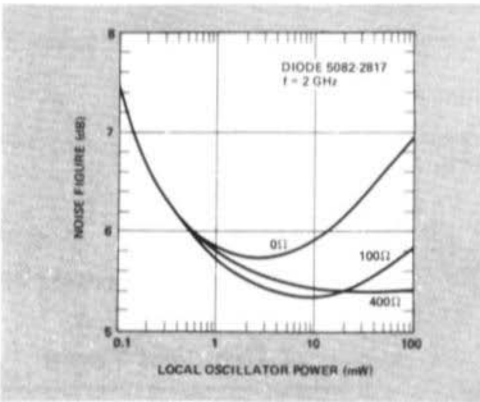


Fig. 16:
Noise figure as a function of oscillator power

In order to achieve the most favorable noise figure, it is necessary for all criteria to be fulfilled simultaneously, or to be optimized. Noise figures of less than 12 dB will not be achieved when using a lossy, printed hybrid ring with inefficient IF-matching and when using a non-state-of-the-art IF transistor, and especially not when no image rejection is provided.

The specifications of the interdigital converters, especially those for the 13 cm and 9 cm bands were classified as good by experts in frequency converter technology. The selectivity of the described converter is sufficiently good that no external filter is required when a pre-amplifier is used at the antenna.

6.2. Receive Converter for the 6 cm Band

Experiments were made to construct the interdigital converter with smaller dimensions a and b for the 6 cm band. The relationship between length and diameter of the resonators, and the spacing between them were, however, very unfavorable. It is necessary for the interdigital filter to be redesigned for this band. Both the author and the publishers would be very interested to hear from a designer that achieves this.

It would probably be more favorable to shift the intermediate frequency to 433 MHz in order to improve the image rejection. However, it is necessary for a state-of-the-art IF-preamplifier to be used whose noise figure is a maximum of 2.5 dB, so that the advantage of the higher IF is not lost. Transistor type BFT 66 could be used for this application (Editorial note).

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Pages 1 to 19, Figure 6, and Pages 1 to 20, Figure 8

RECEPTION OF THE METEOSAT WEATHER SATELLITE

by T. Bittan, DJ 0 BQ / G 3 JVQ

Weather satellites have been used for many years to improve the accuracy of weather forcasting and as additional information to the ground weather data collection system. Uptil the launch of satellites like METEOSAT, weather satellites were usually on a polar orbit at an altitude of 800 to 900 km. Such satellites of the NOAA (USA) and Meteor (USSR) series transmit their data in the VHF satellite band between 136 MHz and 138 MHz. Since these satellites were orbiting relatively fast due to their low altitude, it was necessary to have some method of tracking them, which was outside the scope of most amateurs.

The first European weather satellite METEOSAT was launched on November 23, 1977 and commenced operation mid December 1977. The reception of this satellite, which is one of a series of five around the world (see **Figure 1**) is more challenging due to the higher frequency used of 1694.5 and 1691.0 MHz, as well as its greater distance from the earth of 35,900 km. However, these satellites are in a geostationary orbit above the equator, in other words, they are orbiting the world at the same speed as the earth's rotation. Since the satellites seem to be stationary in space, no antenna tracking will be required.

Of course, due to the higher path loss at the higher frequencies and greater distance from the earth, higher gain antennas must be used than at 137 MHz.

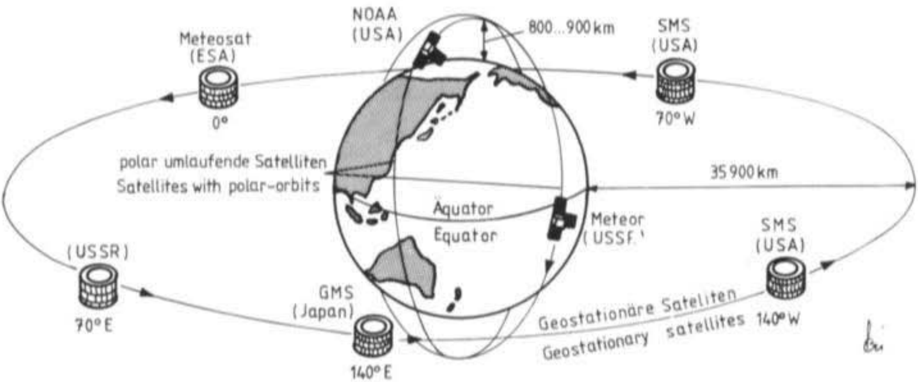

Fig. 1: A system of 5 geostationary weather satellites is planned

1. CLOUD COVER SURVEY AND PROCESSING

The METEOSAT »photographs« the cloud cover of the earth both in the visual and infrared ranges. In the visual range, the satellite produces one image every 30 minutes. This image comprises 5000 lines of 5000 points each, corresponding to a resolution of 2.5 km.

The infrared pictures are also made every thirty minutes in the 10 to 12.5 micron band. Each image comprises 2500 lines of 2500 points each, corresponding to a resolution of 5 km. Water vapour pictures are also made in the 5.7 to 7.1 micron band at the same resolution, but less frequently.

These analog images are digitalized and transmitted to a central ground station in pulse-code modulation. This information is then processed in a computer, divided into sectors, and retransmitted to the satellite together with additional information such as date, time, type of image, sector of the original image, as well as other coded information. A transponder aboard METEOSAT then converts the processed images to the two frequencies of 1691 MHz or 1694.5 MHz which are retransmitted for reception on the earth. It is especially the reception of these processed images that is of interest to us.

2. SPECIFICATIONS

2.1. Satellite Specifications

Carrier frequencies:	Channel 1: 1694.5 MHz Channel 2: 1691.0 MHz
Effective radiated power:	48 dBm
Polarization:	linear
Modulation mode:	FM, with the subcarrier modulated AM
FM-subcarrier:	2.4 kHz
Max. frequency deviation of subcarrier:	9 kHz
Max. transmission bandwidth:	26 kHz
Video bandwidth:	1.6 kHz

2.2. Specifications of a Typical Professional Receiver System

Antenna:	26 dBi, linear polarization	
Preamplifier:	Gain	30 dB
	NF	1.5 dB
Receive converter:	1691 MHz / 137.5 MHz	
	Gain	20 dB
	NF	6 dB
VHF receiver:	137 MHz	
	NF	12 dB
	Bandwidth	25 kHz
	Demodulation	FM

3. AMATEUR RECEPTION

Of course, the above specifications for a professional receiver system must allow enough reserve to provide high-quality images even under worst-case conditions. Amateur systems do not require such reserves, and if so they can be also achieved, using an expensive preamplifier and/or large parabolic dish.

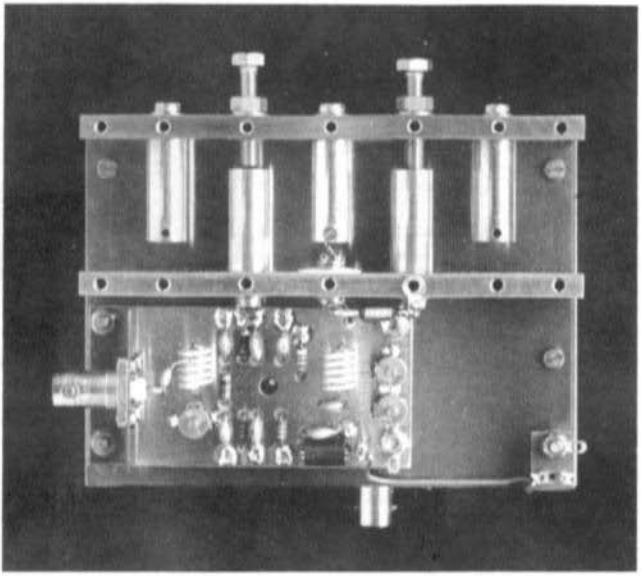

Fig. 2

3.1. Basic Requirements

It is assumed that a 137 MHz FM receiver and APT-display unit is available. Of course, it is not absolutely necessary to use 137 MHz, and 144 MHz would also be suitable. Required is then a converter to transpose the frequency of 1691 MHz or 1694.5 MHz to the required frequency of 137 MHz (or 144 MHz). This converter should be sensitive enough that reception is possible without an expensive preamplifier when using a managable parabolic dish of 1 m to 1.5 m diameter.

