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A Publication
for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 15 · Autumn · 3/1983 · DM 6.50

V-MOS FETs
for the 2 m band
can deliver
up to 100 W

Very
simple
biasing.
No thermal
run-away.
High linearity –
but 28 V
required!

VHF communications

A Publication for the Radio Amateur
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Volume No. 15 · Autumn · Edition 3/1983

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Contents

In the Focus

OSCAR 10 is functioning admirably! QSOs are being made between Europe and America, Africa, Australia, Japan and Asia! To assist you in joining the user family, here are the frequencies:

U-Transponder:
Input 435.025 – 435.175 MHz
Output 145.978 – 145.828 MHz
Engineering beacon: 145.987 MHz
General beacon 145.810 MHz

L-Transponder:
Input: 1269.050 – 1269.850 MHz
Output. 436.950 – 436.150 MHz
Engineering beacon 436.020 MHz
General beacon: 436.040 MHz

About antenna polarisation for OSCAR 10 please read pages 182/183 in this edition. A suitable transmit antenna for the L-transponder is to be found on pages 184 – 189

Early problems with the L-transponder (approx. 30 dB of sensitivity were missing!) seem to be solved, at least to some extent. According to DL1BU the sensitivity on 1269 MHz is near the projected value now, after switching the antenna relais many times by telecommand.

Now, how about you?

Best 73 es 55 de

Robert E. Lentz

DL3WR

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H. Braubach, DL 1 GBH

V-MOS Transistors in Power Amplifiers for 144 MHz

Several V-MOS power transistors have become available on the market in the last few years. This article is to describe the differences and advantages of these transistors with respect to conventional MOS and bipolar transistors, and is to give information regarding their applications. Three power amplifier stages are to be described that match one another. They allow one to carry out one's own experimentation and to gain experience using these components. On being connected together, a three-stage linear amplifier results that will provide an output of 80 to 100 W at 144 MHz with a drive of less than 50 mW. The article is completed by describing measuring aids and a harmonic filter which is able to handle this output power level.

1. V-MOS, A SHORT INTRODUCTION

In the case of conventional MOS-FETs, the current flows horizontally in the substrate, as shown in the diagram given in **Figure 1**. The resistance between source and drain is dependent on the potential at the gate. With a given spacing source – drain and a given gate surface, it can only be reduced to a minimum $R_{DS\ on}$. If the MOS-FET is to produce power, it is necessary for $R_{DS\ on}$ to only amount to a few Ω, otherwise, the dissipation power of the transistor will be too high. Of course, there are conventional MOS-FETs having a low $R_{DS\ on}$, however, they require a large gate surface and therefore exhibit a large gate/source capacitance. For this reason, such transistors are only suitable for low-frequency applications.

In the RF-range, conventional MOS-FETs can produce powers of up to 1 W. Higher power levels – up to approximately 120 W at frequencies of up to approx. 300 MHz – have only become possible using MOS technology since the development of the V-MOS-FETs. The "V" stands for their vertical structure, the current flows vertically in the substrate, as shown in **Figure 2**.

Due to the V-structure, the channel length is only

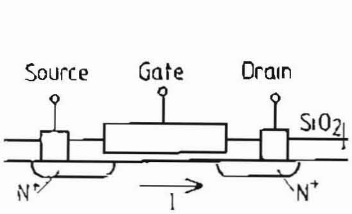

Fig. 1:
Cross section through a conventional MOS-FET (SILICONIX data book V-MOS Power FETs, 1980)

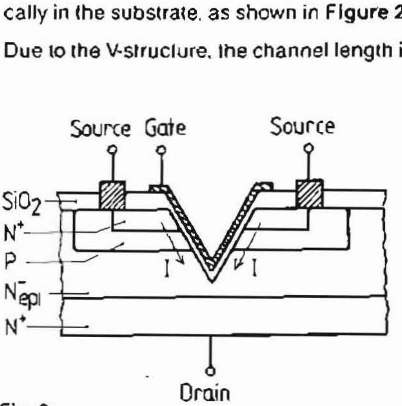

Fig. 2:
Cross section through a V-MOS FET (SILICONIX)

approximately 1/3 that of conventional MOS structures, which reduces the ON-resistance to approximately one third. Furthermore, two current channels are formed, which allows the ON-resistance to be halved again. The other drain structure also allows the ON-resistance to be reduced so that considerably higher currents can flow than in the case of a conventional MOS-FET for the same input capacitance.

Further very important advantages of V-MOS transistors should also be mentioned briefly:

- They exhibit a low feedback capacitance
- The intrinsic feedback capacitance is virtually static in contrast to bipolar transistors, which leads to a very much lower tendency to parametric oscillation.
- V-MOS transistors have a negative temperature coefficient, which means that the drain current is reduced with increasing temperature. This ensures that the transistor cannot destroy itself.
- V-MOS transistors exhibit an extremely linear characteristic. The required quiescent current is simply aligned by connecting a DC-voltage, and a temperature compensation is not required.

2. BASIC CALCULATIONS IN THE CASE OF V-MOS POWER AMPLIFIERS

The maximum obtainable power that can be achieved with a transistor can be approximated with the aid of equation 1.

$$P = \frac{(U_{DD} - U_{DS\,on})^2}{2 \times R_L} \quad (1)$$

whereby

U_{DD} = Operating voltage

$U_{DS\,on}$ = Drain/source residual voltage at full drive

R_L = Load resistance at the transistor

The residual voltage $U_{DS\,on}$ can be calculated with the aid of equation 2.

$$U_{DS\,on} = I_{max} \times R_{DS\,on} \quad (2)$$

One is now able to see the effect of the transistor impedance $R_{DS\,on}$ which was discussed in section 1. A higher resistance allows the residual voltage $U_{DS\,on}$ to be high and also only allows a lower output voltage. The maximum output power P_{out} is mostly given in the data sheets. The required load resistance R_L at the transistor can then be calculated according to equation 3.

$$R_L = \frac{(U_{DD} - U_{DS\,on})^2}{2 \times P_{out}} \quad (3)$$

Finally, a fourth equation is to be given which allows the possible gain (in dB) of a transistor to be calculated:

$$G_p = 10 \log \left[\frac{g_m^2 \times R_s}{R_L\,(1/R_{out} + 1/R_L)^2} \right] \quad (4a)$$

After two transformation steps one will obtain

$$G_p = 10 \log \left[\frac{g_m^2 \times R_s}{\frac{R_L + 2R_{out}}{R_{out}^2} + \frac{1}{R_L}} \right] \quad (4b)$$

where:

g_m = Forward slope

R_s = Source impedance of the input

R_L = Load resistance at the drain

R_{out} = Output impedance

G_p = Power gain in dB

As is shown by equation 4b, the power gain is very strongly affected by $1/R_L$ in the nominator, which means that when R_L increases, this will also cause the gain to increase. However, since R_L is also present in the nominator of equation 1, the maximum output power will drop at the same time. This means that the maximum values of gain and output power are not achievable at the same time

2.1. V-MOS Transistors Used

SILICONIX types DV 2805, DV 2810, and DV 2880 have been used. They belong to a series of N-channel enhancement FETs that comprise six types: DV 2805, DV 2810, DV 2820, DV 2840, DV 2880, DV 28120. The number "28" in the designation shows a nominal operating

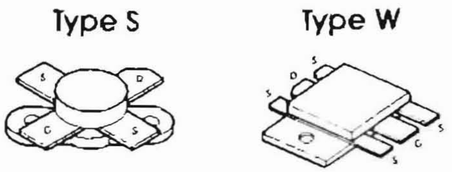

Fig. 3:
Types of cases of the V-MOS FETs used (Siliconix)

voltage of 28 V, whereas the last two or three digits refer to the maximum output power. The designation "DV" designates types for the VHF frequency range in the order of 175 MHz.

All these transistors that exhibit a low noise figure are able to handle mismatch conditions, do not run away thermically, and can be designed to operate with simple bias circuits for class A, B, or C with a high dynamic range. Their nominal power gain when connected in a common source circuit amounts to 10 dB. They are available in cases type "W" and "S" (see **Figure 3**): the higher-powered transistors are available in cases "W", "U" (as "S", but larger), and "T". The cheapest and thus most interesting types for radio amateurs are the transistors mounted in a "W" case.

A few important specifications of the three transistor types used are given in **Table 1**.

3. DESIGN OF THREE AMPLIFIER STAGES

The requirement was to design a power amplifier with an output power of approximately 100 W and with a drive power of a maximum of 100 mW. Of course, a gain of 30 dB cannot be obtained in a single transistor stage. V-MOS FETs have a typical gain of 10 dB per stage, which means that three amplifier stages are required. All three stages are designed so that they possess an input and output impedance of 50 Ω. This simplifies alignment and allows each stage to be used individually. Special features of the circuit are now to be described briefly, beginning with the output stage. The overall circuit diagram is given in **Figure 4**.

3.1. The 100 W Stage

The transistor type DV 2880 is used in this stage; this is able to provide approximately 100 W PEP at 145 MHz with a gain of 10 dB at an operating

Characteristics	DV 2805	DV 2810	DV 2880
Absolute limit values at 25°C			
Gate-source voltage	40 V	40 V	40 V
Drain-source voltage	80 V	80 V	80 V
Drain-gate voltage	80 V	80 V	80 V
Drain current	0.5 A	1 A	8 A
Dissipation power at 25°C case temp	10 W	20 W	160 W
Operating values at U_{DS} = 28 V, f = 175 MHz			
Output powr P_{out}	5 W min	10 W min.	80 W min.
Drain efficiency	60% typ	60% typ.	65% typ.
Slope g_m at 0.5 $I_{D\,max}$	65 mS typ.	130 mS typ.	800 mS min.
Input capacitance at U_{GS} = 0 V	12 pF typ	22 pF typ.	210 pF max.
Output capacitance C_{oss} at U_{GS} = 0 V	11 pF typ	21 pF typ.	175 pF max.
Feedback capacitance C_{rss} at U_{GS} = 0 V	1.5 pF typ.	3 pF typ.	25 pF max.
Noise figure at 0.05 $I_{D\,max}$	6.8 dB typ	6.8 dB typ	4.0 dB typ
Transistor impedance $R_{DS\,on}$	1 Ω	1 Ω	0.5 Ω

Table 1: Several important specifications of the V-MOS transistors used

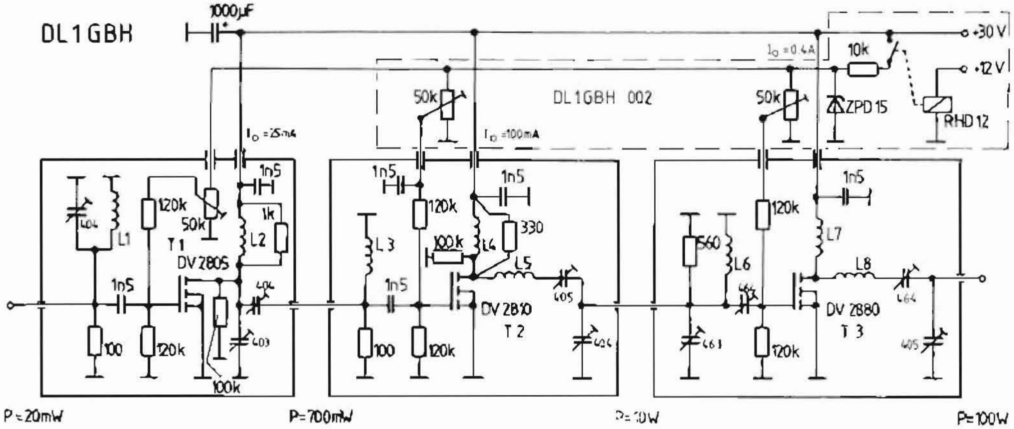

Fig. 4: Overall diagram of the three amplifier stages for 144 MHz

voltage of 30 V In order to calculate the required load resistance R_L, the following specifications are required

$R_{DS\,on}$ = 0.5 Ω
I_{max} = 5 A
U_{DD} = 30 V
P_{out} = 100 W

This can be calculated according to equation 2 as:

$$U_{DS\,on} = 0.5\,\Omega \times 5\,A = 2.5\,V$$

and according to equation 3:

$$R_L = \frac{(30\,V - 2.5\,V)^2}{2 \times 100\,W} = 3.28\,\Omega$$

This means that the output impedance must be transformed to 3.28 Ω as load impedance for the transistor, which is carried out with the aid of a circuit as shown in Figure 5. Inductance L_p is designed so that it forms a parallel resonant circuit together with capacitance C_{DS} of the transistor so that it is able to neutralize this capacitance. The load of 50 Ω present on the output socket is transformed with the aid of L_s, C_s, and C_p to an impedance of 3.28 Ω at the transistor.

The input matching is made with a similar network (Figure 6). In order to avoid parasitic oscillations, a resistance of between 100 and 680 Ω is connected in parallel with the input. In the case of this amplifier stage, a value of 560 Ω has been found suitable.

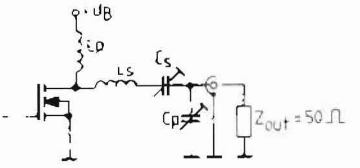

Fig. 5: Output transformation of the 100 W stage

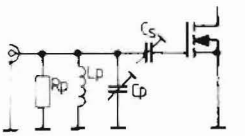

Fig. 6: Input matching of the 100 W stage

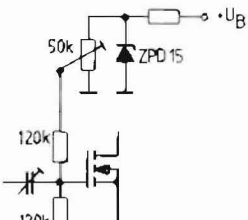

Fig. 7: Operating point adjustment of V-MOS FETs

The quiescent current is determined in a simple voltage divider which can be high-impedance – as in **Figure 7** –, since a very high input impedance exists for DC-voltage For this reason, there is a danger of the transistors being destroyed by static charge, as is the case with all MOS-semiconductors. One should therefore firstly solder in the resistor to ground before mounting the transistor.

3.2. The 10 W Stage

The driver stage must provide 10 W into 50 Ω A type DV 2810 is used whose load impedance can be calculated as follows.

$$U_{DS\,on} = 0.5\,\Omega \times 1\,A = 0.5\,V$$

$$R_L = \frac{(29.5\ V)^2}{2 \times 10\ W} = 43.5\,\Omega$$

The input and output matching are realized with similar networks to that of the 100 W stage The input matching is somewhat simpler as can be seen in the circuit given in Figure 4.

3.3 The 1 W Stage

Since the 10 W stage provided a gain of approximately 12 dB instead of the assumed 10 dB, the first stage equipped with the V-MOS type DV 2805 only has to provide an output power of approximately 700 mW For this first stage the following specifications can be calculated according to the previously mentioned equations:

$$U_{DS\,on} = 1\,\Omega \times 0.3\,A = 0.3\,V$$

This voltage drop can be neglected

$$R_L = \frac{(30\ V)^2}{2 \times 0.7\ W} = 642\,\Omega$$

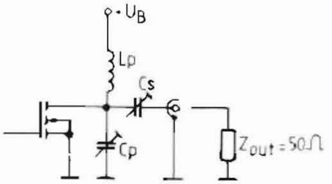

Fig 8:
Output matching of the high-impedance 1 W stage

This means that the transistor input impedance of 50 Ω must be transformed to 642 Ω at the transistor with the aid of a matching network This is achieved with the circuit given in **Figure 8**. The input matching is made with a similar network to that used for the order two higher-powered stages.

3.4. Special Components

- L 1: 2 turns of 1 mm dia. silver-plated copper wire wound on a 5 mm former, pulled out to a coil length of 15 mm, self-supporting.
- L 2: 1 turn of 2 mm dia. silver-plated copper wire wound on a 5 mm former, pulled out to a coil length of 5 mm, self-supporting.
- L 3: 1.3 turns of 1 mm dia. silver-plated copper wire, wound on a 5 mm former.
- L 4: 9 turns of 1 mm dia silver-plated copper wire wound on a 5 mm former, pulled to a coil length of 15 mm, self-supporting
- L 5: 4 turns of 1 mm dia silver-plated copper wire, 5 mm former, 5 mm long, self-supporting
- L 6: 2 turns of 2 mm dia. silver-plated copper wire, 5 mm former, 15 mm long, self-supporting
- L 7 3 turns of 1 mm dia. silver-plated copper wire, 5 mm former, 7 mm long, self-supporting
- L 8: 0.5 turns of 2 mm dia silver-plated copper wire. 15 mm former, self-supporting

9 pcs. mica trimmer capacitors as shown in Figure 4.
The coupling capacitors of 1.5 nF in the input and driver stage are mica types.

Ceramic multi-layer capacitors of approximately 1 nF should be used for bypassing the drain voltage; a 10-turn helical trimmer potentiometer should be used for adjusting the gate bias voltage. The feedthrough capacitors should be for screw fitting, and have capacitance values of between 1 and 2 nF.

4. CONSTRUCTION OF THE AMPLIFIERS

A universal PC-board is provided for the various amplifier stages (**Figure 9**). It is 105 mm x 54 mm and is constructed from 1.5 mm thick epoxy glassfibre board

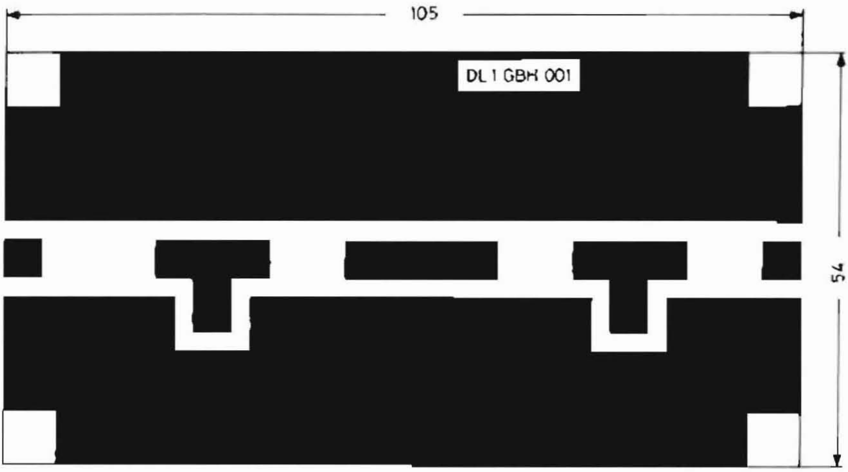

Fig. 9: Single-coated universal board DL 1 GBH 001

Each board is provided with a cutout for the transistor so that the transistor itself can also be directly screwed to the base plate of the case. The components are directly soldered to the surface of the PC-board.

Each amplifier stage is installed in a cast aluminium case. The cases are prepared as shown in Figure 10, whereby the two M3 threaded holes in the narrow side panel of the case are only required in the case of the 100 W stage, and only on the output side, since an N-connector is to be used here that has a conventional BNC-flange. The central hole at the bottom of the case is used to mount the transistor; this position is marked after the PC-board has been placed into the case. The other four holes in the base plate are used for

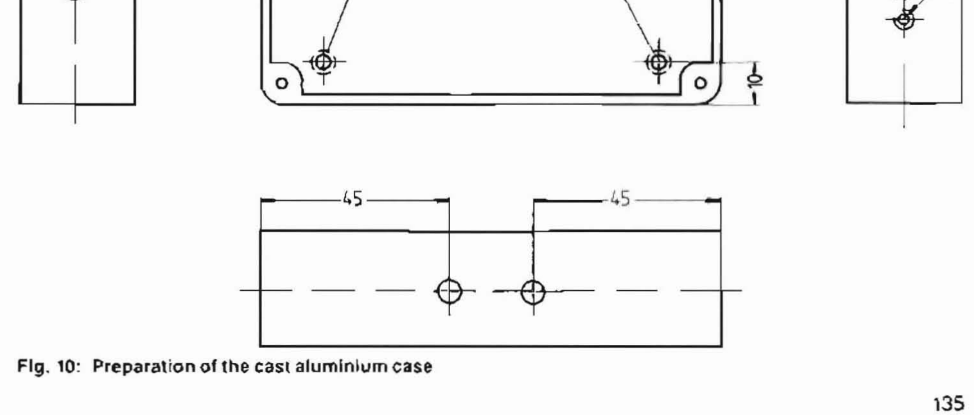

Fig. 10: Preparation of the cast aluminium case

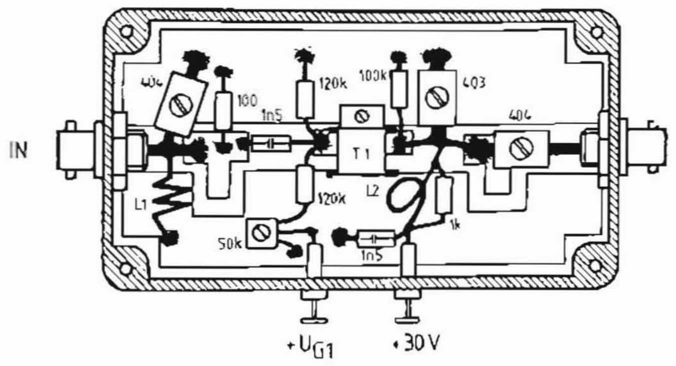

Fig. 11: Construction of the 1 W amplifier

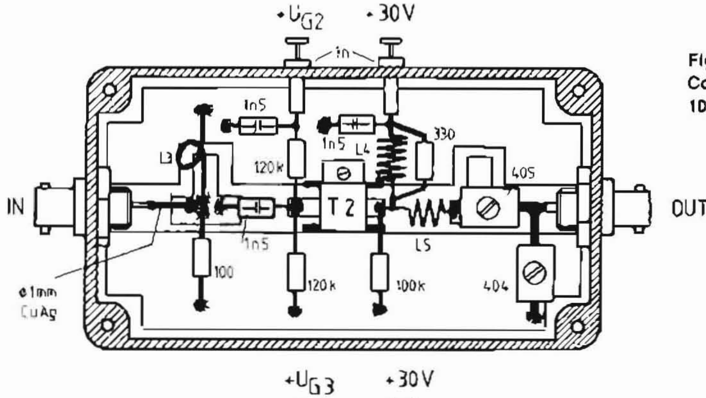

Fig. 12: Construction of the 10 W amplifier

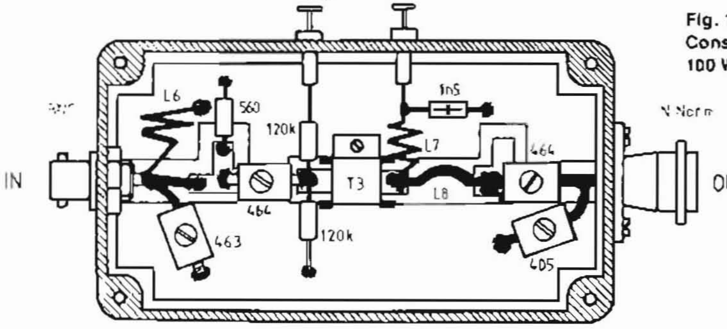

Fig. 13: Construction of the 100 W amplifier

mounting the heat sinks, which are only used for the two higher-powered stages; they are countersunk for M 2.5 screws. These are not required for the 1 W stage. The holes in the long-side panels are for two feedthrough capacitors, each, and those on the narrow-side panels are for BNC connectors for single-hole mounting.

