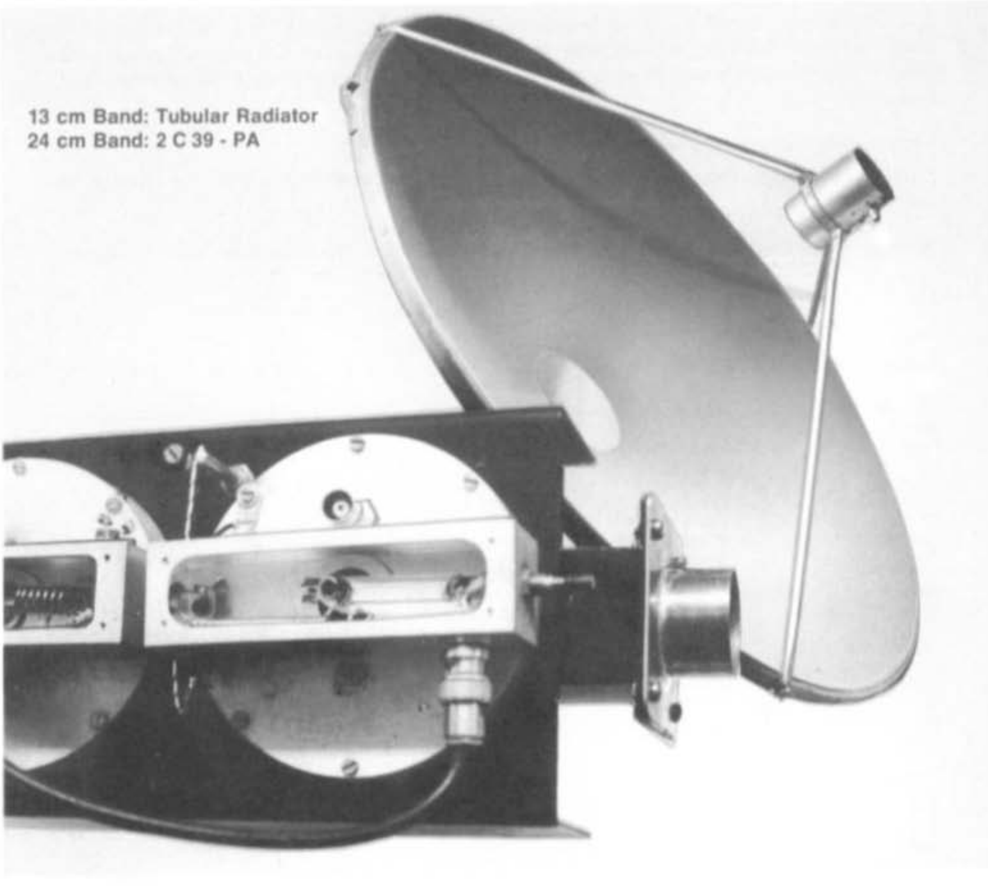
J 20 419 F

VHF COMMUNICATIONS

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

VOLUME NO. 8 WINTER 4/1976 DM 4.50



13 cm Band: Tubular Radiator
24 cm Band: 2 C 39 - PA

VHF COMMUNICATIONS

Published by:
Verlag UKW-BERICHTE · Hans J. Dohlus oHG · Jahnstraße 14 · D-8523 BAIERSDORF
Fed. Rep. of Germany · Telephones (0 91 91) 9157 / (0 91 33) 855, 856.

Publishers:
T. Bittan, H. Dohlus.

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

Advertising manager:
T. Bittan.

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

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

We would be grateful if you would address your orders and queries to your representative :

VERTRETUNGEN: REPRESENTATIVES:

Austria	Hans J. Dohlus, Creditanstalt Bankverein WIEN Kto.17-90.599; PSchKto.WIEN 1.169.146
Australia	WIA PO Box 150, TOORAK, VIC.3142, Tel. 24-8652
Belgium	see Germany, PSchKto. 30455-858 Nürnberg
Canada	see USA
Denmark	E. Halskov, OZ 7 LX, Sigersted Gl. Skole, DK-4100 RINGSTED, Postgiro 7 296 800
France	Christiane Michel, F 5 SM, F-89 PARLY, Les Pillés
Finland	see Sweden
Germany	Verlag UKW-BERICHTE H. Dohlus oHG, Jahnstr.14, D-8523 BAIERSDORF, Tel.09133-855,856 Konten: Postscheckkonto Nürnberg 304 55-858, Commerzbank Erlangen 820-1154, Deutsche Bank Erlangen 76-403 60
Holland	see Germany, Postscheckkonto Nürnberg 304 55-858
Israel	Z. Pomer, 4 X 4 KT, PO Box 222, K. MOZKIN 26100, Tel. 974-4-714078
Italy	STE s.r.l. (I 2 GM) Via Maniago 15, I-20134 MILANO, Tel. (02) 215 7891,Conto Corr.Post.3/44968
Luxembourg	P. Wantz, LX 1 CW, Télévision, DUDELANGE, Postscheckkonto 170 05
New Zealand	E. M. Zimmermann, ZL 1 AGQ, P.O.Box 56, WELLSFORD, Tel. 8024
Norway	Henning Theg, LA 4 YG, Postboks 70, N-1324 LYSAKER, Postgirokonto 3 16 00 09
South Africa	SA Publications, PO Box 2232, JOHANNESBURG 2000, Telephone 22-1496
Spain + Portugal	Julio A. Prieto Alonso, EA 4 CJ, MADRID-15, Donoso Cortés 58 5°-B, Tel. 243.83.84
Sweden	Sven Jacobson, SM 7 DTT, Gamlakommungarden 68, S-23501 VELLINGE, Tel. 040-420430 Postgiro: 43 09 65 - 4
Switzerland	Hans J.Dohlus, Schweiz.Kreditanstalt ZÜRICH, Kto.469.253-41; PSchKto.ZÜRICH 80-54.849
UK	ARBBG, 20 Thornton Cres. OLD COULSDON, CR 3 1 LH
USA - East Coast	VHF COMMUNICATIONS Russ Pillsbury, K 2 TXB, & Gary Anderson, W 2 UCZ, 915 North Main St., JAMESTOWN, NY 14701, Tel. 716 - 664 - 6345
Yugoslavia	Tito Cvrković, YU-56000 VINKOVCI, Lenjinova 36

A PUBLICATION FOR THE RADIO AMATEUR
ESPECIALLY COVERING VHF, UHF AND MICROWAVES
VOLUME NO. 8 WINTER EDITION 4/1976

A CONVERTER FOR THE 13 cm BAND EQUIPPED WITH TWO PREAMPLIFIER STAGES AND AN ACTIVE MIXER

by J. Dahms, DC 0 DA

1. INTRODUCTION

The previously described converters for the 13 cm amateur band (2304 - 2306 MHz) have been equipped with diode mixers (passive mixers) which have been followed by a subsequent low-noise intermediate preamplifier (Ref. 1, 2, 3). Only the converter described in (2) was provided with a preamplifier stage. It is described in (5) how difficult it is to obtain the theoretically possible sensitivity in practice when using external preamplifiers (4). It is therefore favorable to replace the passive mixer by an active transistor mixer, since a conversion gain will be present in this case instead of a conversion loss.

Only few amateurs have made experiments with transistor mixers. Virtually nothing is to be found even in the leading US-magazines such as »ham radio« and »QST«. Since low noise transistors with sufficient gain at 2300 MHz are now available economically, diode mixers are not absolutely necessary for applications on the 13 cm band.
The Siemens transistor type BFR 34 A possesses very favorable specifications: The manufacturer gives a noise figure of 4 dB at 2.0 GHz at a collector current of 3 mA and a collector emitter voltage of 10 V. The values 4.4 dB noise figure and 10 dB gain at 2.3 GHz were given in (5). Of course, these values are only obtained with an optimized input and output matching.

A receive converter 2304 MHz / 144 MHz is to be described, which possesses two selective preamplifier stages and a mixer equipped with the BFR 34 A. The oscillator circuit is constructed in a similar manner to that described in (6), and is designed as a separate module. **Figure 1** shows the photograph of the converter and oscillator module.

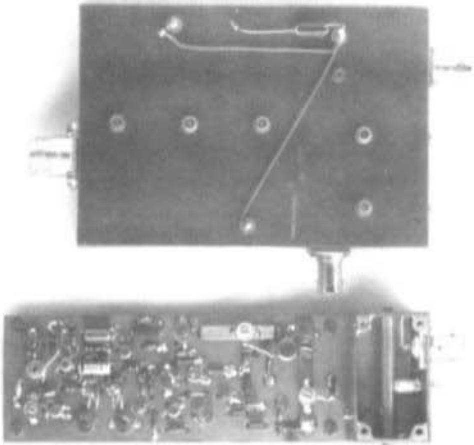


Fig. 1: The 13 cm converter with its oscillator module (below)

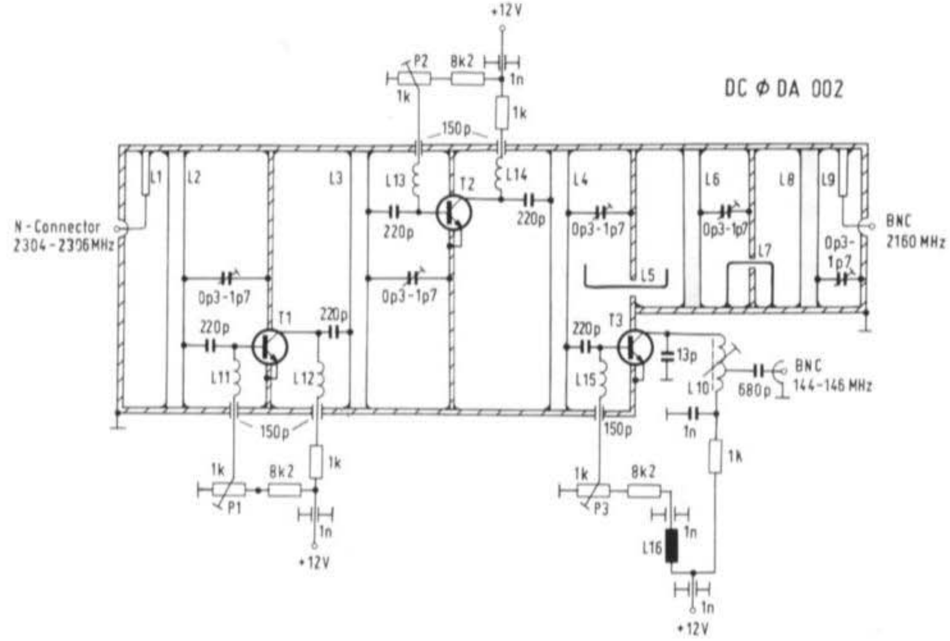


Fig. 2: 13 cm converter with 2 preamplifier stages and an active mixer

2. CIRCUIT OF THE CONVERTER

The circuit diagram of the receive converter is given in **Figure 2**. It will be seen that five coaxial λ/2 lines are used that are capacitively tuned at the center (max. voltage). The first three circuits (L 2, L 3, L 4) are aligned to the required frequency, the other two (L 6, L 8) to the oscillator frequency.

All three transistors (BFR 34 A) operate with the emitter directly grounded, which means that the operating point adjustment made by trimmer potentiometers is not stabilized with respect to voltage and temperature variations. Whereas the operating voltage of the whole converter can easily be stabilized, the gain and noise figure will deteriorate with extreme temperature fluctuations, for instance, as encountered in portable and mobile operation. If this is the case, a simple resistor network can be calculated for the base voltage as was given in (7) so that the collector current is sufficiently stable with respect to temperature. In the author's case, this bias voltage circuit was not used since it required a higher operating voltage, e.g. 20 V.

As was recommended in (5), transistor T 1 is aligned for the lowest noise figure, whereas transistor T 2 is adjusted for maximum gain. Siemens publishes the following values:

U_{CE} = 10 V; I_C = 3 mA for the lowest noise figure at 2 GHz
U_{CE} = 6 V; I_C = 15 mA for maximum power gain (transit frequency)

The author operates T 2 only with an I_C of 6 mA; in order to obtain higher currents, it is necessary for the collector resistor to be reduced from 1 kΩ to 470 Ω.

Mixer transistor T 3 is aligned to Class B without oscillator signal (collector current just visible on the meter, in other words several µA). The oscillator signal will increase the collector current to several mA, which represents a good indication for the tuning of the oscillator bandpass filter and for the coupling to L 5. Any UHF-components in the collector AC-voltage of the mixer will be shorted out at the collector by using a disc capacitor of approx. 10 - 15 pF (value uncritical and only type of capacitor of importance). This capacitor forms, at the same time, the circuit capacitance for 145 MHz. The output signal is fed via a 50 Ω tap of the IF inductance L 10 to the output socket. An IF preamplifier is not required in this concept since the three stages possess sufficient overall gain of 25 - 30 dB.

The bandpass filter for the local oscillator frequency comprising L 6 and L 8 is possibly not ideal with respect to the insertion loss. The author is quite sure that some of the oscillator energy is lost due to the coupling from L 5 to the 2300 MHz circuit comprising L 4. However, since the local oscillator circuit provides at least 10 mW at 2160 MHz this is not important. It is more important that the mixer is provided with a clean local oscillator signal.

3. CONSTRUCTION OF THE CONVERTER

Partly single-coated and partly double-coated PC-board material of 1.5 mm in thickness was used for the prototype. The constructional diagram given in **Figure 3** and the subsequent photographs give both the dimensions and construction details. The author recommends that 0.7 mm thick brass plate be used for construction, since this eases the problem of making the »through contacts« for the emitter connections, and means that it is only necessary to solder on one side. The case can also be silver-plated before mounting the electronic components (although the author's prototype was not). The following table gives the dimensions of the 1.5 mm thick PC-board pieces.

Parts required	Dimensions	Coating
2	138 x 27 x 1.5	single
2	138 x 95 x 1.5	single
2	92 x 27 x 1.5	single
1	92 x 27 x 1.5	double
2	84 x 27 x 1.5	single
4	49 x 27 x 1.5	double

As can be seen from the dimensions, the inner height of the converter is 27 mm. Double-coated PC-board material is only used for the intermediate panels which are adjacent to RF levels. No metal surfaces are present on the outside of the converter in order to avoid that ground connections are made on the outside instead of the inside.

The signal from the antenna is injected via a N-connector and the coupling link L 1. The outer conductor of the connector is soldered around the edge with the inside of the case in order to ensure a no-inductance ground connection at the correct impedance. The coupling link L 1 is 16 mm in length (+ 5 mm bend), 5 mm wide and spaced approximately 1.5 mm from L 2. Since the resonance of this first circuit has been found to be very sharp, it is possible that the input coupling is not quite ideal; however, this can also be caused by the coupling to the first transistor. Unfortunately, the author does not possess the extensive

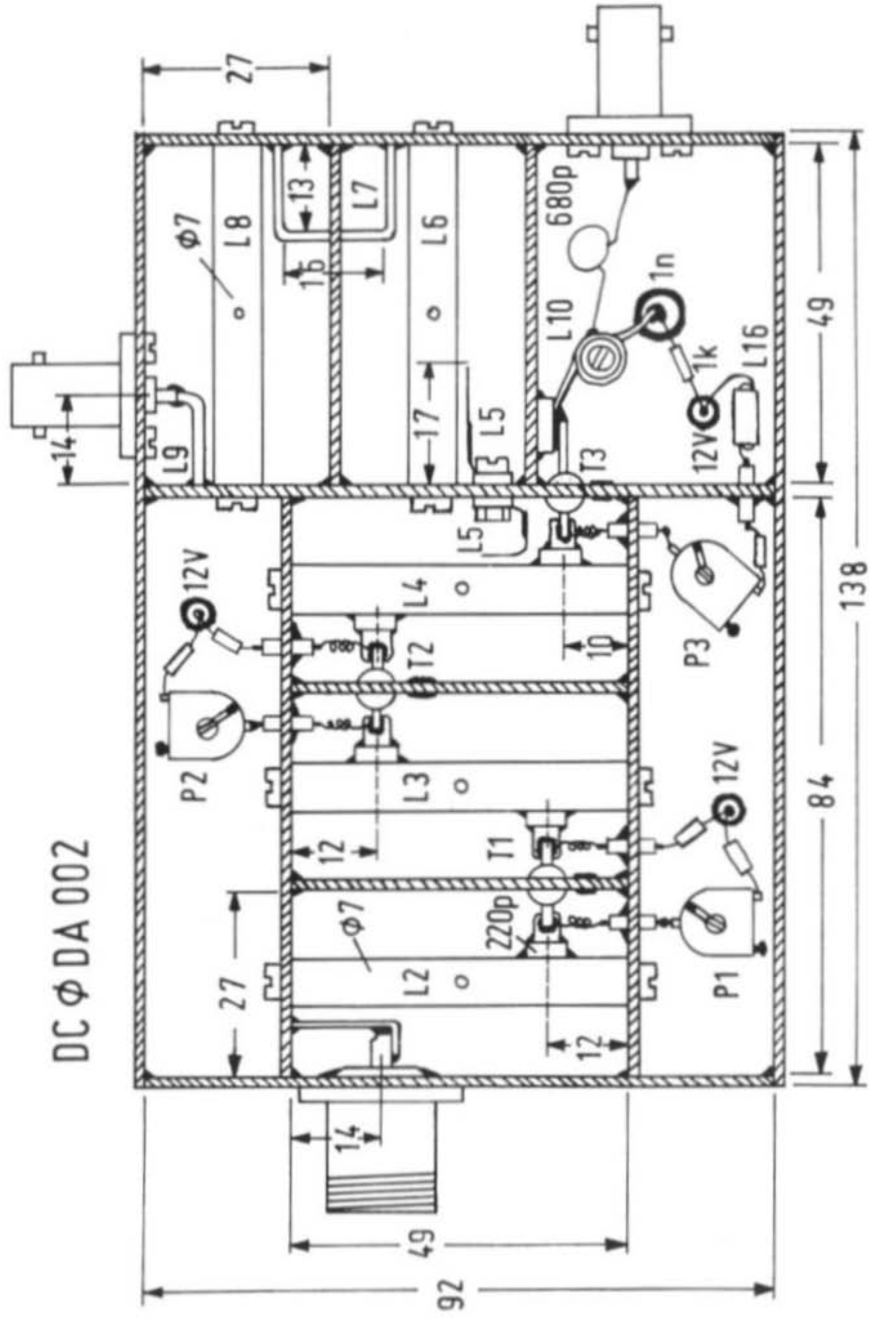


Fig. 3: Constructional drawing of the 13 cm converter

measuring equipment necessary to optimize the converter completely. The author and the editors would be grateful if one of our readers could determine the ideal dimensions so that they can be published.

All five resonant cavities have a square cross section (27 mm) and are 49 mm long. The half-wave inner conductor is made from 7 mm dia. brass tube. M 3 (3 mm) brass nuts are pressed into the open ends of the tube and soldered into place. The nuts should be placed deep enough into the tube so that the edge of the tube protrudes approximately 1 mm further. After filing the ends of the tube flat, the edge of the tube will be forced onto the tinned coating on the inner surface of the chamber and form a good ground contact all around the edge of the tube.

High-quality ceramic spindle trimmers with a capacitance of 0.3 - 1.7 pF are used as tuning capacitors. The trimmers are soldered to the bottom of the case, and their hot ends are placed into the center holes of the $\lambda/2$ circuits, but not soldered. If no trimmers are available with such a low capacitance, the metal surface of higher capacitance spindle trimmers can be removed and the ceramic part placed into the inner conductor. The capacitance between the spindle and the conductor will then be sufficiently small. Of course, it is also possible to use a fine screw and metal disc to achieve the required capacitance.

Holes should be provided in the intermediate panels to provide a tight fit for the transistors. A thin, silver-plated copper foil strip is soldered to each emitter, which is folded down on both sides and soldered to ground surface after the transistor has been inserted into place. This ensures the shortest possible ground connection and a satisfactory »through contacting« at this position.

The tapping point for the base and collector of the preamplifier stages is approximately 12 mm from the cold end. In the case of the mixer, it is approx. 10 mm from the cold end. A ceramic disc capacitor of approx. 220 pF is soldered to the resonant line and the connection to the transistor is made using a 4 mm wide silver-plated strip of copper foil. This means that the narrow transistor connections have virtually no effect. The operating voltages are fed via $\lambda/4$ chokes to the disc capacitors. The cold end of the chokes is connected to the feed-through capacitors of 100 to 200 pF.

A teflon feedthrough should be used for the coupling from L 6 to L 4, which is then provided with 4 mm wide coupling strips bent as shown in **Figure 3**.

The collector of the mixer transistor is fed into a separate chamber for the IF resonant circuit. A disc capacitor of 10 to 15 pF is provided directly adjacent to the transistor connection on the intermediate panel. The variable IF inductance is connected to this position. The output coupling is made 2 turns from the cold end via a capacitor of 680 to 1000 pF. A disc capacitor of approx. 1 nF is soldered to the bottom of the chamber for bypassing the IF circuit.

The base voltage dividers comprising the trimmer potentiometers for alignment of the quiescent currents are accommodated in their own chamber and are provided with the individual operating voltage via feedthrough capacitors of 1 nF. This ensures that no unwanted coupling is made from chamber to chamber, and that no tendency to oscillation exists. After the converter has been completed, M 3 (3 mm) brass nuts are soldered to the corners of all chambers and are flattened off. The cover should be tinned around all holes and then screwed into place with the aid of 22 screws.