Such a converter was obtained using a similar construction to the converter described by J. Dahms, DC 0 DA. The METEOSAT converter possesses an interdigital filter with five resonators which provides the required selectivity. **Figure 2** shows a photograph of a laboratory prototype. The screened oscillator module is on the other side. A block diagram of the converter is given in **Figure 3**.

The first resonator from the left is the input coupling to the antenna, the second is tuned to the required frequency at 1.7 GHz, and will attenuate all other frequencies such as the image frequency. The first resonator from the right is the input coupling for the local oscillator signal and the second will attenuate any unwanted spurious signals. The center resonator couples out both the input and local oscillator signals and feeds them to the mixer diode D 1. The intermediate frequency of 137 MHz (144 MHz) is fed via a selective, low-noise preamplifier to the FM receiver. The operation of the local oscillator module can be checked by measuring the diode mixer current across the 100 Ω resistor. A voltage of 100 mV should be present (permissible tolerance ± 50 %).

Figure 4 shows the METEOSAT converter in its ready-to-operate form. This converter is available ready-to-operate in an aluminium case from UKW-TECHNIK.

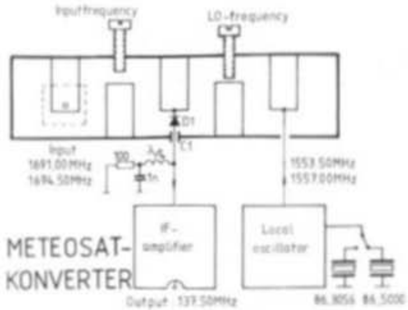

Fig. 3: Block diagram

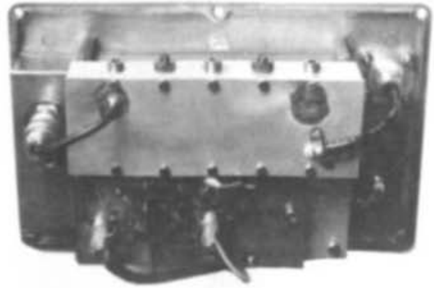

Fig. 4: Ready-to-operate module

3.2. Specifications of the UKW-TECHNIK METEOSAT Converter

Receive frequencies (switchable via DC-lines):	1691.00 and 1694.50 MHz
Intermediate frequency (144 MHz on request)	137.5 MHz
Noise figure (single sideband):	typ. 9 dB
Gain:	typ. 18 dB
Microwave bandwidth:	typ. 5 MHz
IF-bandwidth:	typ. 5 MHz
Input and output connectors:	BNC (N on request)
Input and output impedance:	50 Ω
Operating voltage:	12 V stab.
Current drain:	$<$ 100 mA

Remote feeding via coaxial cable possible.
Accommodated in weatherproof cast aluminium box.

3.3. Practical Reception

Experiments made using this converter together with a parabolic dish of 1.2 m diameter at ground level showed that satisfactory reception (virtually noise-free) was possible. If a greater signal-to-noise ratio is required for professional applications, this can be obtained using a low-noise preamplifier and/or larger dish.

Several such converters are in operation at observatories and professional companies.

3. REFERENCES

(1) J. Dahms: An Interdigital Converter for the GHz-Range
in this Edition of VHF COMMUNICATIONS

(2) Programme METEOSAT: Dissemination Mission European Space Agency

(3) Rohde and Schwarz Magazine 82, Summer 1978 edition, pages 11 - 14 (in German)

CALCULATION OF THE ELEVATION AND AZIMUTH OF THE ANTENNA FOR METEOSAT RECEPTION

by R. Lentz, DL 3 WR

The METEOSAT weather satellite is in a geostationary orbit over the zero degree meridian. It is located in the same plane as the equator approximately 36 000 km from the surface of the earth. The antenna direction can be calculated relatively simply for any geographical location of longitude λ and latitude β using simple geometric equations.

Since the beamwidth of amateur radio antennas for this application will hardly be less than 5°, it is possible for the calculation to be simplified considerably. For instance, the calculation of the elevation was simplified by assuming that the receive location was on the zero degree meridian. The latitude of the receive location is only roughly taken into consideration during the calculation of the azimuth.

The following terms are used in the calculation:

- r = Radius of the earth = 6400 km
- s = Distance Satellite – Ground surface = 36 000 km (and $m = s + r$)
- λ = Longitude of the receive location
- β = Latitude of the receive location
- ε = Antenna elevation
- α = Antenna azimuth

A calculation based on the city of Augsburg, West Germany, (β = 48.4° and λ = 11°) is to be used as an example.

Figure 1 shows the calculation of the elevation graphically.

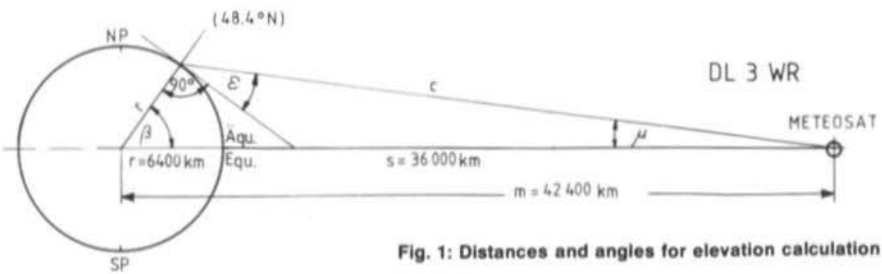

Fig. 1: Distances and angles for elevation calculation

It is made in three steps rather than in one large equation in order to simplify the calculation.

$$c = \sqrt{m^2 + r^2 - 2\,mr \times \cos\beta} \qquad (1)$$

$$\mu = \text{arc cos}\,\frac{c^2 + m^2 - r^2}{2\,mc} \qquad (2)$$

$$\varepsilon = 90° - \beta - \mu \qquad (3)$$

Example for $\beta = 48.4°$

$$c = \sqrt{42400^2 + 6400^2 - 2 \times 42400 \times 6400 \times \cos 48.4°} = 38\,450 \text{ km}$$

$$\mu = \text{arc cos}\,\frac{38450^2 + 42400^2 - 6400^2}{2 \times 42400 \times 38450} = \text{arc cos } 0.9922 = 7.15°$$

$$\varepsilon = 90° - 48.4° - 7.15° = 34.45° \approx \mathbf{35°}$$

Figure 2 shows the calculation of the azimuth angle graphically. This diagram views the earth from »above« from a point over the North Pole.

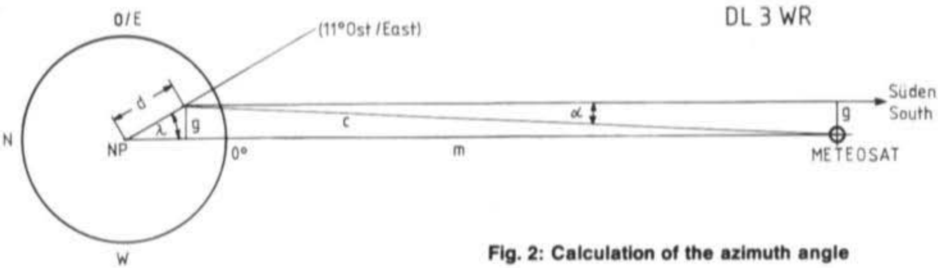

Fig. 2: Calculation of the azimuth angle

The calculation is also made in three steps:

$$d = \frac{\beta}{90} \times r \tag{4}$$

$$g = \sin \lambda \times d \tag{5}$$

$$\alpha = \text{arc sin}\,\frac{g}{c} \tag{6}$$

Example with $\lambda = 11°$ and $\beta = 48.4°$

$$d = \frac{48.4}{90} \times 6400 \text{ km} = 0.54 \times 6400 \text{ km} = 3442 \text{ km}$$

$$g = \sin 11° \times 3442 \text{ km} = 0.1908 \times 3442 \text{ km} = 657 \text{ km}$$

$$\alpha = \text{arc sin}\,\frac{657 \text{ km}}{38450 \text{ km}} = \text{arc sin } 0.0171 = 0.979° \approx \mathbf{1°}$$

This means that the METEOSAT has an elevation of approximately 35° at Augsburg, and is approximately 1° west of geographical South (181°).