It is now possible for the amplifiers to be constructed according to **Figure 11** (1 W stage), **Figure 12** (10 W stage), or **Figure 13** (100 W stage).

The construction is commenced with the resistors, after which the transistor is soldered into place before the inductances and capacitors. A photograph of a completed 1 W stage is given in **Figure 14**, and the two "higher

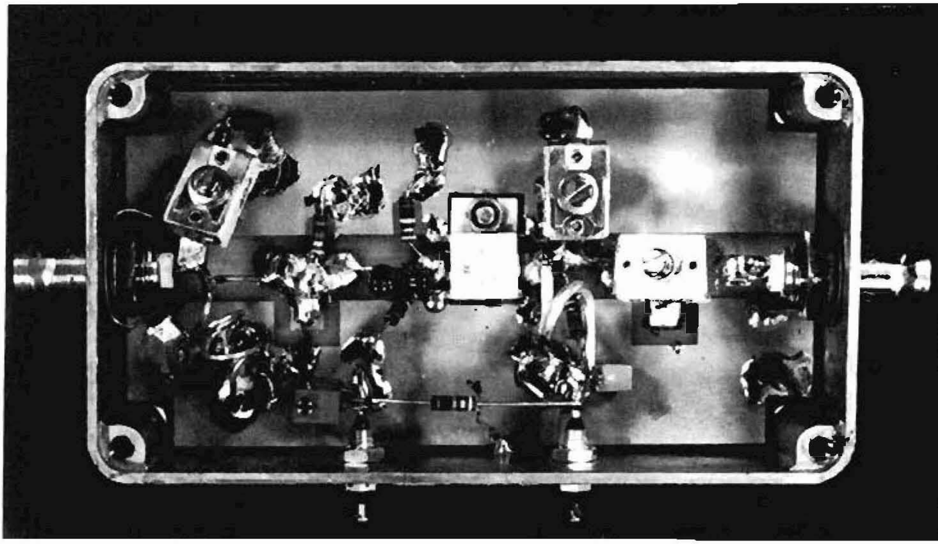

Fig. 14: Photograph of the author's prototype 1 W amplifier

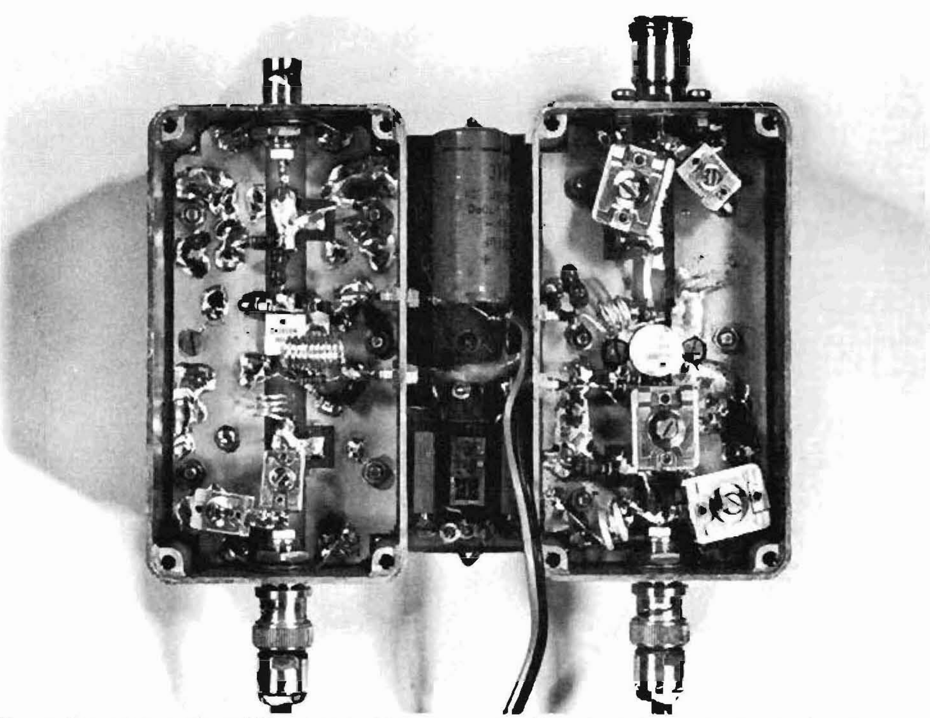

Fig. 15: The 10 W and the 100 W power amplifiers are mounted together with the bias voltage board DL 1 GBH 002 onto a common heat sink

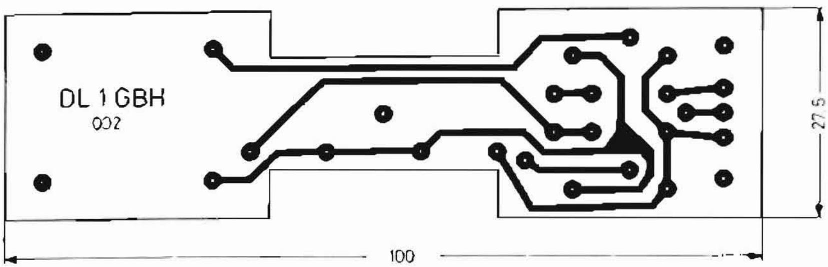

Fig. 16: The single-coated bias board DL 1 GBH 002

powered" stages, together with a small board to provide the bias voltages, are shown in Figure 15. The 100 mm x 30 mm board DL 1 GBH 002 (Figure 16) accommodates the two spindle trimmers, the relay, the zener diode, the 1 mF electrolytic, and a few solder points. Figure 17 shows the component location plan of this PC-board, together with the overall construction and intermediate connections. Finally, it should be noted that the two lower-power amplifiers are provided with covers, whereas this is not the case with the 100 W amplifier; in the case of the latter, the

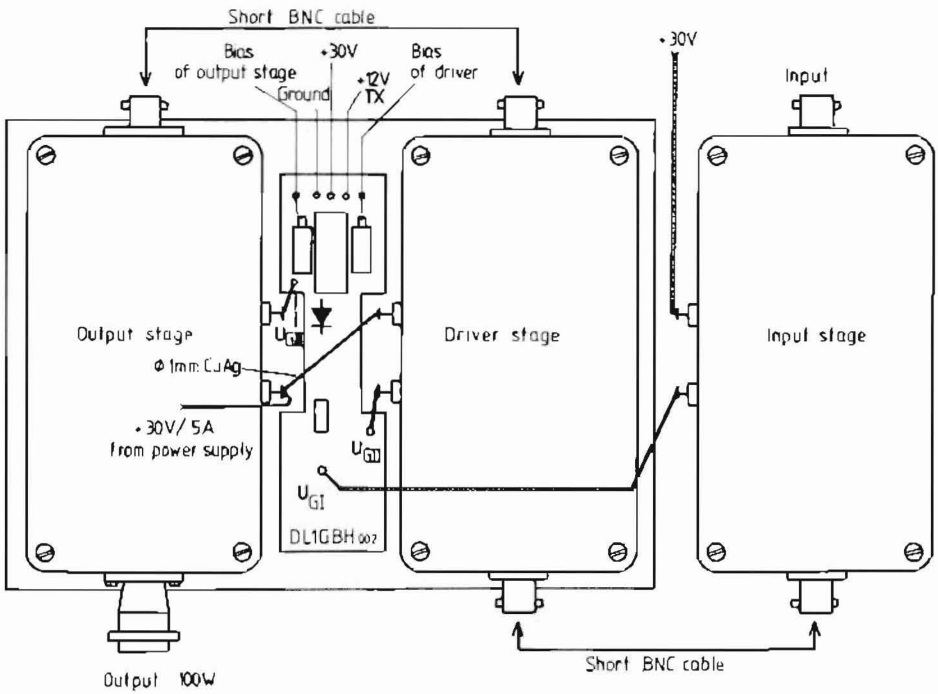

Fig. 17: Mounting and wiring of all three amplifiers to form a common linear amplifier with 37 dB gain and 100 W output power.

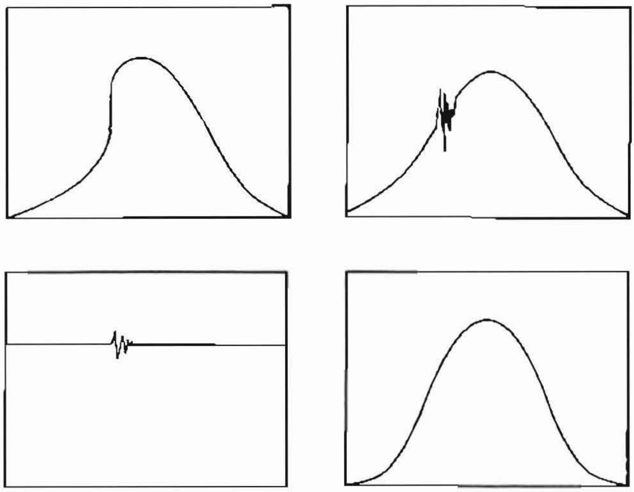

Fig. 18:
The following swept-frequency responses can be seen:
Upper left: Tendency to oscillation due to extra coupling;
Upper right: Slight oscillation of the amplifier;
Lower left: Constant strong oscillation;
Lower right: Correct swept-frequency curve

losses due to induced RF-currents in the cover are so high that the full output power could no longer be achieved.

5. ALIGNMENT OF THE AMPLIFIERS

Any amateur who has constructed a power amplifier with an expensive transistor and is about to align it, knows the danger of a sudden self-destruction of the transistor. The actual danger are parasitic oscillations which cause a steep increase of current. The resulting exeeding of the limit values can be avoided by not providing a continuous signal at one frequency, but driving the stage with a swept-frequency signal over a wider frequency range. If frequency deviation and speed are large enough, the transistor will only be in critical operating conditions for a very short period, which usually does not endanger the transistor. Tendency to oscillation and other instabilities can easily be seen on the swept-frequency curve. **Figure 18** shows a few examples of this. It is also favorable to monitor the input matching by placing a standing-wave meter (reflectometer) between generator and the amplifier stage to be aligned.

5.1. Adjustment of the Quiescent Currents.

For alignment of the quiescent current, each amplifier stage is terminated with 50 Ω at the input and output. Furthermore, it should be ensured that the potentiometers are placed in their fully ground position before connecting the gate bias voltages. This is followed by connecting the voltages of +30 V (max. 5 A) and +12 V for the relay; the following quiescent currents should be adjusted

1 W stage: I_o = 25 mA
10 W stage: I_o = 100 mA
100 W stage: I_o = 400 mA

The current should increase continuously on increasing the gate voltage; any sudden increase indicates unwanted oscillations. In this case, reduce the gate voltage and alter the alignment of the capacitors somewhat. After this, realign the quiescent current.

5.2. VHF Alignment

This alignment requires measuring equipment that should be connected as shown in **Figure 19**. Firstly only connect the 1 W stage and align it for maximum output power. A smooth, and continuous swept-frequency curve should be obtained, similar to that shown in the lower part of

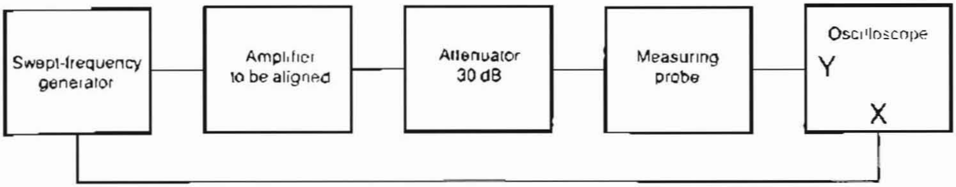

Fig. 19: Measuring equipment for alignment: the attenuator must be able to handle the full output power

Figure 18. If the first stage is operating correctly, the 10 W stage is connected to it and aligned in a similar manner to that of the 1 W stage (to which no further alignment should be made) After completing the alignment of the 2nd stage, the 100 W stage is connected and aligned in a similar manner. After this, it is possible for corrections to be made to all alignment capacitors in order to obtain the maximum output power together with the most favorable frequency response. The author's prototype obtained an output power of 100 W with a drive power of 20 mW; an output power of 80 W should always be achievable

The measured values of the prototype are given in **Table 2**.

In one of the next editions of VHF COMMUNICATIONS, we will bring the concluding parts.

6 Home-made measuring aids
7 Harmonic filter

For those requiring more information on V-MOS transistors, we would like to point out that SILICONIX publish a booklet entitled "V-MOS Power-FET Design Catalogue".

P_{in}/dBm	P_{in}/mW	P_I/dBm	P_I/mW	P_{II}/dBm	P_{II}/W	P_{III}/dBm	P_{III}/W
5	3.2	19 4	87	35 0	3.16	42 5	17.8
6	4	20.2	105	36.0	3 98	43 3	21.4
7	5	21 2	132	36.8	4.79	44 3	26.9
8	6.3	22.1	162	37.6	5 75	45 5	35 5
9	7 9	23.1	204	38 3	6.76	46.8	47.9
10	10 0	23.9	245	38 8	7.59	48.1	64 6
11	12.6	24.8	302	39 3	8 51	48.8	75.9
12	15.9	25 8	380	39.7	9 33	49 3	85.1
13	20 0	26.8	479	40.0	10	50.0	100

Table 2: Measured values of the 1 W stage (I), of 1 W and 10 W stage (II), as well as of all three stages connected together (III)

István Szabó, op. of HA 5 KFV
Sándor Nagy, HA 5 GH

Input Filters For Receive Applications In The 144 MHz Range

Most active VHF amateurs will have noticed that many non-amateur signals are audible in the 2 m band. These can be caused by many things, and can also be generated in the receiver itself. For example, two or more legal out-of-band signals can generate in-band signals within the amateur band due to intermodulation in the input stage of the receiver or preamplifier. These can be mixed with the "wanted" signals and be demodulated with them.

A further problem can be caused by individual out-of-band signals from very strong transmitters that shift the operating point of the first stage and thus reduce the sensitivity of the receiver (desensitization). If the strong, out-of-band signal disappears, the (weak) amateur signal will be audible again. Such surprising variations are not caused by tropospheric propagation, but by the insufficient large-signal handling capability of our receiver.

Such interference can be reduced, or even completely suppressed by using a selective filter at the input of the receiver that only allows the required frequency range to pass.

A good filter will also reduce the sum power of ignition interference and other man-made noise (also a form of contamination!). Of course, the remaining interference in the band will still be bad enough, and one could assume that this problem could be solved by simply increasing the transmit power in order to ease reception. This may bring a temporary improvement, however, if all other transmitters operate with higher power, the same old situation will prevail. For this reason, attempts are being made throughout the world to improve the large-signal capabilities of receivers. The described bandpass filter represents one of these possibilities.

1. BANDPASS FILTER

As is probably known, the losses of a resonant circuit are dependent on the Q of the inductance L and the capacitance C. If the Q of a circuit is high, this will mean that the resonant curve is sharp. This curve can be made sharper – or its slope steeper – when two resonant circuits are coupled together. For quantitative calculations, which can be made with computers nowadays, one requires values of Q and degree of coupling. Calculations need not be made individually since the results can be taken from well-known tables and diagrams (1) and (2).

For the designing process, one requires the required frequency response with attenuation values at certain frequencies to form the basis of the calculation (**Figure 1**). The insertion loss (loss in the passband range) a_0 can fluctuate between a minimum and maximum value within the passband range (number of maximums corresponds to the number of coupled circuits). The corner frequencies of the passband range are to be designated f_{p1} and f_{p2}. The attenuation a_{c1} will appear at frequency f_{c1}, and the attenuation should obtain the value a_{c2} at f_{c2}.

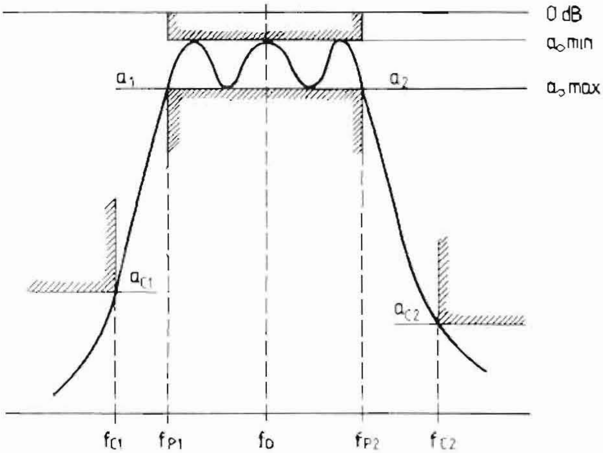

Fig. 1: Typical frequency response of a bandpass filter together with its specifications

The slope steepness of the filter is dependent on the number of resonant circuits and on the Q-values. The shape of the curve, that is its bandwidth, as well as the value of the insertion loss can be varied with the aid of the degree of coupling. If the coupling is fixed, the passband curve will become wider and the fluctuations of the attenuation (ripple) within the passband range will become greater. In the case of so-called critical coupling, the relationships are more favorable, but the bandwidth is lower. A loose coupling, finally, results in the lowest bandwidth, but also the highest insertion loss.

In other words, each filter design requires compromises and the task is to obtain a required bandwidth with a minimum of insertion loss, and at the same time to obtain the required attenuation values for out-of-band signals (ultimate attenuation).

2. 4-STAGE FILTER FOR 144 MHz

A filter is now to be described for home-construction whose specifications are given in **Table 1**:

Center frequency:	f_o = 145 MHz
Insertion loss:	a_o = 1.6 – 1.8 dB
Lower corner frequency	f_{p1} = 144 MHz
Attenuation at f_{p1}	a_1 = 2.0 dB
Upper corner frequency	f_{p2} = 146 MHz
Attenuation at f_{p2}	a_2 = 2.0 dB
Lower cut-off frequency	f_{c1} = 140 MHz
Attenuation at f_{c1}	a_{c1} = 36 – 40 dB
Upper cut-off frequency	f_{c2} = 150 MHz
Attenuation at f_{c2}	a_{c2} = 36 – 38 dB

Table 1: Filter specifications

The filter comprises capacitively coupled LC-circuits. In the case of 144 MHz, high-Q, air-spaced coils with compact dimensions can be used. When using silver-plated wire and not too many turns, one can obtain Q-values of over 100. This is, of course, no comparison to the Q of coaxial resonators (in the order of Q = 1000), but the manufacture is considerably simpler. Inductances for 144 MHz can be made from 1 mm dia. wire with a coil diameter of 6 to 8 mm.

Air-spaced or PTFE trimmers of 8 to 15 pF can be used for tuning, however, there are some other solutions. The construction of special coupling capacitors is also to be discussed in this article, since they contribute to the tuning elements.

Input and output are inductively coupled since they allow simple transformation from 50 Ω to the resonance impedance of the circuits which are usually in the order of kΩ. In order to achieve the required truely capacitive coupling, it is necessary for the individual inductances to be accommodated in screening chambers.

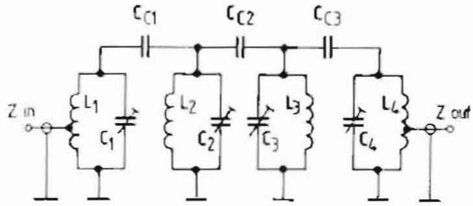

Fig. 2:
Equivalent circuit diagram of a 4-stage filter with capacitive coupling

3. CONSTRUCTION

The equivalent diagram of the 4-stage filter is shown in **Figure 2**. The wire length of the inductive taps amounts to 12 to 15 mm; they are connected to the coil at 0.25 to 0.75 turns from the cold end. Silver-plated copper wire of 1 mm diameter should be used, and this wound around a 6 mm former.

Special attention and care was paid to the construction of the coupling capacitors C_{c1} to C_{c3}. The idea was that two neighbouring conductive surfaces on a PC-board provided well-reproducible capacitance values in the required order of 0.2 to 1 pF. **Figure 3a** shows this arrangement and its equivalent circuit diagram. **Figure 3b** allows one to carry out one's own designs

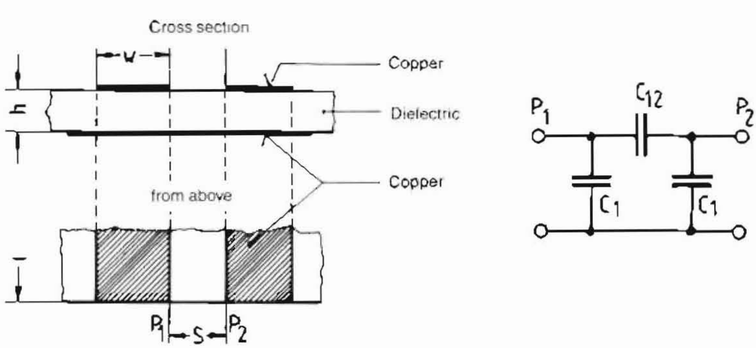

Fig. 3a: Schematic arrangement and equivalent circuit diagram of a coupling capacitor in stripline technology.

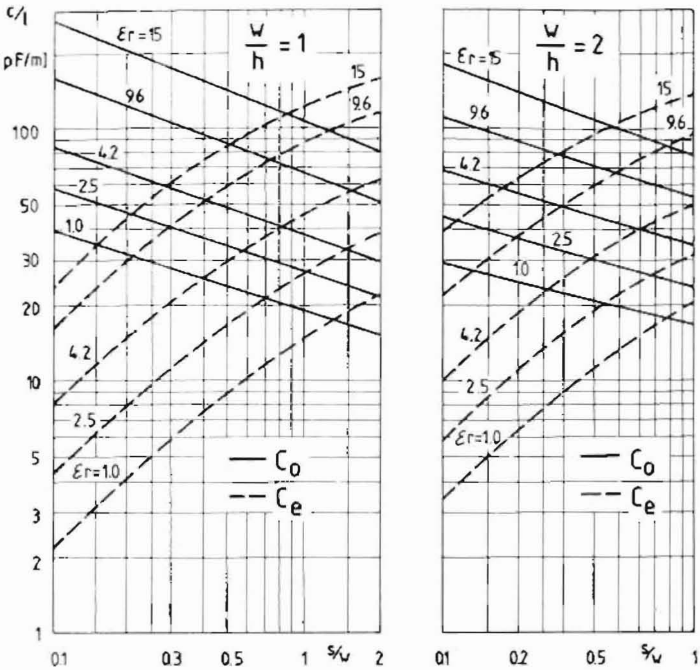

Fig. 3b: Design diagrams for stripline coupling capacitors

Our filter, comprising concentrated inductances and trimmers, is complemented by distributed capacitances using this easy-to-use and wide-spread stripline technology. These capacitances are realized on a double-coated PTFE-glass fibre PC-board with the dimensions 100 mm x 50 mm (**Figure 4**). The arrangement is symmetrical.

The inductances are constructed in a coaxial manner in chambers made from PC-board material as shown in **Figure 5**. All solder joints – also those of the two coaxial connectors (BNC) – must be on the **inside** so that completely RF-tight chambers without gaps result having a good electrical contact to the input and output connectors. It is only the coupling capacitors that conduct the RF-current from chamber to chamber. Tronser air-spaced trimmers are used that have four connections for which holes have been provided on the board.