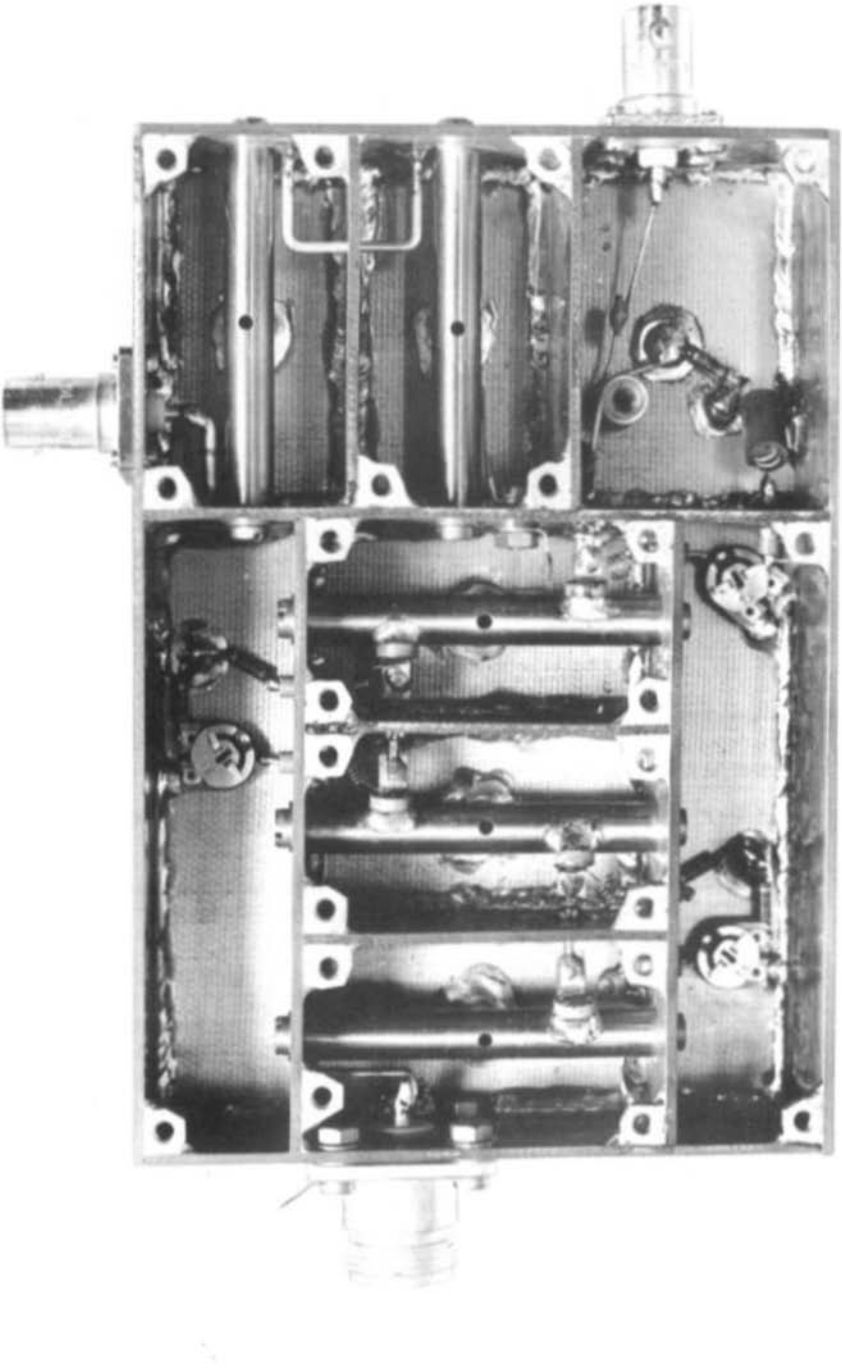


Fig. 4: 13 cm converter with two preamplifier stages and an active mixer

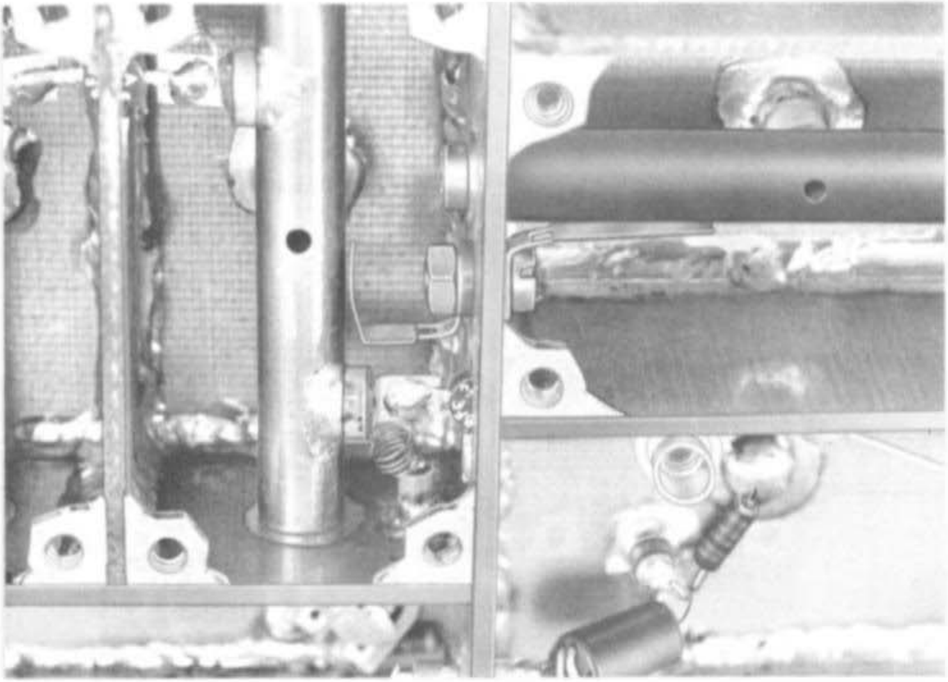


Fig. 5: Further details regarding the mixer stage

3.1. Special Components

T 1 - T 3:	BFR 34 A (Siemens)
L 1:	Brass plate, approx. 14 mm long, 4 mm wide, 1.5 - 2 mm spacing to L 2
L 2 - L 4, L 6, L 8:	see text
L 5:	copper foil appr.4 mm wide, 12-15 mm long, soldered to teflon feedthrough
L 7:	1.5 mm dia. (15 AWG) silver-plated copper wire, 14 mm high, spaced approx. 2 mm to L 6 and L 8
L 9:	1.5 mm dia. (15 AWG) silver-plated copper wire, 13 mm long, spaced 1 - 2 mm from L 8
L 10:	6 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm coil former with VHF core, coil spacing 1 mm, tap approx. 2 turns from cold end
L 11 - L 15:	$\lambda/4$ chokes of 0.3 - 0.4 mm dia. (27 AWG) enamelled copper wire, 3 turns wound on a 4 mm former slightly pulled out, self-supporting
L 16:	6-hole ferrit choke (Philips)

5 ceramic disc capacitors without connections of approx. 220 pF

1 ceramic disc capacitor without connections of approx. 12 pF

1 ceramic disc capacitor without connections of approx. 1 nF

5 ceramic spindle trimmers of approx. 0.3 - 1.7 pF

6 ceramic feedthrough capacitors for solder mounting, as small as possible of approx. 150 pF

3 ceramic feedthrough capacitors for solder mounting, approx. 1 nF

3 trimmer potentiometers, 1 kΩ, spacing 10/5 mm.

4. ALIGNMENT AND OPERATION

The alignment is commenced by adjusting the quiescent currents of the three transistors as was described in Section 2. After this, the local oscillator module is connected and the resonant circuits comprising L 6 and L 8 are aligned for maximum current of the mixer transistor. After aligning the IF circuit at 145 MHz, it should be possible to hear a considerable increase in noise on the 2 m receiver. It is necessary for the cover to be mounted into place for the final alignment of the $\lambda/2$ circuits and a stable signal should be present at the antenna input. In the case of the author's prototype, an overall gain of approximately 28 dB was provided at an operating voltage of 12 V; the overall gain dropped by approximately 2 to 3 dB at the band limits.

The resonance of the input circuit and the last 2160 MHz bandpass filter had been found to be very sharp. However, no tendency to oscillation was found even after the cover had been removed. Inspite of the author's good VHF location, no strong repeater or SSB stations on the 2 m band had been heard to break through. No unwanted conversion products had been observed from a television transmitter located only 3 km from the author.

The described converter was compared by the author to the converters described in (1) and (2). Both a strong local beacon transmitter and a weak beacon signal at a distance of 60 km at the band limit were available. The results were satisfactory, even if the most favorable operation has still not been obtained due to the non-availability of the measuring equipment. The author is sure that the sensitivity can be improved still further by providing a more favorable matching between the transistors and the $\lambda/2$ circuits. In this respect, the author recommends the following modifications:

The base and collector connections of the transistors are placed on small teflon supports provided with a small screw, and provided with copper foil strips. The chokes will also be fed to the supports. It is possible, using a plastic tool, for the coupling to be optimized. This recommendation is similar to that given in (4). The author would like to hear from readers regarding their experience on further development of this converter.

5. REFERENCES

(1) R.E. Fisher: Interdigital converters for 1296 and 2304 MHz
QST, Volume 58 (1974), Edition 1 (January), Pages 11 - 15

(2) K. Hupfer: A stripline converter for the 13 cm band
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 238 - 245

(3) A. Schädlich: A Receive converter for the 13 cm band with diode mixer
VHF COMMUNICATIONS 7, Edition 3/1975, Pages 161 - 167

(4) N.J. Foot: Narrow-band solid state 2304 MHz preamplifiers
ham radio magazine 7 (1974), Edition 7 (July), Pages 6 - 11

(5) D. Vollhardt: Preamplifier and mixer noise at SHF
In this edition of VHF COMMUNICATIONS

(6) K. Hupfer: A 2160 MHz local oscillator for 13 cm converters
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 246 - 247

(7) K. Richter: Design DC stability into your transistor circuits
MICROWAVES, Volume 12 (1973), Edition 12 (December), Pages 40 - 46

A CONVERTER FOR THE 13 cm BAND EQUIPPED WITH 2 PREAMPLIFIER STAGES AND AN ACTIVE MIXER

PART 2: THE LOCAL OSCILLATOR MODULE

by J. Dahms, DC 0 DA

In the 13 cm band, the frequency range from 2304 to 2306 is used for crystal-controlled voice communications. In order to convert this frequency band to an intermediate frequency range of 144 to 146 MHz, a crystal-controlled local oscillator signal of 2160 MHz is required. A local oscillator module providing this frequency is now to be described. When used as local oscillator module, a 90 MHz crystal is used. If this crystal is replaced by a 96 MHz type, and the resonant circuits tuned slightly higher, the resulting frequency will be 2304 MHz. This means that the module is also suitable for use as low-power transmitter or beacon, since its output power amounts to at least 5 mW with a DC input power of 12 V x 50 mA. **Figure 1** shows a photograph of the author's prototype.

1. CIRCUIT DESCRIPTION

The electrical circuit is similar to the local oscillator circuit DJ 4 LB 003 (1), and that of DJ 1 EE (2). The 90 MHz crystal-oscillator frequency is tripled to 270 MHz, and finally doubled to 540 MHz. This is followed by two straight amplifier stages. Sufficient power is now available in order to drive a frequency quadrupler equipped with a Schottky diode. The power at 540 MHz should be between 50 and 200 mW. The output power at the required frequency will be more than 5 mW even when less efficient transistors are used. The circuit was designed to have more power than necessary in order to have a sufficient reserve in the case of transistor gain spread and to allow for losses in filters or long, thin coaxial cables between the local oscillator module and mixer.

The circuit diagram given in **Figure 2** shows that a bandpass filter is used after each frequency multiplication process so that unwanted harmonics of the crystal frequency etc. are not present in the output spectrum. The two 540 MHz amplifier stages have only been provided with relatively simple resonant circuits in order to keep losses at a minimum. These are made in the form of striplines constructed from silver-plated brass strips. They possess a sharp resonance.

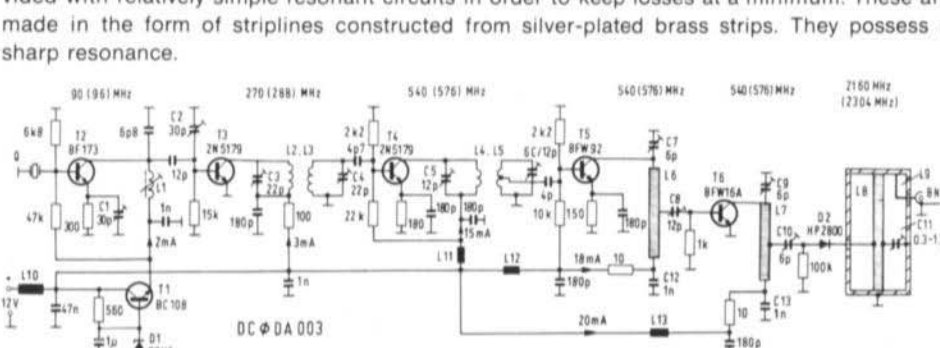


Fig. 2: Circuit diagram of the local-oscillator module DC 0 DA 003

Fig. 1: Photograph of the author's prototype of the local-oscillator module for 2160 or 2304 MHz

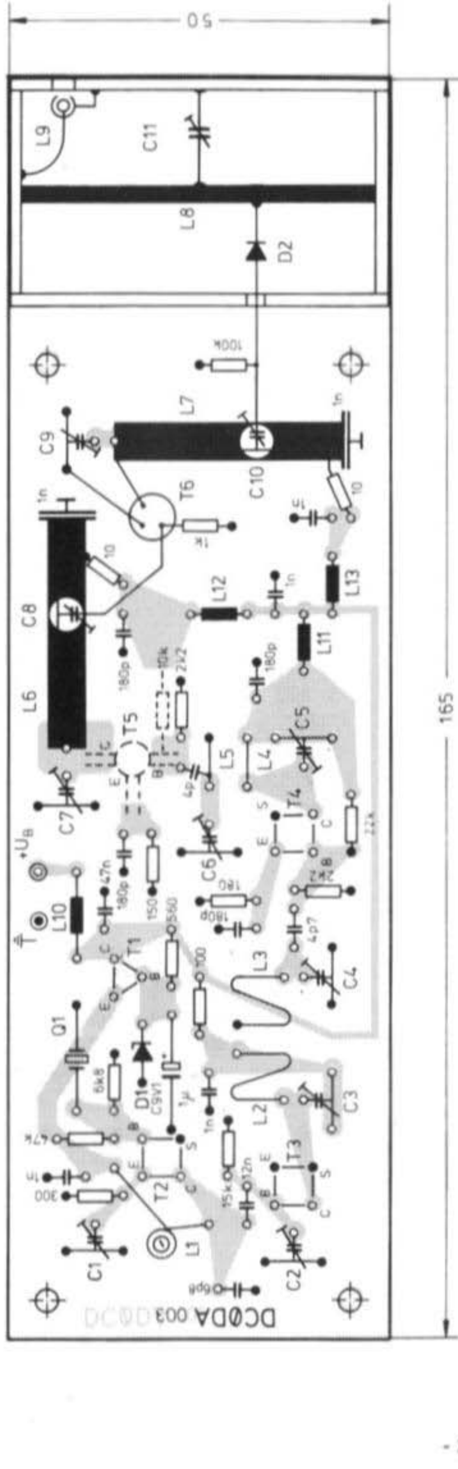


Fig. 3: Component locations and PC-board DC 0 DA 003

The subsequent quadrupler is followed by a single resonant circuit, but since this is in the form of a coaxial λ/2 circuit, the resonance is extremely sharp. An additional bandpass filter for the local oscillator frequency is provided in the converter module.

The current values given in the circuit diagram for the individual stages are given for orientation. These values were measured in the author's prototype when using the transistors given in the circuit diagram. They were measured as a voltage drop across the collector dropper resistors, or across the emitter resistors. The operating points of transistors T 4 and T 5 can be changed by exchanging the emitter resistors.

Such a local-oscillator module will only operate satisfactory when the individual stages have been consequently choked, decoupled, and bypassed. The higher the working frequency, the lower should be the capacitance values of the bypass capacitors so that the series resonance frequency formed by the capacitance and the inductance of the leads is lower than the frequency to be bypassed. Values of 100 pF and less should be used at frequencies in excess of 250 MHz. For example, the reactive impedance of a 100 pF capacitor at 435 MHz is only 3.66 Ω. This means that the values given in the circuit diagram are not too critical. The resonant circuits should be tuned with plastic-foil trimmers of 7.5 mm diameter. Ceramic trimmers are less favorable for several reasons. Either λ/4-coils or ferrite beads with 3 turns of enamelled copper wire should be used as chokes.

2. SPECIAL COMPONENTS

T 1:	BC 108 or other AF transistor
T 2:	BF 173, BF 224, BF 199
T 3:	2 N 5179, BFX 89, BF 224, BF 199
T 4:	2 N 5179, BFX 89
T 5:	BFW 92 (Philips, Siemens), BFR 90, BFR 34
T 6:	BFW 16 A (Philips, Siemens), 2 N 3866
D 1:	C 9 V 1 Z-diode
D 2:	HP 2800, 2835, 2817, 2811, 2900 (Hewlett Packard)

The inductances are made from 1 mm dia. (18 AWG) silver-plated copper wire.

L 1:	6 turns wound on a 5 mm coil former with core, turns spaced 1 mm, hot end facing towards the board
L 2, L 3:	1.5 turns wound on a 5 mm former, self-supporting, spaced 4 to 5 mm above the ground surface, fit into the holes on the board and solder into place.
L 4, L 5:	U-shaped wire bent around a 6 mm former, spaced 8 mm above the ground surface.
L 6, L 7:	Silver-plated metal strip, 0.5 mm thick, 5 mm wide, 30 mm long, spaced 4 mm from the surface of the board.
L 8:	Brass tube 7 mm dia., 47 mm long, M 3 nuts soldered to each end
L 9:	11 mm long wire as for the coils, spaced approx. 2 mm from L 8.

Tapping points:

L 6:	13 mm from cold end
L 7:	10 mm from cold end
L 8:	diode: 10 mm from cold end; hole for trimmer 21 mm from cold end

Connection of trimmer C 8 to the base of T 6:	
	12 mm silver-plated copper wire of 1 mm dia.
L 10:	6-hole ferrite core choke
L 11 - L 13:	see text
C 1, C 2:	approx. 30 pF plastic-foil trimmer, 7.5 mm dia. (Philips, colour red)
C 3, C 4:	approx. 22 pF plastic-foil trimmer, 7.5 mm dia. (Philips, colour green)
C 5, C 6, C 8:	approx. 12 pF plastic-foil trimmer, 7.5 mm dia. (Philips, colour yellow)
C 7, C 9, C 10:	approx. 6 pF plastic-foil trimmer, 7.5 mm dia. (Philips, colour grey)
C 11:	0.3 - 1.7 pF ceramic spindle trimmer
C 12, C 13:	approx. 1 nF ceramic disc capacitor, without leads
Crystal:	90 (96) MHz, fifth overtone, 20 pF, HC-25/U or HC-6/U

Other capacitors ceramic disc types for 5 mm spacing.
All resistors are approximately 2 mm in diameter, and approx. 7 mm long.

3. CONSTRUCTION DETAILS

A double-coated PC-board is used to provide good ground conditions. This board has been designated DC 0 DA 003 and its dimensions are 165 mm x 50 mm **(Figure 3)**. The copper coating on the component side of the board remains intact except for a small area around the holes that it removed with the aid of a 3 mm drill. The actual conductor lanes are to be found on the lower side of the board. On installing the PC-board into the case, it is necessary for it to be spaced at least 5 mm from the nearest ground surface.

Transistor T 5 and its base resistor of 10 kΩ are mounted on the lower side of the board on the larger copper surface provided for this purpose. Caution should be taken since soldering is to take place very near to the case of the transistor. The soldering process should be carried out quickly and made with a small, 15 W soldering iron.

The cold ends of the two stripline circuits for 540 MHz are connected to round bypass capacitors of 6 to 10 mm diameter, whose value is, as has been previously mentioned, relatively uncritical. The copper surface of the PC-board and the contact surface of the capacitors should be tinned before soldering these capacitors into place.

The frequencies of the crystal oscillator and the multipliers or amplifier stages can be checked with the aid of an absorption wavemeter. Further details regarding the alignment should not be necessary since the circuit of this frequency multiplier chain is fairly uncritical, and construction will probably not be attempted by beginners.

3.1. Construction of the Quadrupler

The $\lambda/2$ coaxial circuit is constructed from 1.5 mm thick single-coated PC-board material on one end of PC-board DC 0 DA 003. The inner dimensions of the box amount to 27 mm x 27 mm x 47 mm. The following parts are required from PC-board material:

Number	Dimensions (mm)	Part
2	27 x 27.0 x 1.5	End pieces
1	50 x 27.0 x 1.5	Side piece on the board side
1	50 x 34.5 x 1.5	Side piece for BNC-connector and trimmer
1	50 x 30.0 x 1.5	Cover

The larger side piece for the BNC-connector is soldered to the front edge of PC-board DC 0 DA 003 and will protrude 6 mm below the lower surface of the PC-board. This ensures that a spacing of 6 mm to the nearest ground surface is provided for the whole local oscillator module. The board is also well supported in this manner.

The λ/2 inner conductor constructed from 7 mm diameter brass tubing with a length of 47 mm is mounted with the aid of M 3 nuts inserted and soldered into place within the tube as was described for the converter module, and with the aid of brass screws placed to the center of the end pieces. The surface of the copper coating should be slightly tinned before mounting the inner conductor.

The BNC-connector is mounted with its whole flange on the inner side of the largest side panel and soldered all around the edge. The hole for this socket is spaced 11 mm from the edge of the circuit.

The coaxial circuit should be tuned with a high-quality spindle trimmer with a low commencement capacitance of max. 0.5 pF. Of course, it is also possible to use a M 3 or M 4 screw to which a disc has been soldered as was described in (2).

The Schottky diode should be connected with the shortest possible leads 10 mm from the cold end of the λ/2 inner conductor. Either a small screw or a quick soldering process can be used. The other end of the diode is fed through a small home-made PTFE (teflon) feed-through in the inside panel to the trimmer on the last 540 MHz stripline. The 100 kΩ resistor is also used as support. This resistor (without end caps) should also have the shortest possible connections.

Finally, four M 3 nuts should be soldered to the corners of the quadrupler box and the cover screwed into place. The output frequency of 2160 or 2304 MHz should be measured with the aid of a coaxial resonator system with spindle tuning and measuring diode as described in (3). The described local oscillator module has been constructed many times and has been found to operate with other suitable transistors. Due to its small, elongated shape, it is very suitable for mounting in available receivers and portable equipment.

4. REFERENCES

(1) G. Sattler: A modular ATV transmitter, part 2
VHF COMMUNICATIONS 5, Edition 2/1973, Pages 66 - 80

(2) K. Hupfer: A 2160 MHz local oscillator for 13 cm converters
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 246 - 247

(3) K. Hupfer: An SHF wavemeter
VHF COMMUNICATIONS 7, Edition 2/1975, Pages 90 - 92

TUBULAR RADIATOR FOR PARABOLIC ANTENNAS ON THE 13 cm BAND

by H.J. Griem, DJ 1 SL

1. APPLICATION

A parabolic reflector is to be illuminated with the aid of a round waveguide. A coaxial-waveguide interface is required if this round waveguide is to be energized with the aid of a coaxial cable. In studying American literature (1 to 6), it was found that such arrangements had been constructed and used but that no general design details were provided.

For this reason, a basic research was carried out to find out the basic radiation geometrics of parabolic antennas and to find suitable feed systems. This was followed by designing an actual tubular radiator for 13 cm to fit an available parabolic reflector of 120 cm dia. with a focal point / diameter (F/D) = 0.375.

The result of these measurements and experiments allow design details to be calculated for all other parabolic antennas, and microwave amateur bands. This means that it is possible with a minimum of mechanical work to illuminate parabolic reflectors for amateur radio applications.

2. RADIATION GEOMETRICS

2.1. Parabolic Reflectors

The beamwidth or focal angle of a parabolic reflector is mainly dependent on the ratio F/D. This relationship is shown in **Figure 1** where F is the focal point of the parabolic, and D the diameter of its aperture. The focal angle is 2 φ, since φ is measured from the center point to one side. If the focal point F is not known, it can be calculated from the diameter D and the depth d according to the following formula 1:

$$F = \frac{D^2}{16\,d} \quad (1)$$

The actual positions where these values can be measured are shown in the drawing given in **Figure 1**.

From the dimensions of the given parabolic reflector, it will be seen that the focal angle amounts to 135°. This angle should be illuminated as completely as possible by the primary antenna. The 10 dB beamwidth (not the 3 dB beamwidth as is usually used) of the primary (or exciter) antenna should amount to just 135° in the case of our example. If the beamwidth is less than this, the parabolic reflector will not be completely illuminated, and the efficiency and antenna gain will be less than the maximum available. If, on the other hand, the beamwidth of the primary radiator is greater, radiation will be made past the edge of the parabolic reflector, which will deteriorate the front-to-back ratio, and also decrease the antenna gain. A favorable compromise is found between antenna gain and clean polar diagram when the 10 dB beamwidth of the primary radiator responds to the focal angle 2 φ of the reflector. If these prerequisites are fullfilled, the gain formulas given for parabolic antennas will normally be valid.