AN INEXPENSIVE POWER AMPLIFIER FOR 24 cm USING 2 C 39

by U. Mallwitz, DK 3 UC

The following article is not to describe a new type of 24 cm power amplifier, but is to describe how an efficient cavity power amplifier can be manufactured using inexpensive material and simple tools. This power amplifier is suitable for use with tubes of the 2 C 39 family. In order to avoid lathing in the manufacture of the cavity, a 200 g coffee tin having a diameter of 80 mm was used. Many such amplifiers have been produced, which have been nicknamed »coffee-tin-PAs«.

Figure 1 shows such a power amplifier before soldering the anode resonator to the base plate. The tuning disk is to be seen to the left of the grid contact ring, and to the right of this the output coupling with PTFE-insulation for the compensating screw. This can be seen on the resonator (right) on the left-hand side of the anode contact ring. The connection for the anode voltage can be seen to the right of the contact ring. The anode plate together with the anode contact ring is insulated from the resonator panel using a PTFE-disk.

Fig. 1: Various parts of the power amplifier before assembly

1. SPECIFICATIONS

Tunable range:	1120 to 1300 MHz
Maximum output power, continuous:	approx. 35 W
Large-signal gain:	approx. 4 to 6 dB
Low-signal gain:	approx. 6 to 9 dB

It has been noticed that the gain of 2 C 39 type tubes varies from type to type, and is very much dependent on the quiescent current. The gain increases by approximately 3 dB at quiescent currents between 60 and 120 mA.

An identical power amplifier was also constructed using a large coffee tin (100 mm diameter). This provided a maximum of 2 dB more gain, but the power amplifier is still not ready for publication. Maybe some readers would like to carry out experiments in this direction. The described 80 mm diameter power amplifier has also been equipped with two 2 C 39 tubes and operated with a voltage of 780 V. At an anode current of 390 mA (!) and with 20 W of drive, an output power of 85 W was achieved in continuous operation.

The described power amplifier can also be used as frequency multiplier from 432 MHz to 1296 MHz or from 384 MHz to 1152 MHz. Further details regarding adjustment of the operating point and design of the cathode circuit are to be found in (1).

If several stages are to be connected in series, it is advisable for each stage to be constructed individually and tested before they are connected together using short cables. In order to amplify the output power of, for instance, a transmit converter designed by DF 8 QK (2) to 35 W, two to three stages will be required. When using a compact construction such as the four-stage amplifier described by DJ 6 UT (3), matching losses will occur; also the whole amplifier will be inoperative if one stage should fail. This means that that type of construction is less favorable.

2. TOOLS AND MATERIAL

The following tools are advisable in addition to the normally available tools like screwdrivers, pliers, vice etc.:

- 80 W soldering iron
- Chassis punchers for the following diameters: 16 / 18 / 21 / 25 / 30 mm
- Drills for the following diameters: 1.9 / 3.0 / 4 / 5 / 6 / 8 / 10 mm
- Round files of 3 mm and 6 mm diameter
- Flat files
- Sandpaper
- Metal shears

The following table gives the required mechanical and electrical parts for the power amplifier.

LIST OF REQUIRED MATERIAL

Part	Designation / Dimensions / Material	Pieces
1	Coffee tin, 85 mm diameter, 110 mm height	1
2	Anode contact ring	1
3	Grid contact ring	1
4	Anode plate, 82 mm diameter, brass or tin plate 0.8 to 2 mm thick	1
5	Insulating disk for anode plate, 85 mm diameter, 0.2 to 0.25 mm thick, PTFE (teflon)	1
6	Insulating disk for the output coupling, 9 mm diameter, 0.25 mm thick, PTFE (teflon)	1
7	Insulating strip for the output coupling, 22 mm x 19 mm, 0.25 mm thick, PTFE (teflon)	1
8	Anode ring, 108 x 10 x 1	1
9	Base plate, brass or tin plate, 1 mm thick, 100 mm x 110 mm	1
10	Tuning disk, brass or tin plate, 1 mm thick, 20 mm diameter	1
11	Tuning screw, M 6 brass screw, 80 mm long, with hextagonal head, approx. 4 mm high	1
12	Spring, wire diameter 0.5 to 1 mm, diameter of spring 8 to 10 mm, spring length 20 to 30 mm	2
13	Mounting piece 10 x 50 x 1 mm, tin plate	1
14	Tube for the output coupling, brass, outer diameter 8 mm, inner diameter 6 mm	1
14a	Brass disk, 6 mm diameter, 2 mm thick, saw from the material of part 15	1
15	Round metal rod for heating/cathode, 27 mm long, 6 mm diameter, 5.5 mm diameter if available	1
16	Brass tube for heating/cathode, 23.5 mm long, outer diameter 10 mm, inner diameter 8 mm	1
17	Double-coated PC-board material for the cathode circuit, 20 mm x 25 mm (expoxy)	1
18	Grid tube, brass ≈ 1 mm wall thickness, inner diameter 21 or 25 mm, according to the contact ring	1
19	Brass or tin plate, 1 mm thick, 60 mm x 30 mm for the cathode chamber	2
20	Brass or tin plate, 1 mm thick, 25 mm x 30 mm, narrow sides of the cathode chamber	2
21	(silver-plated) brass plate for the input inductance, 0.3 to 1 mm thick, 34 mm long, 10 mm wide	1
22	Heater chokes, enamelled copper wire of 0.2 to 0.4 mm dia., 65 mm long, wound on a 4 mm dia.former	2
23	BNC socket, single-hole mounting, threading at least 10 mm long, with 2 nuts (for output)	1
24	BNC socket, square flange (for input)	1
25	Ceramic disk capacitor without connection leads for the input, approx. 25 pF	1
26	Feedthrough capacitors for the heater, 500 pF to 2.7 nF	2
27	Tubular trimmer for the cathode circuit, approx. 0.5 to 4 pF, soldered into place	1
28	Brass screws M 3, 10 mm long, with cylinder head, for the anode plate	4
29	Washers, spring washers, brass nuts (M 3)	4 each
30	Insulating disks, 9.5 mm dia., if possible PTFE, similar to those used for mounting power transistors	4
31	Brass screw M 5, threading 30 mm long, compensating screw for the output coupling	1
32	Brass nut M 5, at least 4 mm high	1
33	Brass nuts M 6 (for part 11), at least 4 mm high	3

