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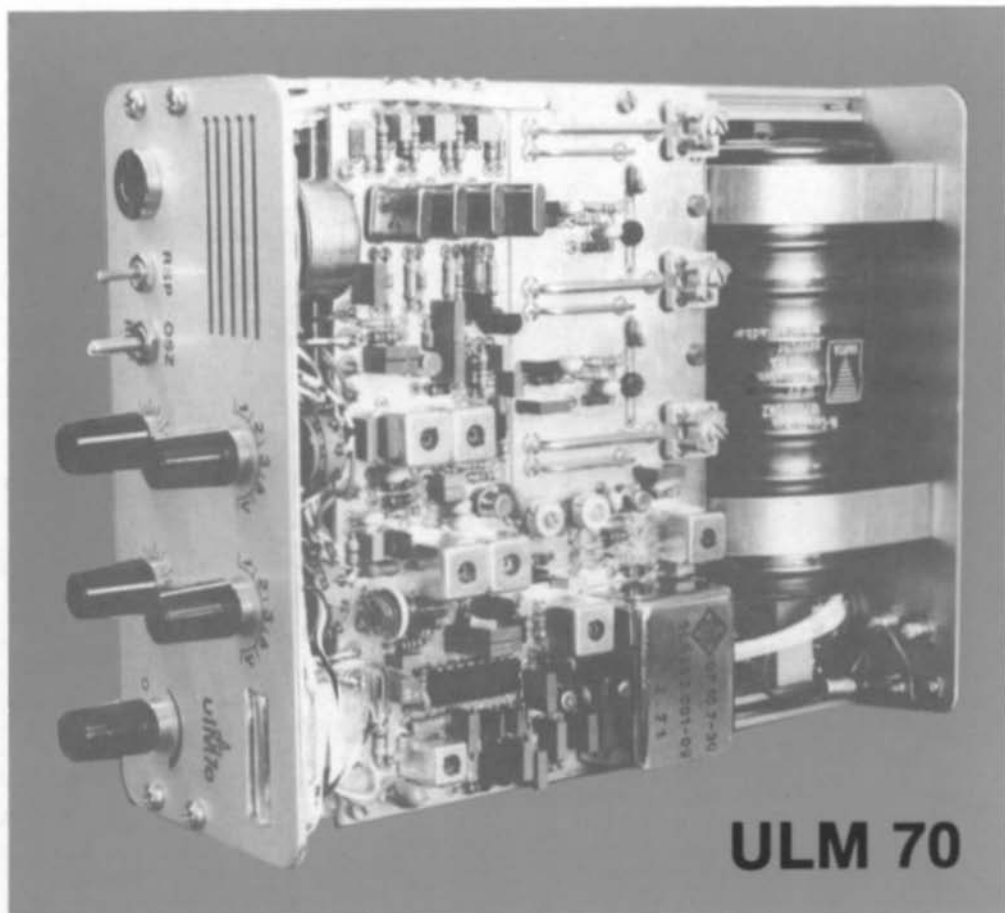
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SUBSCRIPTION RENEWAL 1978

Another year draws to a close, and also another volume of VHF COMMUNICATIONS. We hope, that you have enjoyed reading the magazine, and have found it informative.

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The publishers DJ 3 QC and G 3 JVQ / DJ 0 BQ and their representatives would like to take this opportunity of wishing you a very prosperous, Happy New Year 1978.

THE 70 cm FM TRANSCEIVER »ULM 70«

Part 3: The Transmitter

by I. Sangmeister, DJ 7 OH · H. Bentivoglio, DJ 0 FW · H.J. Franke, DK 1 PN

6. THE TRANSMITTER CIRCUIT

Figure 17 shows a photograph of the completed transmit board. Since the transmitter possesses its own oscillator chain, it can also be used on its own as in the case of the receiver. The modulator is included on the board. The circuit details are now to be discussed in conjunction with the overall circuit diagram of the transmitter given in **Figure 18**. This is followed by construction details and the alignment of the transmitter module. The construction of the complete transceiver is to follow in part 4.