3.1. Required Material

PC-board HA 5 KFV 001: 100 x 50 mm, constructed from double-coated glasfibre PTFE material (ε_r = 2.2), 1.5 mm thick (3M · 250 GX 15)

The following eight pieces are made from single-coated, normal PC-board material of 1.5 mm thickness:

3 pcs. intermediate panels, 47 x 25 mm
2 pcs side pieces, 97 x 25 mm
2 pcs panels with BNC connector, 50 x 25 mm
1 pc cover 100 x 50 mm

Furthermore one will require:
2 BNC connectors (Radiall R-141554)
4 air-spaced trimmers
(Tronser 10-1111-20014-000)
4 air-spaced coils: 7 turns of 1 mm dia. silver-plated copper wire wound on a 6 mm former, self-supporting, connected between ground and the Tronser trimmer.

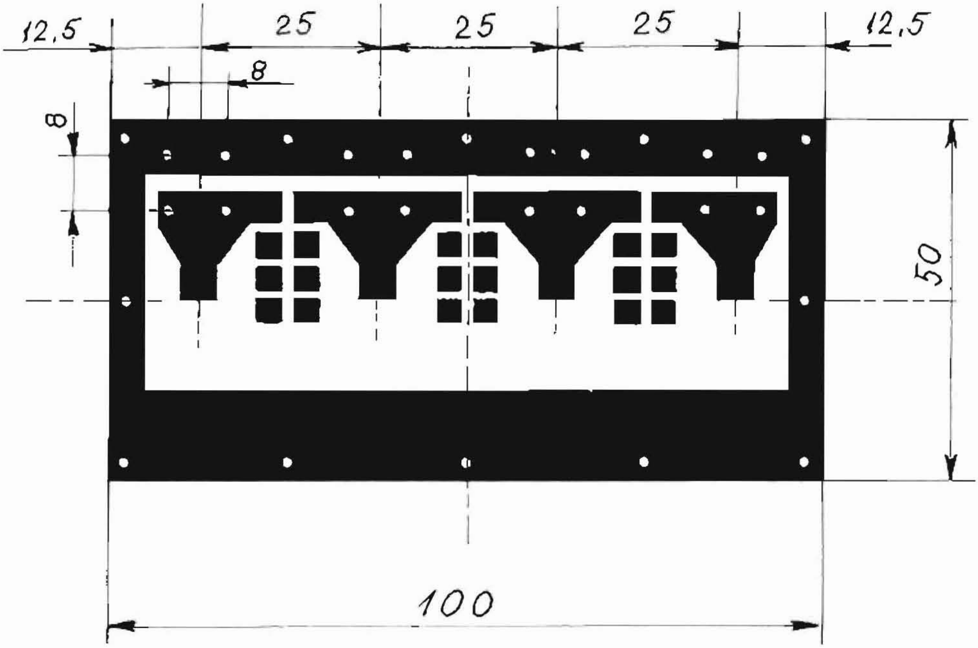

Fig. 4: PTFE PC-board HA 5 KFV 001 for a 4-stage bandpass filter for 144 MHz

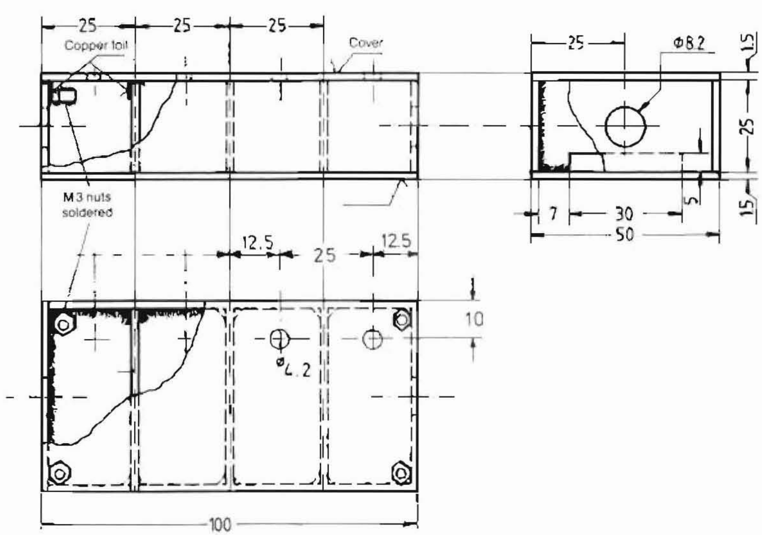

Fig. 5: Outside and intermediate panels, as well as the cover for the 4-stage filter
These can be made from single-coated PC-board material.

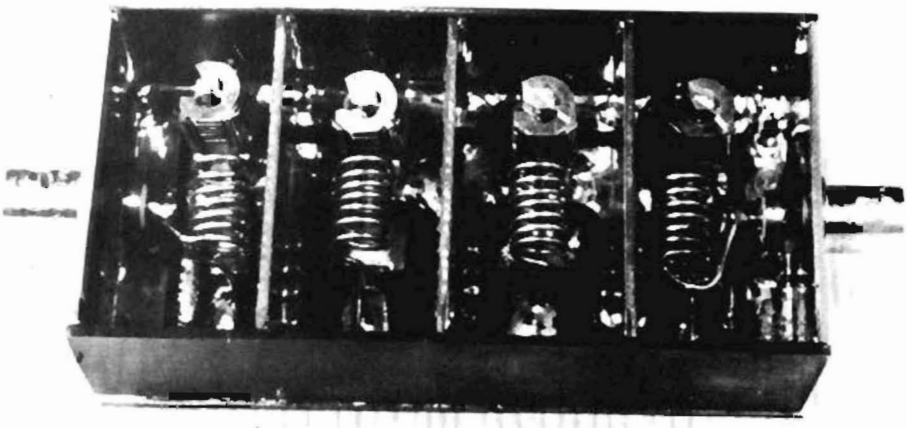

Fig. 6: Photograph of prototype filter, but without cover

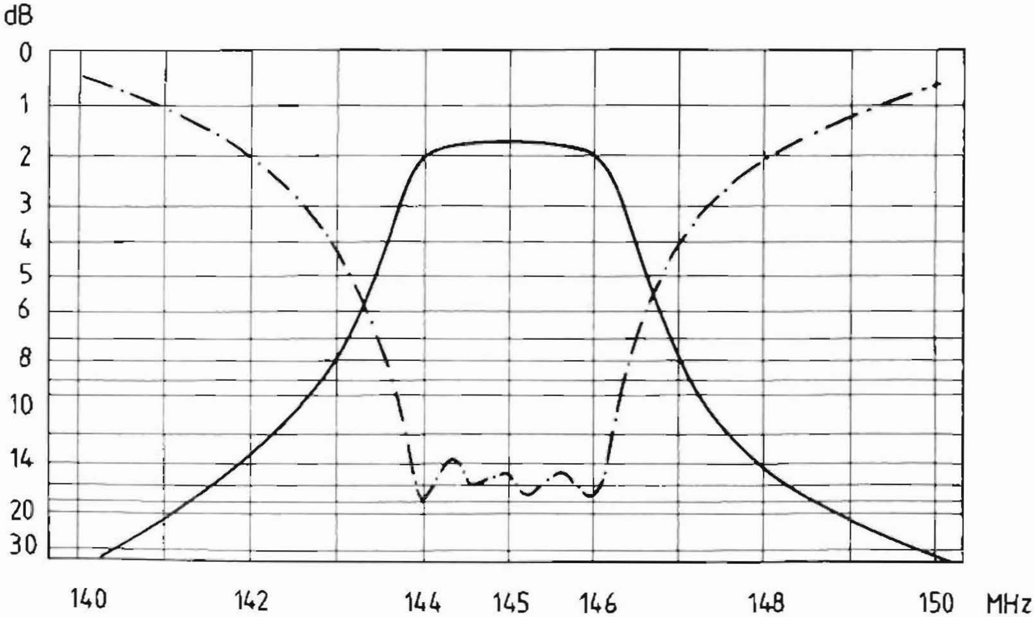

Fig. 7: Frequency response of attenuation (continuous line) and return loss of the input (dashed-dotted line)

Thin copper foil (approx. 0.1 mm thick) to solder the cover to the case and as metallic contact between cover and upper edge of the intermediate panels. A completed filter, but without cover, is given in **Figure 6**.

4. ALIGNMENT

Although one can exactly calculate all values, the manufacturing tolerances of the inductances, and the interaction of resonant circuit and coupling capacitances requires some form of alignment. For this reason, trimmers have been provided as well as several "capacitance islands" for the coupling. The less of these "capacitance islands" that are located opposite to each other, the looser will be the coupling. The optimum alignment of each 4-stage filter is a rather complicated process for which one requires a swept-frequency system or an RF network analyzer, and lots of patience.

The first step is to align all circuits to the center frequency of the band (145.0 MHz). The coupling is then improved in steps by soldering several islands together, until the required bandwidth is achieved. The input and output coupling should, on the other hand, be as loose as possible, what is achieved by connecting the tapping points as near to the cold end as possible. **Figure 7** shows the frequency response of amplitude and input matching after alignment.

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Gerd Otto, DC 6 HL

A Variable Crystal Oscillator (VXO) with a Pulling Range of Approximately 200 kHz at 144 MHz

A variable crystal oscilator (VXO) is to be described that has been especially designed for use in conjunction with the mini-SSB transceiver for 144 MHz described by the same author in (1). This oscillator provides a very clean signal with a level of approximately 7 dBm. This can be tuned from 135.15 to 135.35 MHz, which corresponds to an operating frequency range of 144.15 to 144.35 MHz for the transceiver, in other words for the SSB-range. Details are to be given regarding calculation of the crystal frequency, which means that this oscillator can also be designed for other frequencies. The dimensions of the screened module are only 74 mm × 37 mm × 30 mm. It will be seen that its length corresponds to the width of the transceiver, which means that the oscillator can be located adjacent to the crystal filter of the transceiver.

1. CIRCUIT

Variable crystal oscillators are preferably used when a relatively narrow frequency range is to be covered continuously – in contrast to channel switching with FM transceivers. The frequency stability corresponds to a value between a conventional crystal oscillator and that of a good VFO (LC-oscillator with variable capacitor or diode tuning). Such variable crystal oscillators (VXOs) have been described several times in VHF COMMUNICATIONS – the last one was a version with eight crystals, whose frequency ranges overlapped (2).

In order to ensure a sufficiently wide pulling range, a fundamental crystal is used at one sixth of the output frequency The crystal oscillates together with the dual-gate FET T 1 (Fig. 1). The pulling inductance L 1 and the output circuit comprising L 2 allow the pulling range and the maximum output level to be adjusted with very slight interaction. Due to the control voltage generated across diode D 1, the output voltage of T 1 remains virtually constant over the whole pulling range.

The push-pull push-push doubler equipped with Schottky diodes D 4 and D 5 is provided subsequent to the oscillator and generates a frequency of 45 MHz. This is followed by a subsequent bandpass filter equipped with inductances L 4 and L 5 which is used to filter the 45 MHz signal, especially to suppress its subharmonic 22.5 MHz, and to supply a clean drive signal for the frequency tripler equipped with T 2.

The 135 MHz signal generated in the tripler is fed to a three-stage filter and is available at the output at a level of at least 7 dBm. This power level is sufficient for driving standard Schottky diode mixers such as SRA-1, IE-500, MD-108.

1.1. Selection of the Crystal

In order to obtain the required pulling range of 200 kHz at the final frequency in the 2 m band, it is necessary – as already mentioned – to use a fundamental crystal at one sixth of the required

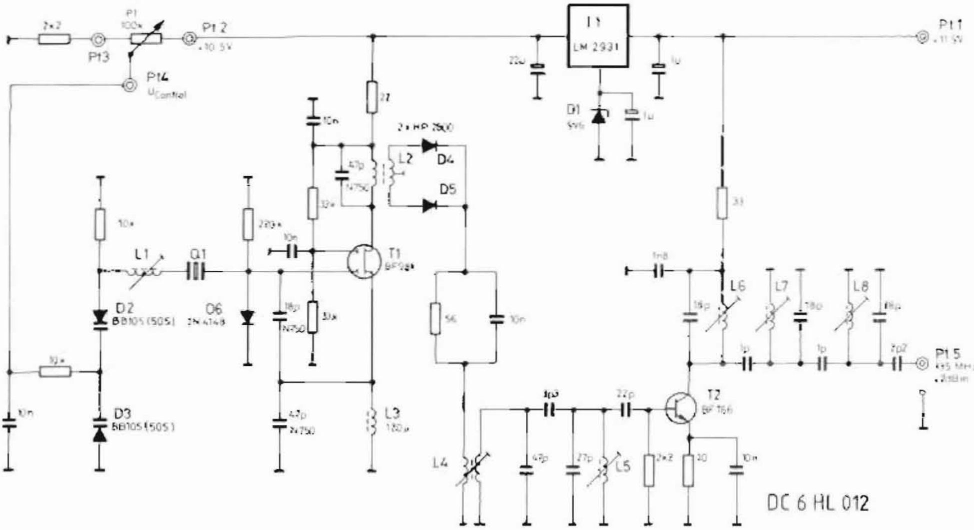

Fig. 1: Variable crystal oscillator frequency multiplication of six-times the crystal frequency

frequency. Since the pulling range is non-symmetric to the nominal frequency (2), the specifications of the crystal should be calculated according to the following equation:

$$f_Q = \frac{f_{ll} + 150\,\text{kHz}}{6} = \frac{f_{ul} - 50\,\text{kHz}}{6}$$

f_{ll} = lower frequency limit
f_{ul} = upper limit of the pulling range

With a pulling range of 135.15 to 135.35 MHz, f_Q will be 22.55 MHz. It is sufficient for one to order a fundamental crystal in a HC-43/U (HC-18/U) for the calculated frequency that is designed for a capacitive load of 30 pF.

1.2. Special Components

T 1: BF 981 (Philips), BF 907 (TI) or similar low-noise DG-MOSFET in a plastic case
T 2. BFT 66 (Siemens) or similar low-noise UHF transistor in TO-18 case
I 1: LM 2931 (National Semiconductor)
D 1: C5V6 zener diode
D 2, D 3: BB 505 B or 505 G
D 4, D 5: HP 2800 (Hewlett Packard) or similar Schottky diode
D 6: 1 N 4148, 1 N 4151 or similar switching diode
L 1 32 turns of 0.2 mm dia. enamelled copper wire in special coil set, (7V 1S) with core (yellow), (previously: blue).
Glue the winding to the coil former with the aid of a dual-component glue without bubbles. Manufacture as shown in **Figure 2.**
L 2: 13 + 2 + 2 turns, wire and coil set as for L 1. Glue the windings into place. Manufacture as shown in **Figure 3.**
L 3: Miniature choke 120 µH
L 4: 2 + 8 turns, wire and coil set as for L 1. Manufacture as shown in **Figure 4.**
L 5. 8 turns, wire and coil set as for L 1. Connection diagram is given in Figure 2.
L 6 – L 8: Ready-wound coil type 05118.

Crystal: see Section 1.1.
Case: Metal case, 74 mm × 37 mm × 30 mm
Tuning potentiometer: 100 kΩ, 10-turn helical potentiometer

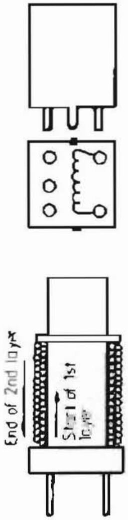

Fig. 2:
Connections of L 1 and L 5 (top), and winding diagram of L 1 (below). Two layers of 16 turns, each!

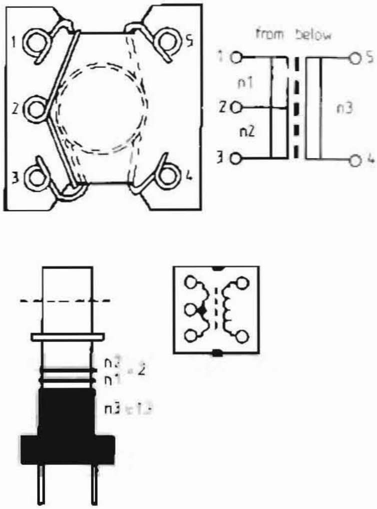

Fig. 3:
Winding and connection diagram of L 2

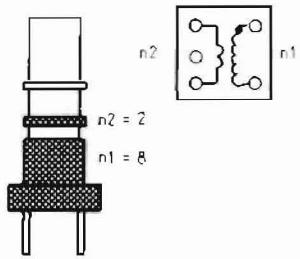

Fig. 4:
Winding and connection diagram of L 4

2. CONSTRUCTION

All components are accommodated on the double-coated PC-board DC 6 HL 012 as shown in Figure 5. This board is 70 mm × 32.5 mm and its ground surface is arranged so that no RF-currents from subharmonics can flow to the output pins.

The component connections designated with crosses should be soldered on both sides of the board (through-contacts)

Inductances L 1, L 2, L 4, and L 5 are wound according to the above-shown illustrations (Fig 2–4). It is important that the winding of L 1 is carried out in two layers using 0.2 mm enamelled copper wire. A single-layer winding using a thinner wire would lead to too low a winding capacitance and too low a Q. Transistor T 1 is solde-

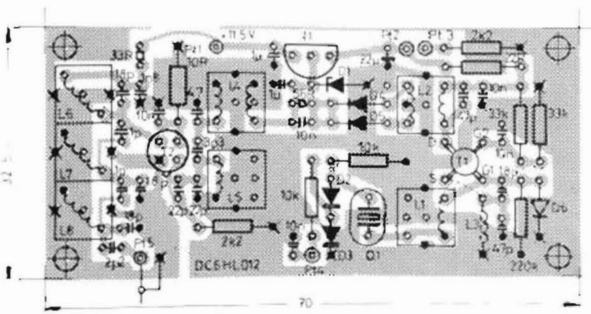

Fig. 5:
The double-coated PC-board is to be soldered on both sides of the board (through-contacts) at the positions designated with a cross.

red onto the PC-board with the marking facing towards the board.

After completing the PC-board, the outer frame of the metal box is soldered around the edge of the PC-board. The operating and tuning voltages are fed in via feedthrough capacitors (short types) of approximately 1 nF (4 pcs.). A thin coaxial cable (RG-174/U or PTFE-cable) is passed through a hole in the case and is directly soldered to Pt 5 and ground - solder pins are not necessary. A photograph of the prototype is given in Figure 6.

3. ALIGNMENT

Connect the operating voltage and the tuning potentiometer. Check the stabilized voltage of I 1 and D 1, it should amount to 10.5 V

Set the potentiometer to the highest tuning voltage and turn out the core of L 1. The oscillator should commence oscillation on tuning L 2. This can be measured with the aid of a (high-impedance) voltmeter at the cathodes of the frequency doubler diodes: the reading should amount to 0.3 to 0.35 V.

Inductances L 4 and L 5 should be aligned for maximum current drain of the complete circuit: it should amount to 15 mA.

Align inductances L 6, L 7, and L 8 for maximum output power; an output power of approximately 10 dBm should be achieved.

Rotate the core of L 1 in, until the output frequency is aligned to $f_q \times 6 + 50$ kHz. The alignment potentiometer is now tuned to the lowest tuning voltage, which should result in $f_q \times 6 - 150$ kHz. If the inductance of L 1 is increased further by inserting the core, this will result in the pulling range to become considerably greater towards lower

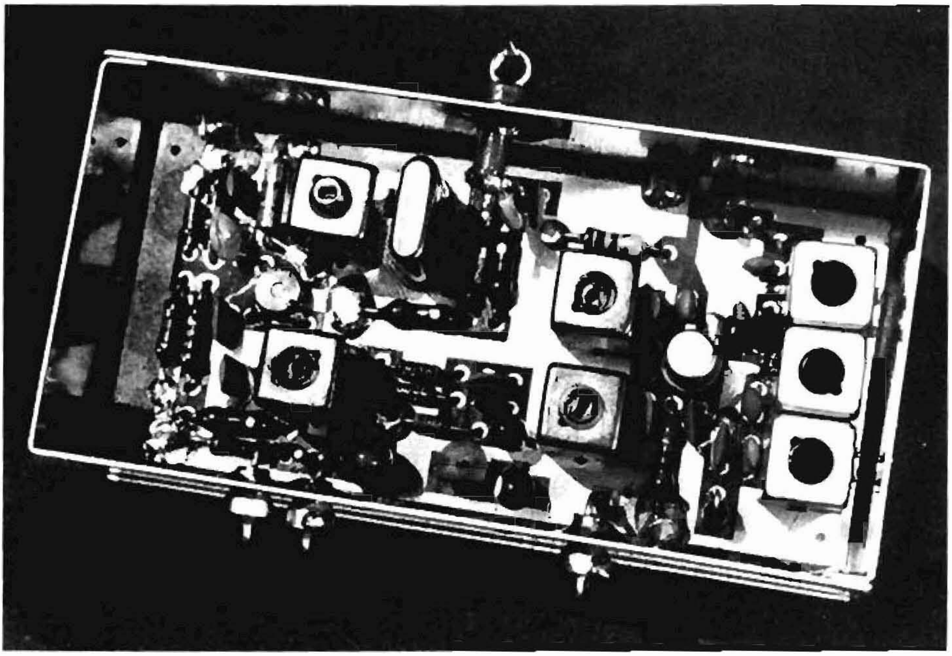

Fig. 6: The construction requires a steady hand, sharp eyes and a soldering iron with a narrow tip

frequencies. However, the frequency stability becomes less and less determined by the crystal on increasing the pulling range. For this reason, it should not exceed 200 kHz (−150 to +50 kHz from the nominal frequency).

4. MEASURED VALUES

Stabilized voltage (using 5V stabilizer and zener diode 5V6): 11.5 V.
Operating current (according to frequency): 16−18 mA
Frequency range: 135.15 to 135.35 MHz
Output power: ≧ 7 dBm (5 mW)
Spurious rejection (f_{ul} + 22.55 MHz), at least 80 dB
All others: at least 80 dB
Harmonic rejection (2nd harmonic): 80 dB
All others, at least 80 dB

Frequency stability for a temperature jump from 20 to 50°C: approx. 2 kHz.

Note:
As is the case of a VFO, this VXO should be mounted in a position in the transceiver or receiver where the lowest amount of heating occurs. In addition to this, it is advisable for the metal case of the oscillator to be surrounded with a layer of at least 5 mm of styrine foam.

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Michael Martin, DJ 7 VY

Extremely Low-Noise Preamplifiers require Low-Loss Antenna Cables!

Wideband Directional Coupler for VSWR-Measurements on Receiver Systems

The use of GaAs-FET technology has brought a considerable increase in receiver sensitivity that was thought impossible several years ago. Both bipolar transistors and GaAs-FETs now offer noise figures of less than 1 dB, and the latter even allow noise figures of less than 0.5 dB to be achieved. This means that system temperatures of 50 K ≙ −223°C are now possible using this technology, which were only possible with cooled parametric amplifiers in the past. The following article is to discuss several special features of the GaAs-FET preamplifiers without which it is not possible to obtain the values given in the data sheets. Furthermore, a wideband directional coupler is to be described that allows VSWR-measurements to be made on the input circuits of receivers and preamplifiers in the frequency range of 2 to 1400 MHz.