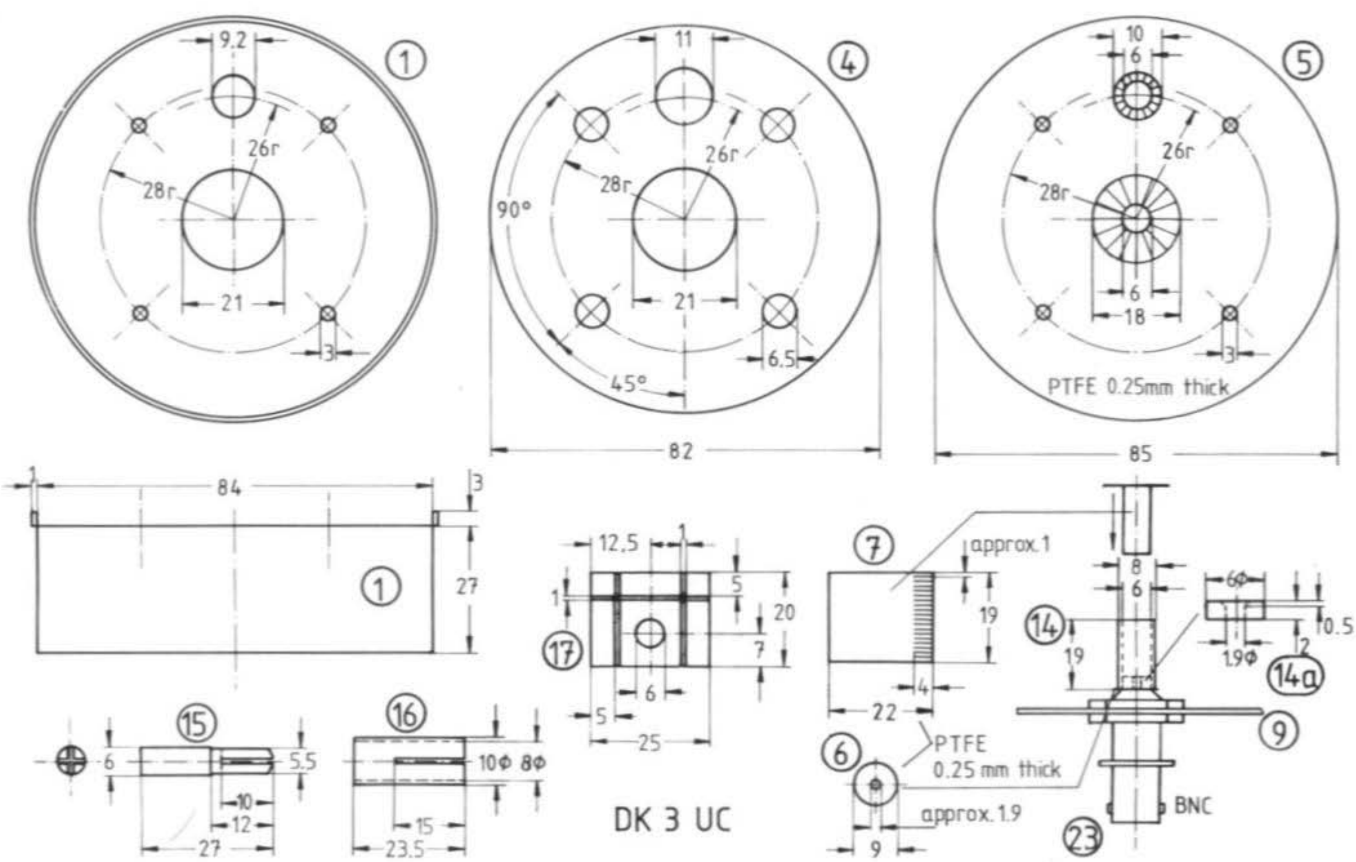

Fig. 2: Individual parts of the power amplifier

3. MANUFACTURE OF THE INDIVIDUAL PARTS, PREPARATIONS

3.1. The Resonator

A view from above of this part is shown in the upper left-hand corner of **Figure 2**, and the side view is shown below this.

A hole of 20 to 21 mm diameter should be punched into the bottom of the coffee tin, or drilled and filed to the correct dimensions. The other holes should be then drilled according to the drawing, and the tin cut down to a height of 30 mm (measured from the fold). All cut surfaces should be debured, and cleaned.

The folded edge in the inside of the tin should be soldered carefully together. After this, the M 3 brass screws (part 28) are placed through the holes from the inside and provided with nuts. It is then possible for the heads of the screws to be soldered into place on the inside of the tin, after which the nuts are removed. The M 5 - screws are now placed into the hole so that they are centered and are also soldered into place on both sides. This means that part 1 is completely prepared.

3.2. Anode Plate (Part 4), Anode Ring (Part 8), and Anode Contact Ring (2)

The anode plate is shown in the center of Figure 2. The 1 to 2 mm thick disk can be manufactured from brass or tin plate. The insulating disks (part 30) should be fitted into the 6.5 mm dia. holes given in the diagram. The anode ring should be fitted with the anode contact ring and can be made as shown in **Figure 4** either from a tube, or from a metal strip, which has been bent into place and soldered together. The anode ring is soldered concentrically to the 21 mm hole on the anode plate. After this, the anode contact ring, that is also shown in Figure 4, should be placed into the anode ring where it is soldered at the top, inside. Correct contact pressure exists when an old tube can be inserted with a slight pressure.

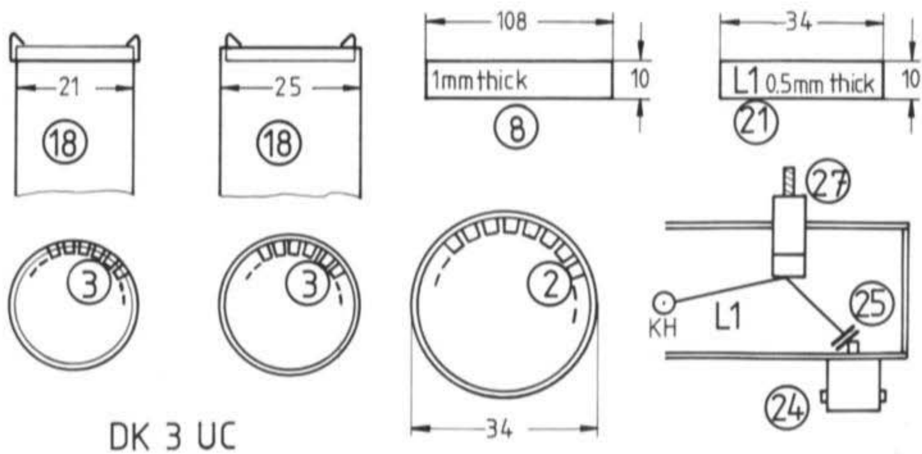

Fig. 4: Anode contact ring and anode ring; the latter will not be necessary when the contact material is of the correct height. Two methods for making the grid tube and the grid contact ring. The input inductance and its construction.

3.3. Insulating Disk for the Anode Plate (Part 5)

The part (right upper corner in Figure 2) insulates the anode plate with its high-tension voltage of up to 1000 V from the resonator pin, which is at ground potential. Due to the heat dissipated by the tube, PTFE material is necessary. The required spring-loaded holes are made with a 6 mm dia. drill and cut around the edge with a pair of shears. An exact fitting is made during the final construction. After this, the disk should not be removed from the resonator.

3.4. Base Plate (Part 9)

This part is shown in the left of Figure 3. The dimensions can be selected so that it is suitable for installation in the cabinet. One M 6 nut is provided at each side of the 6 mm holes and held into place using a screw. The nuts are then subsequently soldered into place. The nut for a BNC socket for single hole mounting is soldered on one side of the 9.2 mm hole. This side of the base plate will then be the inside of the cavity resonator.

All parts should then be deburred and cleaned.

3.5. Tuning Parts (Parts 10, 11 and 12)

Figure 3 shows this assembly when assembled. The disk is firstly made, deburred and cleaned, which is followed by filing the head of the hextagonal screw. After this, the disk is soldered to the screw. It is important that the disk is mounted centrically and is at right angles to the axis of the screw. A 3 to 5 mm deep slot is sawn into the other end of the screw so that it can be adjusted with the aid of a screwdriver.

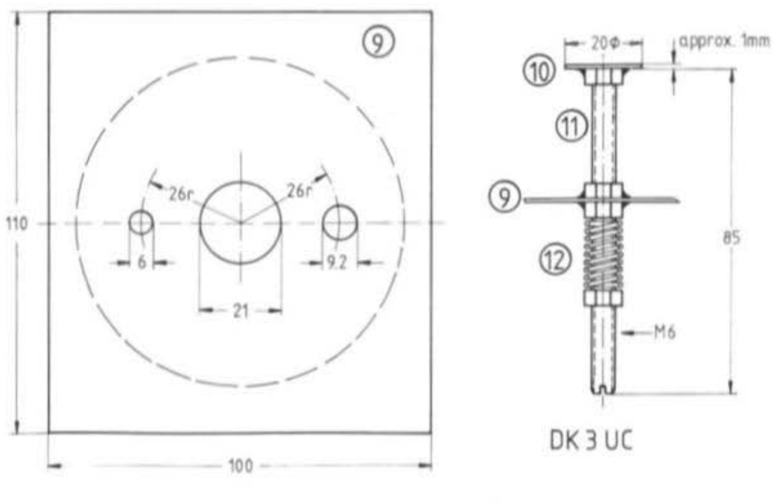

Fig. 3: Base plate and tuning system

3.6. Grid Tube (Part 18) and Grid Contact Ring (Part 3)

These two parts are shown in Figure 4. The dimensions are 21 or 25 mm dia., height 20 mm.

When using a tube diameter of 21 mm, the grid contact ring is fitted tightly into place and soldered. In the case of a 25 mm tube, the contact ring should be inserted into the inside and soldered into place. Finally, all parts should be deburred and the residual solder removed.