6.1. Oscillator and Multiplier

The switchable VXO and the three subsequent frequency multiplier stages are built up in a similar manner to the oscillator of the receiver. The circuit is described in detail in section 2.1. in edition 3/77 of VHF COMMUNICATIONS.

In the case of the transmitter, it is not only necessary for the crystal frequency to be pulled over four channels, but also to be frequency-modulated. This is done by feeding the modulating voltage via the adjustment potentiometer R 11 to varactor diodes D 5 to D 8.

The gain of the last three stages of the transmitter (straight-through amplifier) is adjusted so that the last tripler transistor T 4 should provide approximately 5 mW. For this reason, some resistor and capacitance values are somewhat different than in the receiver, and a more efficient transistor has been used for T 4 (BFY 90).

6.2. Driver and Power Amplifier Stages

The power amplifier (T 7) is designed to provide an RF output power of approximately 1 W. Since the output from the frequency multiplier chain is approximately 5 mW and since approx. 8 dB gain can be obtained per amplifier stage, this means that two driver stages will be required. The UHF stripline transistor type 41008 is used in all three stages. The first driver transistor T 5 will operate with a quiescent current of approx. 30 mA in class A; the second driver transistor T 6 is aligned for class B, and the power amplifier operates in class C. In order to align T 6 exactly for class B operation, it is necessary for the value of resistor R 27 to be found experimentally.

The type of matching between the first and second driver stage should also be mentioned: a multiple transformation is made using also a parallel resonant circuit comprising L 7 and C 37. Unwanted frequencies generated in the oscillator chain are suppressed well at this position. The tuning is really sharp and must be made carefully.