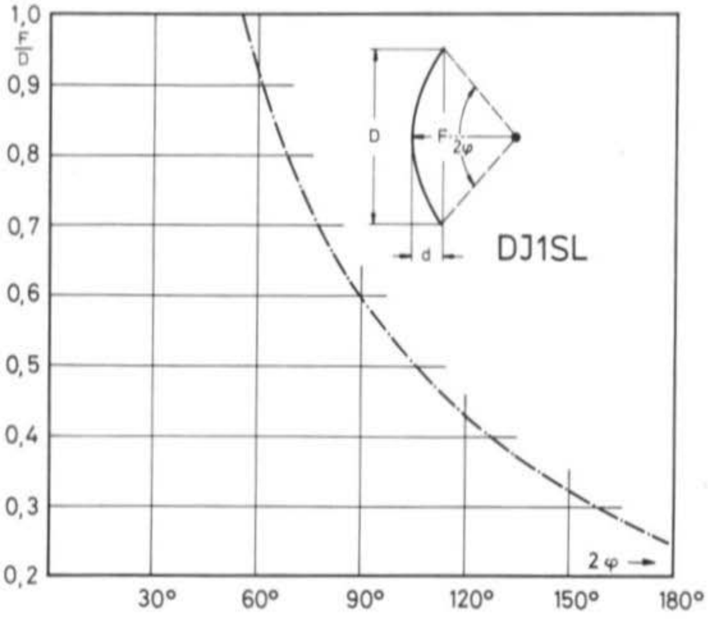


Fig. 1: Focal angle of parabolic antennas as a function of F/D

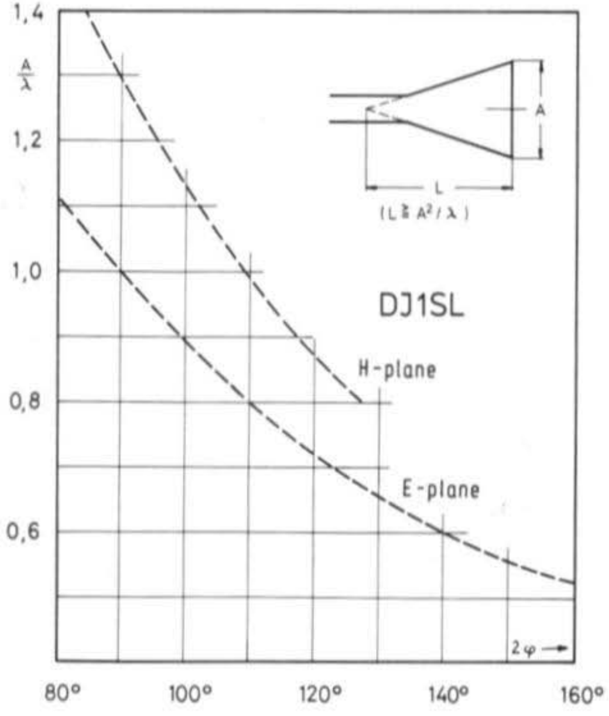


Fig. 2: 10 dB beamwidth 2 φ of horn radiators

A dipole with reflector is often found as primary antenna for parabolic reflectors. It is not easy when using such an arrangement to obtain the required matching of the 10 dB beamwidth to the parabolic reflector to be illuminated as when using radiating surfaces such as given in (8). For this reason, waveguide radiators such as horn antennas should be used, since the beamwidth can be easily determined with the aid of the dimensions of the aperture.

2.2. Horn Radiators

Figure 2 gives the 10 dB beamwidth of a so-called horn radiator as a function of its aperture. It will be seen that the beamwidth decreases on increasing the aperture. This diagram, which is also given in (7), is valid both for round (A = diameter) and square (A = side length) cross sections. In the case of a parabolic antenna with a focal angle of 135°, a horn radiator also having a beamwidth of 135° will be required; this corresponds to a horn radiator having A/λ corresponding to 0.63. The differing values for the H and the E-planes which were mentioned in (8) are not of interest here. With λ = 130 mm, »A« will correspond to: 0.63 x 130 = 82 mm.

3. TUBULAR RADIATOR

Due to its ease of construction, and since tubes are easier to obtain than rectangular profiles, a round waveguide as so-called tubular radiator is to be used. Brass tubes are usually available in diameters with steps of 5 mm. It has been seen in the previous section that a diameter of approx. 82 mm is required. The dimensions of the round waveguide are now to be established.

3.1. Diameter of a Round Waveguide

Without going too deeply into the theoretical relationships in waveguides, the permissible diameters of a round waveguide are to be given which is to be operated in a single-mode range. This range is between the limit wavelengths of the H_{11} and the E_{01} wave. The diameter range for this is given in equation: 2, and the first number of the nominator is valid for the H_{11}-wave and the second number for the E_{01}-wave:

$$D = \frac{\lambda}{1.71 \text{ to } 1.31} \quad (2)$$

It will be seen that the diameter in which only the H_{11}-wave is possible, is in the range of 76 to 99 mm for λ = 130 mm (f = 2.31 GHz).

Due to the increasing loss in the vicinity of the wavelength limit, the diameter of the waveguide should be as large as possible, without going into the range of the next higher oscillation mode, in our case the E_{01}-wave. For this reason, a diameter of 88 mm (outer diameter 90 mm, wall thickness 1 mm) was selected for the 13 cm band. The H_{11}-wavelength limit then amounts to 151 mm corresponding to 1.99 GHz.

3.2. Waveguide Wavelength

The wavelength of a waveguide is not identical to the wavelength of a RF signal in free space. The waveguide wavelength λ_W is greater than in free space. The relationship between these two values is given in equation 3:

$$\lambda_W = \frac{\lambda}{\sqrt{1 - (\frac{\lambda}{\lambda_l})^2}} \qquad (3)$$

Where λ_l is the limit wavelength of the type of wave under consideration.

A value of 151 mm was obtained from equation 2 as the limit wavelength for the H_{11}-wave. In the case of the selected diameter of 88 mm for the tube, the calculated wavelength of the waveguide amounts to 258 mm (free-space wavelength: 130 mm). This value is required later for calculating the length of the tubular radiator.

3.3. Construction of the Tubular Radiator

It was mentioned in section 2.2. that the horn radiator should have an aperture of 82 mm. According to section 3.1., a waveguide diameter of 88 mm would be favorable. A diameter of 88 mm is within the permissible range between 76 and 99 mm, and a horn-shaped aperture of the radiator will not be required. For this reason, the radiator has been called: tubular radiator.

The question was now which length should be selected for the radiator since various conflicting information was to be found in the various sources. The various lengths varied from 2 λ down to considerably lower values, which meant that the most favorable length had to be established with the aid of measurements. A length of 2 λ was chosen for the start of the measurements which corresponded to a length of 260 mm. As will be seen in section 4, the measurements showed that the optimum length is λ/2, when referred to the waveguide wavelength λ_W. This results in a mechanical length of 129 mm for the 13 cm band.

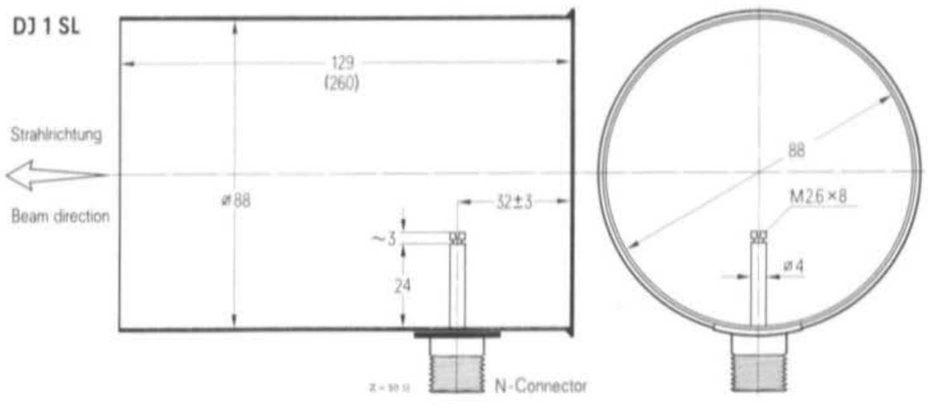


Fig. 3: Dimensions of a tubular radiator for the 13 cm band

Figure 3 gives the dimensions of the final tubular radiator. The interface between the coaxial cable (N-connector) and the waveguide mode is clearly shown. This can be classed as a type of vertical antenna (ground plane), however, its length is far shorter than $\lambda/4$. The spacing of this radiating element to the rear panel of the tubular radiator, and the length of the actual element are adjustable. The given dimensions have been found to be most favorable for the 13 cm band in the following measurements.

4. MEASUREMENTS

The following measurements were made to establish the two most important characteristics of the tubular radiator: the radiation diagram and the matching to the coaxial cable. The radiation diagram should fulfill the demand of illuminating the parabolic reflector so that 10 dB power reduction is present at the edge of the reflector.

These measurements were carried out using an amplitude reflection measuring system HP 8755 having an impedance of 50 Ω. This measuring system allowed both the insertion loss and the return loss to be measured simultaneously. The return loss could be read off directly, whereas the insertion loss was measured by measuring the attenuation from the tubular radiator to an auxiliary dipole placed several meters away. If the spacing between these two antennas is kept constant, the radiation diagram can be obtained by rotating the tubular radiator in its axises, and giving the loss as a function of the various angles.

4.1. Matching

Figure 4 gives the diagram of the return loss over the frequency range of 2 GHz to 2.5 GHz. The amateur band is from 2.3 to 2.35 MHz. The measuring range for the return loss amounted to 0 to 40 dB which corresponds to the VSWR values given on the right side of the diagram. The reference impedance is 50 Ω.

The »vertical antenna« was shifted with respect to the rear panel of the tubular radiator and the length of the element varied with the aid of a screw in order to obtain the lowest VSWR within the amateur band. It was interesting to find that a high return loss could be obtained at 2.3 GHz that could be shifted in frequency somewhat using the described means; however, a peak in the return loss was also found at 2.08 GHz that could not be affected by these adjustments. By calculation, it was found that this resonance was caused by the length of the tubular radiator; since the length of 260 mm corresponds to a waveguide wavelength of $\lambda/2$ at a frequency of 2.08 MHz. This dimension indicated a natural matching of the tubular radiator to space, and further experiments were made to establish further details.

In order to do this, the length of the tubular radiator was calculated so that it corresponded to exactly $\lambda/2$ of the waveguide wavelength in the 2.3 GHz amateur band. This resulted in a mechanical length of 129 mm. The measurements made on this radiator, and the results were as expected. A resonance condition existed at 2.3 GHz which was not affected by alignment of the energizing element, whereas a second resonance point could be varied. By combining these two matching points, it was possible to obtain a bandpass characteristic which allowed matched conditions to be obtaines over a far wider frequency range than was the case with

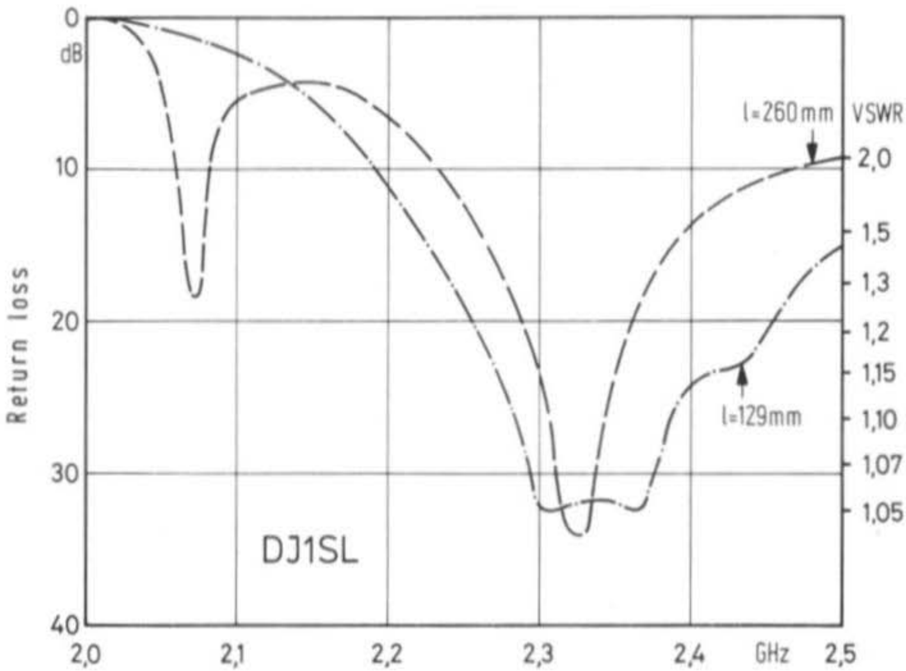


Fig. 4: Matching of a tubular radiator in a frequency range from 2.0 to 2.5 GHz

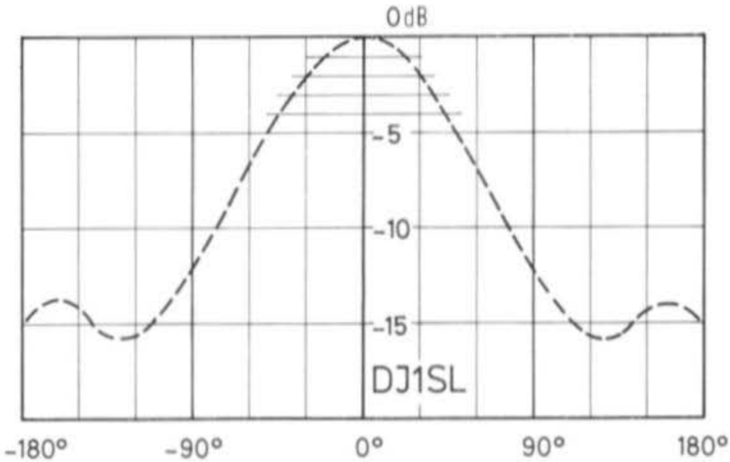


Fig. 5: Radiation diagram (E-plane) of the tubular radiator

the 260 mm long tube. The return loss amounted to approximately 32 dB over the whole amateur band, which corresponds to a VSWR of 1.05 (Figure 4). This is a very good value when one considers that the tolerance in the impedance of coaxial cables amounts to 5 %, and thus is in the same order of magnitude.

The described axperiments were made in a large room where the radiator was either able to radiate freely through a window, or against the ceiling (2 m). In order to establish what effect the parabolic reflector would have on the matching, a metal plate of approximately one square meter was moved towards the radiator. It was found that a slight fluctuation of the return loss was indicated, which increased the nearer the plate came to the aperture of the tubular radiator. These fluctuations remained in the order of 30 dB (min. 25 dB) up to a spacing of approximately 40 cm. This corresponds to a VSWR of approximately 1.1, which is still an excellent value that would not be measurable with amateur means. For those readers that have access to such a measuring system, it is possible for the complete antenna including parabolic reflector to be measured and the optimum matching obtained by varying the coaxial element.

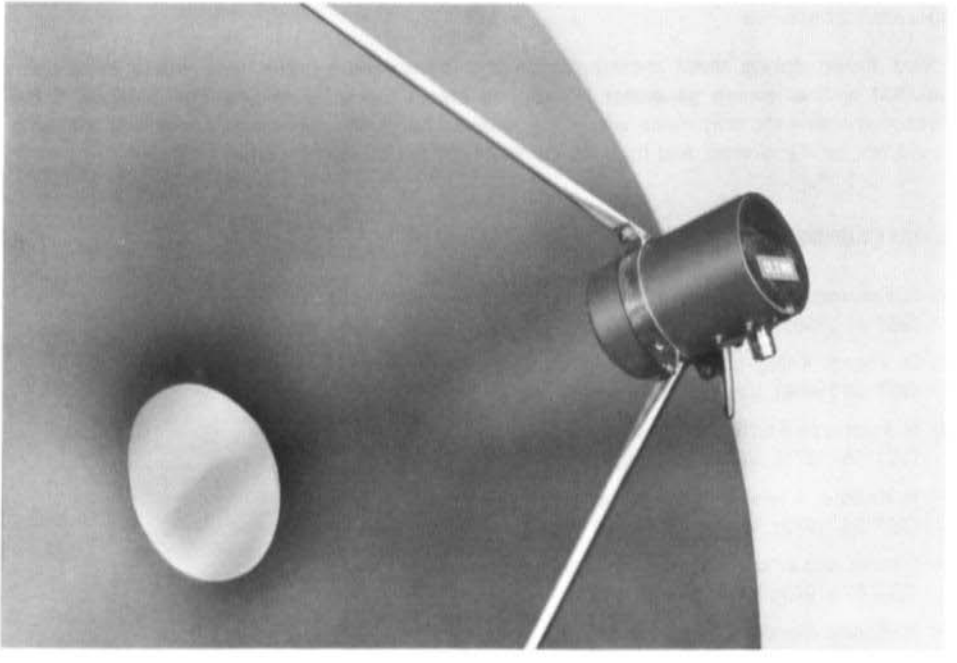


Fig. 6: 1.2 meter parabolic with tubular radiator for the 13 cm band (G_{dipol} approx. 24 - 25 dB)

4.2. Radiation Diagram

The second task was to measure the radiation diagram of the tubular radiator, which of course could not be carried out indoors due to reflections which were found to considerably distort the radiation diagram. On the other hand, a small spacing between the tubular radiator and the auxiliary dipole could cause adjacent effects. For this reason, the measurements were made out-of-doors.

The measuring dipole was mounted 4 to 5 meters away from the tubular radiator and the loss between the tubular radiator and auxiliary dipole was measured with the measuring system. The tubular radiator was rotated in its vertical axis. A second series of measurements was made after the tubular radiator had been rotated by 90°. The vertical plane diagram was obtained finally by rotating the tubular antenna by 360°.

The result of these measurements is given in **Figure 5**. It will be seen that the signal drop of 10 dB with respect to the maximum value is obtained at an angle of approximately 80°. This corresponds to a total beamwidth of 160° which is somewhat more than the ideal value required for the parabolic reflector of 135°. At 135°, this value amounts to 9 dB which is tolerable. It should also be taken into consideration that this measurement of the radiation diagram was not made at highest precision since the measuring arrangement was too simple for this. Due to a lack of time, it was not possible to repeat these measurements under laboratory conditions.

It was found during these measurements that considerably more time would have been required without sweep generator in order to obtain the same results. For instance, if the measurements were only made within the amateur band, the resonance position at 2.08 MHz would not be discovered, and thus the cause of any matching difficulties.

5. REFERENCES

(1) K. Peterson: Practical gear for amateur microwave communications
QST 47 (1963), Volume 6, Pages 17 - 20

(2) D. Vilardi: Easily-constructed antennas for 1296 MHz
QST 53 (1969), Volume 6, Pages 47 - 49

(3) N. Foot: WA 9 HUV 12-foot dish for 432 and 1296 MHz
QST 55 (1971), Volume 6, Pages 100 - 101

(4) R. Knadle: A twelve-foot stressed parabolic dish
QST 56 (1972), Volume 8, Pages 16 - 22

(5) Simple and efficient feed for parabolic antennas
QST 57 (1973), Volume 3, Pages 42 - 44

(6) R. Kolbly: Simple microwave antenna
ham radio 2 (1969), Volume 11, Pages 52 - 53

(7) Reference Data for Radio Engineers
5th edition, Pages 25 - 37

(8) H. Berner: The most important features and characteristics of GHz antennas
VHF COMMUNICATIONS 8, Edition 3/1976, Pages 130 - 141

INTERDIGITAL FILTERS
for the 24 cm and 13 cm band

by R. Griek, DK 2 VF

Interdigital bandpass filters have been described on a number of occassions in this magazine. A three-stage filter for the 24 cm band has already been described by the same author. Various references were given with respect to the theory of operation and calculation. The practical design is obtained, however, most favorably from the »Tabellenbuch Mikrowellenbandpässe« (1). This book of tables provides all dimensions in standardized form that are required for the construction of interdigital microwave bandpass filters with 2 to 12 resonant circuits, providing a Chebichev behaviour of the loss in the passband range. The filter parameters are provided in fine steps: relative bandwidth (1 % to 20 %), minimum return loss in the passband range (8 to 40 dB), and electrical length of the line sections (7.5° to 90°). This allows an easy transition from the design to construction phases. Frequency dependent orientation values for the Q simplify the determination of the most favorable dimensions of the case for each application.

If high demands are to be placed on the bandwidth, ripple, return loss, and slope, an alignment on a swept frequency system will be required. In our case, however, two relatively simple filters have been selected that provide satisfactory passband curves as long as they are carefully constructed and aligned at the required frequency.

The interdigital filter for the 24 cm band consists of three resonant circuits and an input and output coupling. No tuning is provided. The author corrected the passband curve on the prototype by bending the resonant circuits towards or from the coupling link; also the lengths were increased slightly by placing drops of solder to the appropriate positions. The filter for the 13 cm band has five resonant circuits; a tuning screw is provided at the hot end of each stage. The resulting passband curves are given in **Figure 1**.

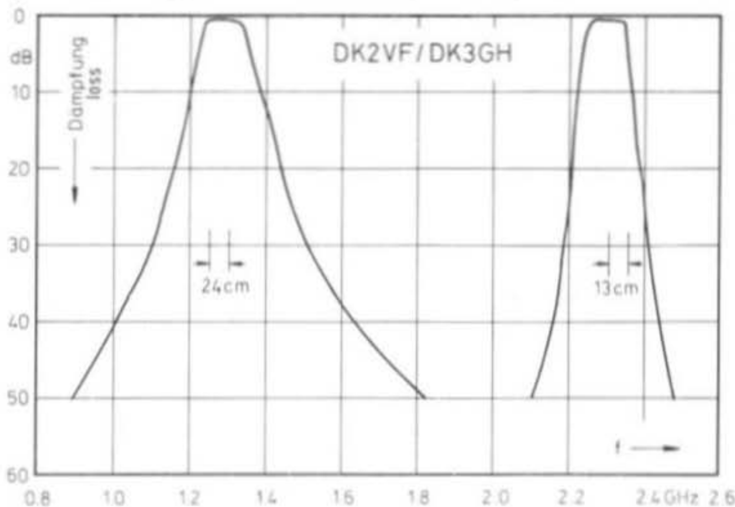


Fig. 1: Passband curves of the described interdigital filters

The 3 dB bandwidth of both filters is in the order of 100 to 120 MHz and the insertion loss amounts to a minimum of 0.6 dB in the case of the 24 cm filter and to 1 dB in the case of the 13 cm filter. This means that these interdigital bandpass filters are very suitable for use in transmitters, especially in the case of frequency multipliers. They are also advisable between receiver preamplifiers and mixers. However, the bandwidth is too large for an effective suppression of the image noise at an intermediate frequency of 30 MHz (2). Once again it will be seen that a higher intermediate frequency of 70 MHz, 100 MHz or 144 MHz has advantages.

1. MECHANICAL CONSTRUCTION

Figure 2 shows photographs of the filters in ready-to-operate condition. In the case of the 13 cm filter, it will be seen that the tuning screws are fixed with the aid of a steel wire (paper clip). The narrow sides of the filters are open, but can be sealed, if required.