3.7. Cathode Circuit (Parts 15, 16 and 17)

The parts for the cathode and heater circuit are given in the lower part of Figure 2. A cross of 10 mm in depth is cut into part 15 using an approximately 0.5 mm thick saw blade. This is followed by filing off the corners to approximately 45° for a length of 1.5 mm. The corners of the other end are also slightly removed. Finally, the diameter of the slotted end should be brought to 5.5 mm for a length of 12 mm. This can be achieved using a flat file. If material of 5.5 mm diameter is available, this will not be necessary since the whole contact pin can be 5.5 mm in diameter.

After filing down to approximately 5.7 mm diameter, the contact should be checked using a 2 C 39 tube so that a good fit is obtained. One should be able to insert it approximately 9 mm into the tube socket only using a slight pressure. If one only presses the slotted ends together and inserts them into the tube, it could be possible for the tube to be damaged.

Part 16 is now manufactured and slotted as shown in Figure 2. The ends should then be pushed together slightly until this part fits tightly to the cathode. It is not necessary to alter the inner or outer diameter.

Part 17 is made from a piece of double-coated PC-board material (epoxy glass fiber, and not Pertinax). The three lines shown in Figure 2 are where the copper coating is removed by etching, filing or sawing. They are made on both sides of the board so that six independent conductive surfaces result. After drilling the hole and removing the copper surface on one side for approximately 0.5 mm around the hole (using an 8 mm drill), it is possible for parts 15, 16, and 17 to be mounted into place.

All parts should now be cleaned and any grease removed. Any oxydized surface should be cleaned with emery cloth. A 2 C 39 tube is now placed on its heat sink, and fitted as follows:

Insert part 15 approximately 9 mm into the heater contact of the tube. Place part 16 with its slotted end on the cathode contact of the tube, approximately 10 mm long, and not over the first step on the cathode contact. Place part 17 onto the end of part 15 in such a manner that the countersunk side of the 6 mm hole faces the tube. The parts must fit tightly. After fitting them exactly at right angles, part 15 should be soldered to the large conductive surface on part 17.

The tube is now laid on its side with the previously mentioned parts and the end of part 16 is soldered to the inner conductive surface on part 17. Part 15 should have no contact to this surface.

The cathode circuit is thus completed. If several tubes are available, it is possible for the module to be checked for correct contact. The described module can be used with any tube of the 2 C 39 family.

3.8. Output Coupling (Parts 6, 7, 14, 14a and 23)

The inner conductor of the BNC socket (part 23) should be shortened to 2 mm length. The brass tube (part 14) should be shortened to 19 mm in length and deburred. The 1.9 mm hole in the brass disk (part 14a) should be countersunk by approximately 0.5 mm on one side using a 4 mm drill. Cut the two PTFE parts.

For construction, the BNC socket is provided with its nut and with the threading facing upwards. This is followed by placing the insulating disk (part 6) and the brass disk (part 14a) into position with the countersunk side facing upwards, and then soldering the disk with a small amount of solder to the inner conductor. The tube (part 14) is now placed onto the disk (part 14a) so that a small edge remains for soldering. Use only a small amount of solder and feed this from the inside of the tube. Pay attention that the tube fits together with the central axis of the socket. Remove any residual solder from this part.

After preparing part 7 (PTFE) with the 4 mm cuts shown in Figure 2, the strips should be wound lengthwise around the 4 mm drill. It is then placed into the tube (part 14) so that approximately 5 mm protrudes, of which 4 mm are the cuts. These are bent 90° outwards with the finger. This is the guide of the compensating screw (part 31) and insulates the output coupling from the resonator. If a good tubular trimmer with a final capacitance of approximately 4 pF is available that is able to fit into part 14 and possesses the required length, it is possible for part 7 to be deleted, and for the trimmer to be placed into part 14.

4. CONSTRUCTION OF THE POWER AMPLIFIER

The PTFE disk (part 5) is now placed over the four M 3 screws of the resonator (part 1). The cuts for the feedthrough of the tube are depressed inwards with the aid of a tube. The teflon disk can be fitted easily when the anode plate (part 4) is placed into position. The disk should be close to the surface and lay flat without folds. The four insulating pieces (part 30) should now be fitted. Their edges should only be as thick as the anode plate, otherwise it will be necessary to file them. After this, it is possible for the washers, spring washers, and M 3 nuts to be screwed into place. Attention should be paid that everything fits correctly when tightening. The fitting of these parts should be checked using one or several tubes.

A tube should remain inserted for the following construction. The grid contact described in section 3.6. is plugged onto the grid contact of the tube and the fitting checked by rotating the tube in the contact. The base plate (part 9) is now placed onto the grid tube until it touches the resonator. The grid tube should now protrude by approximately 3 mm. It is possible by rotating the resonator unit to check whether all parts are centered (coaxial). After the required corrections, the grid tube is soldered to the base plate on the outside. After this, the resonator including the tube should be removed.

The tuning pin described in section 3.5. is now screwed into place on the base plate using nuts. The output coupling is also screwed into place using the associated nuts and it is locked into place at a depth of approximately 18 mm into the resonator cavity. The components are now as shown in Figure 1, but without compensating screw.

The resonator unit is finally plugged onto the base plate unit and adjusted so that the compensating screw is inserted centrically into the output coupling tube. After checking once more, it is possible for the resonator to be soldered to the base plate.

The compensating screw is now removed once again, is provided with the spring (part 12), and screwed back into position until the screw is depressed to approximately 10 mm in length. The counter nut of the BNC socket is also released and the output coupling inserted into the resonator until the PTFE touches. This completes the construction of the resonator.

A tube is reinserted for the construction of the cathode circuit. The tube should fit tightly. The cathode circuit described in section 3.7. is now fitted and the four side panels shown in **Figure 5** (parts 19 and 20) are made. Before soldering these parts onto the base plate, the cathode circuit should be rotated so that it fits as shown in **Figure 6**. Part 17 can now be soldered on the upper side to the panels of the input coupling chamber.

The second spring is now fitted to the tuning pin and depressed using a nut. This nut is secured using a strip of tin plate (part 13), which is soldered to the cathode case. This ensures the important contact pressure between the tuning pin and the nut soldered to the resonator. The tuning pin should be adjustable at relatively low pressure using a screwdriver.

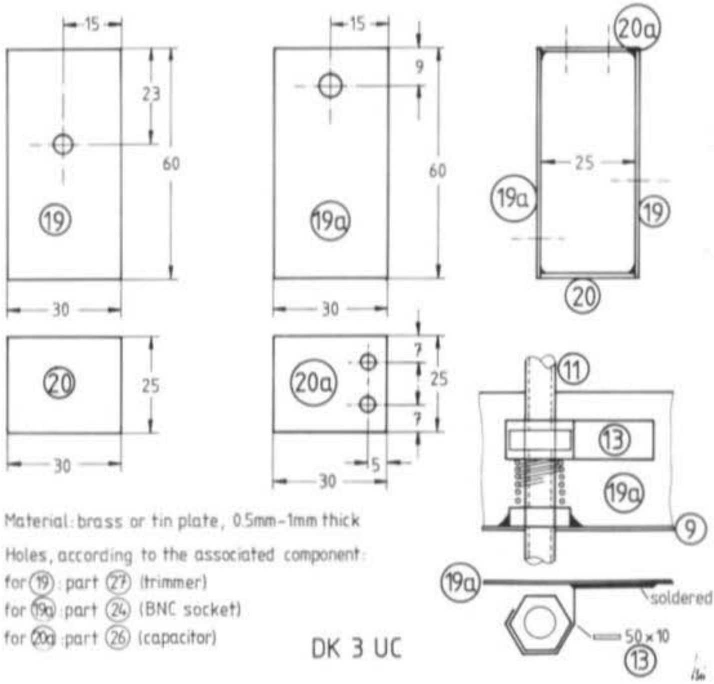

Fig. 5: Parts of the cathode chamber and mounting bracket, part 13

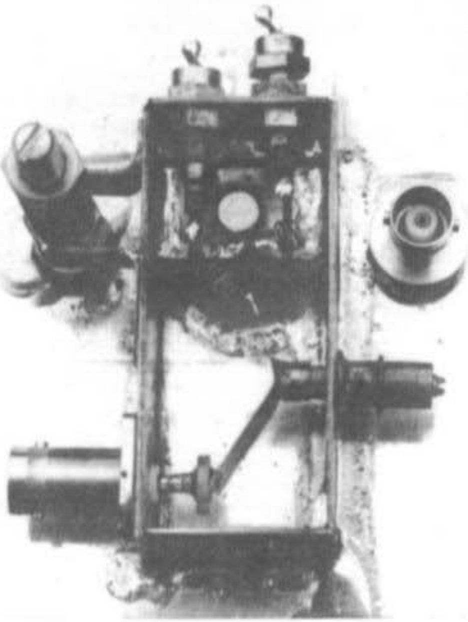

Fig. 6: The completed cathode circuit of a power amplifier (old version of part 13)