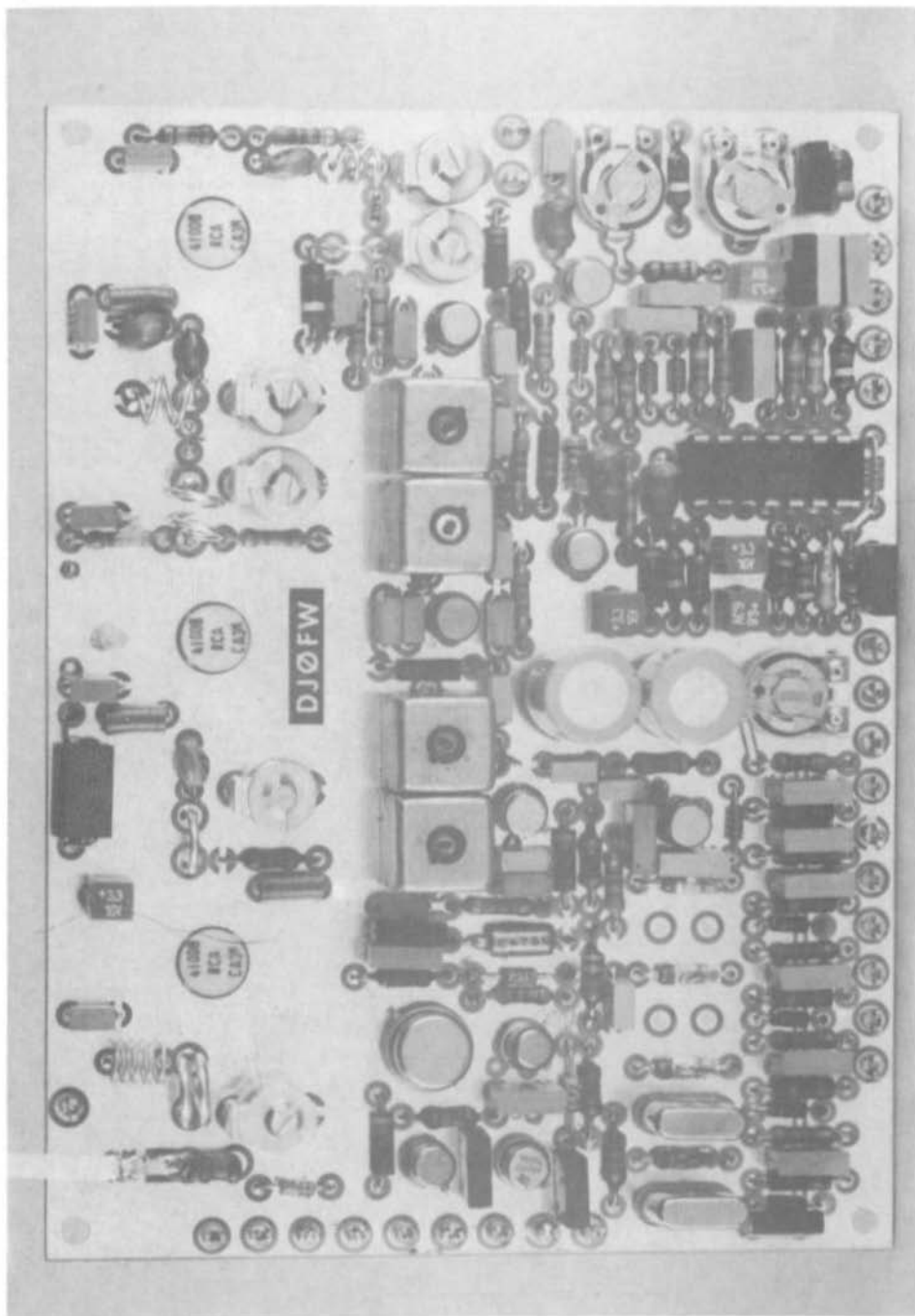


Fig. 17: The completed transmit board

6.3. Universal AF Amplifier

The AF amplifier is the only module of the transceiver that is used both for transmitter and receiver. Since this amplifier is to be found on the transmit board, it is necessary for the receiver to be connected to a small AF amplifier if the receiver is to be used on its own.

Although the AF amplifier only consists of an integrated circuit, two small transistors and a few cheap components, it is able to fulfill all audio tasks required in a transceiver. It can be used as modulator with clipper and calling tone, or as receive audio-amplifier, by simply switching the operating voltage. Before describing the operation of the individual stages of the amplifier, the special features of the integrated circuit used should be mentioned.

The integrated circuit CA 3401 (or LM 3900) contains four current-fed operational amplifiers. The advantage of this integrated circuit over conventional operational amplifiers using a differential amplifier in the input is the low number of external connections required since the voltage dividers have been replaced by simple resistors. Differential amplifiers process the voltage difference between the non-inverted (+) and the inverted (-) input. In contrast to this, the input circuits of the operational amplifiers in the CA 3401 amplify the current difference. The current in the non-inverted (+) input causes approximately the same value of collector current in the first transistor. The difference between this current and the current at the inverted (-) input is fed to the second transistor and to the subsequent amplifier.

When no current is fed to the inverted (-) input, the amplifier will be switched off. This allows several parts of the circuit to be switched, although all four amplifiers are connected to a common operating voltage. In our application, the inverted input is connected via a feedback resistor to the output so that an additional current drives the amplifier into saturation. This is sufficient to block the amplifier at low AC-voltages, but not at the higher voltages that are present in the AF amplifier. In order to block the amplifier, the non-inverted (+) input is over-driven which brings the first transistor into saturation, and shorts out the inverted (-) input.

The individual stages operate as follows:

The first amplifier stage is designed for direct connection of a dynamic microphone. A trimmer potentiometer is provided to match the input to various microphones and voice levels. The second amplifier is in the form of a limiter in order to avoid overmodulation. The limiter diodes are to be found in the feedback links since the operational amplifier of the CA 3401 should not be driven into limiting. This would cause unwanted effects in the internal current distribution and cause instability. Without limiting (when the wiper of R 30 is grounded), the gain of this stage will be approximately ten times. If, on the other hand, the wiper is at the other stop, the modulating voltage will be limited to a P-P value of 1.2 V.

The third stage is in the form of an active low-pass filter with a gain of 1. This stage is also in operation in the receive mode where it is used for attenuating high-frequency noise components. This is followed in the transmit mode by the trimmer potentiometer for adjusting the frequency deviation (R 11). This trimmer is to be found in the DC-circuit of the tuning diodes, so that any adjustment of the modulating voltage has a direct effect on these.