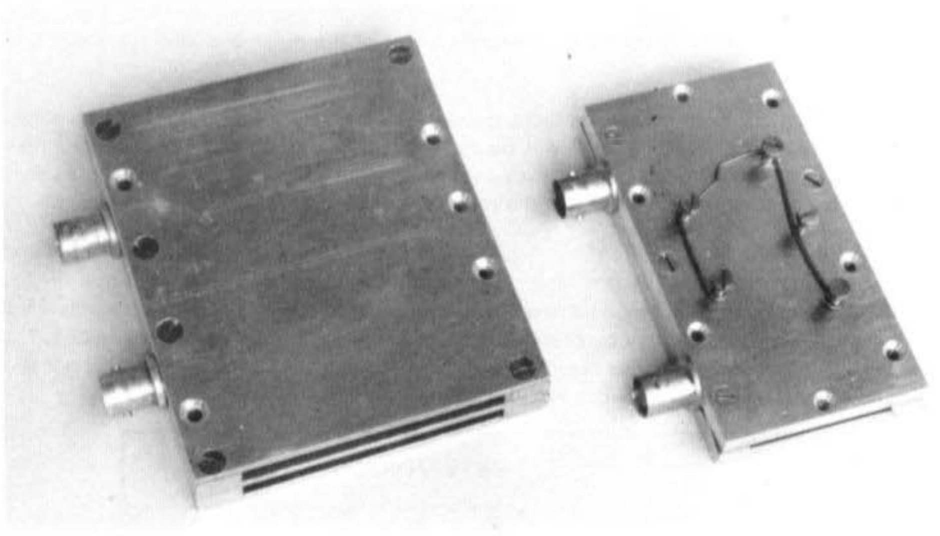


Fig. 2: Photographs of the two filters when completed

Figures 3 and 4 show the construction steps of the 24 cm filter. The material is always aluminium plate. There are two different thicknesses required: 2 mm and 4 mm. The cuts (fret-saw, file) are placed one above the other in the form of a sandwich and are drilled firstly at the corners (2.4 mm dia.). The holes in the base plate are provided with 3 threads, all other holes are drilled out to 3 mm. The holes in the cover plate are countersunk suitably for accommodating the M 3 countersunk screws. After this, the filter is screwed together and the seven other holes are drilled in the same manner. After the filter has been completely screwed together, it is possible to drill the two holes for the BNC sockets and to provide them with a UNEF 3/8" thread. Finally, the BNC sockets are screwed into place and the cover is removed. The inner conductors of the connectors are soldered to the wide coupling links. Special solder and flux should be used.

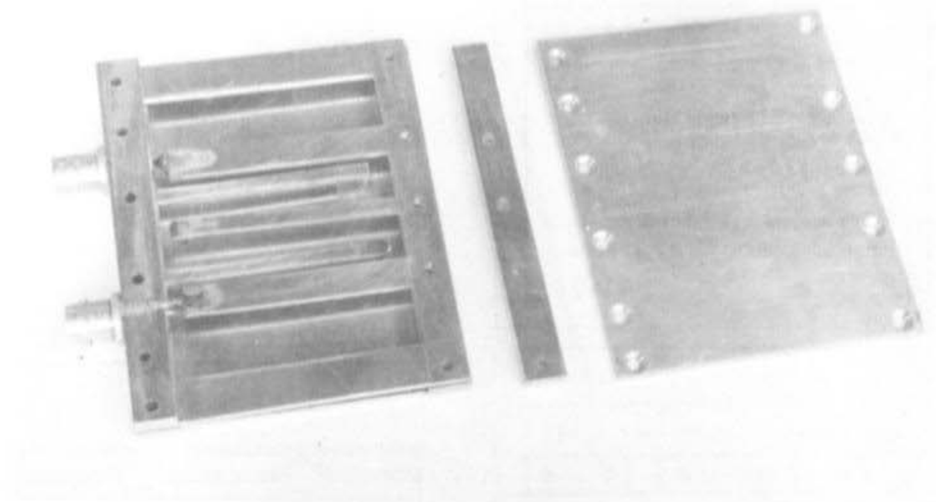


Fig. 3: Photograph of the open 24 cm filter

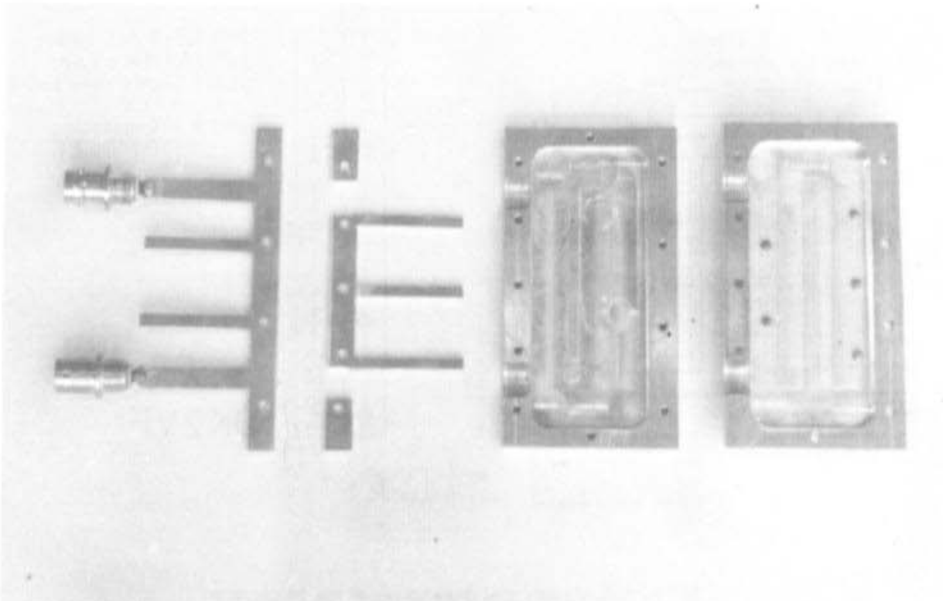


Fig. 5: Various parts of the 13 cm filter

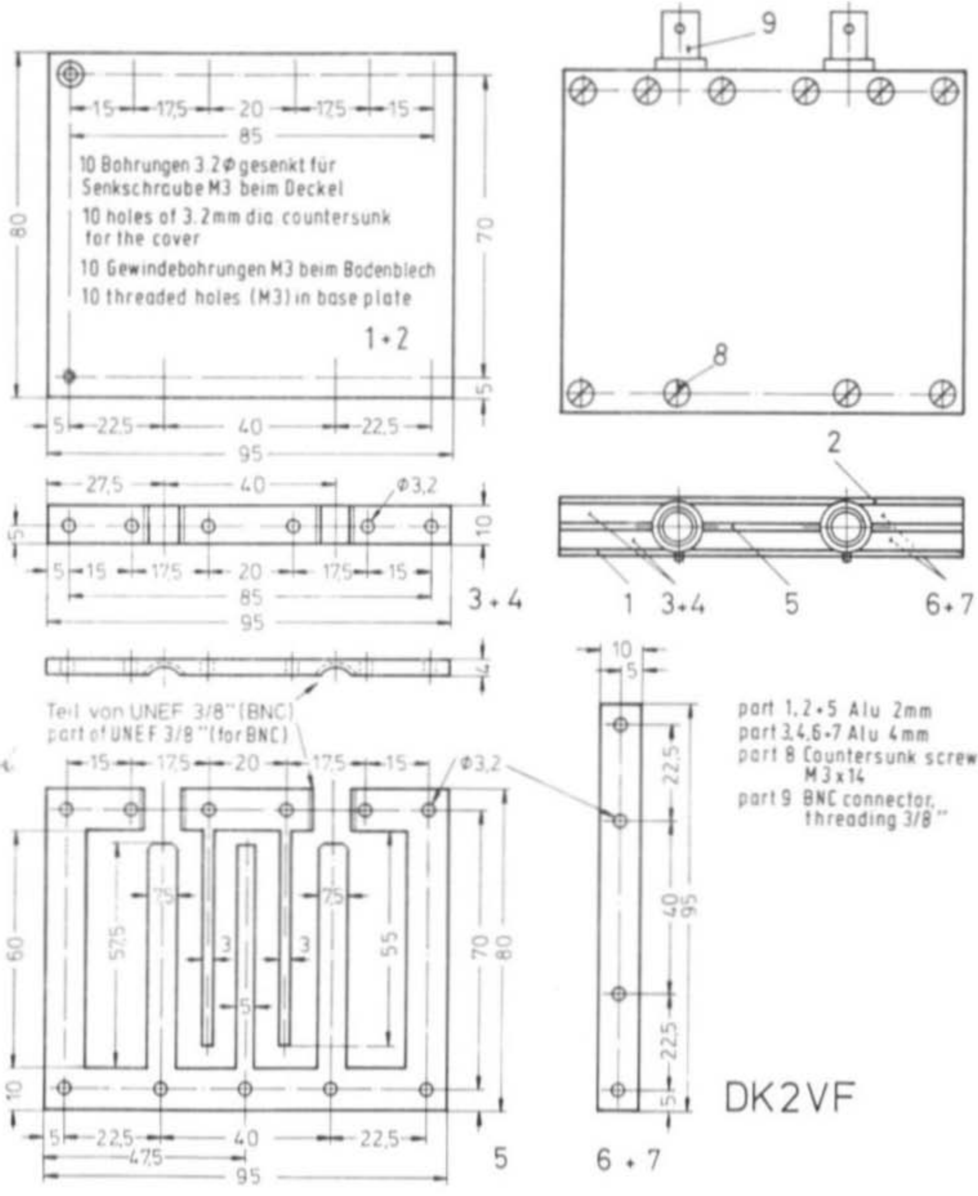


Fig. 4: Individual parts and dimensions of the 24 cm filter

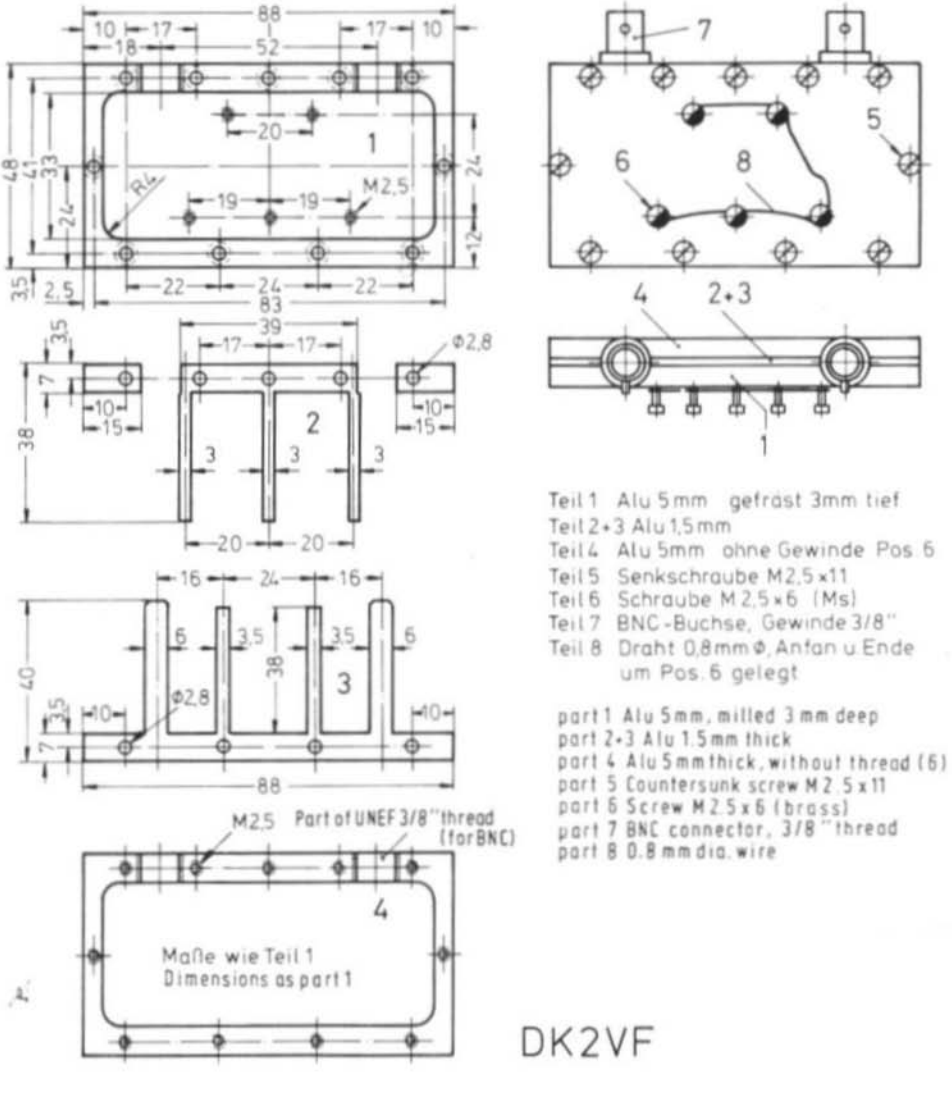


Fig. 6: Individual parts and dimensions of the 13 cm interdigital filter

The internal parts of the 13 cm filter are made from 1.5 mm thick aluminium plate **(Figures 5 and 6)**. The base and cover comprise 5 mm thick aluminium plate which has been milled out to a depth of 3 mm. If it is not possible to make this filter in this manner, it is also possible to use a sandwich construction in the same manner as for the 24 cm filter. In this case, a 2 mm aluminium plate is used for the base and cover plates together with a 3 mm spacing plate. Otherwise, the filter should be constructed in the same manner as for the 24 cm filter.

The most astounding feature of the described filters is the construction of the resonators using aluminium plate. If these plates are made with the aid of a saw and file, rectangular striplines with a relatively rough surface will result which need not be silver-plated. This seems to be very much in contrast to the previously described filter and to the demands made in (1). However, the author wished to establish whether it is possible to construct efficient filters for amateur radio application using »hobby construction« methods. The given measured values confirm that this is clearly the case. The filters are also so light that they can be used in portable stations, which is also very important for amateur radio on the microwave bands.

2. REFERENCES

(1) G. Pfitzenmaier: Tabellenbuch Mikrowellenbandpässe.
Normierte mechanische Abmessungen von Interdigitalfiltern.
Siemens AG, 1972.

(2) D. Vollhardt: Preamplifier and mixer noise at SHF.
In this edition of VHF COMMUNICATIONS.

BALUN TRANSFORMERS for 23 and 13 cm from SEMI-RIGID CABLE

Editorial staff

Coaxial λ/2 balun transformers are required for transforming a balanced impedance of 200 Ω to 50 Ω unbalanced. This is required, for instance, when connecting a folded dipole to coaxial cable. Unfortunately, the mechanical length is so small for the 23 cm and 13 cm UHF band that it is difficult, if not impossible, to bend normal coaxial cable to form such baluns. For this reason, semi-rigid coaxial cable with PTFE (teflon) dielectric and copper tubing as shield is normally recommended for UHF antennas. Such semi-rigid cables are available in various diameters, and a suitable type for our application has the following specifications:

Outer diameter:	3.58 mm	Impedance:	50 Ω ± 1 Ω
Inner conductor:		Capacitance:	96 pF/m
Silver- and copper		Propagation speed:	4.80 ns/m
plated steel:	0.91 mm dia.	Velocity factor:	0.695
Min. bending radius:	6.5 mm	Max. operating voltage:	5000 V DC

This cable is designated SR 3, and corresponds approximately to type 421-668 manufactured by Amphenol, or is similar to RG-141. The loss and RF power values are to be compared to the best known RG-cables in the following table:

Cable designation	Attenuation per 100 m at (MHz)					Max. Power (W) at (MHz)				
	100	200	400	1000	3000	100	200	400	1000	3000
RG-8, RG-213	5.7	8.1	12.3	24	48	975	685	450	230	115
RG-141 (SR 3)	9.9	14	21	39	78	1700	1200	830	450	220
RG-58, RG-58 B	14	21	32	53	113	300	200	135	80	40

When comparing the very small diameter of this semi-rigid cable, it is easy to see which advantages this more expensive cable has for our application.

The following dimensions are required for λ/2 transformers for the three UHF amateur bands after taking the velocity factor into consideration:

70 cm band: 239.7 mm
23 cm band: 80.4 mm
13 cm band: 45.3 mm

1. AVAILABILITY

VHF COMMUNICATIONS offers a kit for manufacturing such balun transformers for 13 cm, 24 cm and 70 cm. This kit comprises one piece of 1 λ (for the feeder), and λ/2 (for the actual balun). The prices are given in the pricelist.

A POWER AMPLIFIER FOR THE 23 cm BAND EQUIPPED WITH THE 2 C 39 TUBE

Editors

A mixer with subsequent three-stage amplifier was described in (1), which was made from metal plate. These metal plates can, of course, be made and soldered together at home, but it is far simpler to manufacture and construct such an amplifier when lathed and milled parts are available. For those readers having such facilities, a linear amplifier for 23 cm is to be described in this article. The amplifier is easy to construct and provides clear and precise tuning, and remains stable even during continuous operation at maximum power.

1. RF-CIRCUIT

The cathode circuit of the amplifier in the form of a Pi-transformer can be designed for 432 MHz or 1296 MHz so that this stage may be used as frequency multiplier or as straight-through amplifier. The plate feeds a round cavity resonator which is tuned with the aid of a capacitive plunger. The output coupling is also made capacitively with the aid of a BNC-connector that can be adjusted into or out of the cavity. A compensation pin is provided on the opposite side to ensure the most favorable matching.

The specifications for this family of tubes given, for instance, in (2) are valid for the operating voltages and power values. In the tripler mode with a plate voltage of 600 V, the output power at 1296 MHz will be approximately that of the drive power at 432 MHz.

Figures 1 and 2 show the photograph of an early prototype using a different system of tuning and compensating screws. The two stages had been mounted 180° with respect to another and exhibit the following specifications at a plate voltage of 570 V.

Tripler 432/1296 MHz: Power gain (at 8 W): approx. 1 (0 dB)
Through-amplifier: Low-power gain (measured at P_{in} = 1 mW or 50 mW):
At I_a = 25 mA: 8 dB / At I_a = 50 mA: 10 dB / At I_a = 100 mA: 12 dB.
High-power gain (at P_{in} = 8 W): output power approximately 30 W
1 dB bandwidth: approx. 8 MHz (sufficient for ATV) / 3 dB bandwidth: approx. 19 MHz

When using both stages as a straight-through amplifier at 1296 MHz, the full output power of 30 W at a plate voltage of 600 V (40 W at 800 V) can be achieved with a drive of max. 1 W.

A principle circuit diagram of the two stages including cathode circuit, heater chokes, and a simple power supply is given in **Figure 3**. A time-delay relay to ensure that the plate voltage is not switched on until one minute has elapsed, has not been provided, and only a switch for the plate voltage supply is shown. In the case of the straight-through amplifier, the wiring of the cathode and heating of the first stage should be carried out in the same manner as for the second stage. Since each stage should be provided with its individual variable grid bias voltage, an individual heater winding on the power transformer should be provided for each tube. The constant-voltage two-pole used in this amplifier was already described in the 70 cm amplifier described in (2). A more detailed description of the »grid bias voltage supply for grounded grid amplifiers« was given in (3).

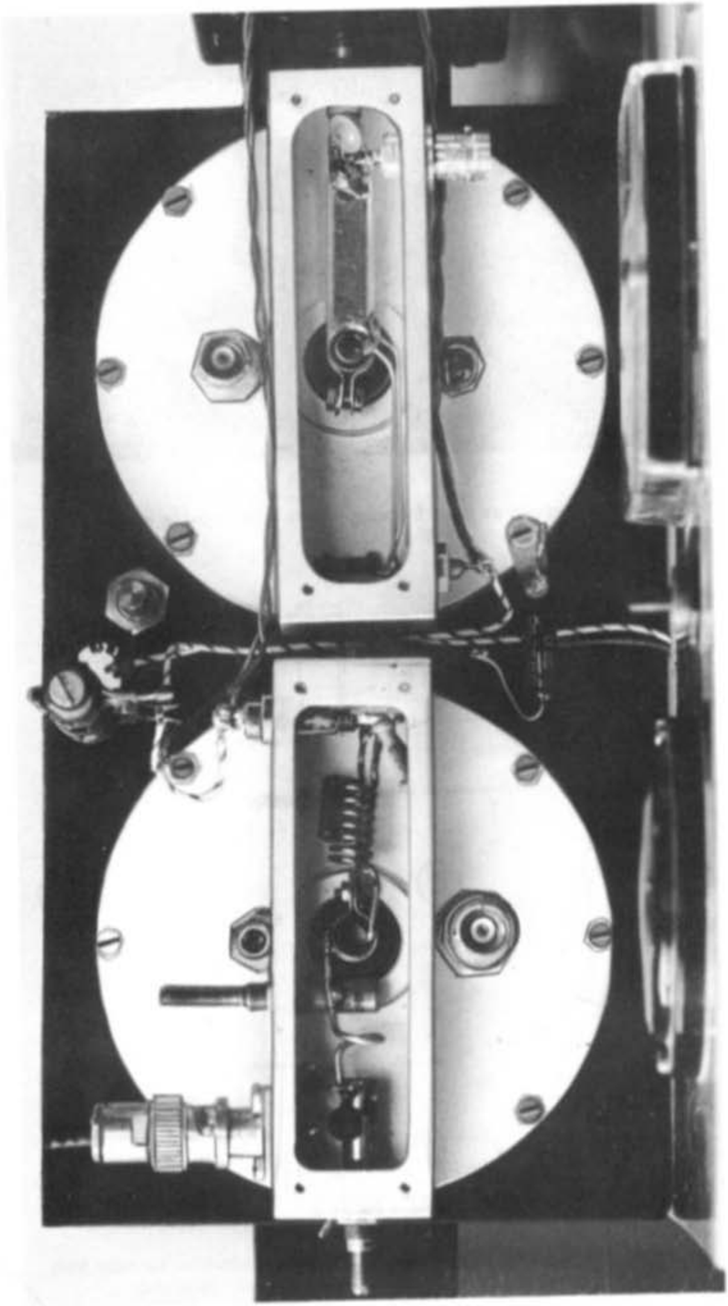


Fig. 1: Tripler 432 MHz / 1296 MHz (left) and amplifier (right) as seen from the cathode side

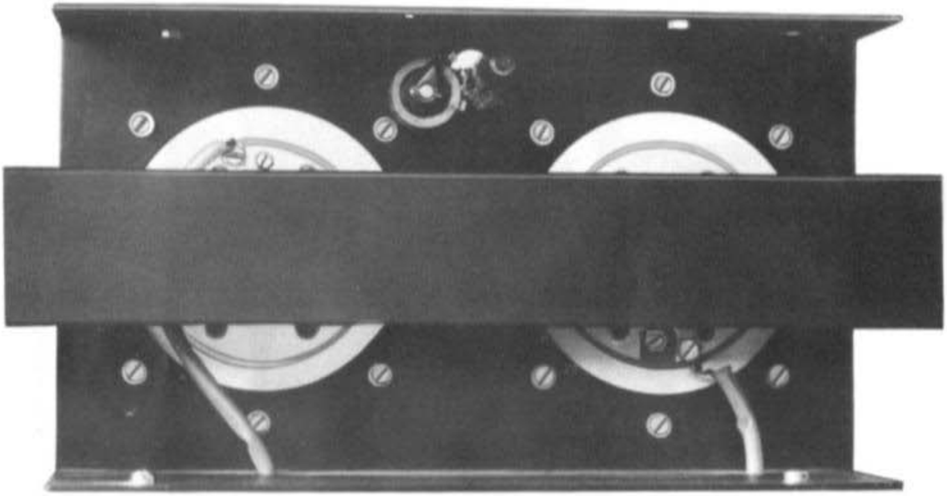


Fig. 2: Two-stage construction as seen from the anode side showing air ducting

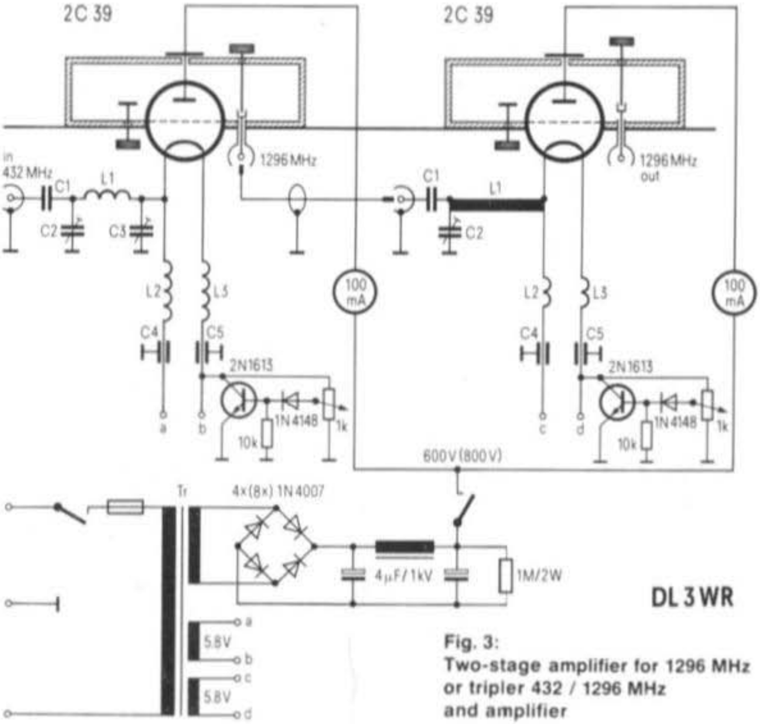


Fig. 3:
Two-stage amplifier for 1296 MHz
or tripler 432 / 1296 MHz
and amplifier

	Tripler 432/1296 MHz	Amplifier 1296 MHz
C 1:	220 pF	47 pF
C 2:	6 pF ceramic tubular trimmer	3 pF ceramic tubular trimmer
C 3:	3 pF ceramic tubular trimmer	deleted
C 4, C 5:	Feed-through capacitor approx. 100 pF, see text	
L 1:	1 turn of 1 mm dia. wire wound on a 8 mm former	part 16
L 2, L 3:	7 turns of 1 mm dia. wire wound on a 8 mm former	wire: 1 mm dia. approx. 45 mm long (see Figure 1)

Transformer: 600 V (800 V), 200 mA DC-current; 2 x 5.8 V / 1 A AC-current.
Choke for the plate voltage filtering.