1. GENERAL

A typical receive system comprises a receive antenna, a short piece of feeder cable between antenna and masthead preamplifier, a second, longer cable to the actual receiver. The task of the antenna is to receive as much energy as possible. This energy should be amplified without any deterioration up to the demodulation level; this is obtained using special low-noise preamplifiers at RF-level.

2. NOISE

2.1. Thermal Noise

Thermal molecule movement causes noise which tends to blanket very low receive-signal levels. The natural limit value is given by the noise of the input impedance of the receiver which amounts to 290 K at an ambient temperature of 17°C. This corresponds to a noise power of −174 dBm per Hz of bandwidth. In the case of a conventional bandwidth of 2.4 kHz, this is equal to an input noise power of −140 dBm, corresponding to an input voltage of 22 nV into 50 Ω. This means that no signals of less than 22 nV can be received, even when using an ideal noiseless receiver that is connected using a lossless cable to the antenna, if the bandwidth is not to be decreased.

2.2. Signal-to-Noise Ratio (Noise Figure NF)

The noise figure is used to define the quality of an amplifier. This definition indicates how much the signal-to-noise ratio at the output of the amplifier has deteriorated with respect to that at the input.

The noise factor F is always > 1 and is obtained according to the following equation:

$$F = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} \quad (1)$$

The noise figure is given in dB and is obtained from the noise factor with the aid of logarithms:

$$NF_{dB} = 10 \lg F \quad (2)$$

An ideal amplifier would have a noise figure of $F = 1 \hat{=} NF = 0$ dB, and the input signal-to-noise ratio would be present at the output without change. A true amplifier with a noise figure of 3 dB will reduce a signal-to-noise ratio of 10 dB coming from the antenna to a value of 10 – 3 = 7 dB at the output. The target of all amplifier developments is to obtain a noise figure that is less than 1 dB. Improvements in the order of 0.5 dB can result in considerable system improvements in the case of EME-communications where the antenna is pointed towards cold space (1). It is possible to achieve noise figures of 0.5 dB using the present state-of-the-art on all amateur bands between 145 and 1296 MHz.

2.3. Cable Noise

Cables, attenuators, input resonant circuits, and all other passive four-poles will cause noise to the value of their insertion loss $NF = a_0$ (dB). A cavity filter having an insertion loss of 0.1 dB will deteriorate the input noise figure by the same value.

2.4. Contribution of the Second Amplifier Stage

The second amplifier stage of our receiver system influences the overall noise figure according to equation 3:

$$F_{tot} = F_1 + \frac{F_2 - 1}{G_1} + \ldots + \frac{F_n - 1}{G_1 \times G_n} \quad (3)$$

It will be seen that the higher the preamplification, the less will be the contribution of the second stage to the overall noise figure F_{tot}.

2.5. Measuring the Noise Figure

The most favorable alignment of preamplifiers with a continuous noise figure measurement is possible with equipment operating according to the PANFI-principle (Precision Automatic Noise Figure Indicator) by which the Y-factor of the amplifier is recalculated into noise figure and is continuously indicated (2). The measuring system for determining the noise figure with the aid of the Y-factor is shown in Figure 1.

According to equation 4 the following results:

$$NF_{dB} = ENR_{dB} - 10 \lg (Y - 1) \quad (4)$$

where

$$Y = \frac{P_2}{P_1} = 10^{\frac{a_z}{10}}.$$

The value a_z should be measured with an accuracy of ±0.05 dB.

ENR = Excess Noise Ratio = Noise power of the noise generator in dB. Y-factor: Ratio of the output noise power of the amplifier with the noise generator switched on and off.

If an additional attenuation of, for instance, a_z = 14 dB must be inserted after switching, in order to obtain the same meter-reading with a noise generator having an ENR of 15 dB, the following will result from equation 4.

$$NF = 15 - 10 \lg (10^{1.4} - 1) = 1.17 \text{ dB}.$$

An a_z-value of 14.5 dB, on the other hand, would result in an NF-value of 0.65 dB. It will be seen from equation 4 that any inaccuracy of the ENR-calibration of the noise source will have an immediate effect on the test result. A further difficulty in the case of absolute measurements is that

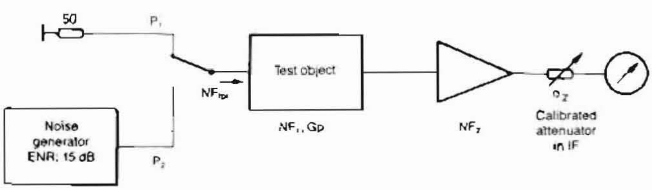


Fig. 1: Determining the Noise Figure NF from the Y-factor

only very good calibration attenuators allow a reproducible accuracy of better than 0.1 dB, which is especially required when low noise figures are to be measured.

mum when the impedance of the antenna corresponds to the impedance Z_0 of the cable, and this to the input impedance Z_{in} of the pre-amplifier

3. CHARACTERISTICS OF ANTENNA CABLES

All cables between the antenna and the preamplifier possess an insertion loss that will deteriorate the system noise figure by at least the value of its insertion loss. This is not only a function of the cable length, but is also dependent on the impedance (Z_{opt} = 75 Ω for CATV), and the terminating resistance. The insertion loss is a minimum when the impedance of the antenna corresponds to the impedance Z_0 of the cable, and this to the input impedance Z_{in} of the pre-amplifier

If there is a difference between Z_0 and Z_{in}, standing waves will be generated in the cable that cause ($I^2 \times R$)-losses at the points of maximum current, and lead to greater dielectric losses at the points of maximum voltage. Cables having gas or PTFE insulation using thick, polished conductors will have the lowest losses. **Figure 2** shows a diagram in which the loss of the most common cables is given for a length of 100 m as a function of frequency. The different slope of the lines shows the different distribution between copper and dielectric losses.

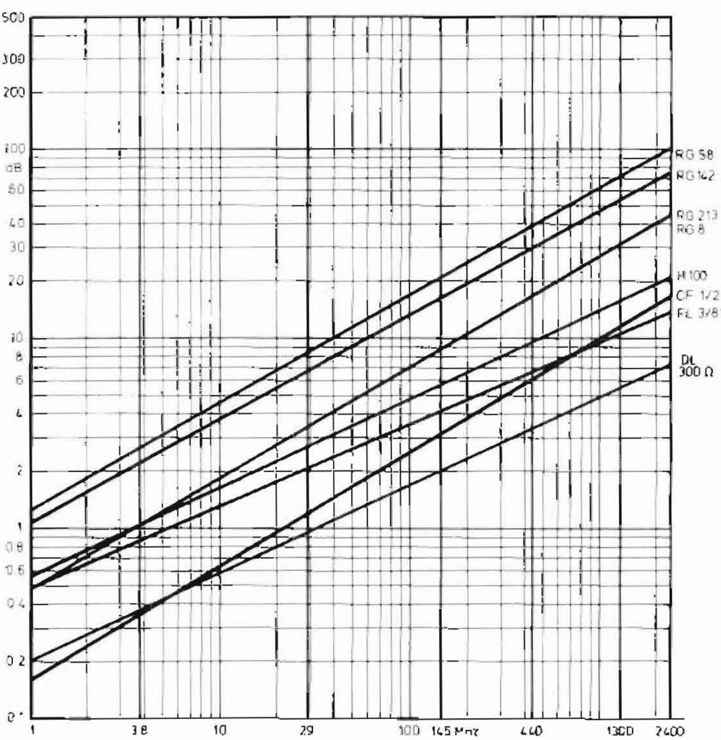
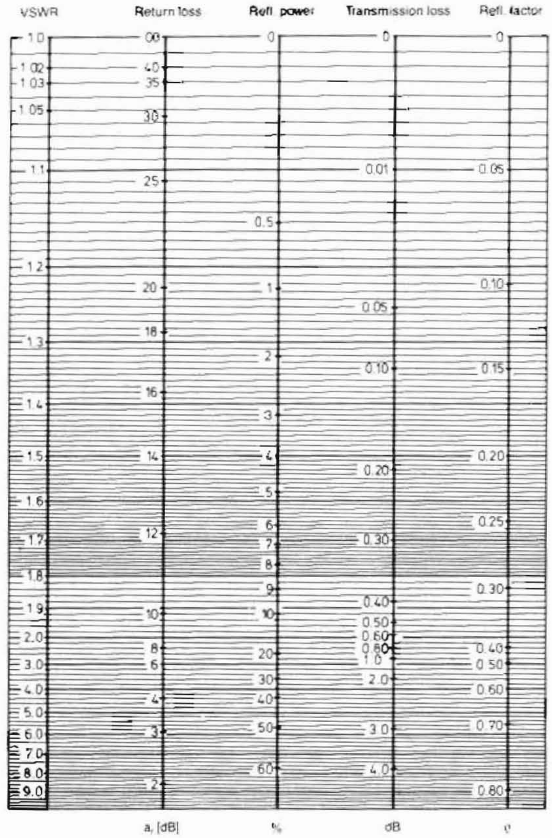


Fig. 2: Attenuation a_0 of various coaxial cables, as well as a 300 Ω twin line shown as a function of frequency for a cable length of 100 m

Fig. 3
Relationship between return loss a_r, VSWR, and reflection factor ϱ



In the ideal, matched case, the following will result, for instance, with two metres of RG-213 cable having a_0 = 0.32 dB at 435 MHz in front of a preamplifier with a NF of 1 dB and an input impedance of 50 Ω and a VSWR of 1.0: Overall noise figure NF_{tot} = 1.0 + 0.32 = 1.32 dB.

4. CHARACTERISTICS OF PREAMPLIFIERS

Both in the case of preamplifiers equipped with bipolar transistors and those equipped with FETs, there is a physically-dependent difference between their input impedance and the generator impedance required for obtaining a minimum noise figure. This is the well-known difference between power and noise matching. It was possible in (3) to obtain an approximation of both values, however, it was necessary to reduce the gain per stage to approximately 10 dB, which resulted in the contribution of the second stage to amount to one ninth of the total value (see equation 3)

Several preamplifiers were examined with the aid of a directional coupler bridge with respect to their input matching. The following relationship exists between VSWR and return loss a_r according to equation 5:

$$VSWR = \frac{1 + 10^{-0.05\,a_r}}{1 - 10^{-0.05\,a_r}} = \frac{1 + \varrho}{1 - \varrho},$$

ϱ = reflection factor

This can be determined graphically using Figure 3.

Bipolar amplifiers with feedback, exhibited values in the order of a_r = 12 dB ≙ VSWR = 1.67 with a NF of 1.2 dB, whereas GaAs-FET amplifiers with a NF of 0.6 dB exhibited a_r-values of only 2 dB ≙ VSWR = 8.72. In the case of special wideband types VSWR-values of between 17 and several hundred were exhibited. Especially low-noise circuits with noise figures of less than 0.5 dB exhibited, sometimes, negative input impedances.

It was found in many measurements of the cable losses in a measuring system comprising noise generator – cable to be measured – preamplifier – PANFI that the determined values differed greatly from the theoretical values (NF_{tot} = NF_{prea} + a_o). This is caused by the fact that residual reactive components of the generator impedance and the resulting impedance transformation in the cable can falsify the noise matching of the amplifier. These values will differ even more, the higher the input VSWR of the preamplifier, and the longer the interconnection cable.

5. EXAMPLES

A few examples are now to be given to show the effect of attenuation on the system noise figure.

5.1.1. Preamplifier equipped with bipolar transistor:

NF = 1.2 dB, a_r = 12 dB, VSWR = 1.67 When using a 10 m length of RG-213 cable at 435 MHz, the following will result from Figure 2.

a_o = 1.6 dB;

NF_{tot} = 1.6 + 1.2 = 2.8 dB

5.1.2. Same amplifier located close to the antenna.

2 m RG-213, a_o = 0.32 dB,

NF_{tot} = 1.2 + 0.32 = 1.52 dB

NOTE:

Low-Noise Preamplifiers should always be mounted in the vicinity of the antenna!

5.2.1. Preamplifier with GaAs-FET:

NF = 0.6 dB

10 m RG-213 a_o = 1.6 dB

NF_{tot} = 0.6 + 1.6 = 2.2 dB; what a waste of a GaAs-FET!

5.2.2. Same amplifier mounted in the vicinity of the antenna:

2 m RG-213, a_o = 0.32 dB

NF_{tot} = 0.6 + 0.32 = 0.92 dB

5.3. Low-Loss Antenna Cable

3/8"-Flexwell cable with inner dia. = 4.2 mm, outer dia. = 16 mm; 0.13 dB/2 m at 435 MHz

5.3.1. Bipolar Amplifier

NF = 1.2 dB and VSWR = 1.67

NF_{tot} = 1.2 + 0.13 = 1.33 dB

5.3.2. GaAs-FET

NF = 0.6 dB and VSWR = 8.72

NF_{tot} = 0.6 + 0.13 = 0.73 dB

6. IMPROVING THE NOISE FIGURE BY USING EXTREMELY LOW-LOSS FEEDER CABLE

Twin-line "DL" is constructed from 2.5 mm dia. enamelled copper wire with PTFE-spacers at a spacing of λ/2, wire spacing 12.5 mm, Z_o = 298 Ω, Velocity Factor VF = 0.95.

Attenuation at 435 MHz: 3.3 dB/100 m,
2 dB/100 m at 145 MHz.

6.1. Bipolar Amplifier with 2 m Twin Line

a_o = 0.07 dB

NF_{tot} = 1.2 + 0.07 = 1.27 dB

6.2. GaAs-FET Preamplifier with 2 m Twin Line

a_D = 0.07 dB
$NF_{(o)}$ = 0.6 + 0.07 = 0.67 dB, which corresponds to a very low NF deterioration!

RESULT:
The high-impedance, low-loss twin line is very suited to the high-impedance input of the GaAs-FET preamplifier!
NF = 0.6 dB ≙ 43 K noise temperature, NF = 0.676 ≙ 49 K

NOTE:
In order to keep the effects of differing cable length in the system noise figure as low as possible, the intermediate cable should be as short and as low-loss as possible. An exact solution is only possible by aligning the preamplifier in conjunction with the antenna and interconnecting cable by injecting a keyed noise power, using a second antenna, into the receive system and aligning it for minimum noise figure on the PANFI.

Since this is very extensive, and is usually not possible at most amateurs' locations, it can usually only be carried out in the "laboratory": The generator impedance which the antenna offers to the preamplifier at the end of the feeder cable must be measured and the amount and phase must be simulated with the aid of a stub-tuner between noise generator and preamplifier during the alignment for minimum noise figure.

A minimum noise figure alignment made in the laboratory in conjunction with a noise generator impedance of 50.0 Ω ±0.5 Ω ≙ a_r ≦ 40 dB is only reproducible in practice if the antenna exhibits the same impedance!

Amateur antennas, on the other hand, sometimes have a_r values between 10 and 20 dB, which means that the results remain somewhat uncertain.

7. MEASUREMENT OF THE INPUT MATCHING WITH THE AID OF A DIRECTIONAL COUPLER BRIDGE

Since preamplifier and receiver input circuits are usually overloaded when using a power of 0 dBm ≙ 1 mW, it is not possible to use conventional VSWR-meters due to their insensitivity.

With the aid of the measuring system shown in **Figure 4**, it is possible for the a_r-value to be determined reliably at levels of less than −40 dBm. A generator feeds a level of approx. −40 dBm at the required measuring frequency to the bridge circuit. The power present at the input of the test object corresponds to the injected power minus 6 dB. A 50 Ω terminating resistor with the best possible matching is connected to the reference port to which the test object is compared. The receiver is now connected via a calibrated attenuator with the output of the bridge. After disconnecting the test object, the receiver is adjusted to a certain S-meter reading corresponding to

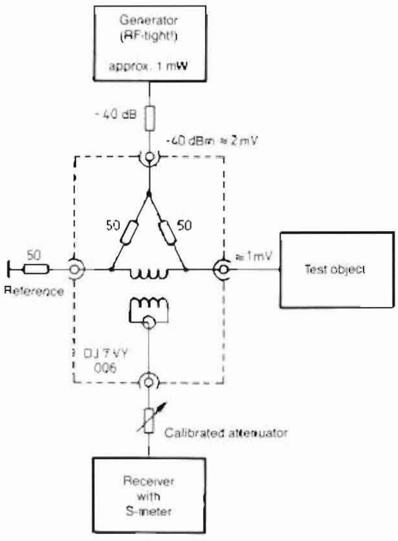


Fig. 4: Measuring system for determining the return loss of preamplifiers

approximately 60 dB with the aid of the calibrated attenuator. If an identical terminating resistor is now connected to the test object port, the reduction of the S-meter reading corresponds to the directional response of the bridge. This should be greater than 30 dB, if possible, when low VSWR-values are to be measured. After connecting the test object, the reduction of the S-meter reading is directly proportional to the a_r-value in dB, if the S-meter is also calibrated in dB. With the aid of the calibrated attenuator, a substitution measurement can be carried out by measuring this attenuation and reading it off on the attenuator.

It is then possible with the aid of equation 5 or Figure 3 to determine the VSWR. In order to be able to establish the overall system noise figure when using differing preamplifiers and cables, it is necessary for the input reflection values to be obtained from the manufacturer of the preamplifiers. A detailed description of further measurements that can be carried out with the aid of the bridge are to be found in (4)

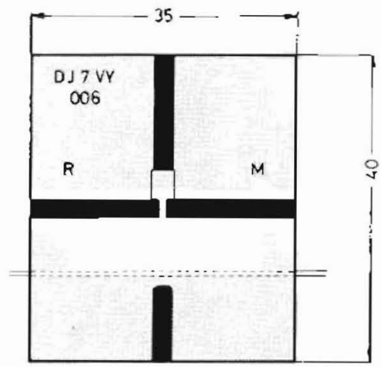


Fig. 5:
The double-coated directional coupler board DJ 7 VY 006 is made from 1.5 mm thick epoxy glassfibre material G 10

8. CONSTRUCTION

The directional coupler bridge can be accommodated on the double-coated PC-board DJ 7 VY 006 and enclosed, as shown in Figure 5, in a metal box having the dimensions 35 mm x 110 mm. The main difficulty of this design is to be seen in the construction of the extremely wideband balun transformer which is similar to a description published by HP in (5)

The case is firstly provided with the four connectors, after their center conductors have been shortened down to approximately 2 mm. After cutting the board at the dashed line, cut-outs should be provided on the PC-board to fit the protruding parts of the connectors and for the resistors. Caution should be paid that not too much material is removed! The PC-board should fit in as well as possible into the top and bottom of the case!

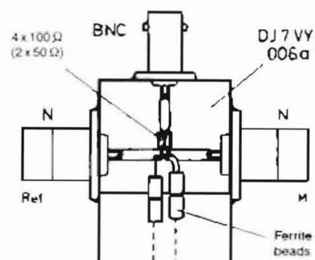


Fig. 6:
Partial PC-board DJ 7 VY 006b must be rotated by 180° and soldered into place

The two parts of the board are now soldered into place as shown in **Figure 6** by firstly soldering the stripline connections after which the ground surface is soldered. The four 1% resistors that form the input resistors are now soldered into place at the location of the cutout, and an approximately 90 mm length of SM 50 PTFE cable is soldered to the output stripline. Place as many ferrite beads onto the cable until approximately 7 mm spacing remains to the balanced center point. The inner conductor of the cable is connected to the reference stripline, and the outer conductor is connected to the measuring port stripline. A copper wire of 0.8 mm diameter is also provided with ferrite beads and is soldered into place between ground and the reference line. The ferrite beads should be glued as shown in the prototype, using a normal adhesive so that the last 7 mm remain free. Important: The balun should run as horizontal as possible to the

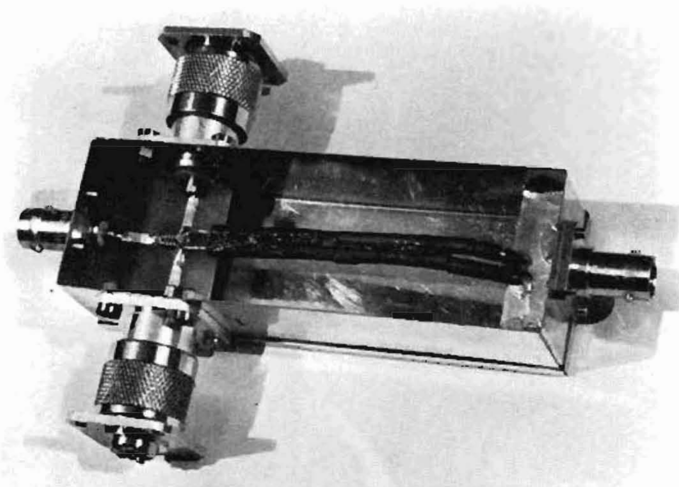


Fig. 7: Photograph of the author's prototype, seen as Fig. 6

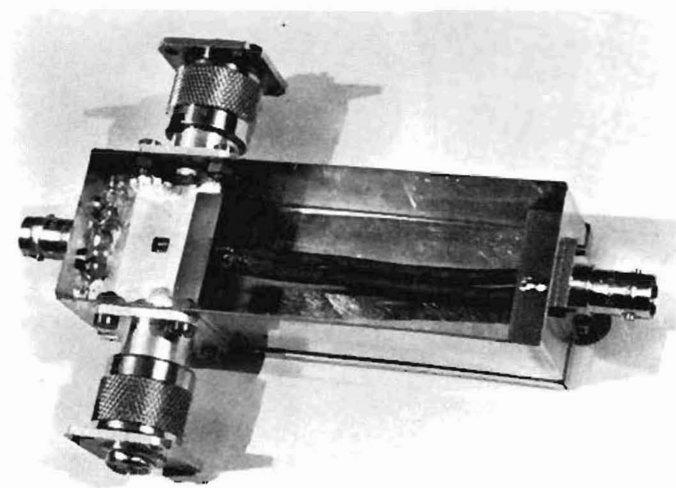


Fig. 8: Photograph of the author's prototype from the other side

balanced point of the bridge! This is shown more clearly than in the text in the photographs of the author's prototype. **Figures 7 and 8** show that two pieces of 51 Ω resistors have been installed instead of the four pieces of 100 Ω.

With a clean, balanced construction, the directivity of the bridge will be greater than 30 dB! The values of the author's prototype are given in **Figure 9**. It is possible to construct one's own terminating resistors instead of the very expensive precision terminating resistors described (see **Figure 10**). Two 1%-resistors of 100 Ω, each, can be soldered onto a flange connector after filing down the center pin to the same height as the outer collar. These terminating resistors exhibited a return loss in the order of 30 dB even at 1296 MHz. This is greatly suitable for amateur measurements, since the test objects to be measured will never be better than 1.2 ≙ a_r = 20 dB.

Since the price of such home-made terminating resistors is low, it is recommended that identical 60 Ω (2 × 120 Ω/1%) and also 75 Ω (2 × 150 Ω/1%) types are constructed so that it is possible to measure VSWR-values at other impedances.

CAUTION:
The input coupling resistors will not handle more than 0.1 W. In the case of the author's prototype, the lowest directivity values amount to:
2 MHz: −22 dB, 4 MHz: −30 dB, 10 MHz: −36 dB, 1400 MHz: −36 dB. The lower cutoff frequency of the bridge can be reduced down to below 1 MHz by using a longer case and thus a longer balun.