In the receive mode, the fourth integrated amplifier drives a complementary output stage equipped with two TO 18-audio-transistors. For an output power of 300 mW, no heat sinks will be required. However, the power output stage is designed for a loudspeaker having an impedance of 32 Ω . Special attention must be paid when using lower impedance loudspeakers (less than 22 Ω).

In order to generate a calling tone, the first two integrated amplifiers are fed back via the calling tone switch and a frequency-determining RC-link. The frequency can be adjusted with the aid of trimmer resistor R 35.

Finally it should be mentioned that the AF voltage of the receiver is connected via connection Pt 122 to the second amplifier. Since this voltage is sufficiently low, the amplifier will not have the effect of a limiter.

6.4. Power Supply

The transceiver is operated from a nominal voltage of 12 V DC. The author's prototype is supplied from a built-in Ni-Cd accumulator having a capacity of 1 Ah, which is usually sufficient to cover a complete contest with the usual transmit-to-receive ratio. Connection point Pt 103, Pt 118 and Pt 127 of the transmit board are directly connected to the accumulator. A stabilizer circuit comprising transistors T 12 and T 13 is connected to Pt 103. This circuit generates a stabilized voltage of 10.5 V for the oscillator circuits and for tuning. A transmit or receive oscillator is switched on alternately via the microphone PTT-button and transistors T 10 and T 11. The antenna relay is also switched from this position (Pt 105).

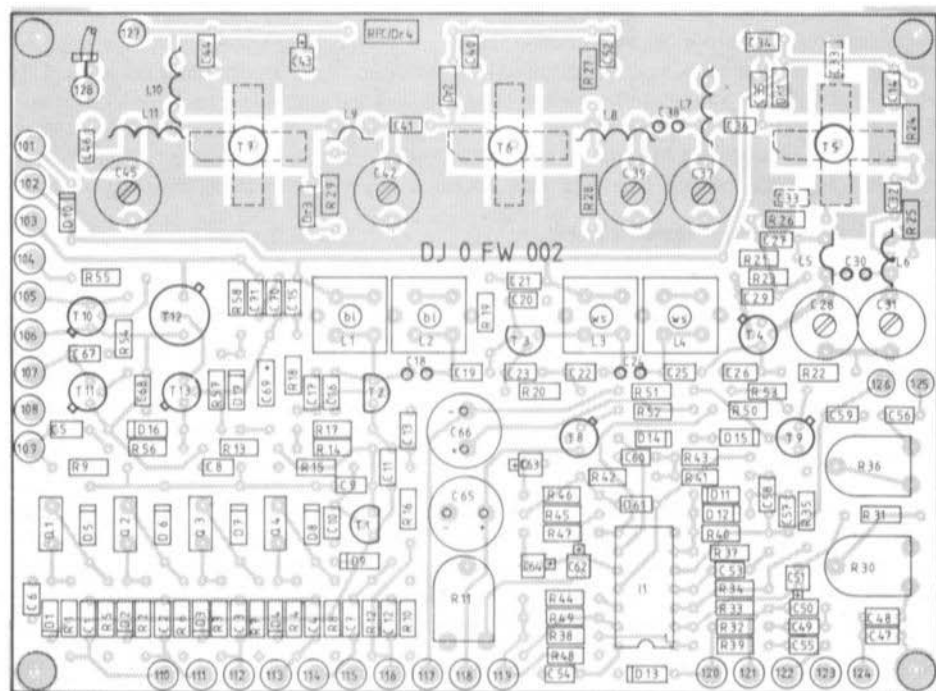


Fig. 19: PC-board and component locations

7. CONSTRUCTION

The transmitter circuit, universal AF amplifier, and the power supply circuit are all accommodated on a 125 mm x 90 mm double-coated PC-board with through-contacts. The component locations and conductor side of PC-board DJ 0 FW 002 are given in **Figure 19**; **Figure 20** shows the component side. With the exception of the small insulated islands, the ground surface on this side remains intact. All connection points are to be found on the outer edges. The three power transistors T 5 to T 7 are placed into holes on the PC-board in order to ensure that their connections can be directly soldered to the associated conductor lanes at low reflection. The two last transistors T 6 and T 7 are provided with small heat sinks that are screwed to the case after installation. The transistors only possess one mounting point; it is necessary for the contact surface to the case to be provided with heat-conductive paste.