3. MECHANICAL CONSTRUCTION

The design of the cathode circuit, the anode resonator and the output coupling was made by DJ 2 LI. The mechanical construction was further developed by DJ 1 JZ and improved in several prototypes so that the tuning and the output coupling could be achieved more easily. **Figure 4** shows the latest design of the anode circuit, and **Figure 5** shows the cathode circuit. The compensating or tuning screw is guided in long tubes so that an easy, and stable alignment is possible. It is advisable for all brass parts to be silver-plated.

The parts of the resonator are shown in **Figure 6.** The surfaces of these parts should be very smooth on the inner side. Part 5 is provided with a groove for the contact spring (part 22). This is shown enlarged in »A«. Part 5 is soldered onto the grid plate (part 2). Part 4, the plate contact ring, is also provided with a groove for the contact spring. Part 4 must be screwed onto the anode plate (part 3) in an insulated manner. This is made using an insulating disc (part 21, not shown) made from PTFE (teflon) foil and 6 insulating parts (part 19). The plate voltage is connected to this contact ring, which is not at RF-potential.

The UNEF 3/8" threaded hole in part 2 is provided for the output BNC socket. The tuning plunger is screwed into the M 7 (7 mm dia.) threading. Part 3 is provided with a hole with M 4 (4 mm) threading for alignment of the compensating pin. The parts required for tuning and compensation are shown in **Figure 7.** Part 11 should be soldered to the inner conductor of the BNC socket and then slid into the insulating cover (part 12). The probe (part 8) then influences part 12, and thus part 11. Both guides are provided with slots at the threaded end so that they provide a certain amount of locking.

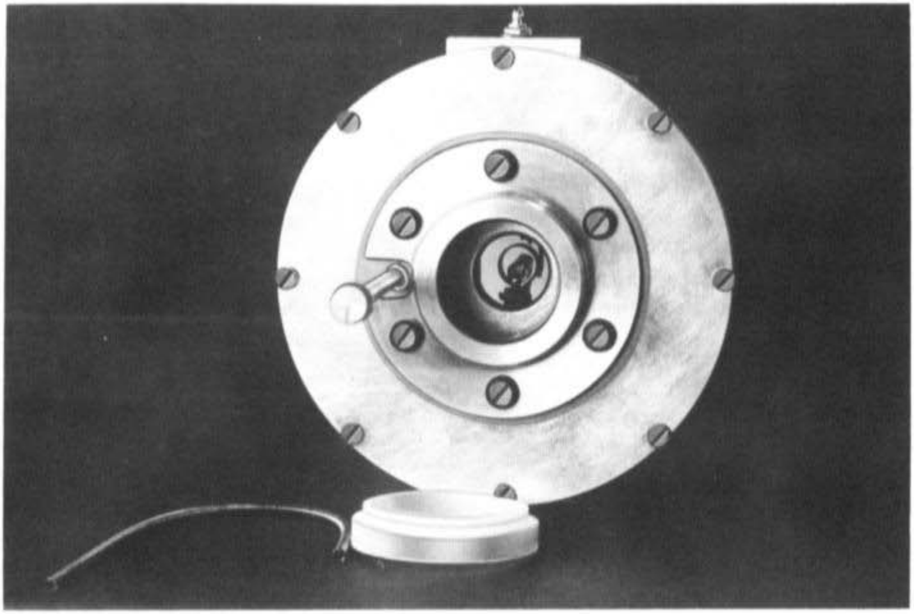


Fig. 4: Second prototype (from anode side)

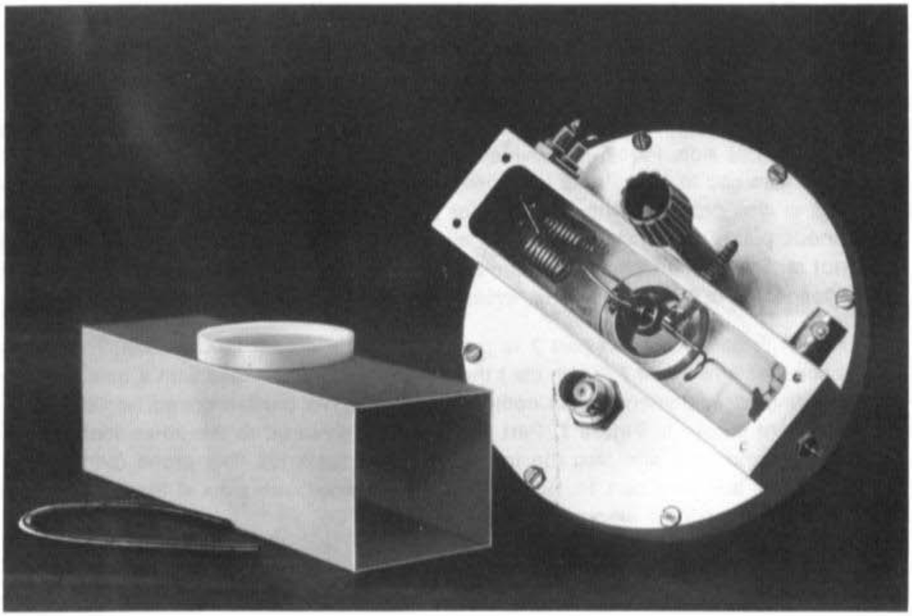


Fig. 5: Cathode side of the second prototype, with air ducting, insulating ring and contact spring

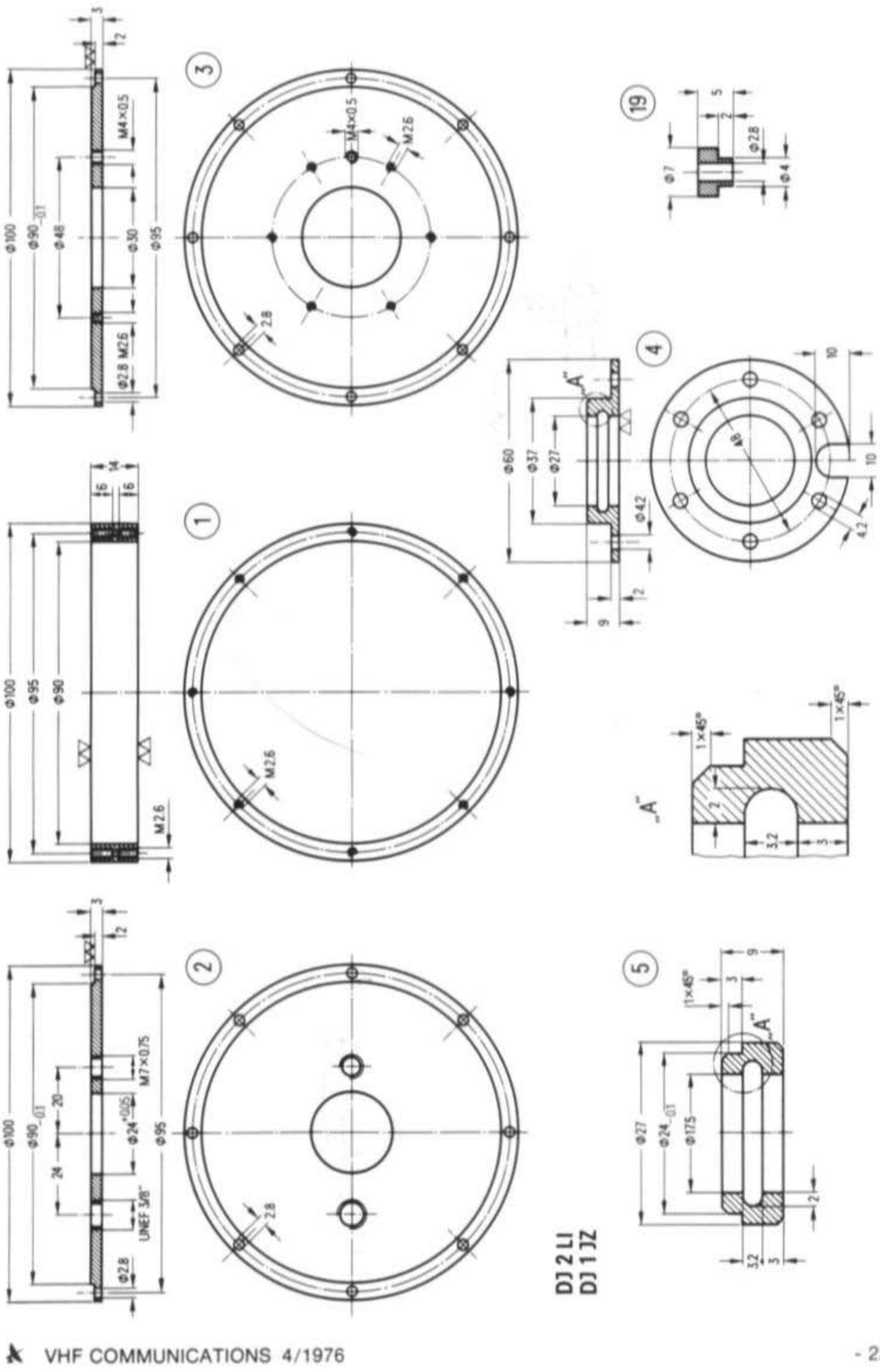


Fig. 6: Parts of the cavity resonator

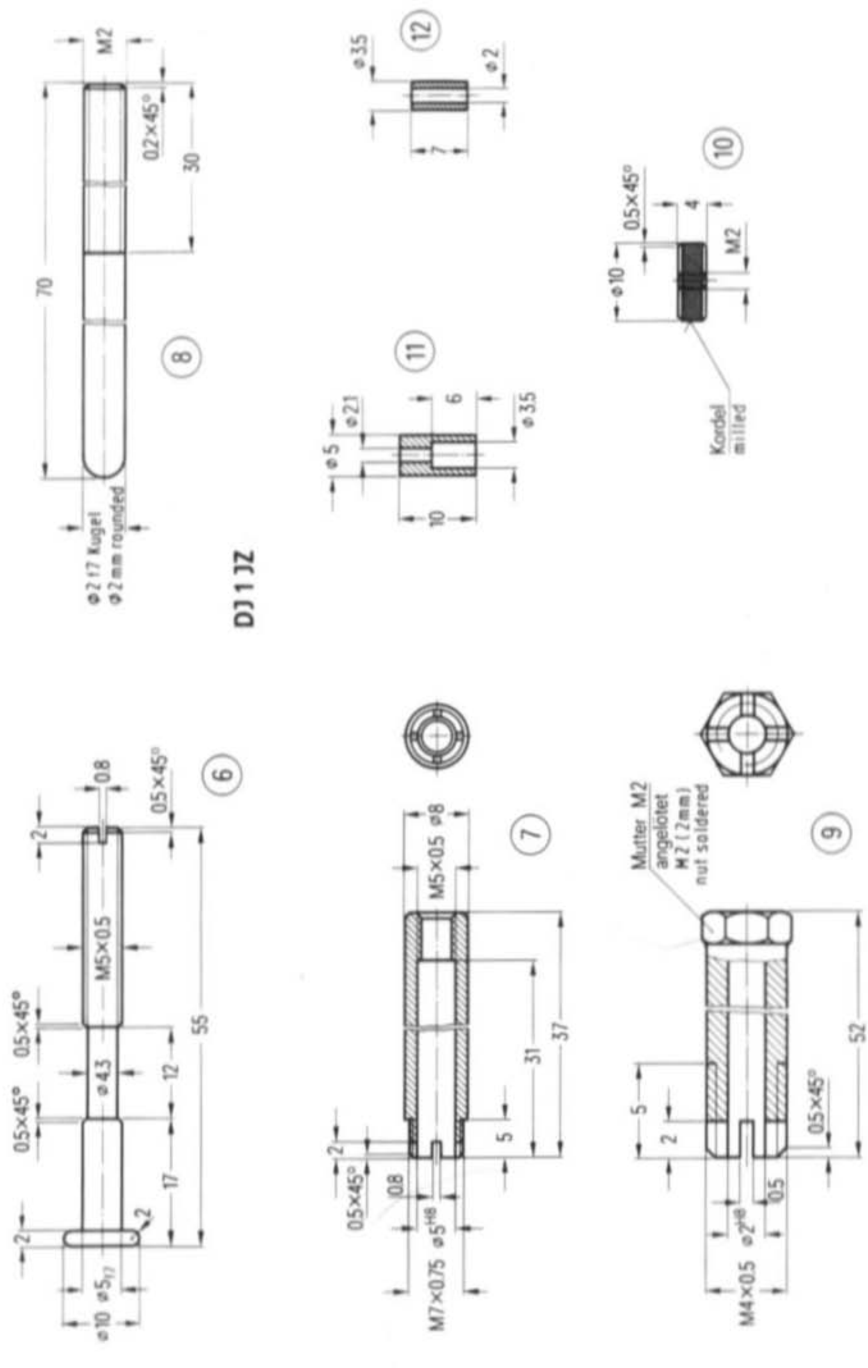


Fig. 7: Parts for tuning and coupling

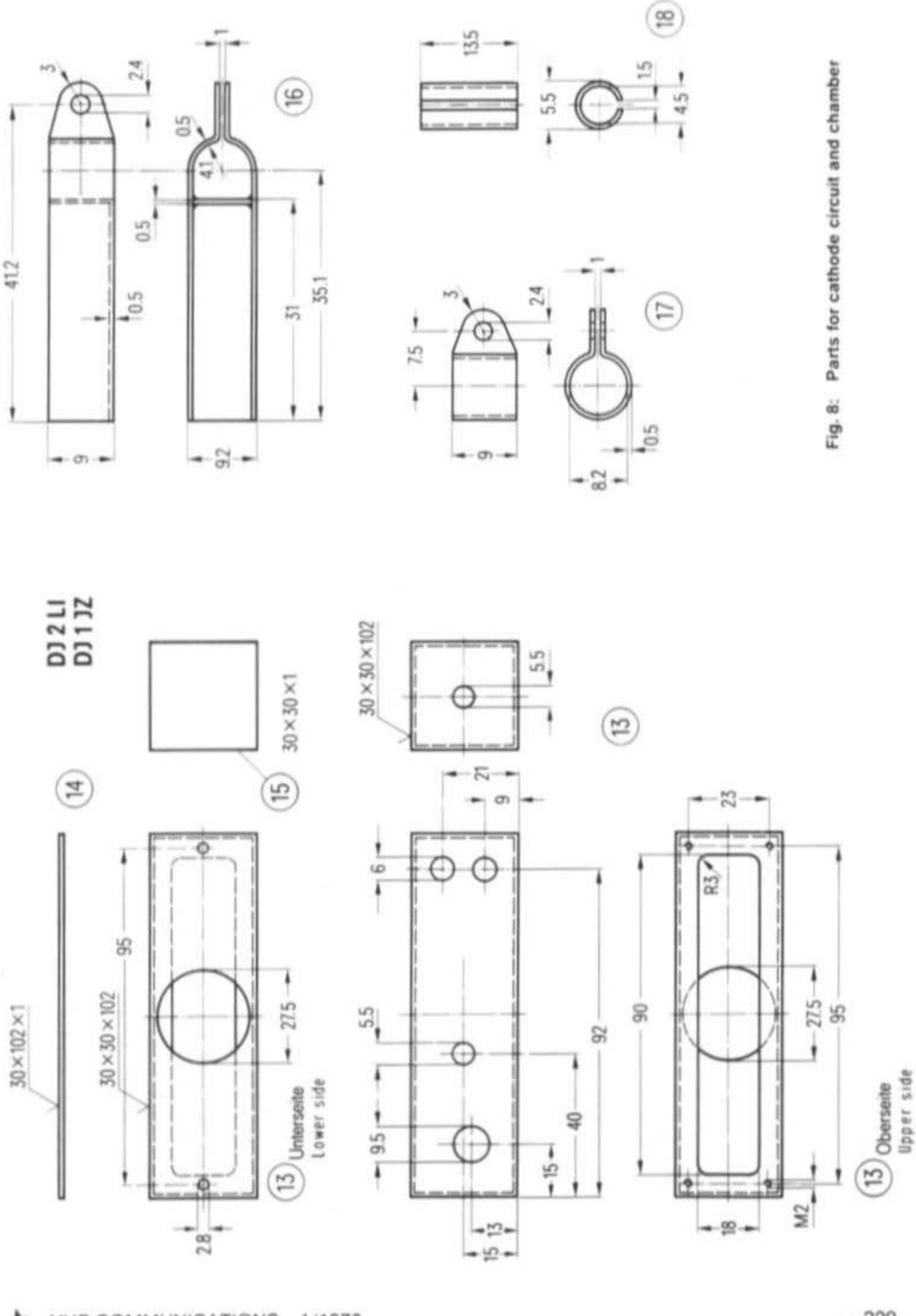


Fig. 8: Parts for cathode circuit and chamber

The cathode chamber for the cathode circuit and heater chokes is shown in **Figure 8**. In the case of the author's prototype, this is made from square-tubing of 30 x 30 x 1 mm (part 13), to which a part 15 was soldered to each end. The cover (part 14) is mounted into place with four M 2 (2 mm) screws. Holes for the input BNC socket, trimmer capacitor C 3 and the feed-through capacitors for the heaters (C 4 and C 5). Trimmer capacitor C 2 is mounted at the end directly adjacent to the input socket. Inductance L 1 (part 16) is soldered to C 2. The bracket (part 17) is provided for the 70 cm input. Chokes L 2 and L 3 are soldered to part 16, (or 17) and part 18, as well as to the feedthrough capacitors. These are not really suitable for 1300 MHz, in other words when their capacitance is more than several 100 pF, ceramic tubular capacitors of approx. 100 pF should be connected in parallel with the shortest possible connections.

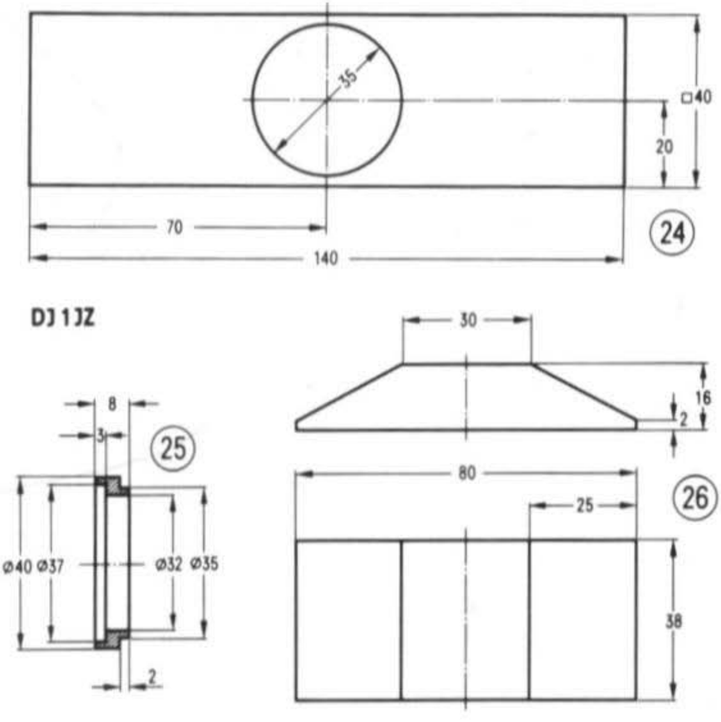


Fig. 9: **Parts for air cooling**

Finally, **Figure 9** shows the cooling ducting made from rectangular material (40 x 40 x 1 mm), the insulating ring (part 25) to ensure that the anode radiator is mounted to the cooling duct in an air-tight manner; the wedge (part 26) is also shown. The latter is glued into part 24 opposite the radiator to ensure that the cooling air must flow through the radiator. In the case of the two-stage amplifier, the cooling air ducting is longer and provided with two cutouts.

All parts required are given in the following parts list.

3.1. Required Parts

Part	Number	Designation	Material
1	1	Resonator ring	Brass 58 tube 100 x 5 mm
2	1	Grid plate	Brass 58, 3 mm thick
3	1	Anode plate	Brass 58, 3 mm thick
4	1	Anode contact ring	Brass 58
5	1	Grid contact ring	Brass 58
6	1	Tuning plunger	Brass 58
7	1	Guide for tuning plunger	Brass 58
8	1	Compensating pin	Brass 58
9	1	Guide for compensating pin	Brass 58
10	1	Knob for part 8	Brass 58
11	1	Output coupling for BNC socket	Brass 58
12	1	Insulating cap for output coupling	Teflon (PTFE)
13	1	Cathode chamber	Brass tube 30 x 30 x 1 mm
14	1	Cover for part 13	Brass plate, 1 mm thick
15	2	Covers for ends of part 13	Brass plate, 1 mm thick
16	1	Cathode inductance L 1 for 23 cm	Brass plate, 0.5 mm thick
17	1	Cathode contact bracket for 70 cm	Brass plate, 0.5 mm thick
18	1	Contact collar for heaters	Brass plate, 0.5 mm thick
19	6	Insulating parts	Teflon (PTFE)
20	2	BNC-connectors UG-1094/U	
21	1	Insulating disc 70 mm dia.x 0.25 mm	Teflon (PTFE)
22		Contact ring for grid and anode	bronce wire 0.2 mm dia. wound on 2 mm former
24	1	cooling-air duct	Brass tubing, 40 x 40 x 1 mm
25	1	Insulating ring	Teflon (PTFE)
26	1	Wedge	PVC
	16	Screws 2.6 mm x 6 mm	
	8	Screws 2.6 mm x 8 mm	

4. REFERENCES

(1) R. Jux and H. Dittberner: A transmit mixer and linear amplifier for 23 cm using four 2 C 39 tubes.
VHF COMMUNICATIONS 7, Edition 3/1975, Pages 146 - 160

(2) A. Tautrim: A stripline power amplifier for 70 cm using a 2 C 39 tube.
VHF COMMUNICATIONS 4, Edition 3/1972, Pages 144 - 157

(3) D. Vollhardt: Gittervorspannungserzeugung für Gitterbasis-Stufen mit geerdetem Steuergitter.
UKW-BERICHTE 12, Edition 2/1972, Pages 105 - 106

(4) W. Rahe: A relatively simple linear transmit-converter from 144 MHz to 1296 MHz.
VHF COMMUNICATIONS 8, Edition 2/1976, Pages 66 - 79

DESIGNATION OF THE MICROWAVE BANDS AND WAVEGUIDES

by R. Lentz, DL 3 WR

In the last 30 years, numerous different systems of designating the microwave bands have resulted using letters. Whereas the designation of the microwave bands L, S and X is virtually common throughout the world, the manufacturers of microwave equipment have developed different designation systems for the other bands. In some cases, there is danger of confusion, for instance, when the same letter is used for different bands. No real form of logic seems to exist in any of the designating systems – it seems that the designation letters have some form of historical origin.