9. COMPONENTS

1 metal case
2 BNC flange connectors
2 N flange connectors
2 N flange connectors for terminating resistors
1 PC-board DJ 7 VY 006
8 resistors 100 Ω/1%
28 ferrite beads
approx. 20 cm 50 Ω teflon cable SM 50 or semi-rigid copper cable

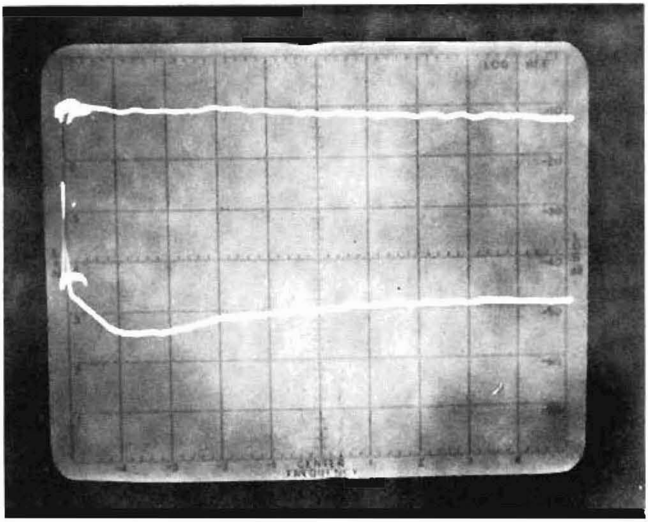


Fig. 9:
Directivity of the directional coupler bridge constructed as Fig. 6 with two precision terminations manufactured by HP. Measured between 0 and 1 GHz
(H: 100 MHz/T;
V: 10 dB/T)
2 MHz: −22 dB
4 MHz: −30 dB
10 MHz: −36 dB
1400 MHz: −36 dB

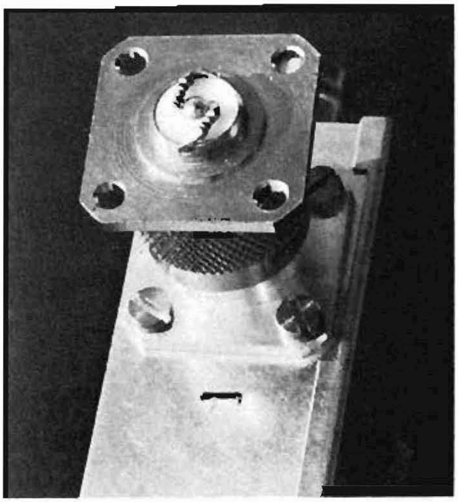


Fig. 10:
Photograph of a home-made termination using an N-flange connector and two 100 Ω resistors

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Using the Vector Impedance Meters

Günther Borchert, DF 5 FC

A 2 m/70 cm SSB Transmitter with High Spurious Rejection

Part I

Recently, various publications have been brought giving circuit descriptions for high-quality receiver input circuits, IF-amplifiers, and oscillators. This means that receive technology is now at a very high level. This article is to give a recommendation for a combined transmit system for 2 m and 70 cm, which is based on an intermediate frequency of 9 MHz. However, other intermediate frequencies are possible. It is only necessary to modify the frequency plan.

During conception of the individual modules, special attention has been paid that they are modular and that they can be used for other applications. It may seem that the modular construction is rather complex, however, it will increase reproducibility and simplify alignment.

1. CONCEPT

The development target was to obtain a 2 m and 70 cm transmit signal from an SSB IF-signal at 9 MHz with the aid of the lowest number of mixer stages and individual frequencies. The considerations that led to this concept can best be seen with the aid of a frequency scale:

As can be seen in Figure 1, the mixing of 135 MHz and 9 MHz will not only generate the required 144 MHz, but will also generate the image frequency of 126 MHz. This shows that the 2 m signal can be generated easily if the 126 MHz signal is suppressed.

A few problems exist in the case of the 70 cm. A direct conversion from 9 MHz is not possible since the required bandwidth should cover the whole amateur band of 10 MHz, which means that the local oscillator would fall in the required range at the band limits. A further processing of the 2 m signal is very unfavorable, since the third harmonic will also fall in the required range. The generation of this harmonic is inavoidable since a conversion can only occur in conjunction with non-linearities, and these will also generate harmonics.

A solution can be found here in several ways. One can, for instance, firstly convert the 2 m

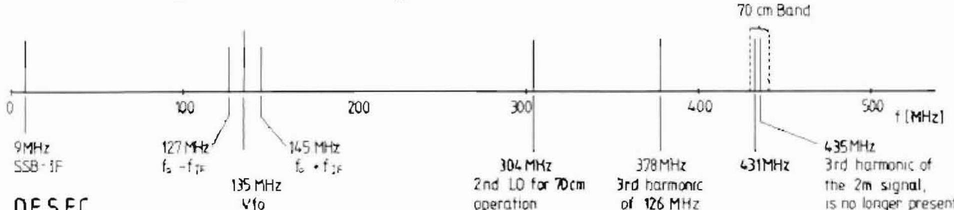

Fig. 1. Individual frequencies appearing in the frequency plan

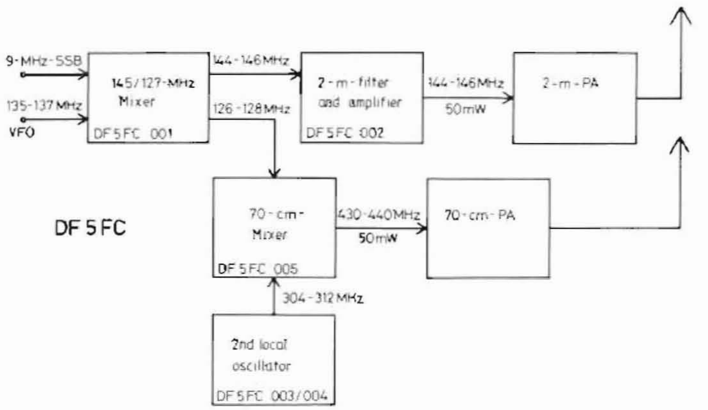

Fig. 2: Block diagram of the concept

signal to a further frequency, after which it is converted further. This, on the other hand, results in further conversion products (and correspondingly more lines on the frequency axis shown in Figure 1).

These frequencies can now form further unwanted conversion products and harmonics in conjunction with further non-linearities. The necessary amount of filtering can then become excessive, since these unwanted products can be adjacent to the required frequency, which means that the filters must be very steep.

As previously mentioned, the output signal of a mixer contains not only the sum signal $f_{osc} + f_{in}$, but also the difference signal $f_{osc} - f_{in}$. The resulting image frequency of 126 to 128 MHz can be used successfully for a further processing to the 70 cm band. The third harmonic of this signal is nearly 50 MHz from the required 70 cm band (see Figure 1) and can be suppressed easily. A similar concept has already been described by DL8ZX in (1).

A disadvantage of this method is that an inversion of the modulation sideband of the SSB signal takes place during the difference conversion. This means that an USB-signal will become LSB and vice versa. This disadvantage is not so important as the increase of the spurious signal rejection. It can be overcome by organizing a corresponding connection of the sideband crystals together with the amateur band selection. A frequency concept resulted from these considerations that can be best described with the aid of a block diagram.

As can be seen in **Figure 2**, the VFO-signal is mixed with the 9 MHz SSB signal in the first mixer, which results in frequencies of 145 MHz and 126 MHz. The 2 m signal is now carefully filtered and amplified so that an output power of approximately 50 mW is available after the selective amplifier for further processing.

The 126 to 128 MHz signal is fed to a second mixer where it is mixed with 304 – 312 MHz to form the required frequency of 430 to 440 MHz. Approximately 50 mW is also available at the output of the mixer module.

The second local oscillator frequency is generated in a 5-stage crystal oscillator and multiplied. The oscillator and the frequency multiplier are constructed in a complex manner, since the signals are also to be used for receive applications.

2. TECHNOLOGY

2.1. SELECTION OF THE MIXER

Schottky diode ring mixers are used in both mixer stages, since they are balanced and thus suppress both input frequencies. The use of a high-level mixer, type SRA-1H, is not absolutely necessary, however, the less expensive IE-500 or SRA-1 have a considerably poorer intermodulation behaviour.

The typical intermodulation behaviour of diode ring mixers can be seen in the data sheets. It is shown that an intermodulation rejection IM_3 of 60 dB results at an RF-level of 2 × −10 dBm ≙ 2 × 0.1 mW, in the case of the SRA-1H, or approximately 45 dB at an RF-level of 2 × 0 dBm ≙ 2 × 1 mW. In comparison to this, only an intermodulation rejection of approximately 20 dB (!!) is obtained from the SRA-1 at 2 × 0 dBm RF-level; the SRA-1 is very similar to the IE-500 with respect to its construction and level values. A reduction of the RF-level to 2 × −10 dBm improves the rejection to 45 dB. These measured values show that the SRA-1H does not react as critically to overload as the less expensive mixers. This is only valid with the oscillator powers given in the data sheets:

SRA-1H + 17 dBm ≙ 50 mW;
SRA-1 + 7 dBm ≙ 5 mW

Furthermore, it is necessary for all ports of the mixers to be terminated with 50 Ω, otherwise different values will result. This 50 Ω matching must be given for all possible frequencies in order to obtain a good intermodulation behaviour, which was discussed in detail by DJ 7 VY in several articles (2, 3). Since the individual frequencies appear discretely in the case of a transmitter, (f_{osc}, f_{in}, etc.), and not as a virtually continuous spectrum as given in the case of a receive mixer, it is not necessary to provide wideband matching, and this is specially valid when the oscillators possess a low-noise, narrow-band spectrum.

2.2. SUITABLE FILTERS

Helicial filters were used exclusively to obtain the main selectivity in this design. They were selected because they can be calculated very easily and are reproducible during construction. The expense may seem high, but the results are excellent. The most important calculation fundamentals for helical filters are to be mentioned briefly.

Helical filters are bandpass filters. The bandwidth and slope of such filters is not only dependent on the coupling between the two circuits but also on the Q of the individual circuits. Normal inductances have non-load Qs of up to approximately 200, and this is reduced considerably on load. A high Q is required for high selectivity. In the frequency range of interest, any coaxial systems can be used, however, a normal λ/4 length would be too long. For this reason, one combines inductance and resonator by replacing the inner conductor of the coaxial circuit with a helical inductance. It is possible in this manner, to obtain non-load Q-values of more than 1000. These circuits can either be tuned with the aid of a capacitor which is built into the chamber, or by a metal plate which will be brought into the vicinity of the hot end of the coil. Due to the field distribution, this acts as a capacitor.

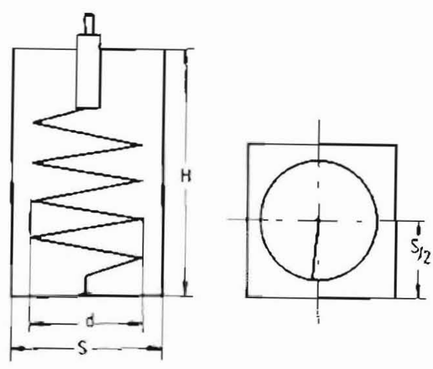

Fig. 3:
Construction of an individual helical circuit

In our case, the circuits were tuned with capacitors, and these were special spindle trimmers in order to obtain a higher reproducibility. The spindle trimmers are not mounted on the front of the chamber, but in one of the side surfaces, which results in mechanical simplifications. The effect of this capacitor is taken into consideration during the design of the filter by deleting several turns from the calculated number (this must be determined experimentally in the case of home-made filters). Actually, such a filter should have a round case, however, square types are simpler and do not have any disadvantages. Figure 3 shows a drawing of such a single-filter chamber.

The following is valid $S = \frac{Q}{24\sqrt{f_o}}$ with

f_o = resonant frequency in MHz
Q = Q of the inductance
H = 1.6 S

The number of turns is found as $N = \frac{4000}{f_o \times S/cm}$

The length of the inductance should amount to S, the coil diameter d = 0.66 S, and the wire diameter D = S/2N. This results in the following for such an inductance:

$$Z_o = \frac{2 \times 10^5}{f_o \times S}$$

In the case of 145 MHz as center frequency, one will obtain the design shown in Figure 4 with the following specifications:

H = 36 mm
S = 20 mm
d = 13 mm
D = 1 mm
N = 8 turns

It should be taken into consideration that the length of the resonator and the spacing from the case are given.

The other bandpass filters are combined from these resonators; neighbouring inductances should have an opposite direction of turns. The spacing between the inductances amounts to S. In the case of the 2 MHz filter, the coupling is made capacitively with the aid of a coupling wire which protrudes from one chamber into the other. In the case of the filter with a bandwidth of 10 MHz for the second local oscillator, and in the 70 cm stages, the filters are coupled by radiation. This allows a higher bandwidth. For this reason, only 2-stage filters are used for the 10 MHz bandwidth filters. Three-stage filters can only be tuned with considerable effort.

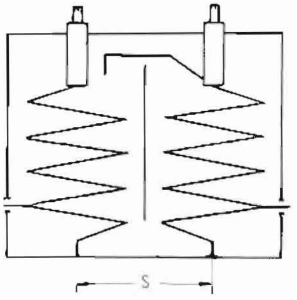

Fig. 4:
A two-stage helical filter with two types of coupling

An example for the selectivity characteristics of these filters can be seen in the case of the 2 m mixer. A total of two three-stage filters are used here. After a careful alignment to obtain a flat passband curve, one will obtain a stopband attenuation in the order of 70 dB, 8 MHz from the required band.

2.3. AMPLIFIERS

Either FETs in a common-gate circuit or amplifiers such as described by DJ7VY in (4) are used in all modules. Both types have been found to work without problems and only show a very low tendency to instability.

In the case of the FETs, it has been found that types P 8000, P 8002, and 2 N 4856 A are equally suitable, however, the first two types are more favorable for power applications since they can be cooled more easily.

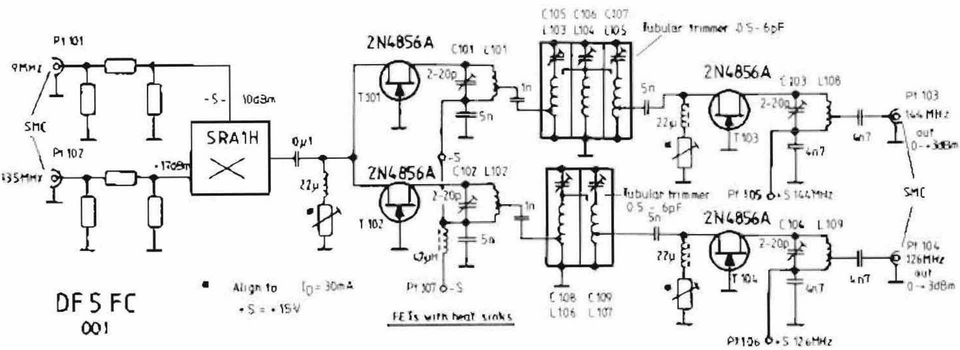

Fig. 5: Circuit diagram of the mixer from 9 MHz and 135 MHz to 144 MHz or 126 MHz, respectively, for 70 cm transmitters

3. CIRCUIT DESCRIPTION

3.1. MIXER 145/126 MHz

The first mixer stage, which is shown in Figure 5, is wideband at the inputs. This means, it is simple to change it to other frequency plans. The attenuators are to match the input level to that required by the mixer and to dampen the ripple of the 50 Ω matching. Attenuation values are selected so that an oscillator power of + 17 dBm (+ 7 dBm) results and an IF-power of − 10 dBm (value in parantesis for IE-500 or SRA-1)

The output of the mixer must now be terminated with approximately 50 Ω for all frequencies present, in order to obtain the required intermodulation rejection. furthermore, the two frequencies must have equal priority for the output coupling.

The matching is made with the aid of FET-stages in a common-gate circuit. In order to couple out two frequencies in parallel, the mixer is followed by two FETs with a drain circuit, each, for the required output frequency. The current is set with the aid of a common source resistor so that the mixer is provided with approximately 50 Ω. If the two FETs have approximately the same characteristics, the current will be divided approximately equally between them. A matching measure-

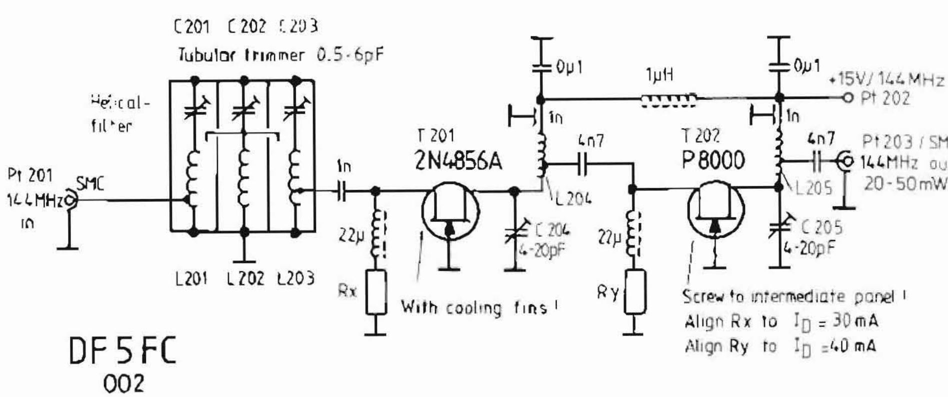

Fig. 6: Selective mixer-amplifier for 144 MHz with a maximum output of + 23 dBm

ment carried out with a network analyzer from HP showed that the resistor value of the input impedance of the FET-circuit varied between 40 Ω and 60 Ω, although the currents were not optimized, and the FETs were not selected. For both resonant frequencies, the reflection factor of the stage was in the order of 0.1, which meant that the impedance values were considered to be permissible, and no further matching was made.

The output coupling stages are followed by helical filters, which are used for the main selectivity, a three-stage filter is used for 2 m, and a two-stage filter for 126 MHz. They are calculated according to the already described criteria. The subsequent FET-amplifiers are constructed as usual.

3.2. ACTIVE FILTER FOR 2 m

The selectivity of a three-stage filter for 144-146 MHz was not found to be satisfactory in our measurements. For this reason, a further module (DF5FC 002, Figure 6) and identically constructed filter with two further FET-amplifier was provided. By correct selection of the most favorable output coupling of the inductances in the drain circuit of the stages (must be made experimentally), it is possible to obtain an overall gain of the module of approximately 20 dB. If the operating voltage is increased to 24 V, it is possible to obtain an output power of more than 100 mW. After changing the filter circuits, it is possible for this module to also be used as a local oscillator amplifier up to an output level of + 23 dBm.

3.3. SECOND LOCAL OSCILLATOR

A further local oscillator signal is required for the further processing of the 126 to 128 MHz signal. In order to cover the full required 10 MHz bandwidth in the 70 cm band with a tuning range of 2 MHz, it is necessary for the oscillator to provide a total of five frequencies in the range of 304 to 312 MHz (similar to that described by DJ3VY in (5)).

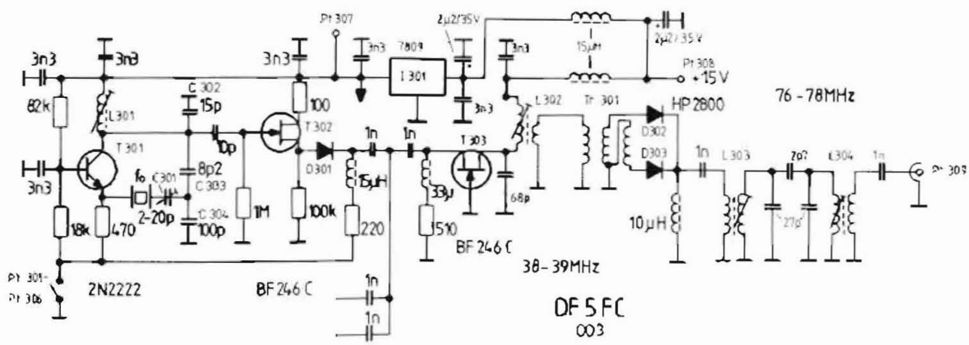

Fig. 7: Crystal oscillator/doubler/filter for the second local oscillator.
The following values are valid for the 76 MHz crystals: C 302: 12 pF; C 303: 6.8 pF, C 304: 82 pF; L 301 and L 302 see section 4.5.

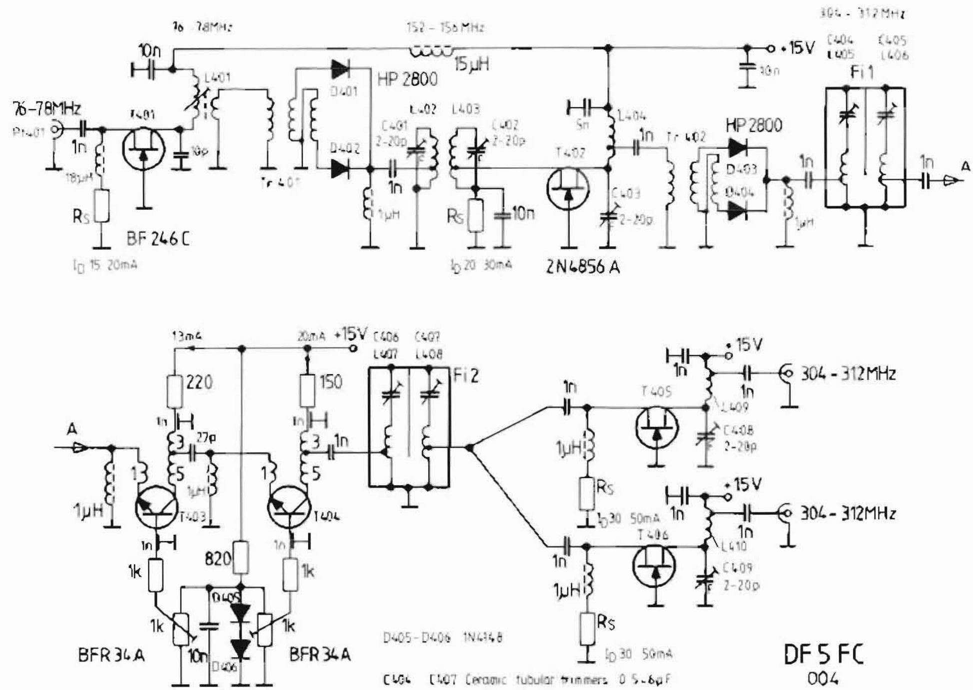

Fig. 8: Circuit diagram of the frequency multiplier for the second local oscillator

In the author's prototype, the fundamental frequency chosen was 38-39 MHz, however, a frequency range of 76-78 MHz would be more favorable, since the frequency multiplication factor would be lower. DJ3VY's PC-board can be used for this oscillator if some of the component values are changed, or, as an alternative, the oscillator module DF5FC 003 shown in **Figure 7** that operates with crystals either in the range between 38 and 39 MHz, or between 76 and 78 MHz. In the first case, a frequency doubler equipped with Schottky diodes is provided on the PC-board, which can be bridged if it is not required.