5. ALIGNMENT

A preliminary alignment is made firstly since the resonance point is very narrow and it could endanger the tube when looking for resonance with the operating voltage connected. The preliminary alignment is made in conjunction with a 1296 MHz receiver, a stable signal at 1296 MHz (harmonic from a 2 m or 70 cm transmitter, or a beacon transmitter), as well as a BNC T-piece.

A tube should be inserted and the test signal fed via the T-piece to the receiver. The T-piece is now connected to the input of the power amplifier and the input circuit aligned with the aid of the trimmer. At resonance, the receive signal will be reduced in strength which will be indicated as a dip on the S-meter. This process is then repeated at the output socket in conjunction with the anode resonator. The test signal should not be stronger than approximately 20 to 30 dB. The S-meter reading will drop by approximately 5 to 20 dB under resonance conditions. A strong signal of 60 dB or more is not favorable since the resonance dip will hardly be noticeable.

After the preliminary alignment, it is possible for the linear amplifier to be brought into operation. It is firstly operated at low drive and optimized in steps increasing the drive. The latest measurements made using new tubes of the YD-series with sufficient cooling produced a continuous output power of 65 W ! The anode voltage amounted to 780 V at 305 mA plate current and a drive power of 12.8 W.

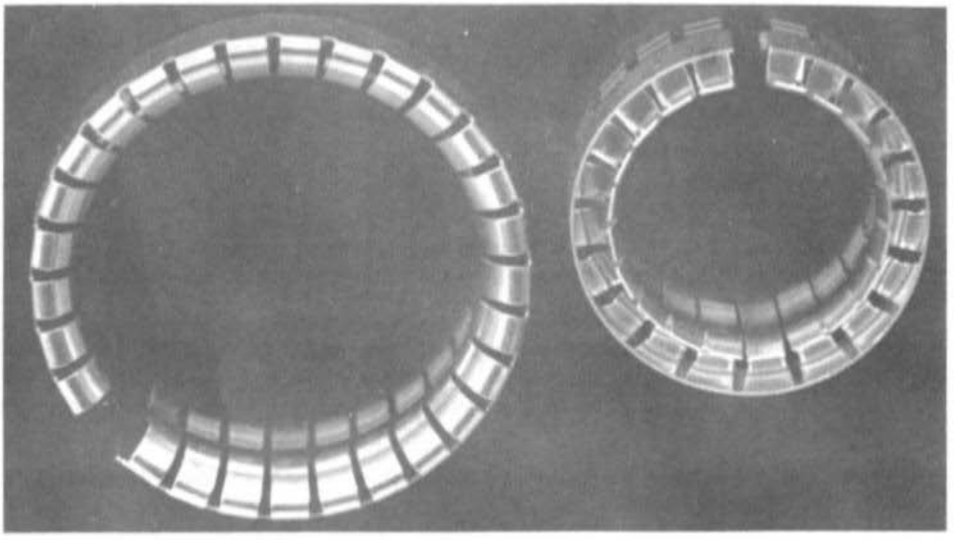

Fig. 7: Suitable contact rings for anode and grid are an important prerequisite for correct operation

6. NOTES

It is always necessary for the tube to be cooled, even when only the heating is connected. Further data on this was given in the description of a 70 cm power amplifier by A. Tautrim. Fundamentally speaking: the more cooling that can be provided the better. The best way of concentrating the cooling air onto the cooling fins of the tube is to use some form of tube that fits onto the blower to be used. It is also possible to construct air ducts for two-tubes.

Regarding part 13: An unsuitable type and mounting of this part can be seen in Figure 6. If the contact between mounting piece, nut and cathode case is too great, the spring pressure will not be present.

Regarding L 1: The input circuit is designed so that only a very low resonance peak is present. This ensures that no tendency to self-oscillation is present, and various exciters and lengths of cables can be used without realignment being necessary.

7. REFERENCES

(1) Editors: A Power Amplifier for the 23 cm Band Equipped with the 2 C 39 Tube
VHF COMMUNICATIONS (8), Edition 4/1976, pages 222 - 231

(2) U. Beckmann: A Linear Transverter for 28 MHz - 1296 MHz with Push-Pull Mixer
VHF COMMUNICATIONS (9), Edition 4/1977, pages 212 - 220

(3) R. Jux and H. Dittberner: A Transmit Mixer and Linear Amplifier for 23 cm Using 2 C 39 Tubes
VHF COMMUNICATIONS (7), Edition 3/1975, pages 146 - 160

(4) I. Nielsson: Homemade Fingerstock
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A FREQUENCY CONTROL LOOP FOR A 433 MHz VCO

by T. Krieg, DK 8 GY

1. REQUIREMENTS

The requirement was for a mobile 70 cm transmitter which would be able to transmit a 20 kHz wide frequency multiplex signal in the FM mode. The accuracy of the output frequency should be better than 10^{-4}.

A conventional crystal oscillator circuit and frequency multiplication could not be used because, firstly, it would hardly be possible to achieve the large frequency deviation, and there was, secondly, a danger that the frequency multiplex signal would be distorted in the frequency multiplication process.

This meant that the solution could only be found when using a self-excited oscillator for 433 MHz with a frequency control loop.

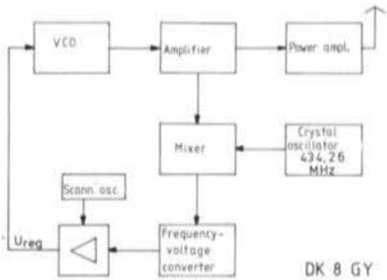


Fig. 1: Block diagram of the frequency control loop

2. PRINCIPLE

The concept shown in **Figure 1** is based on experiments made by DL 9 FX (1).

The VCO frequency (output frequency of the transmitter) is mixed down to a lower IF frequency in conjunction with a crystal controlled frequency. The IF signal is then converted to a DC-voltage U_{reg} in a frequency-voltage converter, which is fed, after amplification, directly to the varactor diode of the VCO. If the VCO frequency increases, U_{reg} will be reduced and the capacitance of the varactor will increase, thus reducing the frequency of the VCO.

3. FREQUENCY-VOLTAGE-CONVERTER

Since this is the most important part of the control loop, various circuits have been studied:

3.1. Schmitt-Trigger

A Schmitt-trigger will convert a sufficiently high input voltage into square wave pulses which can be fed to an integrating RC-link where it is converted to a frequency-dependent DC-voltage.

A conventional circuit using transistors exhibited a considerable dependence of the output voltage on the amplitude of the input signal, and thus not only on its frequency.

Another circuit using the integrated Schmitt-trigger SN 7413 N was also found to be unfavorable due to its low slope, i.e. the dependence of the output DC-voltage on the frequency of the input AC-voltage. Furthermore, the shape of the square wave pulses was dependent on the input amplitude, which means that U_0 was also dependent on the amplitude at the input U_{in}.

3.2. Monoflop

A conventional transistor circuit exhibited such a low slope that it was completely unusable.

The integrated monoflop SN 74121 N possessed a constant slope of 1.02 mV/kHz in the range of 500 kHz to 1500 kHz. The circuit is given in **Figure 2**.

However, this slope is still not sufficient for a control circuit, although the large linear range would be very favorable. It should be possible to connect a subsequent control amplifier.