Also for waveguides, several different designating systems are used. With the exception of the American JAN numbers, the numerical designations used at least show some form of systematic order. The numerical designations according to the American EIA and IEC contain both information about the waveguide even when this is coded: the EIA number provides the longest inner dimension of the waveguide in one-hundredth of an inch, whereas the IEC number provides the mean frequency of the fundamental wave range in GHz times ten.

The following table that is derived from the magazine »COMMUNICATIONS INTERNATIONAL« – edition February 1976 – shows the various designation systems. It also provides the inner dimensions of rectangular waveguides converted into millimeter. It is recommended that these dimensions be measured when it is possible to obtain waveguide components, especially on the surplus market. It should then be possible with the aid of the table to evaluate the individual parts.

Waveguide - inner dimensions (mm)	(inches)	Frequency (Ghz)	I.E.C	Official designations U.S. (EIA)	U.S. (JAN) W/G	U.S. (JAN) Flansch	U.K. (RCSC)	Company designations U.K.	D-B	FXR	H-P	MRI	Narda	Philips	Sperry	TRG
584.2 x 279.4	23.0 x 11.0	0.32 - 0.49		WR 2300			WG 00									
533.4 x 266.7	21.0 x 10.5	0.35 - 0.53		WR 2100			WG 0			J						
457.2 x 228.6	18.0 x 9.0	0.41 - 0.625		WR 1800	RG-201/U		WG 1									
381.0 x 190.5	15.0 x 7.5	0.49 - 0.75		WR 1500	RG-202/U		WG 2									
292.1 x 146.1	11.5 x 5.75	0.64 - 0.96		WR 1150	RG-203/U		WG 3									
247.7 x 123.8	9.75 x 4.875	0.75 - 1.12		WR 975	RG-204/U		WG 4					P				
195.6 x 97.8	7.7 x 3.85	0.96 - 1.45		WR 770	RG-205/U		WG 5									
165.1 x 82.6	6.5 x 3.25	1.12 - 1.7	R 14	WR 650	RG- 69/U	UG-417 A/U	WG 6	L		L		L	L	L(25cm)	L	L
129.5 x 64.8	5.1 x 2.55	1.45 - 2.2	R 18	WR 510			WG 7							D		
109.2 x 54.6	4.3 x 2.15	1.7 - 2.6	R 22	WR 430	RG-104/U	UG-435 A/U	WG 8			R		M	LS			
86.4 x 43.2	3.4 x 1.7	2.2 - 3.3	R 26	WR 340	RG-112/U	UG/533-U	WG 9A									
72.1 x 34.0	2.84 x 1.34	2.6 - 3.95	R 32	WR 284	RG-48/U	UG-53/U	WG 10	S	L	S	S	S	S	S(10cm)	S	S
58.2 x 29.1	2.29 x 1.145	3.3 - 4.9	R 40	WR 229			WG 11A							A(7.5cm)		
47.5 x 22.1	1.872 x 0.872	3.95 - 5.85	R 48	WR 187	RG-49/U	UG-149 A/U	WG 12	C	K	H	G	C	C	G(6cm)	C	C
40.4 x 20.2	1.59 x 0.795	4.9 - 7.05	R 58	WR 159			WG 13				C			C		
34.8 x 15.8	1.372 x 0.622	5.85 - 8.2	R 70	WR 137	RG-50/U	UG-344/U	WG 14		J	C	J	XB	XN	J(4-5cm)	G	
28.5 x 12.6	1.122 x 0.497	7.05 - 10.0	R 84	WR 112	RG-51/U	UG-51/U	WG 15		H	W	H	XL	XB	H(3-5cm)	H	XL
22.9 x 10.2	0.9 x 0.4	8.2 - 12.4	R 100	WR 90	RG-52/U	UG-39/U	WG 16	X	G	X	X	X	X	X(3cm)	X	X
19.05 x 9.53	0.75 x 0.375	10.0 - 15.0	R 120	WR 75			WG 17		FA		M					
15.80 x 7.90	0.622 x 0.311	12.4 - 18.0	R 140	WR 62	RG-91/U	UG-419/U	WG 18	J	F	Y	P	KU	KU	P(2cm)	U	KU
12.95 x 6.48	0.510 x 0.255	15.0 - 22.0	R 180	WR 51			WG 19				N	K				
10.67 x 4.32	0.420 x 0.170	18.0 - 26.5	R 220	WR 42	RG-53/U	UG-595/U	WG 20		E	K	K		K		K	K
8.64 x 4.32	0.340 x 0.170	22.0 - 33.0	R 260	WR 34			WG 21									
7.11 x 3.56	0.280 x 0.140	26.5 - 40.0	R 320	WR 28	RG-96/U	UG-381/U UG-599/U	WG 22	Q	D	U	R		V	Q(8mm)	V	A
5.69 x 2.84	0.224 x 0.112	33.0 - 50.0	R 400	WR 22	RG-97/U	UG-383/U	WG 23		C	Q			Q		Q	B
4.78 x 2.39	0.188 x 0.094	40.0 - 60.0	R 500	WR 19			WG 24									U
3.76 x 1.88	0.148 x 0.074	50.0 - 75.0	R 620	WR 15	RG-98/U	UG-385/U	WG 25		B	M			M			V
3.10 x 1.55	0.122 x 0.061	60.0 - 90.0	R 740	WR 12	RG-99/U	UG-387/U	WG 26	O	A	E			E	E(4mm)		E
2.54 x 1.27	0.100 x 0.050	75.0 - 110.0	R 900	WR 10			WG 27									W
2.03 x 1.02	0.080 x 0.040	90.0 - 140.0	R 1200	WR 8	RG-138/U		WG 28		W	F			N			F
1.65 x 0.83	0.065 x 0.0325	110.0 - 170.0		WR 7	RG-136/U		WG 29							B(2mm)		D
1.30 x 0.65	0.051 x 0.0255	140.0 - 220.0		WR 5	RG-135/U		WG 30			G			A			G
1.09 x 0.55	0.043 x 0.0215	170.0 - 260.0		WR 4	RG-137/U		WG 31									H
0.86 x 0.43	0.034 x 0.017	220.0 - 325.0		WR 3	RG-139/U		WG 32						R			J

MIXER AND PREAMPLIFIER NOISE AT SHF

by D. Vollhardt, DL 3 NQ

Semiconductor technology has made tremendous advances in the SHF range in the last few years. The result of this is that low-noise transistors are available to the SHF-amateur at reasonable prices. However, a large number of such transistors are offered on the market, and it is difficult to select a transistor with the best price-to-performance relationship.

Table 1 provides information which should be able to aid selection of a suitable type. However, it is only possible to select a really satisfactory transistor when the individual noise components of each stage are known and taken into consideration. The following information can be classed as a continuation of the information given in (11).

It can be classed as being self-explanatory that the required receiver sensitivity can only be obtained using a frequency conversion system, and that this conversion from high to low frequencies is made in a manner that the original frequency band is not inverted. This will be the case when the local oscillator frequency is spaced below the required frequency range to the value of the first intermediate frequency.

1. THE MIXER

It is Schottky diodes that are mainly used for mixers at SHF, and such diodes are available at economical prices. The upper frequency limit of transistor mixers is in the order of the 13 cm band. When compared with the silicon diodes of the 1 N 21/23 family, Schottky diodes have the advantage of being more reliable and robust. Numerous measurements and experiments have shown that Schottky diodes always exhibit the noise figures given in the data sheets, whereas only some individual diodes of the »E« and »F« 1 N 21/23 diode types are able to offer such good values. It should, however, be mentioned that such silicon diodes then usually only require a relatively low diode current of 30 - 100 µA, whereas Schottky diodes usually require 300 - 800 µA, and sometimes upto 1.5 mA, in order to operate satisfactorily. In a recent article (5), it was mentioned, however, that a portion of the required diode current can be impressed by a positive DC voltage, without disadvantage.

Hewlett-Packard diodes of the 5082-.... series, including the special mixer diodes 2213 (stripline) and 2713 (cased) exhibit a noise figure 5.5 dB at 2300 MHz (13 cm band), which seems to be a lower limit for mixer diodes. Cheaper types such as 2215, 2350, and 2817 (glass case) are in the order of 6 to 7 dB; however, even cheaper types not actually designed as mixer diodes such as the 2810, 2811 etc. are usually not more that 1 to 2 dB inferiour. If the case shape of the 1 N 21 is required for replacement reasons, the BAW 95 series manufactured by Philips can be used.

There are usually two footnotes, given in the data sheets, that are often overlooked. One of these indicates that this value indicates the »single sideband noise figure«, which will be discussed later, and the other states that this noise figure will only be valid when the mixer is followed by a correctly matched low-noise IF stage with a total noise figure of max. 1.5 dB.

In the case of a cheap mixer diode, poorly matched to an IF amplifier with a noise figure of 3 dB, it is easily possible for the total noise figure to be 10 to 12 dB, even when the design and matching of the RF-circuitry is ideal. Since diode mixers have a conversion loss, any deterioration of the noise figure of the first IF stage will be fully effective on the overall noise figure. On the other hand, the use of a suitable transistor in the stage directly after the mixer can bring an improvement of 2 to 3 dB.

On the 24 cm band, it is possible to increase the level of the required signal using a relatively inexpensive preamplifier so that the signal is stronger than the mixer noise. Such preamplifiers are expensive for the 13 cm band, and virtually out of the question at present for the higher frequencies. Very low-noise, high-gain transistors are now available on the market for the intermediate frequency, as well as the amateur bands upto 500 MHz. A good example of these is the BFT 66 (see first part of Table 1).

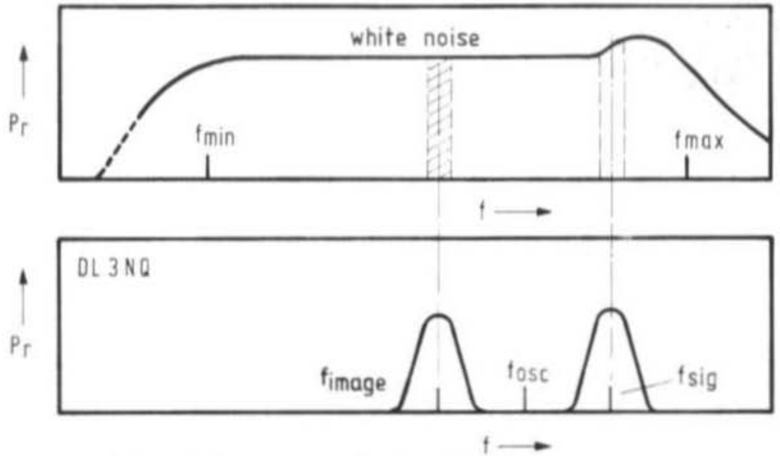


Fig. 1: Noise power as a function of frequency

Above: Noise generator with compensation to increase the upper frequency limit (f_{min} and f_{max}: frequency limits)

Below: Noise power taken by receiver in the required and image frequency passband range

We are now to discuss the term »single sideband noise figure«. It is known, that an image frequency is always generated in a frequency conversion process. In our case, the following frequencies will be received:

$$f_{in} = f_{osc} + f_{IF} \quad \text{as well as} \quad f_{in} = f_{osc} - f_{IF}$$

In the case of 2 m and 70 cm receivers, the image rejection is usually sufficiently high due to the use of selective preamplifier stages and an intermediate frequency of at least 10 MHz. Therefore, a strong image signal can produce a signal in the IF-range, however, it is not able to deteriorate the noise figure to any degree. This is different at the higher GHz bands where the anetnna is often directly connected to the mixer. Let us consider the conditions present when measuring the noise figure of a mixer **(Figure 1)**:

If no special selectivity is provided, the mixer will exhibit a passband curve both for the required and image frequency. This means two equal powers are taken from the white noise spectrum, in which case it is completely immaterial whether single-diode or balanced (push-pull) mixers are used.

This means that the measurement is 3 dB »too good«, and it is the double sideband noise figure that has been measured, and not the single sideband measurement at the required frequency. This shows that it is easy to obtain a noise figure of 6 to 9 dB for the required frequency. There are mixers that use a suitable phase relationship (similar principle to phase SSB), or use a high Q of the mixer circuit (similar to filter SSB) to suppress the unwanted sideband, or at least attenuate it to a great degree. Such mixers are used in professional communications, but it is virtually impossible for them to be obtained by radio amateurs. Usually, wideband mixers are used for amateur applications (References 1 to 5). However, mixers have been described that can be constructed using amateur means (Ref. 6 to 8).

2. IMAGE NOISE AND PREAMPLIFIER STAGES

An image rejection of at least 10 dB should be provided by SHF converters, not only due to the suppression of interference signals, but in order to improve the single sideband sensitivity threshold. However, this selectivity should not be provided directly between antenna and preamplifier since the image noise component of the preamplifier would still be present at the mixer. It is better when this selectivity is made in the mixer circuit, since if no real impedance match is provided at the image frequency, no image noise will be generated. This is important when the mixer is used without preamplifier. It is more favourable even to loose 1 to 1.5 dB on the input side, if it is possible to improve the noise figure by 2 to 3 dB by suppressing the image noise. If preamplifier stages are provided, the mixer noise becomes less important as will now be seen.

The image rejection is very difficult when an IF of 28 MHz is used for 23 cm or even 13 cm converters. The spacing between required and image frequencies will then only amount to 56 MHz. Even the use of the higher frequency of 144 MHz with all its disadvantages (breakthrough of strong stations, or provision of too much preamplification) is not able to provide any considerable improvement of the image rejection, especially when a wideband concept having Pi-networks (2) is used. Measurements on such converters have shown that the image rejection only amounts to approximately 2 dB when using an IF of 144 MHz. The bandwidth of the low-noise preamplifiers described in (9) and (10) amounts to more than 500 MHz and is therefore not able to provide the required single sideband noise figure, if no additional filters are employed. Upto 3 dB of sensitivity is still lost.

Before spending money on low-noise transistors, it is important that the »weakest« point in the chain is found. This was handled in (11), and the following have been taken from section 4 of that article:

$$F_{tot} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \times G_2} + \dots \quad (1)$$

Corresponding to:

$$F_{tot} = 1 + \frac{F_{a2}}{G_1} + \frac{F_{a3}}{G_2 \times G_1} + \dots \quad (2)$$

Where F is the noise factor of the stage; G is the gain factor of the stage and F_a is the additional noise factor $F_a = F - 1$.

In this manner, it is easy to establish the noise component of the terminating resistor, which can be less than 1 in the case of radio astronomy and satellite antennas. This was also discussed in detail in (11).

Furthermore, it is possible to see the noise component of the first and each subsequent stage. It should be remembered that the noise factor or additional noise factor, as well as the gain and gain factor are numerical values and are not identical with the dB values that are often used and which are used in Table 1.

Conversion between noise factor and noise figure, as well as between gain and gain factor where discussed in detail in (11).

However, the above form does not allow us to determine the noise components from the signal and image channels. We only know that the total noise factor F_{tot} is factor 2 greater than that indicated on the noise generator when a double sideband noise figure is present instead of the single sideband value required. For this reason let us rewrite equation 2 for two preamplifier stages and use a block diagram **(Figure 2)** to explain the relationships:

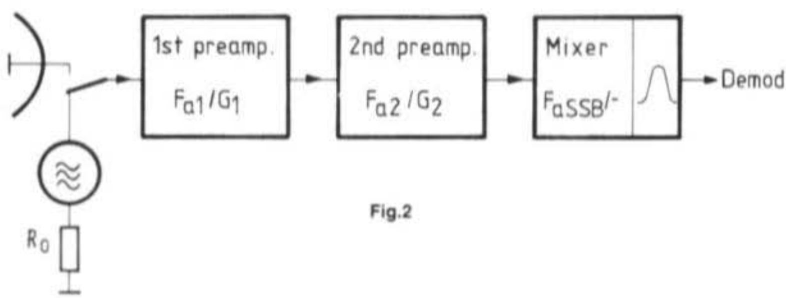


Fig.2

Signal channel: $F_{a1}\ (f_{sig}) + \dfrac{F_{a2}\ (f_{sig})}{G_1} + \dfrac{F_{aSSB}\ (f_{sig})}{G_1 \times G_2}$

Image channel: $1 + F_{a1}\,(f_{image}) + \dfrac{F_{a2}\ (f_{image})}{G_1} + \dfrac{F_{a\,in}\ (f_{image})}{G_1 \times G_2}$

In other words: the total noise factor is comprised of the noise component of the termination resistor (Generator or antenna) at the ambient temperature plus additional noise factor of the first stage, plus additional noise factor of the second stage divided by the gain of the first stage, plus the additional noise factor of the mixer divided by the gain factors of both preamplifier stages. In the case where no selectivity is provide in the image channel, this will mean that this will occur twice so that the total noise figure is twice this value.

These considerations show clearly why the image selection must be made before or in the mixer stage. If a filter is placed between the antenna and the first preamplifier stage, the image noise will be suppressed that originates from the terminating resistor, however, the preamplifier will generate noise in the image channel which will be passed to the mixer. This can be avoided if a filter is used in front of the mixer, and when a selective preamplifier is employed.

Type	Manufacturer	IF – VHF – UHF Frequency (MHz)	NF (dB)	G (dB)	1300 MHz NF (dB)	1300 MHz G (dB)	2300 MHz NF (dB)	2300 MHz G (dB)	One-off price (Oct. 1975) DM net
K 6001	KMC	30	1.0	33	-	-	-	-	120.—
K 6001	KMC	150	1.2	22	-	-	-	-	120.—
K 6007	KMC	30	1.0	35	-	-	-	-	150.—
K 6007	KMC	150	1.2	27	-	-	-	-	150.—
K 6007	KMC	450	1.6	20	-	-	-	-	150.—
BFR 34 A	Siemens	30	2.0	28	see below				14.—
BFR 34 A	Siemens	150	2.1	20	see below				14.—
BFR 34 A	Siemens	450	2.2	15	see below				14.—
MT 2116	Fairchild	450	1.5	18	-	-	-	-	120.—
0817	Avantek	450	1.7	16	-	-	-	-	67.—
BFT 66	Siemens	30	1.0	28	-	-	-	-	43.—
BFT 66	Siemens	150	1.2	24	-	-	-	-	43.—
BFT 66	Siemens	450	1.6	20	-	-	-	-	43.—
MRF 901	Motorola	150	1.6	19	see below				35.—
GAT 1	Plessey	-	-	-	3.5	11	-	-	78.—
GAT 2	Plessey	-	-	-	1.8	11	3.5	9	180.—
BFR 34 A	Siemens	-	-	-	3.2	12	4.4	10	14.—
BFR 14 A	Siemens	-	-	-	3.0	13	4.0	11	115.—
MRF 901	Motorola	-	-	-	2.3	10	3.1	7.5	35.—
DC 5403	AEI	-	-	-	2.4	13	3.2	9.5	280.—
AT 2645 A	Avantek	-	-	-	2.3	15	3.0	12	185.—
AT 4642	Avantek	-	-	-	2.2	14	2.7	10	245.—
AT 4641	Avantek	-	-	-	1.9	15	2.3	11	350.—
AT 4631	Avantek	-	-	-	1.8	15	2.3	11	464.—
BFR 49	Valvo	-	-	-	3.0	12	4.3	9	98.—
FMT 4578	Fairchild			no longer available					
FMT 4575	Fairchild	450	1.3	22	2.0	16	2.8	12	139.—
FMT 4005	Fairchild	-	-	-	1.8	16	2.7	12	221.—
FMT 4000	Fairchild	450	1.0	22	1.5	16	2.3	13	487.—
2 SC 1236	Toshiba	-	-	-	2.3	13	3.2	10	100.—
358...	Hewl.-Pack.								
29 E/Option 100		-	-	-	2.5	12	3.3	8.5	84.—
21 E/22 E/26 E -		-	-	-	2.7	11	4.2	6.5	78.—/89.—/57.—
61/62 E/Opt.100		-	-	-	2.2	14	2.8	10	310.—
66 E/Opt.100		-	-	-	2.2	13	2.8	10	139.—
NE 02103	NEC	(= NE 02107 = V 021)			2.9	11	3.8	8	66.—
NE 41705-2	NEC	(= NE 41703-2 = V 912 B)			2.8	14	3.7	10	82.—
NE 41703-1	NEC	(= 2 SC 1336 = V 912 A)			2.6	12	3.5	9.5	102.—
NE 57803-1	NEC	(= /05-1 = /07-1 = V 913 A)			2.3	14	2.8	10.5	260.—
NE 22207-3	NEC	(= V 222-3) at 4 GHz: 4-5			6	-	2.7	11	Grade D: 139.—
NE 22207-1	NEC	(= V 222-1) at 4 GHz: 3.5			7	-	2.7	11	Grade D: 450.—
NE 24406	NEC	(V 244 C) at 4 GHz: 3.0			10	-	2.2	14	1040.—
BFR 91	Valvo	450	1.9	13	3.6	10	6.5	8	9.—

3. EXAMPLES

Several practical examples are now to be calculated in order to show how this theoretical information is put into practice. A wideband mixer is assumed. Its image noise component is taken into consideration by adding the mixer-noise component twice. It is also assumed that 2 dB are lost between preamplifier and mixer due to an efficent selectivity. This is taken into consideration by using a factor of 1.6 in the numerator. This is simplified in equation (3) to:

$$F_{total} = 1 + F_{a1} + \left(\frac{F_{a2}}{G_1} \right) + \frac{2 \times 1.6 \times F_{a\,SSB}}{G_1 \times (G_2)} \quad (4)$$

If two preamplifier stages are used, the whole of equation (4) will be required; if only one preamplifier stage is provided, the parts given in brackets can be deleted.

It should be taken into consideration that the operating point for minimum noise does not usually coincide with that for optimum gain. This is not only valid for the collector current, but also for the input and output matching. It is dependent on the specific combination of the stages, which is the most favourable configuration to obtain the lowest total noise factor. It is sometimes advantageous to use a mixer with an inferior noise factor but higher gain, especially when it is cheaper. When using two preamplifier stages, it is always advisable to align the first stage for noise matching (min. noise) and the second stage for power matching (max. gain). The noise and gain values given in Table 1 are valid for noise matching, and have been extrapolated for the 13 cm and 23 cm bands from the typical data given by the manufacturer. Although they vary from transistor to transistor, it can be classed that the noise figure will increase by 1 dB and the gain by 2 dB in the power matching mode. In order to avoid confusion, the additional noise factor will be given as F_a whereas the noise figure in dB will be given as usual as NF. NF_{SSB} is the noise figure of the signal path (without image). The gain in dB is given as G, and the numerical gain factor as GF.