Further details regarding this are given in part 4 of this description.

The two chip capacitors C 33 are soldered to the left and right from each emitter connection to ground as shown as a dashed line in Figure 19.

Of course, all component leads should be kept as short as possible as is good UHF-practice; however, special care should be taken with the black flat capacitors so that they do not touch the ground surface.

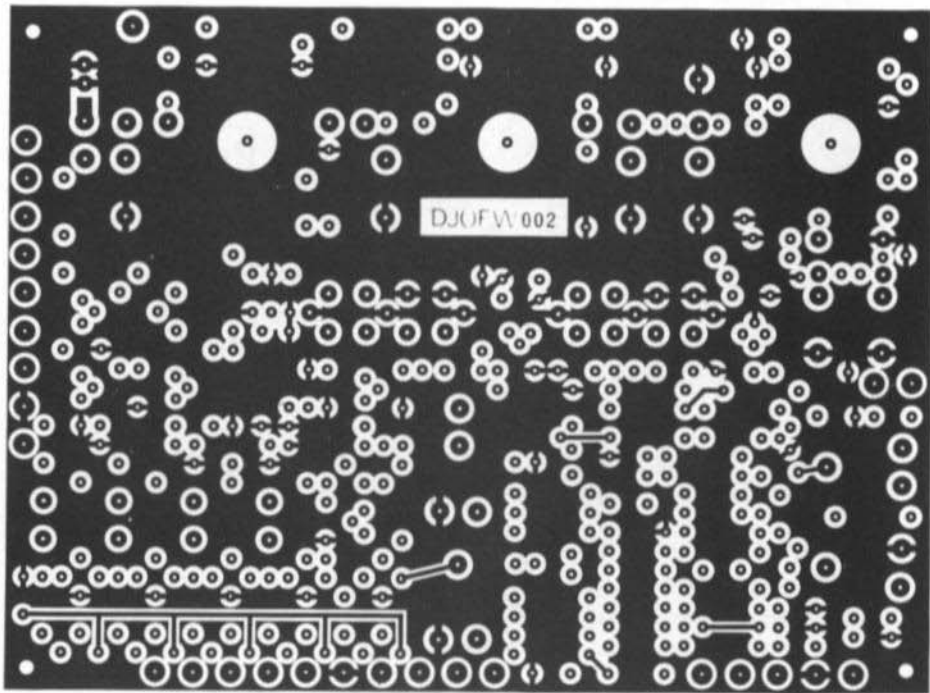


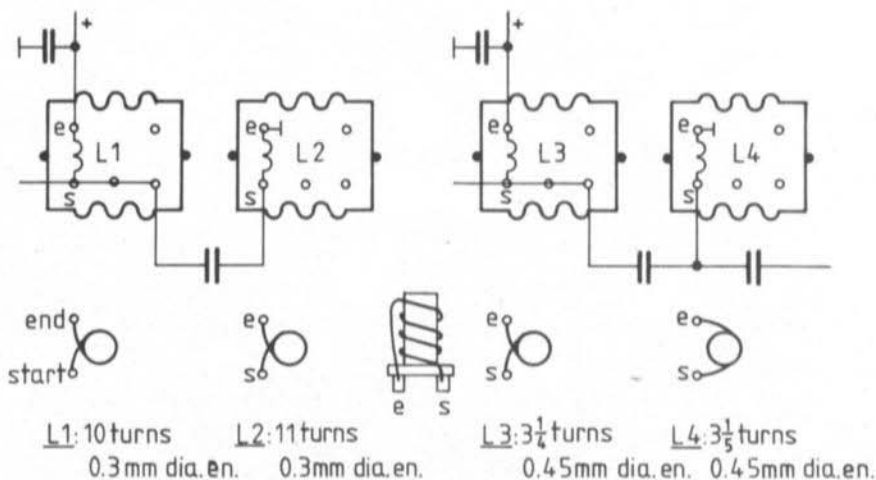
Fig. 20: Component side of PC-board DJ 0 FW 002