The other two doublers are found on the second board DF5FC 004, as can be seen in **Figure 8**; they are similar to those published by DJ3VY and DJ7VY, and the information given in these articles is also valid. The bandpass filter at 152 MHz was necessary in order to suppress the subharmonics sufficiently.

The helical filters are calculated according to the previously mentioned criteria, but are provided for radiation coupling in order to obtain the required 10 MHz bandwidth without complicated alignment processes. Two filters have been provided in order to obtain a sufficient spurious rejection despite of the two-stage filter. The amplifier between these is a "standard" circuit as designed by DJ7VY, and the instructions given in his article (4) are valid. The output stage comprises two FET amplifiers that allow a low-reactive, simultaneous decoupling of the oscillator frequencies, for instance, for transmitter and receiver.

A maximum of + 17 dBm can be coupled out at an operating voltage of 24 V when using P 8000 or P 8002 with a drain current of 40 to 50 mA. This is sufficient for driving high-level mixers. When used exclusively in the transmit mode, the two amplifier stages can be deleted from the board

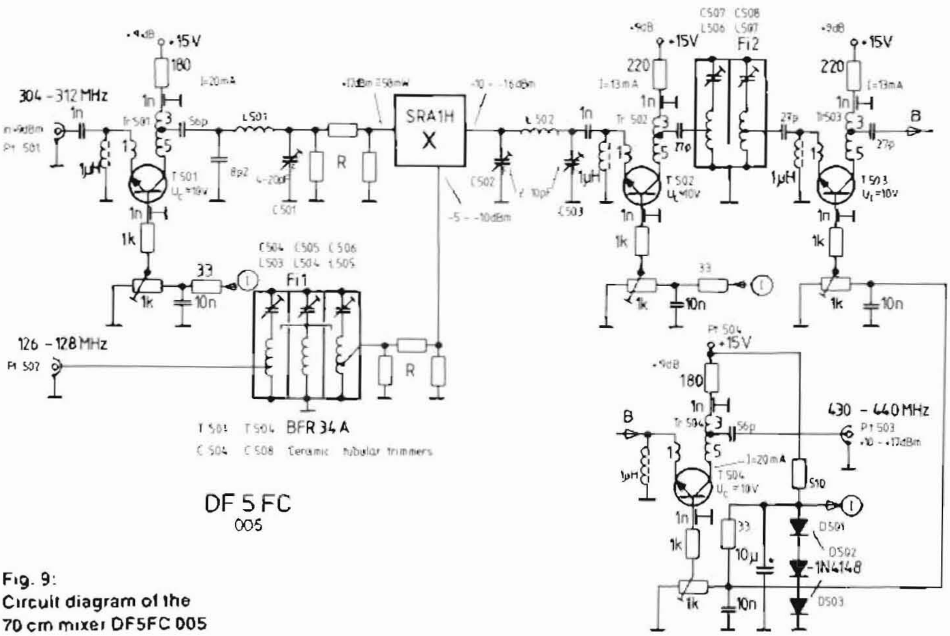

Fig. 9:
Circuit diagram of the
70 cm mixer DF5FC 005

Measurements made on the frequency multiplier have shown a virtually spurious-free signal: only a slight component of 152 MHz could be measured at 75 dB down on the output signal.

3.4.
70 cm MIXER

The same is valid for the matching of the mixer of module DF5FC 005 (**Figure 9**) as was valid for the first mixer The attenuators are once again used for power matching and to reduce the ripple The three-stage filter on the 126 MHz side is to remove all traces of the critical 145 MHz signal, and it will also keep any harmonics of the amplifier away from the mixer For this reason, a bandpass filter is provided subsequent to the oscillator amplifier (T 501). This subsequent amplifier is necessary if the oscillator signal is present with a level of approximately 7 to 8 dBm If a power level of + 17 dBm is already available, this amplifier can be deleted and the signal is sent directly to the lowpass filter.

The matching of the mixer to the subsequent amplifier is made with the aid of a Pi-filter. This, in turn, is followed by the 432 MHz two-stage helical filter which is the only selectivity provided for the output frequency. It is radiation-coupled in a similar manner to the oscillator filters in order to obtain the required bandwidth of 10 MHz. Approximately + 17 dBm are available at the output of the module, and the author's prototype even provided + 20 dBm. When only one filter was used, the spurious waves could be reduced to − 60 dB. This high spurious rejection with only one filter is the result of the large spacing between the frequencies used.

Hans Joachim Senckel, DF 5 QZ

A 13 cm Fully Transistorized Transverter

All previously described and published 13 cm transmitters operated exclusively with high local oscillator powers. Power mixers, such as with a diode BXY 28 (1), or with a tube 2C39 (2) require local oscillator powers in the order of 1.5 to 2.5 W. In order to achieve such power levels, several varactors, as well as a power amplifier in the range of 350 to 500 MHz are used. The mechanical but also the financial expense are very considerable. For this reason, a transmitter was designed that could be manufactured at low cost and with low mechanical needs.

BLOCK DIAGRAM

As can be seen in Figure 1, the transverter comprises five modules, the interdigital mixer and the IF-preamplifier are also given.

The local oscillator module (DF5QZ 001) provides an extremely low spurious signal to the transmit mixer (002) The oscillator signal for the receive converter is taken at this point via a further BNC-connector The receiver comprises an interdigital mixer with IF-preamplifier as described in (3).

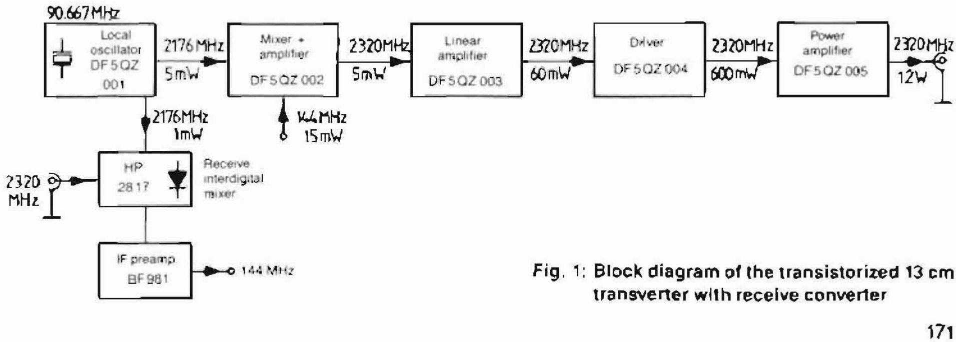

Fig. 1: Block diagram of the transistorized 13 cm transverter with receive converter

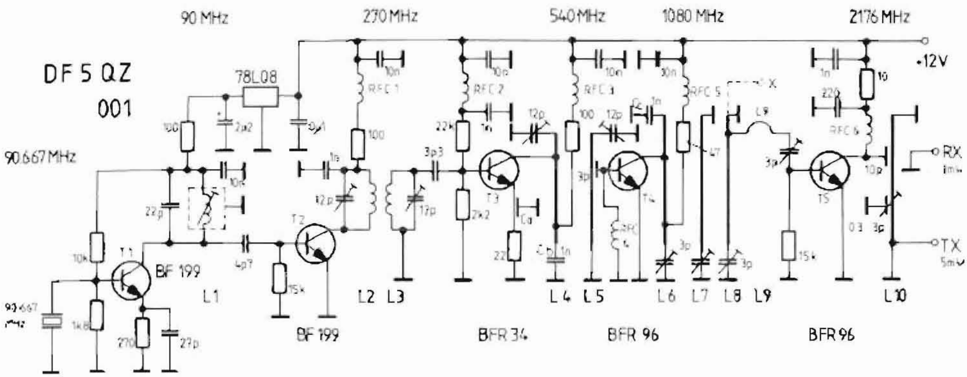

Fig. 2. Local oscillator module DF5QZ 001

The transmit mixer is in a push-pull circuit, which converts up the oscillator signal and the 144 MHz drive signal to the required frequency range of 2320-2322 MHz. Two linear amplifier stages are provided to amplify the required signal by 10 dB before feeding it to the third stage.

DF5QZ 003 is equipped with a further two-stage linear amplifier, which supplies sufficient power for driving the driver module (004). This module is followed by a low-power amplifier (005), that provides an output power of 1 to 1.2 W.

1. LOCAL OSCILLATOR MODULE

The crystal oscillator of this module **(Figure 2)** oscillates at 90.666 MHz. This is multiplied by twelve to provide 1088 MHz (the frequencies given in the circuit diagram are rounded down). This is followed by a printed 3-stage filter, which ensures a clean spectrum. The following frequency doubler stage is driven via the series resonant circuit comprising L 9 and generates the required local oscillator frequency of 2176 MHz. The frequency doubler transistor T 5 operates into a $\lambda/2$ air-spaced stripline circuit (L 10). In order to obtain very high efficiency and good selectivity, a printed inductance was not used here.

The output signal for the transmit mixer was tapped off at the 50 Ω point of the circuit, whereas the coupling for the receive mixer (RX) is achieved via a coupling link in the direct vicinity of the output circuit.

This local oscillator can also be used for the 23 cm band by replacing the crystal with a 96.000 MHz type and leaving out the last stage comprising T 5. The output frequency of 1152 MHz is tapped off at the 50 Ω point of L 8 (which is usually soldered to L 9). The 3-stage filter comprising L 6 – L 8 is highly selective, which means that one is able to obtain a very clean local oscillator signal for 1296 MHz equipment having an IF of 144 MHz.

1.1. Components

L 1:	Coil set blue/brown
L 2:	2 turns of 1 mm dia. silver-plated copper wire wound on a 6 mm former, self-supporting
L 3:	as L 2
L 4 – L 8:	printed on DF5QZ 001
L 9:	Bent wire, link, made from 1 mm dia. silver-plated copper wire, bent around a 6 mm former, 10 mm high
L 10:	Brass-plate strip, 6 mm wide, 35 mm total length, each end bent down by 4 mm

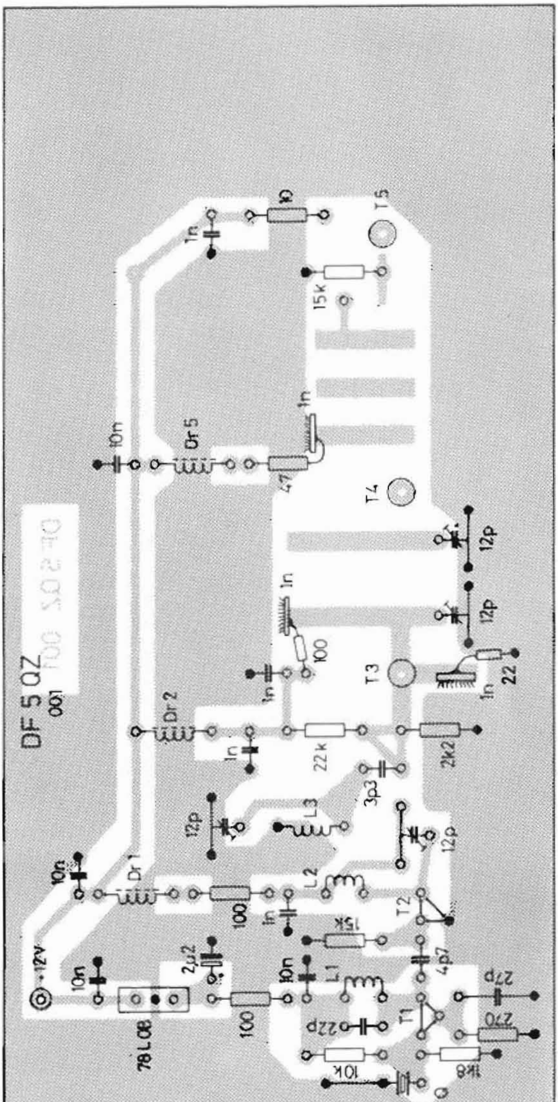

Fig. 3:
PC-board DF5QZ 001 showing component locations

C_a, C_b, C_c: Chip capacitors, approx. 1 nF; press into slots on the board!

Trimmers: 2 – 12 pF: plastic foil trimmer, yellow (Philips)

0.5 – 6 pF: plastic foil trimmer, grey (Philips)

0.3 – 3 pF: ceramic spindle trimmer (Philips)

RFC 1 – RFC 5: 2 turns of approx. 0.4 mm dia enamelled copper wire wound in a ferrite bead

RFC 6: 2 turns of 1 mm dia. silver-plated copper wire wound on a 2.5 mm former, self-supporting

Metal case, 74 x 148 x 50 mm (cover dimensions x height)

1.2 Construction

A double-coated PC-board DF5QZ001 was designed for accommodating the local oscillator module (Figure 3). The dimensions of this board are 146 mm x 72 mm and it can be mounted in the given metal box.

The upper surface is in the form of a ground surface, which is only removed around those connections that are not grounded. Since the ground surface of the base can affect the output power, it is necessary for the cover on the conductor lane side to be soldered all around the edge. The component side of the board does not require a cover.

The board is firstly completed from the component side of the board, and the 3 pF spindle trimmers (horizontal) are soldered from the conductor side. It is important that the ground connections of these trimmers be connected with the ground surface of the component side of the board. The output circuit L 10 is also mounted from the conductor side. The cold ends of L 6, L 7, and L 8 must be "through-contacted" to the component side of the board using a copper strip. Figure 4 shows this critical part of the board.

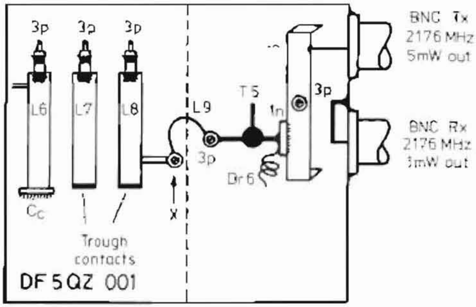

Fig. 4: Construction details regarding the last frequency doubler on DF5QZ 001

1.3 Alignment Details

The frequency is checked after the crystal oscillator has locked in. The subsequent frequency tripler stage is checked with the aid of a dipmeter or frequency counter since it is possible for the second harmonic to be aligned instead of the

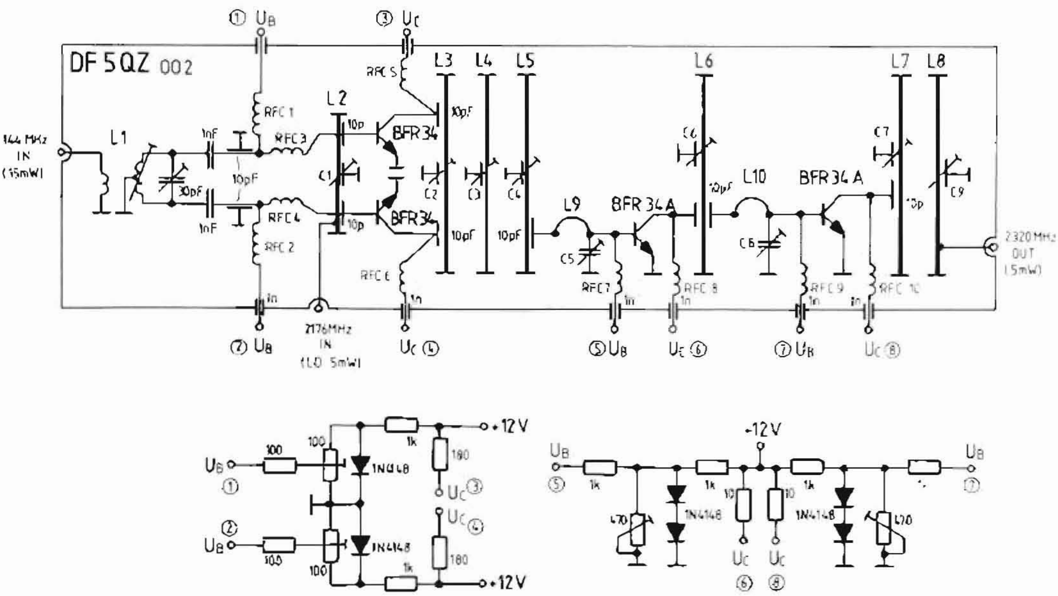

Fig. 5: Push-pull mixer, 2-stage linear amplifier, and bias voltage supply

third. If the required frequency is present at this position, it is possible for the subsequent stages to be aligned for maximum output power. An incorrect alignment is virtually impossible since these circuits have been designed exactly. The alignment of the λ/2 circuit to the final frequency does not present any problems and can be carried out easily.

2. TRANSMIT MIXER

The mixer module comprises two BFR34A as push-pull mixer and two further transistors of this type as selective linear amplifiers (see Figure 5). It is also enclosed in a metal box, and the stages are built up in separate chambers using intermediate panels.

Various attempts to use these SHF standard transistors, type BFR34A, in printed circuits on epoxy glassfibre boards have not been successful. Unfortunately no attempts were made using PTFE board material due to the high costs envolved. It seems, however, that the expected performance is not worth such expense.

The construction recommended here uses λ/2 air-spaced striplines (Figure 6), and it seems that the maximum values of the semi-conductors have been achieved at 2300 MHz. In addition to this, the excellent selectivity of air-spaced stripline circuits together with the chamber construction ensures a clean output signal from this module.

The mixer version recommended here with its specific tuning of the collector circuit to the required frequency corrects many old opinions that a conversion to 2300 MHz with a BFR34A is no longer possible. The problem was not the BFR34A, but the printed construction of the modified DF8QK type. In those cases, due to the undefined selectivity, the 16th harmonic of the 2 m drive signal was radiated after being amplified in the transmitter.

This problem was not present here, and the subsequent filter improves the output signal further so that spurious waves are more than sufficiently suppressed. Furthermore, an important point is the exact matching to the subsequent transistor input with the aid of the series circuit comprising L 9.

2.1 Components

L 1: 5 turns of 1 mm dia. silver-plated copper wire wound on a 5 mm dia. coil former with VHF-core; Input coupling: 2 turns of insulated wire wound symmetrically to the center of L 1

L 2 - L 8: Brass strips, 6 mm wide, length before bending = 35 mm, bent down 4 mm at both ends at an angle of 90°

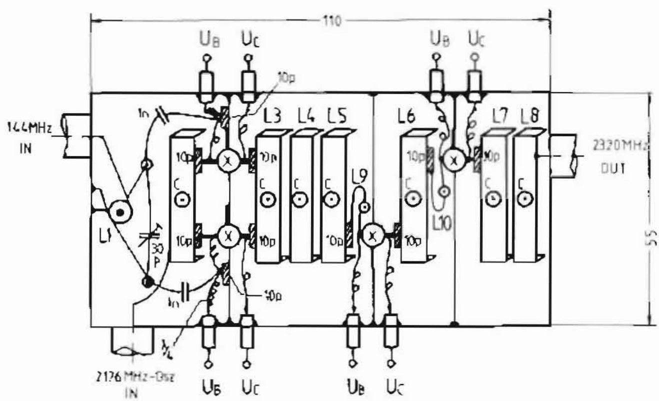

Fig. 6: Construction of module DF5QZ 002 (mixer and linear amplifier)

L 9, L 10. Bent wire from 1 mm dia. silver-plated copper wire, bent over a 6 mm former

RFC 1, RFC 2: λ/4 choke for 145 MHz = 50 cm. of 0.4 mm dia. enamelled copper wire wound on a 4 mm former, self-supporting

RFC 3 - RFC 10 2 turns of approx. 0.4 mm dia. enamelled copper wire wound on a 2.5 mm former, self-supporting

C 1 - C 9: Ceramic miniature spindle trimmer 0.5 - 3 pF (Philips)

Coupling capacitors Ceramic disks of 10 - 3 pF
Feedthrough capacitors: approx. 1 nF
Metal box: 57 x 111 x 30 mm

2.2 Construction Details

Construction is commenced by making the holes for the spindle trimmers in the base of the box as shown in **Figure 6**. After this, the λ/2 circuits are soldered into place with the holes for the trimmers at the center, after which the intermediate panels are also soldered into place. The side panels are provided with the holes for the BNC-connectors and feedthrough capacitors. After assembling the base plate with the side panels, it is possible for the transistors, chokes, disk capacitors, as well as the 144 MHz circuit to be mounted and then soldered into place.

The network for adjusting the mixer balance and the operating points is provided outside of the box on a small Vero boards. This can be srewed below the mixer module afterwards. **Figure 7** shows a photograph of the author's prototype.

2.3 Alignment

Firstly connect the operating voltage and local oscillator signal; measure the voltage drop across the collector resistor of one of the mixer transistors (between +12 V and point 3 or 4). If the output circuit of the oscillator module DF5QZ001 and the input circuit of the mixer are in resonance, the voltage drop will increase clearly. The circuits should be aligned for maximum voltage drop.

This is followed by injecting a 144 MHz drive signal (max. 15 mW!), and measuring the voltage drop across the collector bias resistor of the mixer transistor. This drive will cause a further increase of voltage. This is brought to maximum

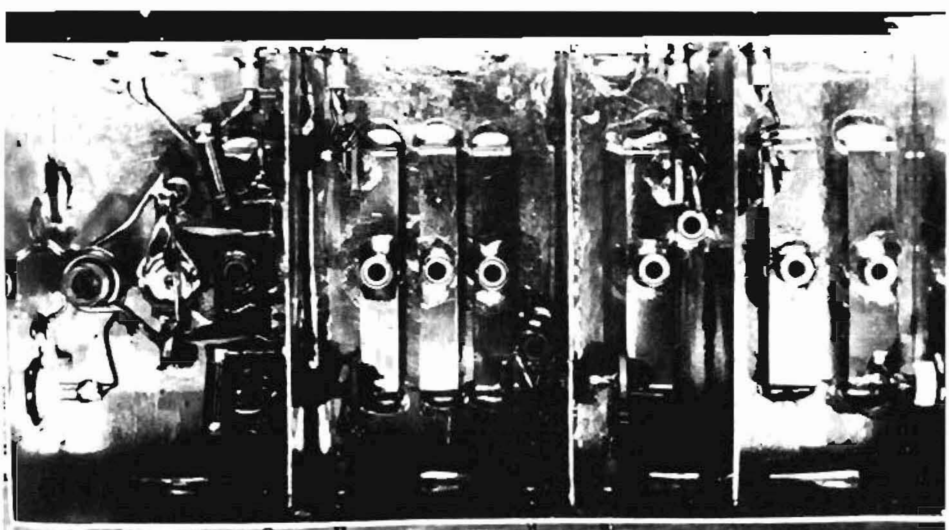

Fig. 7: Photograph of the author's prototype of module DF5QZ 002

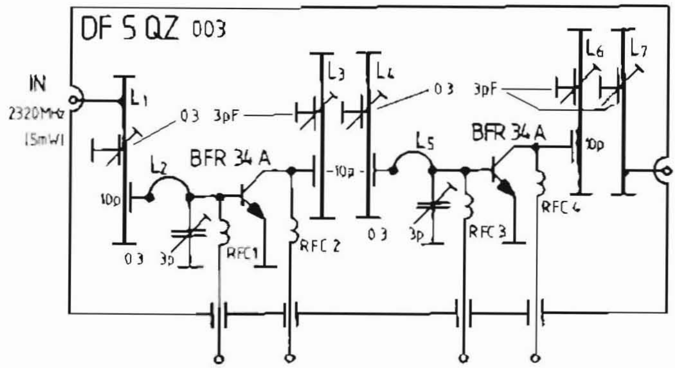

Fig. 8: Two-stage linear amplifier for approx. 60 mW in the 13 cm band

with the aid of the core of inductance L 1 and the circuit capacitance. Finally switch off the 2 m carrier and adjust the quiescent currents of the transistors to approximately 10 mA.