3.1. Examples for 1296 MHz

a) Mixer NF_{SSB} = 10 dB ($F_{a\,mix}$ = 9); one preamplifier stage with transistor MRF 901 noise-matching, NF = 2.3 dB (F_a = 0.7); G = 10 dB (GF = 10):

$$F_{total} = 1 + 0.7 + \frac{2 \times 1.6 \times 9}{10} = 4.57 \mathrel{\hat{=}} \mathbf{6.6\ dB}$$

b) Mixer as a) but using power matched preamplifier BFR 34 A with NF = 4.2 dB (F_a = 1.6) G = 14 dB (GF = 25):

$$F_{total} = 1 + 1.63 + \frac{2 \times 1.6 \times 9}{25} = 3.78 \mathrel{\hat{=}} \mathbf{5.8\ dB}$$

c) Mixer with NF_{SSB} = 8 dB ($F_{a\,mix}$ = 5.3); noise-matched FMT 4005 preamplifier with NF = 1.8 dB (F_a = 0.5); G = 16 dB (GF = 40):

$$F_{total} = 1 + 0.5 + \frac{2 \times 1.6 \times 5.3}{40} = 1.924 \mathrel{\hat{=}} \mathbf{2.85\ dB}$$

d) Mixer as c) but with two-stage preamplifier. First stage noise-matched MRF 901, second stage power matched BFR 34 A :

$$F_{total} = 1 + 0.7 + \frac{1.63}{10} + \frac{2 \times 1.6 \times 5.3}{10 \times 25} = 1.927 \triangleq \mathbf{2.85\ dB}$$

e) Mixer with NF_{SSB} = 6 dB ($F_{a\ mix}$ = 2) with two stage preamplifier. First stage noise-matched FMT 4575, NF = 2 dB (F_a = 0.58); G = 16 dB (GF = 40); second stage power-matched MFR 901 with NF = 3.3 dB (F_a = 1.14); G = 12 dB (GF = 15.8) :

$$F_{total} = 1 + 0.58 + \frac{1.14}{40} + \frac{2 \times 1.6 \times 3}{40 \times 15.8} = 1.624 \triangleq \mathbf{2.1\ dB}$$

f) Finally the mixer described in (7) combined with the two-stage preamplifier version given in (9). The mixer exhibits a high image rejection which means that the image noise component as well as the loss of 2 dB in the assumed filter between preamplifier and mixer can be deleted. The mixer NF is given NF_{SSB} = 5.5, which seems to be rather optimistic. The preamplifier is said to exhibit a overall gain of 20 dB with a NF = 2.3 dB, which can therefore be classed as a single stage.

$$F_{total} = 1 + 0.7 + \frac{2.55}{100} = 1.725 \triangleq \mathbf{2.4\ dB}$$

3.2. Examples for 2304 MHz

a) Mixer with NF_{SSB} = 10 dB with one noise-matched BFR 34 A, without filter in front of mixer so that full image noise is present. NF = 5.4 dB (F_a = 2.46), G = 12 dB (GF = 15.8) :

$$F_{total} = 2 + 1 + 2.46 + \frac{9}{15.8} = 8.12 \triangleq \mathbf{9.1\ dB}$$

b) Mixer as a) but with filter in front of mixer and power-matched BFR 34 A. NF = 5.4 dB (F_a = 2.46); G = 12 dB (GF = 15.8) :

$$F_{total} = 1 + 2.46 + \frac{2 \times 1.6 \times 9}{15.8} = 5.26 \triangleq \mathbf{7.2\ dB}$$

c) As b) but with additional preamplifier stage with noise-matched BFR 14 A. NF = 4 dB (F_a = 1.5); G = 11 dB (GF = 12.6) :

$$F_{total} = 1 + 1.5 + \frac{2.46}{12.6} + \frac{2 \times 1.6 \times 9}{12.6 \times 15.8} = 2.84 \triangleq \mathbf{4.55\ dB}$$

d) Mixer with NF_{SSB} = 7 dB ($F_{a\,mix}$ = 4); second preamplifier power matched BFR 34 A. First preamplifier MRF 901 with NF = 3.1 dB (F_a = 1.04); G = 7.5 dB (GF = 5.6) :

$$F_{total} = 1 + 1.04 + \frac{2.46}{5.6} + \frac{2 \times 1.6 \times 4}{5.6 \times 15.8} = 2.65 \;\hat{=}\; \mathbf{4.2\ dB}$$

e) Mixer as d); second preamplifier power-matched MRF 901; first preamplifier noise-matched V 222-3 with NF = 2.7 dB (F_a = 0.86); G= 11 dB (GF = 12.6) :

$$F_{total} = 1 + 0.86 + \frac{1.57}{12.6} = \frac{2 \times 1.6 \times 4}{12.6 \times 8.8}\ 2.1 \;\hat{=}\; \mathbf{3.2\ dB}$$

It will be seen in these examples and in Table 1 that it is possible to obtain good sensitivity values when using two preamplifier stages, careful selection of the transistors, and that this need not be too expensive.

4. FURTHER DETAILS

A discussion of individual design parameters is not to be included in this article. However, it should be remembered that the given noise figure values for noise matching will only be obtained, when no selective circuits are provided before the first preamplifier stage (selectivity always brings losses). Such losses must be kept to a minimum as was clearly stated in (11). However, the selectivity should be provided as soon as possible for cross modulation reasons, if possible in the collector circuit of the first preamplifier stage.

Let us reconsider example e) in 3.2. and assume that the image rejection previous to the mixer were only 10 dB. This means that the values of the image channel would still be effective to the value of 10 %, and this would cause the noise figure to deteriorate from 3.2 to 3.6 dB. In the case of an image rejection of 20 dB, the result would be 3.25 dB. The better the preamplifier, the better must be the image rejection. One may question the validity of the assumed 100 % rejection of the image channel previous to the mixer. However, the possible errors due to this assumption are very small so that they are of no practical significance. On the other hand the assumption that the image noise of the mixer is fully effective is rather too pessimistic since the mixer will »see« considerable reactive components of the assumed filter at the image frequency.

Furthermore, one must remember that the spread of individual transistors of one type can be as great as ± 0.5 dB from the given values (usually towards higher values !). The same is valid for the gain.

We should also consider the behavior of noise generators at their upper frequency limits. Very often compensation networks are provided to ensure that a usable noise spectrum is available at higher frequencies. This often causes the actual noise power to rise before falling off at the upper frequency limit. More expensive noise generators often provide a frequency-dependent correction table, but such information is usually lost, not only on the surplus market. This means that care must be taken in the vicinity of the upper frequency limit, otherwise measuring errors of several dB can occur. The same is valid for the provision of highly selective filters between the noise generator and test object. On one hand, unfavourable lengths of cable (multiples of $\lambda/4$) can cause considerable falsification of the terminating impedance of the noise generator, on the other hand, such a filter would probably provide considerable different conditions than when connecting the object later to to the antenna. It is always adviseable to check the results of measurement made with noise generators by measuring the signal-to-noise ratio of a weak signal. One is often confronted with surprises in this respect.

It should be self-explanitory that the following information should be always provided, and not only for SHF-converters: Noise figure, overall gain, 3-dB-bandwidth, image rejection, and if possible also cross modulation rejection and large-signal handling.

Finally a tip on aligning for best signal-to-noise ratio: A receiver with FM discriminator should be used as IF. Tune in a weak carrier at the noise threshold. In this range an RF signal-to-noise improvement of 3 dB will bring an improvement of more than 12 dB of the AF signal. This means that one has a clear indication of every improvement of the sensitivity.

5. REFERENCES

(1) Information Booklet »Betriebsfertige Geraete«:
Description of the Converters MMC 1296/28 and MMC 1296/144
Available from the Publishers of VHF COMMUNICATIONS

(2) K. Hupfer: A Stripline Converter for the 13 cm Band
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 238 - 245

(3) A. Schädlich: A Receive Converter for the 13 cm Band with Diode Mixer
VHF COMMUNICATIONS 7, Edition 3/1975, Pages 161 - 167

(4) L. May and B. Lowe: A Simple and Efficient Mixer for 2304 MHz
QST 58 (1974) Heft 4, Pages 15 - 19 and 31

(5) P.C. Wade: High-Performance Balanced Mixer for 2304 MHz
ham radio 8 (1975), Edition 10, Pages 58 - 62

(6) E. Hunecke: Ein 1297/145-MHz-Konverter mit Halbleitern
UKW-BERICHTE 8, Edition 2/1968, Pages 61 - 80

(7) R.E. Fisher: Interdigital Converters for 1296 and 2304 MHz
QST 58, Edition 1/1974, Pages 11 - 21

(8) P. Shuch: How to use double-balanced Mixers on 1296 MHz
ham radio 8, Edition 7/1975, Pages 8 - 15

(9) P. Shuch: Microstripline Preamplifiers for 1296 MHz
ham radio 8, Edition 4/1975, Pages 12 - 27

(10) P. Shuch: Low-cost 1296-MHz-Preamplifier
ham radio 8, Edition 10/1975, Pages 42 - 46

(11) R. Lentz: Noise in Receive System
VHF COMMUNICATIONS 7, Edition 4/1975, Pages 217 - 235

VHF SERVICES SUITABLE FOR USE AS PROPAGATION INDICATORS

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

1. INTRODUCTION

One often reads of VHF broadcasting stations in the 88 MHz to 100 MHz band being monitored for indicating improved conditions for long distance VHF communications. Unfortunately, this does have a number of disadvantages since each of these VHF-FM channels has been allocated to a large number of different transmitters spread throughout the continent. Under improved conditions usually a number of stations are received at the same time so that identification is difficult; even if this is not the case it usually takes some time to identify the country concerned let alone the transmitter. The same is true for Band III TV-transmissions. Also the lower VHF-FM channels are too far frequency-wise from our two meter band.

However, there are a large number of aeronautical beacons and meteorological broadcast stations in continuous service just below the two meter band that could be very useful for indicating improved conditions. These are to be split into two categories: Firstly the high power aeronautical and meteorological broadcast stations.

2. TYPES OF RADIO SERVICES

2.1. Meteorological Broadcasts

Most of these broadcast stations operate on a 24 hour basis, but their actual geographical location is not published exactly. A list of the European meteorological broadcast stations for the aeronautical service are as follows:

Area	Identification	Frequency (MHz)	Hours of operation
Amsterdam	Met broadcast	126.20	0608-2208 GMT
Athens	Volmet	127.80	continuous
Barcelona	Volmet	127.60	continuous
Belgrade	Volmet	112.30	0600-1800
Bordeaux	Radio	126.40	continuous
Brindisi	Volmet	127.60	continuous
Brussels	Brussels	127.80	continuous
Copenhagen	Volmet	127.00	continuous
Frankfurt	Volmet	127.60	continuous
Geneva	Met broadcast	126.80	continuous
Hannover	Bremen Volmet	127.40	continuous
Helsinki	Volmet	128.40	0400-2200
Jonkoping	Volmet	127.20	0530-2100
Lisbon	Volmet	126.40	continuous

Area	Identification	Frequency (MHz)	Hours of operation
London South	Volmet	128.60	0600-2130
			Mon-Fri: 0800-1600
			Mon-Fri: 1600-0800
			Fri-Mon: 1600-0800
London North	Volmet	126.60	0600-2130
London	Volmet	128.60 / 126.60	21.30-0600
Luxembourg	LUX	114.00	0450-2150
Madrid	Volmet	126.20	continuous
Marseille	Radio	127.40	continuous
		119.75	continuous
Milan	Volmet	126.60	continuous
Oslo	Volmet	128.60	0415-2330
Paris	Radio	126.00	continuous
		125.15	continuous
Patscherkofel	Volmet Radio	127.00	continuous
Rauchenwarth	Wetterrundsendung Vienna	122.55	0630-1800
Rome (Ciampino)	Volmet	126.00	continuous
Seville	Volmet	127.00	continuous
Shannon	Radio	127.00	continuous
Stockholm	Volmet	127.60	0530-2100
Sundsvall	Volmet	127.80	0600-2100
Zagreb	Volmet	127.80	0630-1900
Zugspitze	Volmet Broadcast	126.40	Apr-Sep: 0900-1630
	Innsbruck		Oct-Mar: 0900-1500
Zurich	Met Broadcast	127.20	continuous

2.2. Aeronautical Terminal Information Service (ATIS)

This service provides aircraft with airfield information but is not usually available on a 24 hour basis. The information is often transmitted via existing radio navigation facilities.

The range is usually less than that of the meteorological stations. Where VHF is given, the indication will be given in voice, whereas the indication will be made in morse code (A 2) when a VOR beacon and three letter callsign are indicated. A list of these services in Europe is given on the next two pages:

2.3. Radio Navigation Facilities

Of the various forms of radio navigation facilities, only the VHF Omni Radio Range (VOR) and Tactical VOR (VORTAC) are to be considered. This is because only these are located in the frequency range of interest. Such transmissions consist of a continuous carrier frequency in conjunction with a second phase modulated carrier, whose phase difference to the reference carrier to the magnetic bearing of the aircraft from the beacon.

The identification is usually made in Morse code in the A 2 mode. Some of these VOR and VORTAC beacons are also amplitude modulated with voice aeronautical information in the A 3 mode.

Location	Identification	Frequency (MHz)	Hours of operation
Amsterdam	VOR/PAM	117.8	0608-2208
Amsterdam	VOR/SPL	108.4	0608-2208
Amsterdam	VOR/SPY	113.3	0608-2208
Amsterdam	VOR/TS	113.9	0608-2208
Athens	VHF	123.40	continuous
Berlin (Tegel)	VOR/TGL	112.3	0500-2300
Berlin (Tegel)	VOR/HVL	113.3	0500-2300
Berlin (Tempelhof)	VOR/TOF	114.1	0500-2300
Bremen	VOR/BMN	111.6	0500-2400
Cologne-Bonn	VOR/KBO	108.8	0500-2400
Copenhagen (Kastrup)	VHF	122.75	0600-2200
Copenhagen (Kastrup)	VHF	122.85	0600-2200
Copenhagen (Kastrup)	VHF	122.75	2200-0600
Copenhagen (Kastrup)	VHF	122.75	continuous
Dusseldorf	VOR/DUS	115.9	0500-2400
Dusseldorf	VOR/BAM	113.6	0500-2400
Frankfurt	VOR/CHA	115.5	0500-2400
Frankfurt	VOR/FFM	114.2	0500-2400
Frankfurt	VOR/MTR	117.7	0500-2400
Frankfurt	VOR/TAU	116.7	0500-2400
Frankfurt	VOT	108.2	0500-2400
Geneva (Cointrin)	VHF	122.75	0500-2300
Goteborg	VHF	122.80	0600-2100
Hamburg	VOT	108.0	0500-2400
Hamburg	VOR/HAM	113.1	0500-2400
Hamburg	VOR/LBE	115.1	0500-2400
Hannover	VOR/HNV	108.2	0500-2400
Hannover	VOR/DLE	115.2	0500-2400
Hannover	VHF	121.85	0500-2400
Helsinki	VOR/HEL	114.8	0400-2200
Helsinki (Malmi)	VHF	122.70	0500-1900
Kuopio	VOR/KUO	113.8	0500-1900
London (Heathrow)	VOR/BIG	115.1	0500-2200
London (Heathrow)	VOR/BNN	112.3	0500-2200
London (Heathrow)	VHF	121.85	0600-2300
London (Gatwick)	VOR/MAY	117.9	0500-2200
Lyneham	ILS/LA	249.4	continuous
Malmo (Sturup)	VOR/SUP	113.0	0525-2300
Manchester (Barton)	VOR/BTN	115.6	0500-2250
Manchester (Barton)	VHF	121.95	0500-2250
Munich	VOR/MUN	112.3	0500-2400
Munich	VOR/ERD	113.6	0500-2400
Munich	VOR/WBU	116.7	0500-2400
Nurnberg	VOR/NUB	114.9	0500-2400
Oslo (Fornebu)	VOR/FBU	112.9	0530-2300
Oslo (Fornebu)	VHF	123.70	0530-2300

Location	Identification	Frequency (MHz)	Hours of operation
Paris (Charles de Gaulle)	VHF	128.00	continuous
Paris (Le Bourget)	VHF	120.0	continuous
Paris (Orly)	VHF	126.50	continuous
Rome (Fiumicino)	VOR/OST	114.9	0700-1900
Rotterdam	VOR/RTM	110.4	0625-2055
St. Yan	VHF	122.7	Mon-Fri: 0600-1800
Stockholm (Arlanda)	VOR/ARL	112.7	0600-2100
Stockholm (Bromma)	VHF	122.45	0600-2100
Stuttgart	VOR/TGO	112.5	0500-2400
Stuttgart	VOR/LBU	109.2	0500-2400
Sundsvall	VOR/SUN	113.1	0600-2100
Toussus-Le-Noble	VOR/TSU	108.2	0700-2100
Vienna (Schwechat)	VOR/WGM	113.5	0600-2100
Vienna (Schwechat)	VOR/BRK	111.2	0600-2100
Vienna (Schwechat)	VOR/TUN	113.0	0600-2100
Vienna (Schwechat)	VHF	122.95	0600-2100
Vienna (Schwechat)	VOR/SNU	115.5	0600-2100
Zurich	VOR/TRA	114.3	0500-2300
Zurich	VOR/KLO	116.4	0500-2300

Due to the large number of these stations even in Europe, it would take too much space in the magazine to bring a list of these facilities. A complete list has, however, been prepared, and such a list with exact geographical location of all such beacons in Europe can be provided for a small fee. If sufficient interest is shown, the editors would be only too pleased to bring this list in VHF COMMUNICATIONS.

3. REFERENCES

The about information has been taken from various official aeronautical publications published in Europe.

FOURTH FRENCH EDITION OF VHF COMMUNICATIONS

There are now four French editions of VHF COMMUNICATIONS available. Please contact the publishers or our French representative for more details:

Mlle. **Christiane MICHEL** · Les Pillés · F-89117 PARLY (France)

Please inform your French speaking friends of these omnibus editions of VHF COMMUNICATIONS.

A SENSITIVE 500 MHz 10:1 PRESCALER AND PREAMPLIFIER FOR FREQUENCY COUNTERS

by J. Grimm, DJ 6 PI

The increasing activity on the 70 cm band means that frequency counters with an upper frequency limit of 500 MHz are becoming more and more interesting. A 500 MHz prescaler was described in (1). However, in this prescaler, the input frequency was divided in three integrated circuits by 2, by 10, and finally by 5. This now seems to be a very extensive manner of obtaining the overall frequency division ratio of 100:1. Relatively inexpensive prescalers are now available that allow a frequency division of 10:1 in a single integrated circuit. The Plessey SP 8515 and Fairchild 11 C 90 are examples of this new generation.

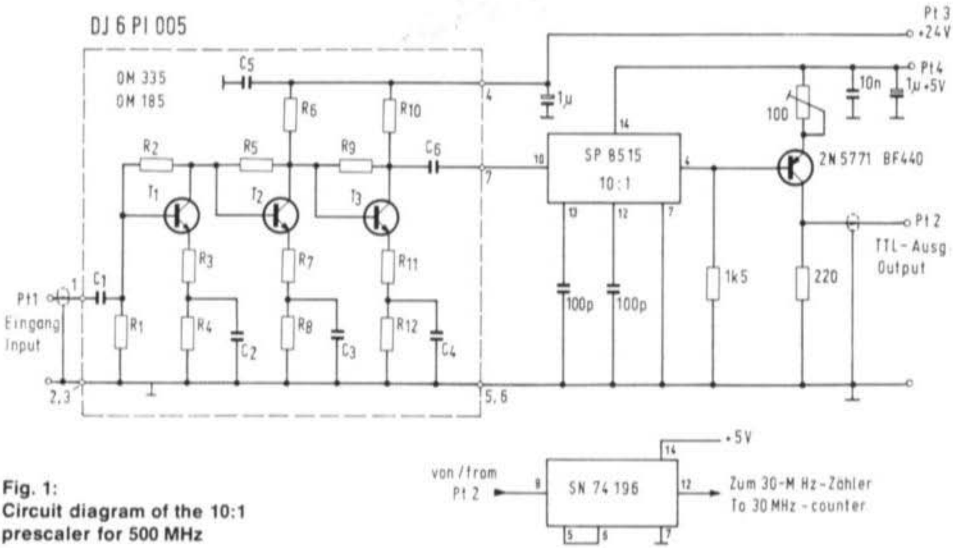


Fig. 1: Circuit diagram of the 10:1 prescaler for 500 MHz

1. CIRCUIT

The described prescaler uses the 10:1 divider SP 8515 which operates typically up to 500 MHz, and some up to 600 MHz. A preamplifier is necessary in front of the frequency divider which is able to increase the RF voltage to be counted to the level required by the divider. The integrated amplifier module OM 185 manufactured by Philips was selected since it is difficult, when using an amplifier circuit with descrete components, to obtain a virtually constant characteristic of the gain from the lower HF-range up into the UHF-range. The OM 185 is a wideband amplifier in thin-film hybrid technology with 27 dB gain up to 860 MHz. The combination ensures that the prescaler module is extremely sensitive.

The circuit diagram of the module is shown in **Figure 1**. The ECL-output signal of the frequency divider SP 8515 is converted to TTL-level with the aid of a level converter stage equipped with a PNP transistor. This means that the output (Pt 2) can be directly connected to the counting gate (usually a SN 74 S 00 N) of the frequency counter. Of course, it is necessary that the counter is able to count up to 50 MHz. For older counters whose frequency limit is in the order of 30 MHz, it will be necessary for a further 10:1 divider equipped with an IC SN 74196 N to be connected between the prescaler and the counter. A suitable circuit is also given in Figure 1.

One disadvantage of the integrated amplifier OM 185 is the required operating voltage of 24 V. It is possible, however, to operate this amplifier at a lower voltage of down to 12 V; however, the sensitivity will decrease. Simple methods of obtaining 24 V from an available power supply are to be discussed in section 5.

2. SPECIFICATIONS

As can be seen in **Figure 2**, a voltage of 0.9 mV is sufficient for measurements at 150 MHz, whereas 2 to 3 mV are sufficient for a reliable count in the 70 cm band. The given values, which are valid for an operating voltage of 24 V / 5 V, can differ slightly. If the amplifier OM 185 is only provided with 12 V (lowest limit), the required input voltage will be approximately doubled.

The input impedance of the amplifier amounts to approx. 75 Ω, and the required operating current is in the order of 35 mA (24 V); approximately 70 mA are required at 5 V.

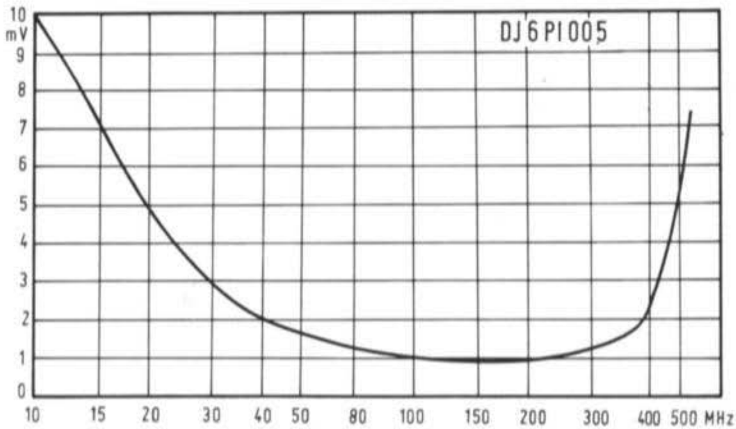


Fig. 2: Input sensitivity of the prescaler module

3. CONSTRUCTION AND COMPONENT INFORMATION

All components of the prescaler module are mounted on a PC-board having the dimensions 60 mm x 40 mm. The PC-board is double-coated in order to ensure good grounding conditions. Through-contacts are provided in order to assist the construction. If a PC-board is used without through-contacts, it is necessary for the connections of the components marked with a »X« in the component location plan to be soldered to the top and lower side of the board. **Figure 3** shows a drawing of this board which has been designated DJ 6 PI 005.