7.1. Component Details

T 1 - T 3:	BF 224 or BF 199
T 4:	BFY or similar high-gain UHF transistor
T 5 - T 7:	41008 (RCA)
T 8, T 13 :	2 N 2222 or BC 108 or BC 413 (NPN)
T 9:	2 N 2907 or BC 213 or BC 415 (PNP)
T 10, T 11:	2 N 2907 or BC 213 or BC 415 (PNP)
T 12:	BFW 72 or 2 N 2905 A (PNP in TO 5)
D 1-D 13, D 15:	BAY 94 or 1 N 4148 or similar silicon planar diode
D 14:	AAZ 10 or AA 119 or similar germanium diode or 56 Ω resistor
D 16:	C 6 V 2 small zener diode
D 17:	C 4 V 7 small zener diode
I 1:	CA 3401 (RCA) or LM 3900 N (National Semiconductor)

Inductances L 1 to L 4 are accommodated in special coil sets available from the publishers.

L 1:	10 turns of 0.3 mm dia. (29 AWG) enamelled copper wire with core
L 2:	11 turns, otherwise as L 1
L 3:	3.25 turns of 0.45 mm dia. (25 AWG) enamelled copper wire with core
L 4:	3.2 turns, otherwise as L 3



Inductances L 5 to L 11 are air-spaced inductances made from silver-plated copper wire.

L 5:	two turns of 0.8 mm dia. (20 AWG), diameter 3.3 mm
L 6:	two turns as L 5, with center tap
L 7:	two turns as L 5, but diameter 4 mm
L 8:	as L 7
L 9:	0.5 turns, 18 mm long, wire 1 mm dia., (18 AWG), bent to 5 mm diameter
L 10:	5 turns of 0.5 mm dia. (24 AWG), diameter 3.5 mm
L 11:	2 turns of 1.5 mm dia. (15 AWG), diameter 5 mm

- Ch 1 - Ch 3: approx. 0.22 μ H: 19 turns of 0.3 mm dia. (29 AWG) enamelled copper wire, 2.5 mm dia. close wound, self-supporting (wire length approx. 187 mm)
- Ch 4: 6-hole core ferrite choke (Philips) or 31 turns of 0.8 mm dia. (AWG 20) with core

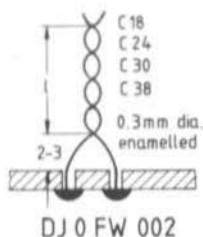
C 18, C 24, C 30, C 38: twisted enamelled copper wire

C 18: 10 turns/l = 11 mm

C 24: 7 turns/l = 9 mm

C 30: 5 turns/l = 6 mm

C 38: 5 turns/l = 6 mm



C 28, C 31, C 37, C 39: approx. 5 pF air-spaced trimmer with two connections

C 42, C 45: approx. 9 pF air-spaced trimmer as above

C 33: 2 chip capacitors of approx. 1 nF for the emitter connections

All capacitors with a value given in pF or nF in the circuit diagram are ceramic disc multi-layer or flat capacitors. Capacitors with μ F values are tantalum or aluminium electrolytics.

The three trimmer potentiometers are designed for horizontal mounting; spacing 10/5 mm

R 11: 10 k Ω , linear R 30: 1 k Ω , linear R 36: 100 k Ω , linear

All other resistors are carbon resistors for a spacing of 10 mm.

7.2. Crystal Specifications

Since no frequency conversion is made in the transmitter, the crystal frequency is one twentyseventh of the output frequency. This means it is only necessary for the center frequency of the required band to be divided by 27 (approx. 16 MHz). When ordering crystals, it is necessary to give the following additional information: Parallel resonance, load 16 pF, holder HC-18/U.

As was also the case in the receiver, the crystal frequency varies due to tolerances of the circuit which is very noticeable due to the high frequency multiplication factor. This means that the crystal frequency should be selected so that the most important channels are not to be found on the edge of the pulling range.

8. ALIGNMENT OF THE TRANSMITTER

8. 1. Before alignment, check all the components on the PC-board and ensure that no residual solder is present.
8. 2. A resistor of 6.8 k Ω should be soldered into place temporarily for R 27, and Pt 127 connected to + 12 V (PT 123 to ground). A resistance value is now to be found that results in a current drain of between 2 and 5 mA.