Switch on the 2 m signal again and measure the voltage drop across the collector bias resistor of the first amplifier transistor (point 6). The collector circuit of the push-pull mixer, the filter circuit, and the input network of the first amplifier should now also be aligned for maximum voltage drop

Also align the second amplifier stage. Attention should be paid that one does not align this to the oscillator frequency. All circuits are sharp enough to be aligned cleanly to the required, local oscillator, or image frequency! The rule of thumb is: The required frequency results with the trimmer virtually at minimum capacitance

A sensitive power-meter or indicator suitable for the frequency should be used for optimizing this module. The balance of the mixer is aligned for best oscillator signal suppression with the aid of the wattmeter. The most favorable values can, however, only be obtained with the aid of a spectrum analyzer. The measured spurious suppression amounted to 45 dB, and the output power was 5 mW

The intermediate frequency of 144 MHz used was a disadvantage. As previously mentioned, the 16th harmonic of this band falls into the 13 cm band. This unwanted effect could be avoided easily by using a different IF (such as the 70 cm band). However, if a spectrum analyzer is available, this harmonic of the 2 m signal can be completely suppressed with the aid of the 2 m circuit and the mixer operating points

With respect to the output power, it is possible using this output power to communicate via a transponder at a distance of 35 km (projected by DCØDA and DF5QZ). The input frequency was 2320.13 MHz, the output frequency 432.75 MHz. The author's signal was 45 dB above noise and was able to drive the transponder fully. The mixer operates stabily without any tendency to oscillation

3. LINEAR AMPLIFIER WITH TWO BFR 34 A

This low-power amplifier is equipped with 2 BFR 34A and is designed to be as selective as possi-

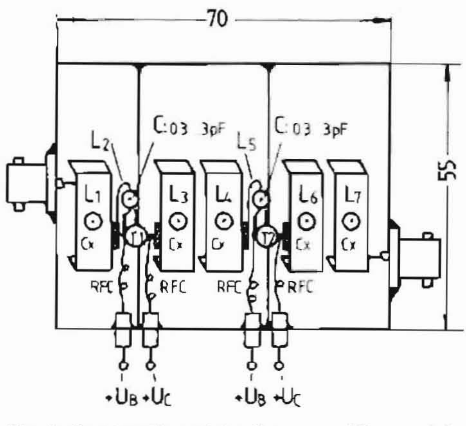

Fig. 9: Construction of the linear amplifier module DF5QZ 003

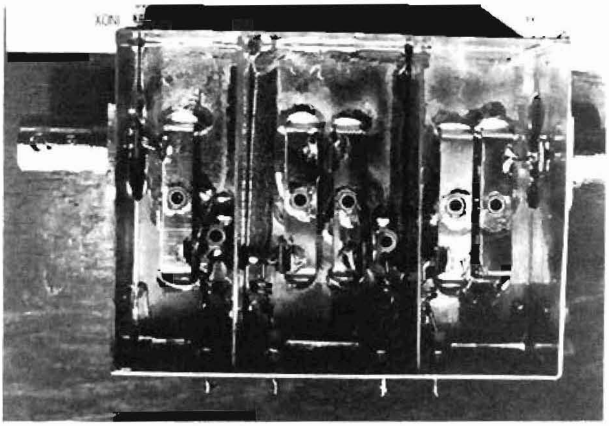

Fig. 10:
Photograph of the author's prototype DF5QZ 003

ble and to simultaneously provide a high, linear gain. These demands can only be satisfied when using chamber-type construction and when using air-spaced striplines. The construction is therefore identical to that of the 2-stage amplifier in the mixer module 002. The SHF-circuit is shown in **Figure 8**.

The bias voltage supply is carried out according to the circuit diagram given in Figure 5. A sketch given in **Figure 9** and the photograph of the author's prototype shown in **Figure 10** give an impression of the construction. The arrangement of the circuits and the chambers is not critical, however, the base connection of the transistors to the trimmers of the associated matching circuit should be kept as short as possible.

3.1. Component Details

L 1, L 3, L 4, L 6, L 7: Brass strips, 6 mm wide, 35 mm total length, bent down 4 mm at both ends (90°)

L 2, L 5: Bent from 1 mm dia. silver-plated copper wire wound on a 3.5 mm former, height 8 mm.

Trimmers (7 pcs): ceramic miniature spindle trimmers, 0.3-3 pF (Philips)

Metal box: 72 × 57 × 30 mm

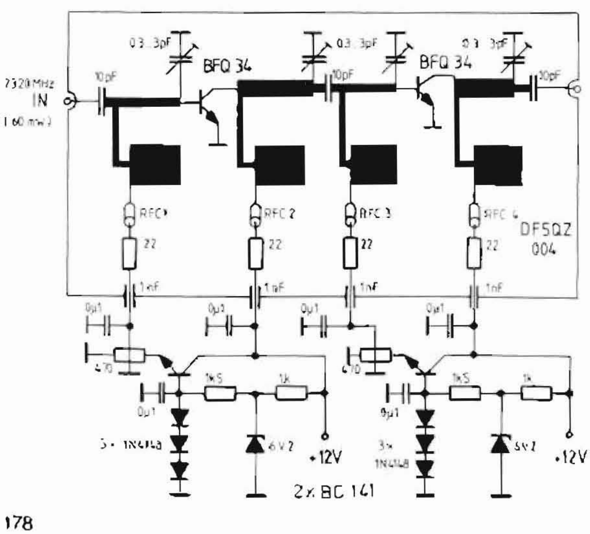

Fig. 11:
Two-stage linear amplifier for approximately 600 mW in the 13 cm band

3.2. Alignment

Firstly, adjust the quiescent currents to approximately 10 mA and connect a power-meter or other indicator to the output. A signal of approximately 5 mW is now connected from the mixer module 002 to the amplifier module 003. The voltage drop across the two collector bias resistors should now be measured again, and the resonant circuits aligned for maximum reading. The wattmeter should finally indicate approximately 60 mW. The gain of this module was measured to be 12 dB, and the 3 dB bandwidth to be 23 MHz.

4. DRIVER STAGE

Unfortunately, there are virtually no cheap power transistors for the frequency range around 2500 MHz. A possible type was given in (4). The layout of the PC-board described there was modified at one position, and then tested together with the transistor types BFQ 68 and BFQ 34. Whereas the BFQ 34 provided positive results, it was found that the BFQ 68 was unsuitable for use on this board. It seems that the S-parameters of this type differed greatly, which meant that matching could not be obtained.

On the other hand, the BFQ 34 provided a gain of 5 to 6 dB at 2.3 GHz, and thus an output power in the order of 0.5 W. It was therefore decided to construct the two-stage driver equipped with two BFQ 34, which provided the stable 10 dB gain as was expected, (see **Figure 11**)

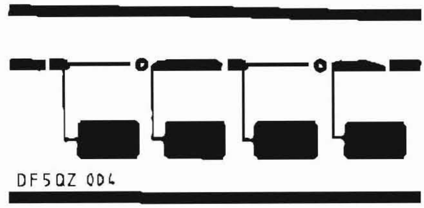

Fig. 12: Single-coated epoxy PC-board for the linear amplifier equipped with two BFQ 34

4.1. Construction

The two-stage driver amplifier for 2300 MHz equipped with two BFQ 34 is accommodated on PC-board DF 5 QZ 004 (see **Figure 12**). This board is made from epoxy glassfibre material (G 10) and its dimensions are 108 mm × 52 mm.

This board can be fitted into a metal box, whose dimensions are 110 × 55 × 30 mm. The two networks for adjusting the quiescent currents are built up on a piece of Vero board outside of the case (see **Figure 13**). The contact of the transistor bolts to the base plate of the case is sufficient

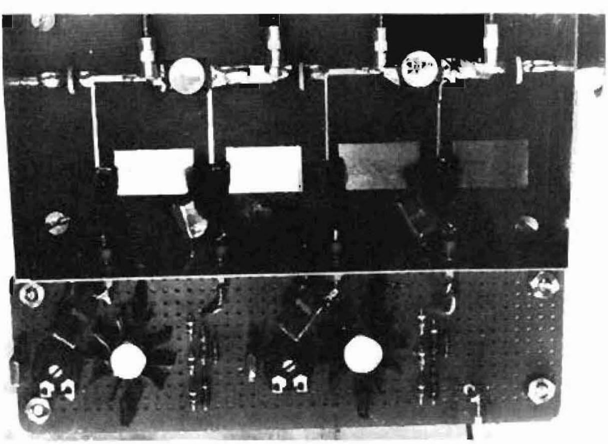

Fig. 13: Photograph of the author's prototype DF5QZ 004

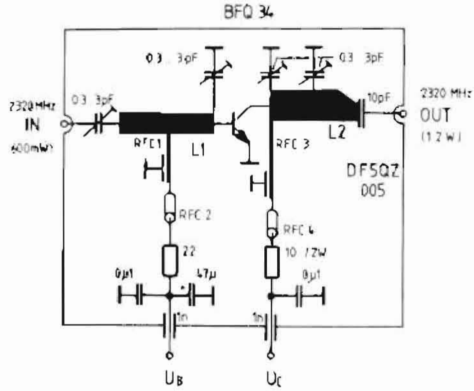

Fig. 14: 1 W power amplifier for the 13 cm band DF5QZ 005

for cooling. The collector and base connections of the transistors should be shortened and cut with the aid of scissors before soldering into place. The four chokes are ferrite beads placed over the connection wires.

4.2. Alignment of the Driver

The quiescent current of both transistors is adjusted to 140 mA. The output should now be connected to a power meter. The input signal is now connected, and the spindle trimmers aligned for maximum output power. With an input power of 60 mW, it is possible to obtain an output power of 600 mW, as was measured in the author's prototype. This allows a tube amplifier such as (2) to be driven to approximately 6 W. Since the construction of such a tube amplifier requires a lot of metal work, an attempt was made to increase the output power further using the relatively inexpensive BFQ 34.

5. POWER AMPLIFIER

The transistor BFQ 34 was examined at higher drive levels with the aid of an adjustable 2320 MHz power source. The result was that the efficiency of this transistor ceases at an output power of 1.2 to 1.5 W. The parallel connection of two such BFQ 34 did not result in any noticeable success.

In the author's opinion, a printed construction is not suitable to obtain the determined 3 dB gain in the power range over 0.5 W. For this reason, the matching links were once again made in air-spaced stripline technology. **Figure 14** shows the circuit diagram, **Figure 15** the construction, and **Figure 16** the photograph of the author's prototype.

The transistor is screwed to the base; a heat sink made out of 5 mm thick aluminium plate having the same dimensions as the case is used for cooling. The hole around the transistor is carefully sawn out with the aid of a fretsaw, so that the base and emitter connections do not touch the base plate.

5.1. Components

L 1: Brass strip 6 mm wide, 15 mm long, spaced 4 mm over the base plate

L 2: Brass strip, 10 mm wide, 25 mm long; collector end bent down 4 mm, other end

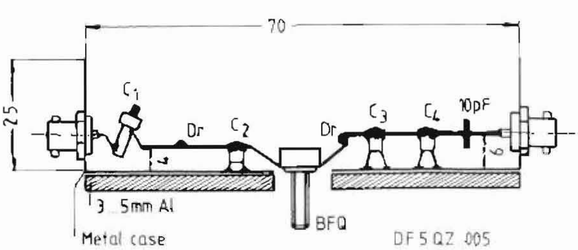

Fig. 15: Construction on the base of a metal case

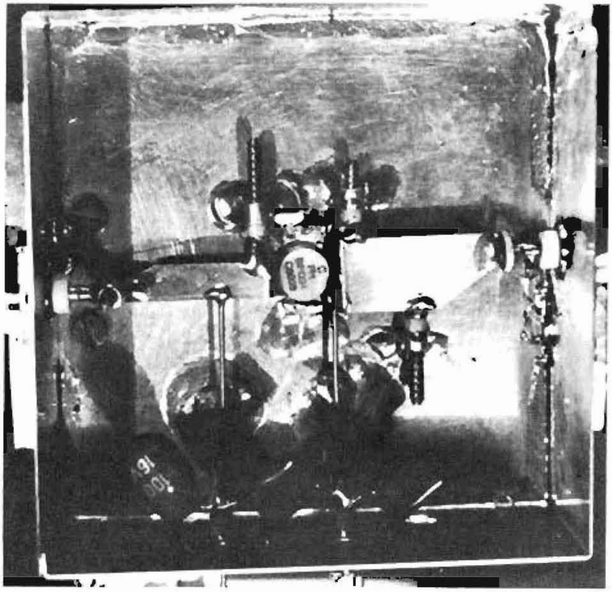

Fig. 16:
Photograph of the author's prototype 1 W power amplifier DF5QZ 005

should be cut as shown in Figures 14 and 16; mounted 6 mm over the base plate.

RFC 1: 1 mm dia silver-plated copper wire, 10 mm long

RFC 2, RFC 4: Ferrite bead

RFC 3: 2 mm dia silver-plated copper wire, 10 mm long

Trimmer: 4 ceramic miniature spindle trimmers, 0.3 – 3 pF (Philips)
C 1 mounted between the connector and L 1 self-supporting; C 2 – C 4: for horizontal mounting.

Metall case: 74 × 74 × 30 mm
Base and collector voltage supply: As in the case of driver DF5QZ 004.

5.2. Alignment of the Amplifier

Connect a power meter to the output, and align a quiescent current to 140 mA. Connect the signal at the input and align for maximum output. In the author's prototype, 1 to 1.3 W were obtained with a drive level of 500 to 600 mW.

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G. Schwarzbeck, DL 1 BU

Antenna Polarisation for OSCAR 10

AMSAT have clearly stated that clockwise circular polarisation will be required for communication via OSCAR 10. However, this was not completely accepted by many users in the first weeks of practical operation via this satellite. Such opinions were due to monitoring the satellite incorrectly or at the wrong time: It will be observed that anticlockwise polarisation can provide a stronger signal when the satellite is just over the horizon (Fig. 1, Fig. 3), however, these are not the most favourable times for working via OSCAR 10.

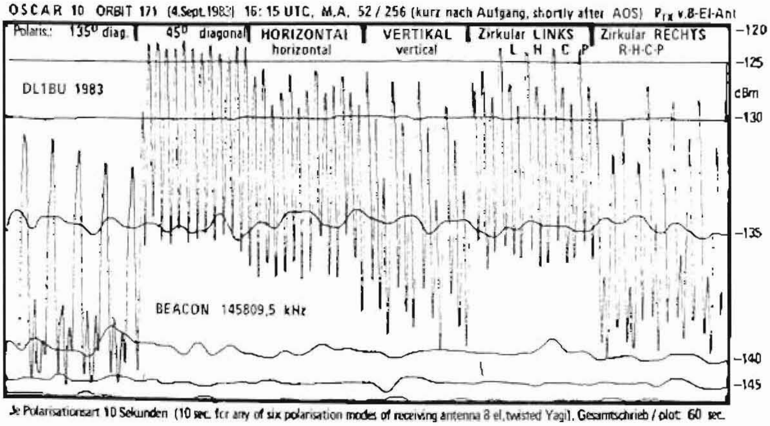
The most favourable time for operation via OSCAR 10 is when it is at its maximum altitude (Apogee), and it will be seen from the measured fieldstrengths that clockwise circular polarisation is far more favourable, both with respect to fieldstrength and fading (Fig. 2).

To prove this, the author plotted the receive fieldstrength on a precision measuring receiver connected to a XY-plotter. The 2 m beacon was received using a 8 element crossed Yagi antenna in conjunction with a polarisation switching unit as offered by UKW-Technik, and described in (1). Measurements were made just after the satellite appeared over the horizon, at maximum altitude, and just before dropping below the horizon. During a measuring period of 60 seconds, each of the following polarisations were selected for 10 s

- Diagonal 135°
- Diagonal 45°
- Vertical
- Horizontal
- Anticlockwise circular
- Clockwise circular

It will be seen clearly that clockwise circular

OSCAR 10 ORBIT 171 (4.Sept.1983) 16: 15 UTC, M.A. 52 / 256 (kurz nach Aufgang, shortly after AOS) P_{rx} v.8-El-Ant



Je Polarisationsart 10 Sekunden (10 sec. for any of six polarisation modes of receiving antenna 8 el. twisted Yagi). Gesamtschrieb / plot: 60 sec.

polarisation is superior at the maximum altitude of approximately 35000 km. The received field-strength was an average of −127 dBm, which amounts to 18 dB above the noise of a GaAs-FET preamplifier. The depth of fading only amounted to ±1 to 1.5 dB.

One can also see from the number of fading periods (spin of the satellite times 3) that one of the antennas is defective. You will notice that the depth of the fading is far more pronounced with linear polarisation than with circular (except at the horizon, where we are probably looking side on to the antenna, ed.).

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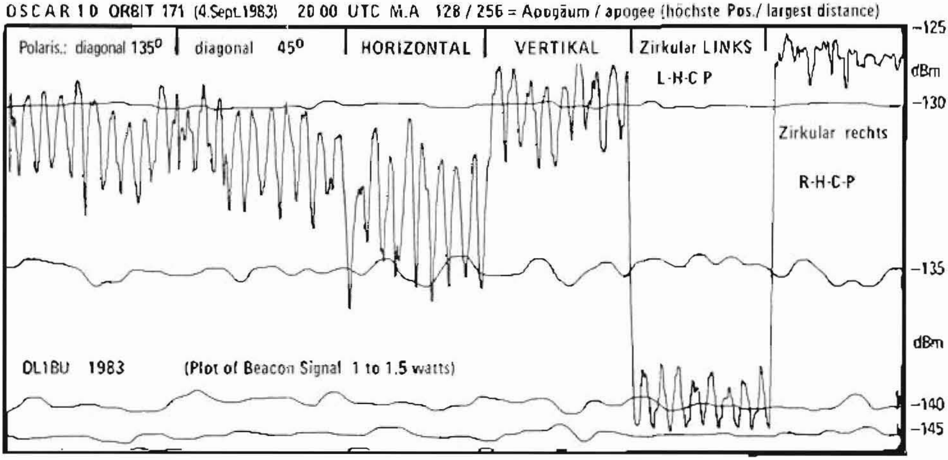
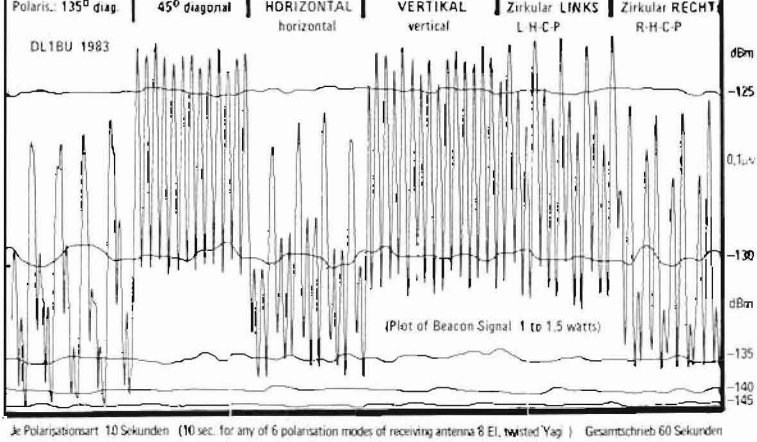
OSCAR 10 ORBIT 171 (4.Sept.1983) 20 00 UTC M.A. 128 / 256 = Apogäum / apogee (höchste Pos./ largest distance)



Je Polarisationsart 10 Sekunden (10 sec. for any of 6 polarisation modes of receiving antenna 2 x 8 El.) Gesamtschrieb / total plot 60 sec.

OSCAR 10 ORBIT 171 (4.Sept.1983) 00:00 UTC, M.A. 216/256 (kurz vor Untergang / almost LOS)



Je Polarisationsart 10 Sekunden (10 sec. for any of 6 polarisation modes of receiving antenna 8 El. twisted Yagi) Gesamtschrieb 60 Sekunden

Hans-J. Griem, DJ 1 SL

A Helical Antenna for the 23 cm Band

Although OSCAR 10 was only considered secondarily during the conception of this antenna, this article describes just the antenna required for operating via the 23 cm/70 cm transponder.

1. INDRODUCTION

The main reason for selecting a helical antenna were the built-in wide bandwidth, and the circular polarization The wide bandwidth of the antenna means that it is not critical to construct, since the resonant frequency is not dependent on fractions of a millimeter! Circular polarization allows one to operate with stations using virtually any polarization

One limit of this antenna is that the direction of the helical winding determines the polarization of the transmit and receive signals and cannot be switched as in the case of a crossed Yagi with two separate feeders (1). Furthermore, the gain of the antenna when receiving a linearly polarized wave can be 3 dB lower than would be the case when receiving the correct circular polarization (2).

These disadvantages were acceptable since the author's geographical location in a shallow valley did not allow long-distance tropospheric communications, and since he required a universal, simple directional antenna for communication on the 23 cm band A recent application for this antenna is for operation via the „L-Transponder" of OSCAR 10 Remember, a clockwise circular-polarized antenna will provide 3 dB more gain in conjunction with OSCAR 10 than a linear-polarized antenna of the same gain!

2. FUNDAMENTALS

There are a large number of publications regarding the mechanical construction of a helical antenna, such as (2) and (3) Three main characteristics were important during the conception of the antenna:

1. Good matching to the feeder cable over the widest possible bandwidth
2. Low sidelobes in the polar diagram
3. Best possible circular polarization

All three demands are required to obtain the highest antenna gain with given dimensions, in other words, to get as near as possible to the maximum theoretical values The author studied all available literature in order to determine the state-of-the-art.

2.1. Matching

Publications (2) and (3) described a matching network for transforming the characteristic impedance of the helical winding (approximately 140 Ω) to the required feeder impedance of 50 Ω using a λ/4 transformer. In (3) the matching obtained in the 70 cm band (2.3% bandwidth) was given as 1.4 VSWR In (4) a type of L/C-matching was described, however, this was only aligned for a descrete frequency In (5) and (6), a stripline circuit with continuously varying impedance (exponential line) was used for transformation where a VSWR of 1.5 and a bandwidth of 25% were obtained in (5), whereas 53% was obtained in (6). In (7), a wideband matching was achieved by using a cone around the commencement of the helical winding instead of a flat plate reflector. This antenna was called "Heli-Cone". The transformation can be influenced within certain limits by varying the angle of the cone.