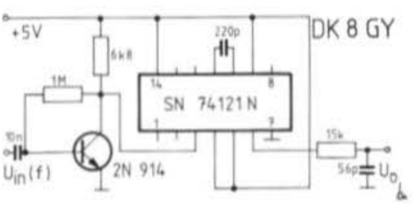


Fig. 2a: Monoflop circuit

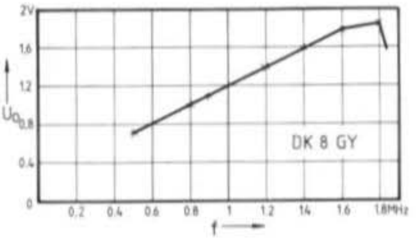


Fig. 2b: Discriminator characteristic of the Monoflop

3.3. FM Demodulator Module TCA 420 A

The TCA 420 A is an integrated IF amplifier with coincidence demodulator and control voltage output. A DC-voltage is present at the two AF output connections that is dependent on the input frequency. The connection circuit is given in **Figure 3**. This voltage is used as control voltage. **Figure 4** shows the characteristic curves for two different values of the damping resistor R.

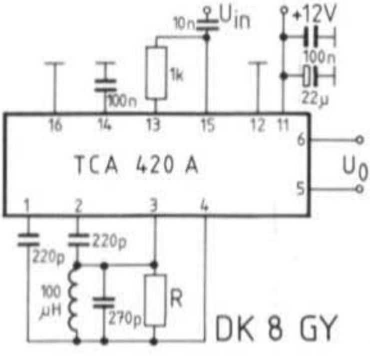


Fig. 3: FM Demodulator TCA 420 A

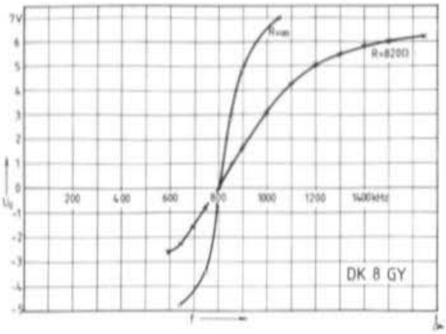


Fig. 4:
Differential DC-voltage U_0 of the TCA 420 A as a function of the intermediate frequency; with damping resistor as parameter

4. OPERATION

A practical circuit built-up using this circuit showed that the VCO was controlled but not quickly enough. It was found that it was possible to cause frequency deviations of several tens of kHz by placing one's hand in the vicinity of the oscillator without the control voltage changing sufficiently to bring the VCO back to the nominal frequency. This means that the control slope is still too low. Of course, an increase of the slope will mean that the lock-in and hold range will be considerably limited. The solution of this problem was found in a control circuit which will be discussed later.

The control slope was increased using the following measures:

a) Firm coupling of the varactor diode to the oscillator circuit, as can be seen in **Figure 5**;

b) Installation of a control amplifier having a gain of 5-times. The zener diode is used to shift the potenital to a favorable input signal for the transistor **(Figure 6)**;

c) Removal of the damping resistor (R in Figure 3) of the discriminator circuit (phase-shift circuit) of the TCA 420 A.

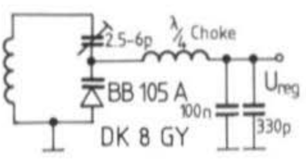


Fig. 5:
Resonance circuit of the 433 MHz oscillator

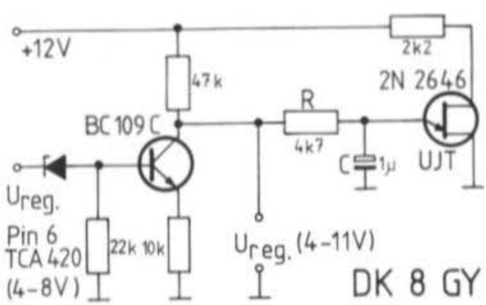


Fig. 6:
Control amplifier and scanning oscillator

After carrying out these modifications, the VCO very seldom locked onto the nominal frequency of 433.4 MHz after switching on, and when it did, it could easily be unlocked by mechanical shock. Instead of this, it oscillated nearly always at a frequency in the vicinity of the crystal oscillator frequency or at any frequency up to 438 MHz. In this case, the DC-voltage at the output of the TCA 420 A sunk to a minimum value and, due to the inverting process, the control voltage at the collector of the control amplifier increased to a maximum value of approx. 10.6 V.

This was used then to switch on a scanning oscillator that drives the VCO with a sawtooth voltage, and sweeps the frequency range from the lower end of the band. The control circuit will lock in as soon as the nominal frequency of 433.4 MHz is reached. The scanning oscillator is shown in **Figure 6**.

Capacitor C is charged via R. As soon as the switching voltage adjusted by R is reached, the E-B_1 path will conduct and discharge C; this will cause the potential present at the collector of the control amplifier to drop immediately due to the high resistance of the collector resistor. This means that the switching voltage threshold will be exceeded so that the unijunction transistor blocks again. The capacitor will then slowly charge itself up to the collector voltage, and the control voltage will slowly increase. As soon as the nominal frequency is achieved (U_{reg} = approx. 6 V), the control loop will lock in. The time constant of the sawtooth voltage is in the order of milliseconds, which means that the control process does not cause any noticeable interference.

A frequency stability of 10^{-5} was achieved using these measures. It was found that the scanning oscillator only made one scan on switching on the unit, and the nominal frequency locked in virtually immediately. The switching voltage was set to approx. 7 V. The trimmer in the oscillator circuit was tuned so that the nominal frequency is present at U_{reg} = 6 V. A fine tuning of the frequency is possible under locked conditions using the core of the discriminator inductance (TCA 420 A).

5. CONTROL CHARACTERISTICS

The control loop between the control amplifier and the VCO was disconnected for measurement of the control curve. The control voltage U_{reg} was then measured as a function of the oscillator frequency. During this measurement, the oscillator was operated with an external voltage of 0 - 11 V which was fed to the varactor diode **(Figure 7)**.

The control range is between 433.3 and 433.5 MHz. Below 433.3 MHz and in excess of 435.1 MHz, U_{reg} is so great that the scanning oscillator will be brought into operation. This is also true at the oscillator frequency of 434.46 MHz where the intermediate frequency is 0.

The VCO will immediately jump to the crystal oscillator frequency, when it is in the vicinity of it.

No stable operation can be made in the range between 433.5 and 434.1 MHz, since the curve increases here and the VCO frequency will not be able to lock in. In the range from 434.5 MHz to 435 MHz, the curve drops, and it would, theoretically, be possible for the VCO to lock in. However, it was found that this did not take place. The frequency always jumped, as previously mentioned, to values at which the control voltage achieved a maximum value and switched on the scanning oscillator.

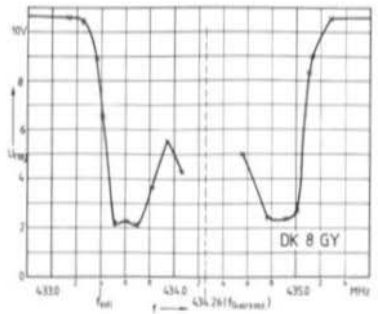


Fig. 7: Control voltage as a function of the oscillator frequency with the control loop open

6. POSSIBLE IMPROVEMENTS

The circuit has, up to now, been operated experimentally, and provided satisfactory results. However, a few points should be noted so that they can be taken into consideration on developing this circuit further:

a) The VCO should be built up in stripline technology in order to improve the mechanical stability.

b) Increase the IF of 850 kHz, using a different crystal frequency, to, for instance, 10.7 MHz, so that the crystal oscillator frequency is further away from the required frequency.

c) A high-Q resonant IF circuit should be provided at or after the mixer. This means that the TCA 420 A will receive too low an input voltage at great deviations from the intermediate frequency, which will mean that the input voltage is too low and the resulting maximum control voltage will cause the scanning oscillator to switch on earlier.

7. RF CIRCUIT

The VCO, amplifier, mixer, power amplifier were conventional circuits using BFY 90 and BFW 16 A (power amplifier = 50 mW).