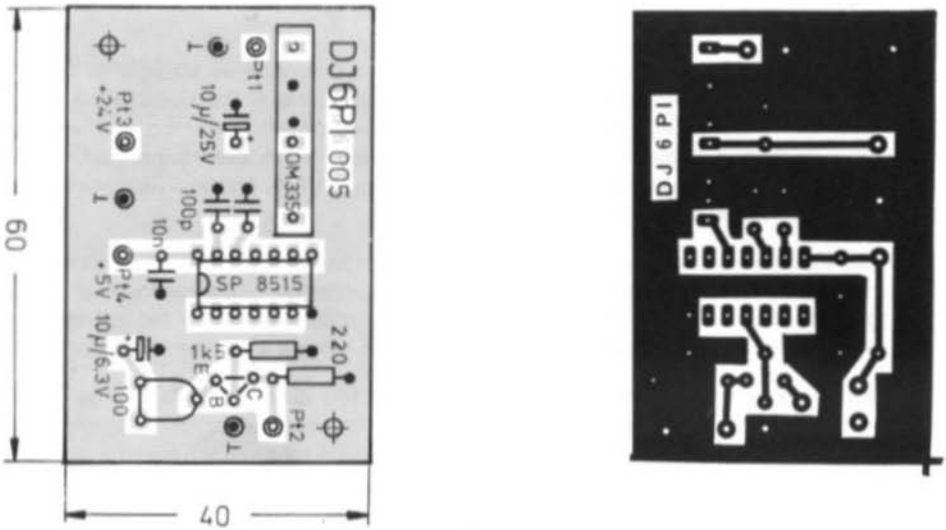


Fig. 3: Double-coated PC-board DJ 6 PI 005

Philips also manufacture another version of this amplifier under the designation OM 335. Both amplifier types can be used in this module. It is not possible to insert the amplifier the wrong way round since they both have an asymmetric connection pin configuration.

The value of the bypass capacitors for the divider SP 8515 is not critical in the range of approx. 10 pF to 1 nF. It is only important that the wire lengths are kept to a minimum.

A large number of silicon PNP transistors can be used in the level converter. The author has used both types BF 324 and BSX 29, but transistors BCY 71 or 2 N 5771, BC 213, or BC 415 should also be suitable. All have the same connection diagram E-B-C. If the connections are changed, it would also be possible to use the BF 440 (B-E-C).

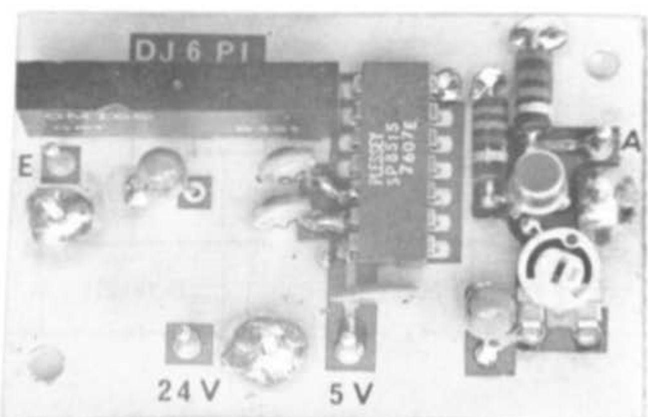


Fig. 4: Photograph of the author's prototype DJ 6 PI 005

4. ALIGNMENT

The input of the prescaler module should be connected as near as possible to the input socket of the frequency counter via a short piece of coaxial cable. The output voltage of the TTL-level converter is aligned with the 100 Ω trimmer potentiometer so that a reliable frequency indication is given at the lowest input voltage. It has been found that the most favorable adjustment is usually approximately 100° from the fully anticlockwise position. The potentiometer is then left in this position.

It is self-explanatory, that the prescaler is only coupled to the frequency to be measured as loosely as possible. In order to protect the input of the preamplifier against too high an input voltage, it is possible for two Schottky diodes (e.g. HP 2800) to be connected in antiphase at the input of the amplifier module. However, it is more reliable when an attenuator is used.

The author also made experiments using the Fairchild 10:1 divider 11 C 90 together with the preamplifier OM 335. However, the results were not satisfactory since the integrated circuit 11 C 90 only operated in a relatively limited input voltage range. No frequency count was made above or below this range, and the fault seemed to lie with the 11 C 90. On the other hand, the upper frequency limit was in the order of 750 MHz.

5. 24 V SUPPLY

A transformer winding of 8 to 14 V is usually available in the power supply for the 5 V operating voltage of the TTL circuits. It is relatively easy to obtain the required 24 V DC by voltage multiplication from this voltage. If a voltage of 12 to 14 V is available from the transformer, a voltage doubler circuit as shown in **Figure 5** should be used: if only 8 to 10 V AC-voltage are available, a voltage tripler circuit as shown in **Figure 6** would be suitable. It is not necessary to stabilize the output voltage. The power zener diode ZL 24 with dropper resistor is only used for limiting the voltage to 24 V.

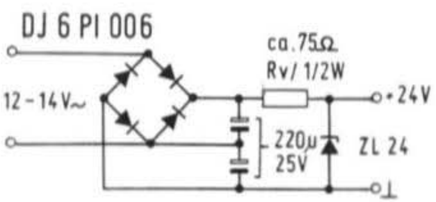


Fig. 5: Voltage doubler circuit

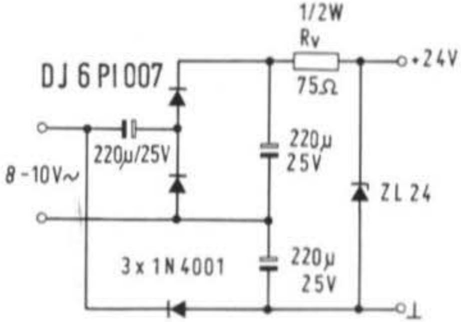


Fig. 6: Voltage tripler circuit

Two PC-boards have been designed for the 24 V supply: DJ 6 PI 006 (65 mm x 50 mm) for the voltage doubler and DJ 6 PI 007 (70 mm x 55 mm) for the voltage tripler. **Figures 7 and 8** give the components locations on these boards, and **Figure 9** shows photographs of the author's prototypes.

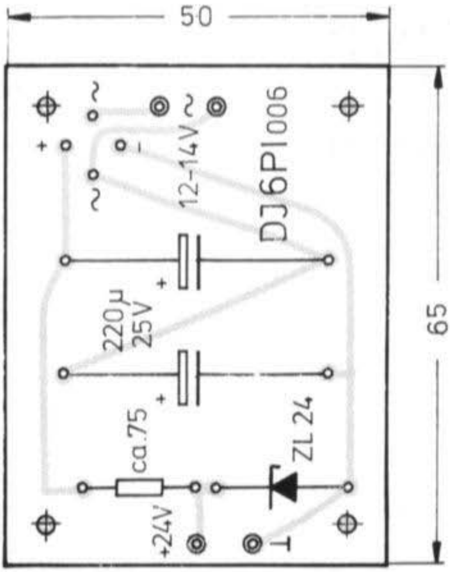


Fig. 7:
Voltage doubler
board DJ 6 PI 006

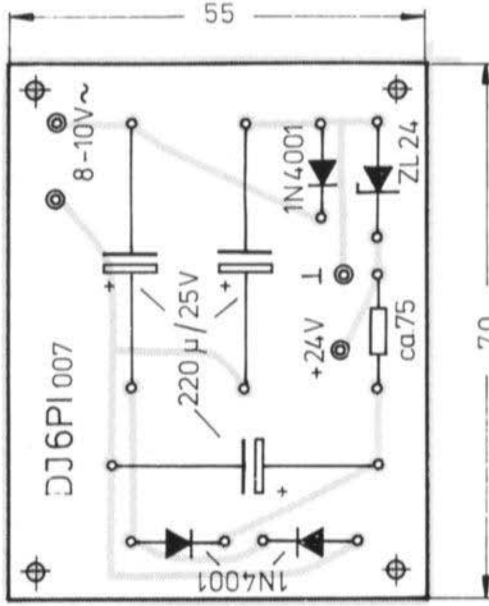


Fig. 8:
Voltage tripler
board DJ 6 PI 007

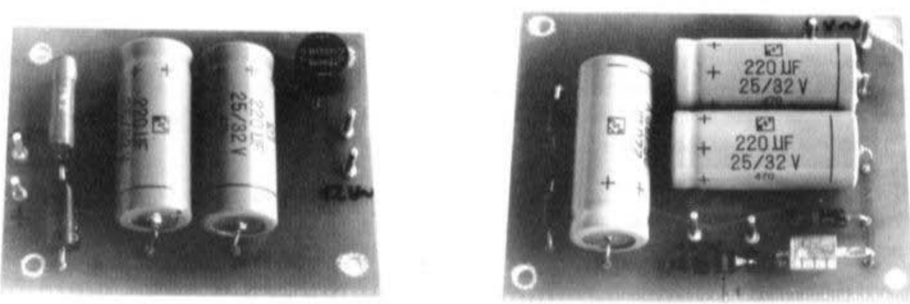


Fig. 9: Photograph of the author's prototypes

6. REFERENCES

(1) G. Bergmann, M. Streibel: A 500 MHz Prescaler and Preamplifier for Frequency Counters
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 238 - 245

CALLING-TONE DECODER AND OSCILLATOR

by R. Reuter, DC 6 FC

Several semiconductor manufacturers offer completely integrated stereo decoders which regenerate the required 38 kHz subcarrier in a phase-locked loop (PLL). The actual oscillators operate at twice the frequency (76 kHz), and the 19 kHz pilot tone provided in the multiplex signal is used as reference frequency at the phase comparator. Such a circuit was described, for instance, in (1). These circuits can also be used for other frequencies, such as for a 1750 Hz calling tone. The external circuitry is really simple if an LC-circuit is not used as in the case of the CA 3090, but a RC-link as in the case of the μA 758 (Fairchild).

A lamp (LED) will indicate when the frequency to which the circuit is aligned is present. The relatively short calling tone can also be stored, which means that the indicator lamp will continue to light until the storage is cancelled. The oscillator of the PLL-circuit can be used as calling-tone oscillator in the transmit mode. A module using this dual-mode as calling-tone decoder and oscillator is now to be described. It is accommodated on a 45 mm x 35 mm PC-board and requires very few components besides the integrated circuit.

1. OPERATION

The principle of operation is now to be described with the aid of the block diagram of the integrated circuit μA 758 **(Figure 1)** and the circuit diagram given in **Figure 2**. The oscillator (VCO) oscillates at four times the required frequency, which is determined by the RC-link connected to pin 15. With the circuitry given in Figure 2, the oscillator can be tuned to frequencies between approx. 5.2 and 9.2 kHz, which corresponds to an actual frequency range of 1.3 to 2.3 kHz. The capacitor C 3 should be a plastic foil type (Styroflex). A helical potentiometer should be used for the trimmer resistor P 1. The frequency is divided twice by 2 and then appears at connection 11 in a square-waveform with a ratio of exactly 1:1. A two-stage lowpass filter at the input and output of transistor T 1 filters out the high frequency components and feeds the amplified fundamental frequency to a buffer stage (impedance converter) accessible at pin 6. The output signal can be taken from pin 5 (sinusoidal) and fed via Pt 4 to the transmit modulator.

Pin 1 of the integrated circuit is connected via Pt 1 to the hot end of the volume control of the receiver. The composite signal (noise, voice, calling tone) is amplified in the integrated circuit and passed from pin 2 to pin 12. It is then fed to both a phase and amplitude detector. If the design frequency is present, the phase-locked loop will look in and the VCO-frequency will coincide to four times the required frequency. This will cause also the amplitude detector to be actuated and will switch on a lamp or a relay via the internal driver. This current is available at pin 7. The lamp or relay driver can be short-circuited and can provide up to 100 mA. In the case of LED indicators, it will be necessary to provide a current limiting resistor.

The calling tone indication can be stored by providing a DC-feedback from pin 7 to pin 10, in other words via the driver. The indication will remain until pin 9 is grounded via a decoupling diode. If the storage facility is not required, diode D 2 to D 4 will not be required.

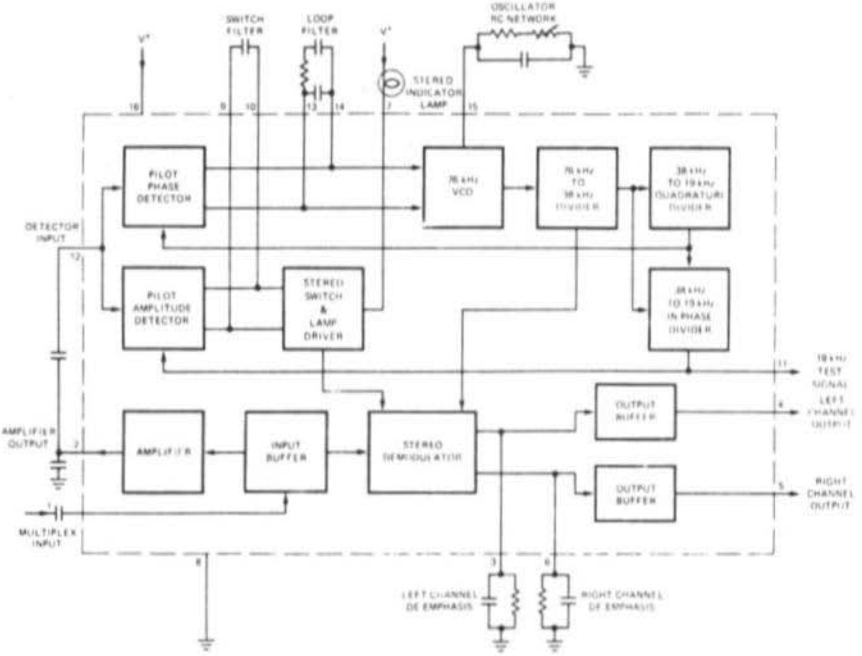


Fig. 1: Circuit diagram of the stereo decoder µA 758

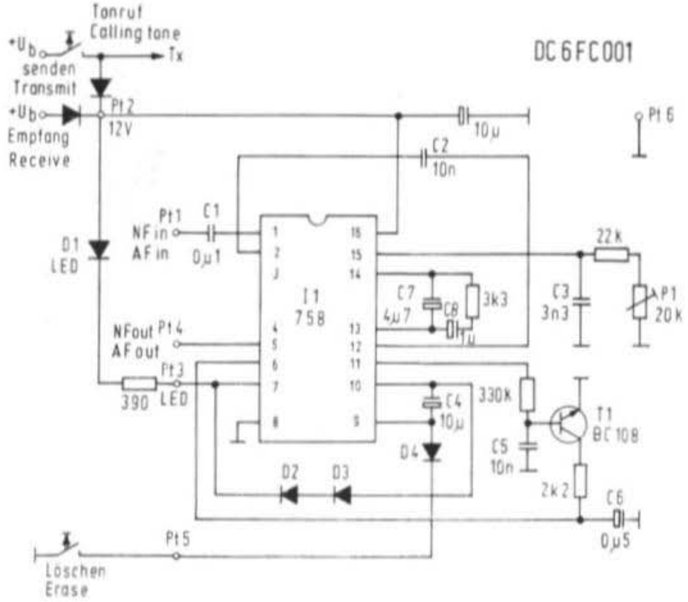


Fig. 2: Circuit diagram of a calling tone decoder and oscillator

The bandwidth of the PLL-circuit only amounts to approximately 100 Hz which means that the frequency of the calling tone of the partner station must be relatively accurate. On the other hand, the low bandwidth reduces the possibility of the circuit being actuated by the voice signal, or unwanted tones. In addition to this, the calling tone must be present for a period of 0.3 to 1 s, so that the circuit is able to carry out a positive evaluation. The actuation time is dependent on the level of the input voltage, and from the value of capacitor C 4.

After the phase-locked loop has locked into place, it is possible for the frequency to vary by ± 500 Hz from the nominal frequency. This means that the calling tone need not be too stable in the frequency, as long as the commencement frequency is correct.

The sensitivity of the decoder circuit can be varied by exchanging capacitor C 2. The given values are valid for a capacitance of 10 nF. At higher values, the sensitivity will be higher.

2. TECHNICAL DATA

Operating voltage:	10 to 16 V
Operating current without lamp:	approx. 30 mA
Sensitivity (RMS):	approx. 100 mV
Input impedance:	35 kΩ
Actuation bandwidth (at 1750 Hz):	approx. 100 Hz
Hold bandwidth (at 1750 Hz):	approx. 1 kHz
Tuning range:	1.3 to 2.3 kHz
Operating range by changing C 3:	1.3 to 50 kHz
Output voltage (1750 Hz, sinusoidal):	approx. 200 mV
Output impedance:	1.5 kΩ

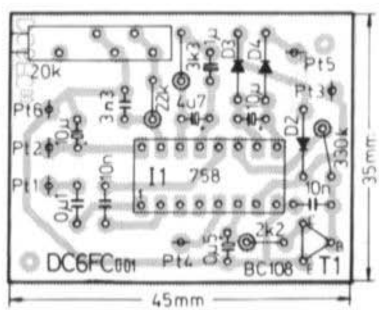


Fig. 3:
PC-board DC 6 FC 001

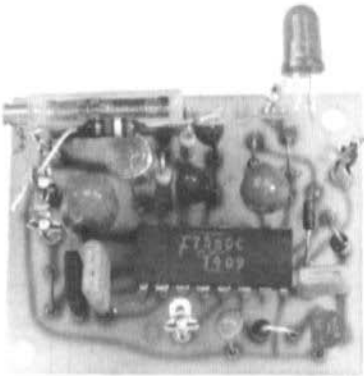


Fig. 4:
Photograph of the author's prototype

3. CONSTRUCTION

The circuit given in Figure 2 can be accommodated easily on a PC-board of 45 mm x 35 mm. The component locations and PC-board DC 6 FC 001 are shown in **Figure 3**. A photograph of the author's prototype is given in **Figure 4**. With the exception of the frequency determining RC-link, the construction is uncritical.

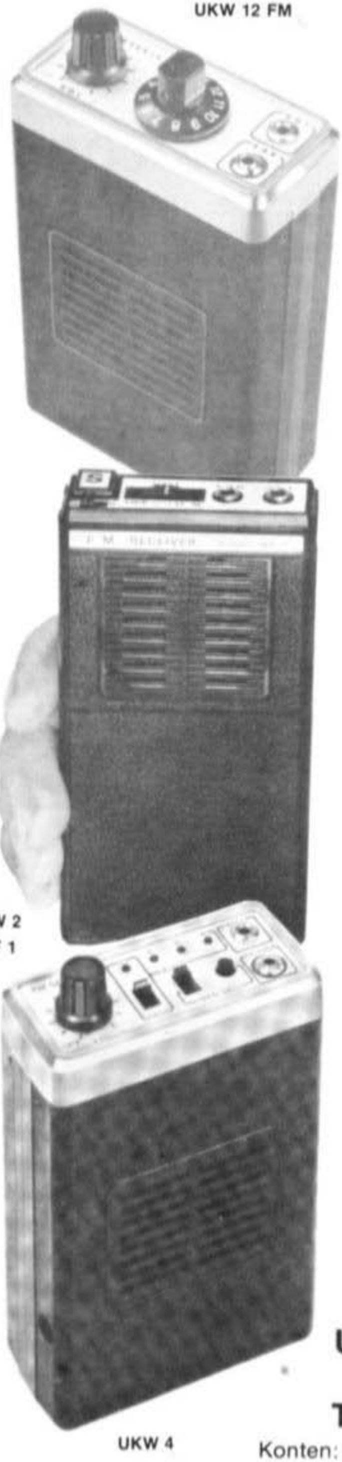
3.1. Components

I 1:	μA 758 (Fairchild)
T 1:	BC 108, BC 413, or other AF transistor
D 1:	Any LED
D 2 ... D 4:	1 N 4148 or other silicon diode
P 1:	20 kΩ, 10 turn helical-potentiometer (Amphenol)
C 3:	3.3 nF styroflex (Siemens)
C 1, C 2, C 5:	Ceramic capacitors
All other capacitors:	Tantalum drop-type electrolytics

4. REFERENCES

(1) J. Kestler: A stereo VHF/FM receiver with frequency synthesizer
VHF COMMUNICATIONS 7, Edition 2/1975, Pages 66 - 77.

UKW 12 AM
UKW 12 FM



UKW 2
UHF 1

UKW 4

MINIATURE VHF-UHF RECEIVERS and Scanners for Both Professional and Amateur Applications

We offer a whole series of miniature, handy VHF and UHF receivers which are suitable for a wide varity of applications. The receivers are so small that they fit in a shirt-pocket, and are hardly larger than a pack of cigarettes. According to type, the receivers are suitable for reception on the various VHF and UHF bands both for radio amateurs and professional users. The receivers are, for instance, suitable for use as paging receivers in existing radio networks, or as small monitoring receivers for transport companies, police, fire services, ambulance services etc. There is also a model for the airband range from 108 to 136 MHz, which is suitable for monitoring the local tower frequency, as emergency receiver, as receiver for the glider and ballon recovery groups, as well as mountain rescue services etc.

These sensitive double-superhet receivers are crystal-controlled, equipped with squelch and are supplied with rechargeable NC-cells and battery charger. Some types (UKW-2 and UHF-1) can be equipped for two-tone selective call. An antenna and earphone are provided, and all receivers are equipped with a built-in loudspeaker.

At present the following receivers are available. Further receivers are under development. Special models can be made to customer specifications if sufficient quantities are required.

MINIATURE VHF-UHF RECEIVERS AND SCANNERS

UKW 12 AM	**12 channel miniature airband receiver**
Modulation mode: AM	Frequency range: 108 - 136 MHz
Dimensions:	112 mm x 69 mm x 33 mm
Accessories:	Antenna, earphone, battery charger
UKW 12 FM	**12 channel miniature VHF-FM receiver**
Modulation mode: FM	Frequency range: 70-86 MHz, 140-170 MHz
Dimensions:	112 mm x 69 mm x 33 mm
Accessories:	Antenna, earphone, battery charger
UKW 2	**2 channel miniature VHF-FM receiver**
Modulation mode: FM	Frequency range: 70-86 MHz, 140-170 MHz
Dimensions:	120 mm x 60 mm x 22 mm
Accessories:	Antenna, earphone, battery charger
Features:	Possibility of installing two-tone selective call
UKW 4	**4 channel VHF-FM Scanner-receiver**
Modulation mode: FM	Frequency range: 70-86 MHz, 140-170 MHz
Dimensions:	112 mm x 69 mm x 32 mm
Accessories:	Antenna, earphone, battery charger
UHF 1	**1 channel UHF-FM miniature receiver**
Modulation mode: FM	Frequency range: 350 - 512 MHz
Dimensions:	120 mm x 60 mm x 22 mm
Accessories:	Antenna, earphone, battery charger
Features:	Also available as two-channel receiver. Possibility of installing two-tone selective call.

Sensitivity of the miniature receivers: 0.5 μV or 1 μV / 20 dB S/N

Distributor enquiries welcomed.

U K W - T E C H N I K · Hans Dohlus oHG
D-8523 BAIERSDORF · Jahnstraße 14
Telefon (09133) - 855,856 · Telex: 629.887
Konten: Postscheck Nürnberg 30455-858 · Commerzbank Erlangen 820-1154