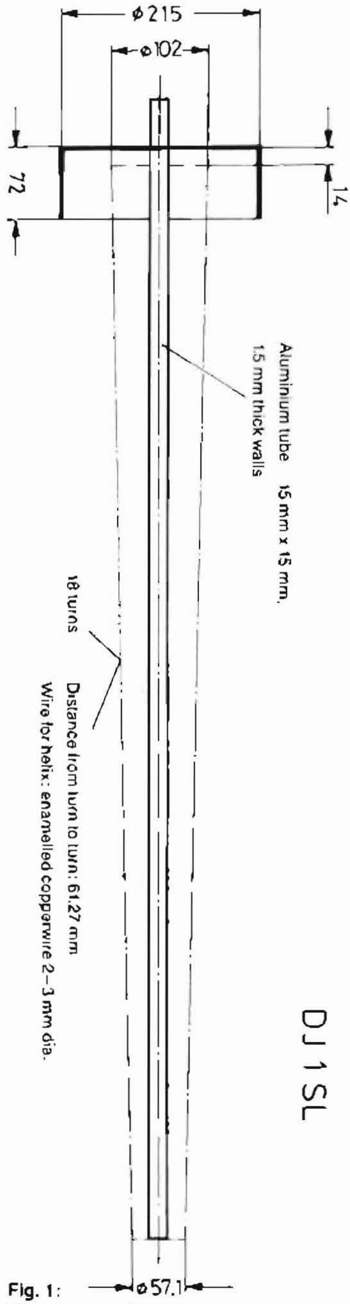


Fig. 1: Dimensions of a helix antenna comprising 18 turns, for 1240–1300 MHz

2.2. Radiation Diagramm

Sidelobes reduce the antenna gain and allow signals to be received from unwanted directions. These unwanted signals can be ignition interference (yes, they can still be received on this band!) or terrestial noise. The latter is important when using the antenna in conjunction with very low-noise receive systems for space communications (EME, satellites) Sidelobes and front-to-back ratio can be improved up to 30 dB by increasing the size of the reflector and forming it into a cone, however, the design of the helical winding can also influence the sidelobes (6), (7).

2.3. Circular Polarization

If a circular-polarized antenna also generates a left-hand circular component or receives it, this represents lost gain, unless the other station has an antenna with the same characteristics. A measure of the polarization error is the axial ratio of the elliptical polarization. This is a true circle with pure circular polarization According to the axial relationship and position of the elliptical polarization of two antennas to another, the reduction in gain due to non-circularity can be between a few hundredth of a dB up to several dB (9), (10).

As can be seen from (6) and (7), the axial ratio can be improved considerably by using a conical shaping of the commencement and end of the helical winding. It seems that this reduces the reflection at the end of the antenna considerably which means that the transition of the UHF-signal from the antenna to space is made without discontinuity.

2.4. Characteristics of the Antenna

As a result of these experiments, an antenna was designed according to (6) with the characteristics listed in **table 1**.

The main characteristic is the conical form of the helical winding. **Figure 1** shows the construction of the antenna in the form of a diagram, and the helical winding is only indicated. **Figures 2, 3, and 4** show the completed antenna as a rear view and from the front of the antenna towards the reflector

Table 1:
Helix antenna
for the 23 cm band

Frequency range	1240 - 1300 MHz
Gain:	15.8 dBi, clockwise circular (RHC)
Sidelobe suppression	- 20 dB
Circularity:	0.6 dB ≙ 1.07
VSWR:	1.2
Overall length	1117 mm
Helical length.	1103 mm
Helical diameter	
at feedpoint:	102 mm
at the front end:	57.1 mm
Helical winding:	3 mm dia. enamelled copper wire
Turns spacing	61.27 mm (calculated value)
Reflector diameter	215 mm
Reflector collar:	72 mm
Boom dimensions (mm)	15 x 15 x 1.5 aluminium

3. CONSTRUCTION DETAILS

The helical winding is held every half a turn using an **insulating support**. This is made by cutting the required length of dielectric insulation from an RG-213/U coaxial cable. A total of 37 pieces are required whose length should be approximately 5 mm greater than the helical radius at the associated position. Of course, half the diameter of the boom should be deducted. These supports are cut and are provided with holes at the correct position which are drilled with the diameter of the helical winding. All supports should now be placed onto the helical winding in the correct sequence, and distributed in approximately the correct spacing of 163 to 95 mm, which decreases approximately linearly.

Enamelled copper wire is used as **helical element**, since it can be bent easily and is protected against corrosion. The diameter is not critical, and 2 to 3 mm diameter was found to be suitable. Approximately 5 m are required.

It is advisable to manufacture **the boom** from a tube with a square cross-section, such as 15 mm x 15 mm aluminium, with a wall thickness of 1.5 mm. The length should be at least 1400 mm, in order to ensure that a mast clamp can be provided behind the reflector. The author reinforced

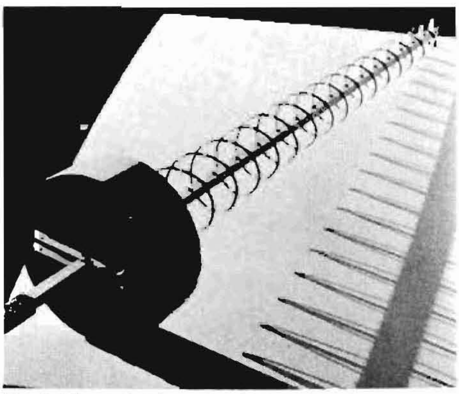


Fig. 2: The author's prototype antenna

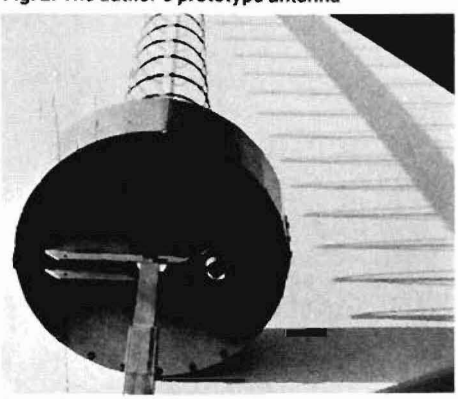


Fig. 3: View onto the reflector

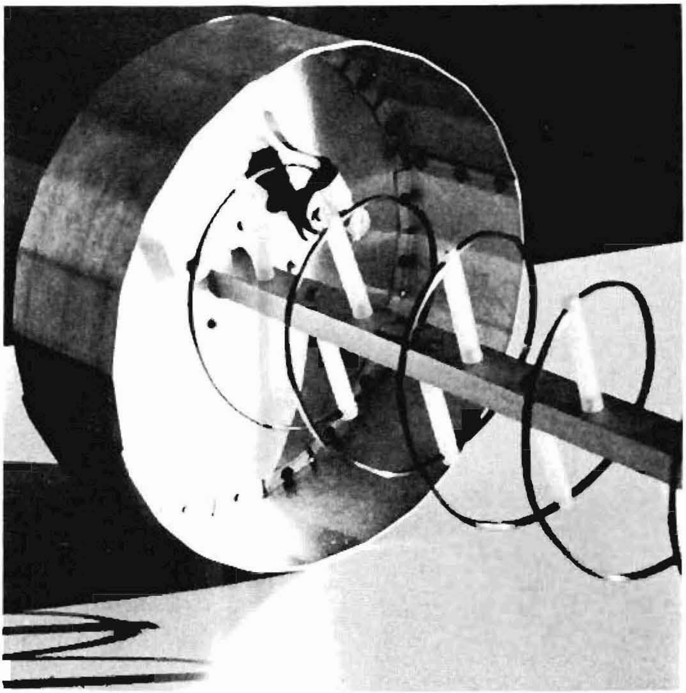


Fig. 4: Matching device inside the reflector

the boom using an additional tube of 20 mm x 20 mm which was glued to the boom with the aid of dual-component glue. Starting from the front of the boom, 37 holes of 3 mm diameter are required for mounting the supports with the aid of brass screws M3 x 20 The spacing is equal to half the spacing of the winding, which amounts to 30.635 mm (calculated value)

The reflector is made from aluminium plate. The circular shape of the collar is approximated using 22 cuts, in order to obtain straight mounting tabs which are then screwed to the reflector disk. The coaxial connector is mounted in the reflector with a spacing of 51 mm from the center. The author used a 50 Ω type 4.1/9.5 manufactured by Spinner (BN 982300), which possesses a mounting flange and a cable connection, onto which the helical winding was later soldered.

The whole reflector plate is mounted to the boom with the aid of two aluminium brackets so that the commencement of the helical winding (longest support) is shifted by 90° with respect to the coaxial socket. The intermediate piece of helical winding is used later for matching The reflector possesses a round hole with a diameter of 30 mm through which the boom is placed. As can be seen in the photographs, it is not necessary to close this gap

The helical winding is mounted onto the boom as follows: The boom is mounted in a vice without reflector and support. The helical winding is laid out so that the shortest support is in the vicinity of the front of the antenna This support is now screwed to the boom; the brass screw can be screwed in tightly into the hole where the inner

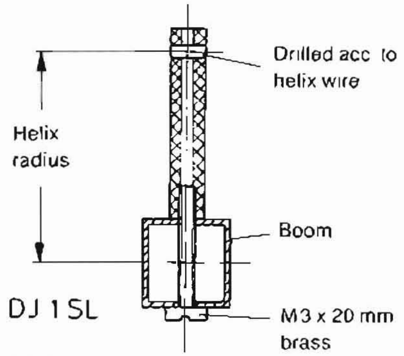


Fig. 5:
One of the 37 insulating supports made from the dielectric of RG213/U

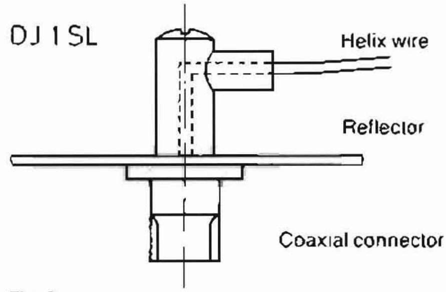


Fig. 6:
Antenna connector and helix wire, without matching device

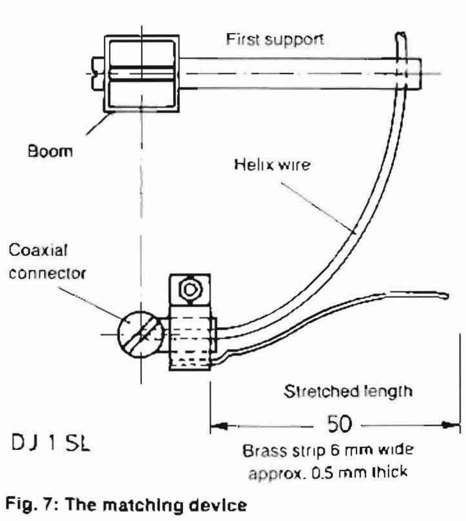


Fig. 7: The matching device

conductor was originally supported (Fig. 5). In order to ensure that the helical winding cannot jump out, bend back the end around the support! Open the vice and turn the boom by 180° and tighten the vice again. Pay attention to the direction of winding: The helical winding should form a clockwise screw. Form the helical winding at the front of the antenna in a semi-circular shape and screw in the next support into place, and so on until the 37th support has been mounted.

Inaccuracies of the helical winding should be compensated for during mounting and after completion by bending and shifting the wire to the various supports. One is best able to see deviations from the circular form by observing the antenna from the front. Finally, the helical winding is fixed to each support by providing a glue into the hole provided by the original inner conductor.

This type of construction ensures that no metal pieces interfere with the operation of the helical winding.

After this, the reflector is mounted so that it is spaced 14 mm from the center of the longest helical support. The end of the helical winding is then soldered to the inner conductor of the coaxial connector. In the author's case, this resulted in an arrangement as shown in **Figure 6**.

The transformation of the characteristic impedance of the helical winding to 50 Ω was made with the aid of an exponential line, which is shown in **Figure 7**. The shape of the brass strip (that is grounded to the body of the coaxial connector) was found experimentally by observing the matching in the frequency range of interest with the aid of a directional coupler, and by bending the strip correspondingly until the return loss was better than 20 dB, which corresponds to a VSWR $\leq$ 1.2.

It is important that the strip is near to the helical winding at the ground end (approx. 2 mm), in order to ensure that the local impedance distribution is in the vicinity of 50 Ω. The matching experiments were made, by the way, with strip lengths of 32 mm to 60 mm, however, 50 mm has been found to be optimum, and corresponds to approximately $\lambda/4$ at a wavelength of 23 cm.

3.1. Mast Mounting

It is advisable to reinforce the boom using an extra tube of 20 mm x 20 mm in order to be able to use conventional mast clamps. The antenna is mounted so that the commencement of the helical winding at the coaxial connector faces downwards, which means that no rain can enter the connector. This means that the helical supports are then vertical, which ensures a minimum deposit of dust and snow. Furthermore, birds are not able to use them as supports.

RG 214/U was used as feeder cable, since it was found to have far less reflection discontinuities when measuring the impedance run up to 1300 MHz, than similar lengths of RG-213/U (obtained from various suppliers). The measurement was made in the same way as measuring the matching of the antenna, however, the test object was a ring of cable (20 m or more) together with a 50 Ω terminating resistor at the far end

4. OPERATING EXPERIENCE

The characteristics of the antenna could only be judged subjectively, in other words, the measuring equipment and a measuring range were not available for gain measurements The good side-lobe suppression was indicated when receiving the 23 cm beacon DBØGP at Goppingen (42 km, non-line-of-sight) with approximately 12 dB over noise, whereas it could hardly be heard when the antenna was rotated to both sides. CW-signals can be heard to far below the noise level.

In experiments carried out locally in conjunction with DL 7 IX it was found that the helical antenna showed hardly any field strength variations when DL 7 IX rotated his indoor transmit antenna, whereas a very strong QSB was exhibited when using the author's vertical, omni-directional antenna. The result is that a far more homogeneous signal strength is obtained than when using a linearly polarized antenna

5. REFERENCES

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Further Notes On:

The Digital Storage Module For Weather Satellite Images YU3UMV 001/002

In the meantime, several hundreds of these digital storage modules are in operation and it has been seen that the 64-k-storage ICs are not so critical as was assumed. It is possible to use (good) sockets if "pull-up resistors" are used on the address multiplex lines. As can be seen in **Figure 1**, a TTL-high line (center) can be modulated by its neighbouring lines (upper and lower in Figure 1) so that an incorrect pulse appears when a high-to-low transition takes place on these lines. This cross-talk is possible with clock or address lines. This can be seen most easily when no video signal is present. Instead of having a continuous black area, the monitor will show a geometric pattern of brighter points. Pull-up resistors as shown in **Figure 2** provide a reliable aid. The storage block is to the right-hand side of Figure 2.

YU 3 UMV

Dietmar Gralz of Oberursel has sent us modification instructions for the digital storage module which are too extensive to be brought here. They describe circuit modifications that allow the readout of the storage to be made in the correct direction, even though it may not correspond to the direction of the satellite path.

This is possible by changing both 74LS161 (I 213, I 214) for 74LS169. This IC is an up/down counter with which all connections are compatible with the exception of pin 1 (U/D). However, pins 7 + 10 of I 213, as well as pins 7 + 15 of I 214 must be connected in an inverted manner. The required inverters are taken from the parallel circuits of I 217.

Unfortunately, this requires more extensive modifications on the board than may be assumed from this description. However, the operating comfort provided by this compensates for the effort made. If there is sufficient interest from readers, the editors are willing to either bring this modification in the magazine, or to translate it and make it available.

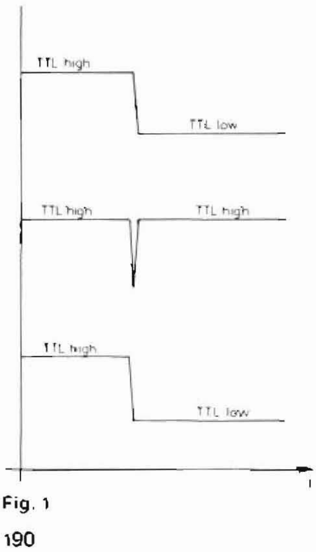


Fig. 1

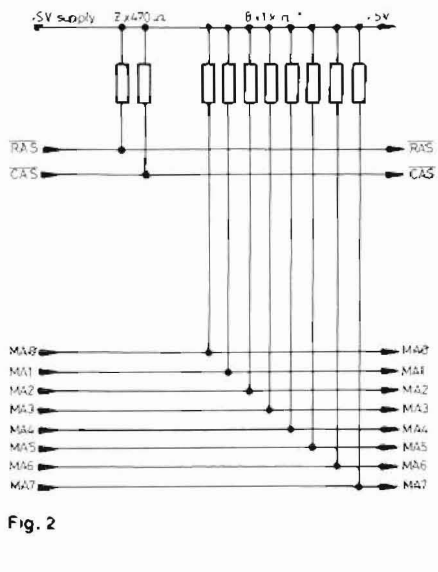


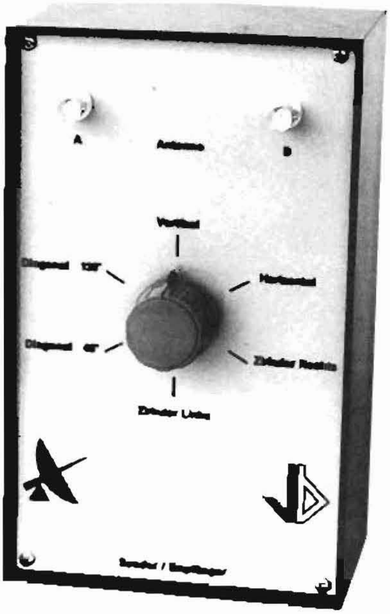
Fig. 2

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 3/83 of VHF COMMUNICATIONS

Power amplifiers for the 2 m band with V-MOS-FETs			Art No.	Ed. 3/1983
1 W stage				
PC-board	DL 1 GBH 001	single coated, without comp. loc plan	6787	DM 14,50
Parts	DL 1 GBH/1 W	1 V-MOS-FET DV 2805, 1 Z-diode 15 V, 3 mica trimmer caps., 2 mica caps 1500 pF, 2 feedthrough caps, 1 m of silver-plated wire 1 mm ∅, 1 alumin diecast case, 2 BNC conn., 1 10-turn pot. 50 kΩ, 6 resistors	6789	DM 132,—
Kit	**DL 1 GBH/1 W, complete with above parts**		**6790**	**DM 145,—**
10 W stage				
PC-board	DL 1 GBH 001	single coated, without comp. loc plan	6787	DM 14,50
PC-board	DL 1 GBH 002	single coated, without comp. loc. plan	6788	DM 12,—
Parts	DL 1 GBH/10 W	1 V-MOS-FET DV 2810, 1 Z-diode 15 V, 2 mica trimmer caps., 2 mica caps 1500 pF, 1 ceram. multilayer 1500 pF, 2 feed-through caps., 1 m of silver-plated wire 1 mm ∅, 1 al. diecast case, 2 BNC conn., 1 10-turn pot. 50 kΩ, 1 relaise RS-12V, 1 electrol. cap. 1000 µF/40 V, 6 resistors	6791	DM 172,—
Kit	**DL 1 GBH/10 W, complete with above parts**		**6792**	**DM 196,—**
100 W stage				
PC-board	DL 1 GBH 001	single coated, without comp. loc. plan	6787	DM 14,50
PC-board	DL 1 GBH 002	single coated, without comp. loc. plan	6788	DM 12,—
Parts	DL 1 GBH/100 W	1 V-MOS-FET DV 2880, 1 Z-diode 15 V, 4 mica trimmer caps., 1 ceram multilayer cap 1500 pF, 1 m of silver-plated wire 1 mm and 2 mm ∅, 1 BNC and 1 Mini-flange N conn., 1 al. diecast case, 1 10-turn-pot 50 kΩ, 1 relaise RS-12V, 1 Alumin. electrol. cap. 1000 µF/40 V, 4 resistors	6793	DM 618,—
Kit	**DL 1 GBH/100 W, complete with above parts**		**6794**	**DM 639,—**
DC 6 HL	**Variable crystal oscillator (VCXO) for the 2 m band**			**Ed. 3/1983**
PC-board	DC 6 HL 012	double-coated, with through-contacts	6784	DM 31,—
Parts	DC 6 HL 012	2 transistors, 1 special regulator, 1 Z-diode, 2 varactors, 2 Schottky and 1 switching diode, 4 coil kits, 3 coils already wound, 1 m ins. wire 0.2 mm ∅, 1 choke, 1 metal case, 4 feedthrough caps., 3 tantalum and 19 ceram. caps., 11 resistors	6785	DM 99,—
Crystal	22.550 MHz	HC-43/U	6795	DM 41,—
Kit	**DC 6 HL 012 complete with above parts**		**6786**	**DM 166,—**
DJ 7 VY	**Directional coupler bridge for VSWR measurements**			**Ed. 3/1983**
PC-board	DJ 7 VY 006	double coated, silver plated	6775	DM 17,—
Parts	DJ 7 VY 006	1 tinned-metal case, 2 N flange-connectors, male and female, 2 BNC connectors, 8 metal-film resistors 100 Ω/1%, 4.5 mm long, 1.5 mm dia., 28 ferrite beads 5 mm long, 20 cm of 50 Ω PTFE cable type SM 50		
Kit	**DJ 7 VY 006 complete with above parts**		**6776**	**DM 95,—**

FOR OSCAR 10 AND NORMAL COMMUNICATIONS



Polarisations Switching Unit for 2 m Crossed Yagis

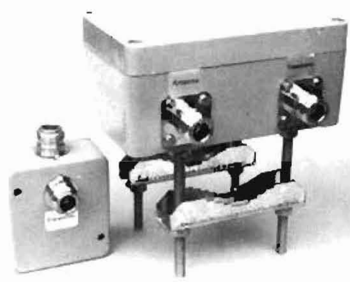
Ready-to-operate as described in VHF COMMUNICATIONS. Complete in cabinet with three BNC connectors. Especially designed for use with crossed yagis mounted as an "X", and fed with equal-length feeders. Following six polarisations can be selected: Vertical, horizontal, clockwise circular, anticlockwise circular, slant 45° and slant 135°.

VSWR	max 1.2
Power.	100 W carrier
Insertion loss:	0.1 to 0.3 dB
Phase error	approx. 1°
Dimensions:	216 x 132 x 80 mm

Low-Noise GaAs-FET Masthead Amplifiers for 144 MHz and 432 MHz MV 144 and MV 432

Selective High-Power Masthead Amplifiers in Waterproof cast-aluminium case with mast brackets. Built-in relay for transmit-receive switching. PTT via coaxial cable using supplied RF/DC-splitter.

- Noise figures: MV 144: 0.9 dB/0.6 dB
  (typ. values) MV 432: 1.0 dB/0.8 dB
- Overall gain. MV 144. 15/20 dB, switchable
  MV 432 . 15 dB
- Insertion loss, transmit: typ. 0.3 dB
- Maximum transmit power:
  MV 144: 800 W SSB, 400 CW/FM
  MV 432: 500 W SSB, 250 CW/FM
- Operating voltage. 12 V via coaxial cable



- Connections:
  N-Connectors
- Dimensions:
  125 x 80 x 28 mm
  (without brackets)

Further details on request.

UKW technik Terry D. Bittan Jahnstr. 14 Postfach 80 · D-8523 Baiersdorf
Tel.: West Germany 9133 / 855 · For representatives see cover page 2