8. REFERENCES

G. Hoffschild: The AFC Loop
VHF COMMUNICATIONS 9, Edition 3/1977, pages 184 - 188

NOTES AND MODIFICATIONS

1. ATV TRANSMITTER DJ 4 LB 001-006

Günter Sattler, DJ 4 LB, has been able to improve the linearity and intermodulation ratio of his ATV transmitter in time-consuming experiments in conjunction with a spectrum analyzer. The following small modifications allow the intermodulation ratio at the output of module DJ 4 LB 007 to be greater than 50 dB, at a maximum output level of – 3 dBm.

1.1. Modifications to Module DJ 4 LB 001 (a)

A transistor type BF 199 should be used for T 105 instead of BF 224, and resistor R 115 should be changed to be 470 Ω (instead of 680 Ω). Since the output impedance at Pt 103 only amounts to 45 Ω, a resistor of 5 Ω (for 50 Ω cables) should be connected in series with the output coupling capacitor C 113 (1 nF). Both measures improve the linearity which is especially noticeable at the modulation peaks.

2.2. Modifications to Module DJ 4 LB 007

The transistor type BF 199 should be used in the RF-stages instead of the original BF 224. In the output coupling stage T 3, transistor type BF 223 = BF 311 will be more suitable due to the higher current (15 mA). The emitter resistor of T 3 should be reduced from 560 Ω to 330 Ω.

Since the output impedance of this stage (Pt 705) only amounts to 22 to 25 Ω, it is necessary for 25 to 39 Ω to be connected in series in order to match it exactly to the subsequent side-band filter. Transistor T 2 which is used for remote level adjustment will cause noticeable modulation distortions and intermodulation products if the control is more than 1/3 open. This can be avoided using the following measures: Reduce the value of the coupling capacitor between the interconnection of 10 Ω/680 Ω at T 2 from 0.1 μF to 150 pF. A FET, type BF 246 C should be used instead of the BF 224 for transistor T 2. The connections are as follows: drain for collector, gate for emitter (ground) and source for base. The source is connected via a dropper resistor of 10 kΩ to potentiometer P 1, whose value should be increased from 220 Ω to 10 kΩ. Diode D 1 is bridged from the potentiometer to + 12 V with 8.2 kΩ instead of the previous 5.6 kΩ.

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 3/1978 of VHF COMMUNICATIONS

DC 0 DA 006	**9 cm LOCAL OSCILLATOR MODULE**		**Ed. 3/1978**
PC-board	DC 0 DA 006	(double-coated, without thru-contacts, w/plan)	DM 24.—
Semiconductors	DC 0 DA 006	(6 transistors, 2 diodes)	DM 52.—
Minikit	DC 0 DA 006	(10 foil, 2 ceramic trimmers, 1 chip, 10 ceramic-caps, 1 tantalum-electrolytic, 1 coilformer w/core, 1 6-hole ferrite core, 5 ferrite beads)	DM 24.—
Crystal	92.000 MHz	HC-25/U	DM 34.—
Kit	**DC 0 DA 006**	**with above parts**	**DM 129.—**
Mixer diodes for the above converter or others for 70 cm to 9 cm, each			DM 29.—

DC 0 DA 007	**145 MHz (IF) PREAMPLIFIER**		**Ed. 3/1978**
PC-board	DC 0 DA 007	(double-coated, without thru-contacts, w/plan)	DM 12.—
Transistor	BF 900	(each)	DM 4.50
Minikit 1	DC 0 DA 007	(1 ferrite choke, 3 foil trimmers, 8 ceramic capacitors)	DM 8.50
Minikit 2	DC 0 DA 007	(6 resistors, 1 m silverplated copper wire)	DM 3.50
Kit	**DC 0 DA 007**	**with above parts**	**DM 27.—**

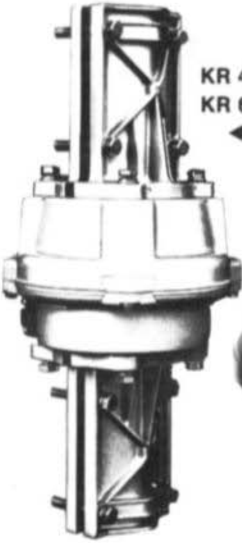
DC 1 QW 001-003	**80-CHANNEL SYNTHESIZER for 2 m**		**Ed. 3/1978**
PC-board	DC 1 QW 001	(double-coated, without thru-contacts,w/plan)	DM 16.—
PC-board	DC 1 QW 002	(as DC 1 QW 001)	DM 16.—
PC-board	DC 1 QW 003	(single-coated, with plan)	DM 11.—
Semiconductors	DC 1 QW 001-3	(13 transistors, 36 diodes, 2 LED)	DM 78.—
Semiconductors 2	DC 1 QW 001-3	(8 C-MOS ICs, 1 Op.amp)	DM 49.—
Minikit 1	DC 1 QW 001-3	(2 chokes, 12 foil trimmers)	DM 14.—
Minikit 2	DC 1 QW 001-3	(36 ceramic capacitors, 20 feedthru-caps., 4 tantalum electrolytics)	DM 42.—
Minikit 3	DC 1 QW 001-3	(Filter) (6 1% resistors, 4 2.5% caps.)	DM 7.—
Kit	**DC 1 QW 001-3**	**with above parts**	**DM 230.—**
Crystal	71.650 MHz	HC-25/U	DM 26.—
Crystal	66.300 MHz	HC-25/U	DM 26.—
Crystal	1.600 MHz	HC- 6/U	DM 35.—

Price reduction DJ 7 VY LOW-NOISE PREAMPLIFIER, 50 - 570 MHz

PC-board	DJ 7 VY 001	(single-coated with plan)	DM 8.—
Minikit	DJ 7 VY 001	(2 ferritecores, 2 ferrite chokes, 2 ferrite beads, 4 disc caps., 1 tantalum electrolytic, 2 trimmer potentiometers)	DM 15.50
Semiconductors	DJ 7 VY 001	(2 transistors)	DM 19.—
Kit	**DJ 7 VY 001**	**with above parts**	**DM 42.—**
if required	4 Schottky diodes		DM 30.—

METEOSAT-Converter, as described, ready-to-operate	**DM 985.—**
Converter 135/28 to convert above to 28 MHz IF	**DM 120.—**

ANTENNA ROTATING SYSTEMS



KR 400
KR 600
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KR 2000
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MXX
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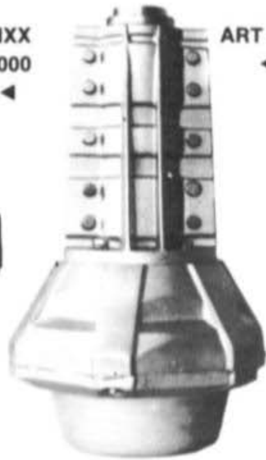

ART 8000
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SPECIFICATIONS

Type of Rotator	KR 400	KR 600	KR 2000	MXX 1000	ART 8000	
Load	250	400	800	1000	2500	kg
Pending torque	800	1000	1600	1650	2450	Nm *)
Brake torque	200	400	1000	1200	1400	Nm *)
Rotation torque	40	60	150	180	250	Nm *)
Mast diameter	38 - 63	38 - 63	43 - 63	38 - 62	48 - 78	mm
Speed (1 rev.)	60	60	80	60	60	s
Rotation angle	370°	370°	370°	370°	370°	
Control cable	6	6	8	7	8	wires
Dimensions	270 x 180 ∅	270 x 180 ∅	345 x 225 ∅	425 x 205 ∅	460 x 300 ∅	mm
Weight	4.5	4.6	9.0	12.7	26.0	kg
Motor voltage	24	24	24	42	42	V
Line voltage	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	
	50	55	100	150	200	VA

*) 1 kpm ≙ 9.81 Nm

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Especially designed for vertical tilting of antennas for EME, OSCAR etc.

Type	**KR 500**
Load	ca. 250 kg
Brake torque	197 Nm *)
Rotation torque	40 Nm *)
Horiz. tube diam.	32 - 43 mm
Mast diameter	38 - 63 mm
Speed (1 rev.)	74 s
Rotation angle	180° (+ 5°)
Control cable	6 wires
Line voltage	220 V/50 Hz 30 VA
Weight	4.5 kg

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