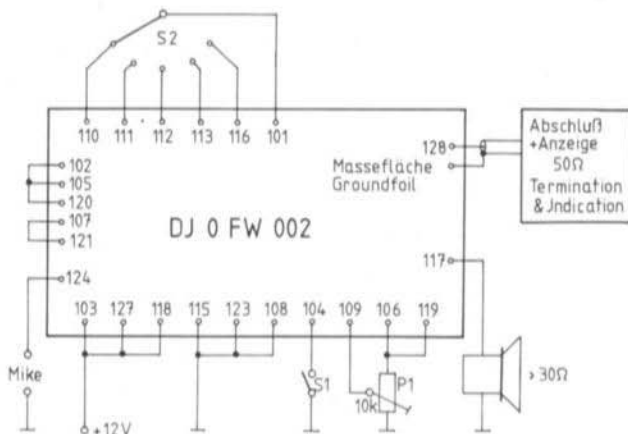


Fig. 21: External connections to the transmit board for alignment

8. 3. It is now possible for the transmit board to be connected as shown in **Figure 21**. This is followed by making the following measurements:

Measured at	S 1 open	S 1 closed
Pt 106	+ 10.5 V \pm 5 %	+ 10.5 V \pm 5 %
Pt 107	+ 10.2 V \pm 5 %	0 V
Pt 105	0 V	+ 10.2 V \pm 5 %

8. 4. Switch on S 1; S 2 to Pt 110

The voltage drop across R 18 should amount to 0 V. After inserting crystal Q 1, a voltage drop should be present. No variation of this voltage drop should be seen on rotating P 1.

Switch S 2 is now switched to Pt 111 and crystal Q 2 inserted. A voltage drop should also be present across R 18, which should not vary on rotating P 1.

The same procedure should be made for crystals Q 3 and Q 4 by placing S 2 to Pt 112 and Pt 113 respectively.

8. 5. Switch S 1 on; insert Q 1; S 2 to Pt 110; measure the voltage drop across R 18.

Align L 1 for dip.

Measure voltage drop across R 20. Align L 2 for maximum reading, L 3 for dip.

Measure the voltage drop across R 23. Align L 4 for maximum reading, C 28 for dip.

Measure voltage drop across R 26. Align C 31 for maximum reading.

Repeat the alignment in the same order, however, only measure the voltage drop across R 26.

8. 6. Switch positions as described for 8.5. Connect ammeter in the connection to Pt 127. Connect a power meter and terminating resistor. Align capacitor C 37 and C 39 alternately for maximum current reading.

Correct C 28 and C 31.

Align C 42 for maximum current.

Align C 45 for maximum output power.

Correct alignment of C 42 and C 45 for maximum output.

8. 7. The output power should remain constant within $\pm 5\%$ on switching crystals. If this is not the case, align C 37 for a constant output power.
8. 8. Switch S 1 off. The output power should drop to 0.
8. 9. S 1 off. Touch Pt 122 with the finger. A hum should be audible in the loudspeaker.
- 8.10. Switch on S 2. Connect microphone.

After monitoring the transmission in a receiver, align the limiting with R 30 and the frequency deviation with R 11.

NOTES WITH RESPECT TO PART 2

On constructing the receiver DJ 0 FW 001 in the laboratory of the publishers, the following notes were made which should be of interest to our readers:

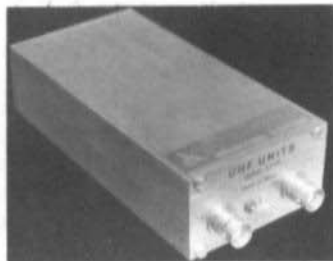
If the oscillator power is not sufficient for frequency conversion, it is possible for the multiplier transistors T 8 and T 7 to be biased for approximately class B operation by soldering a resistor of $10\text{ k}\Omega$ each between the base and the plus-pole of the operating voltage (on the conductor lane side of the board). In section 3, T 6 should read: BC 307, BC 308, BC 213 or BC 415 (PNP). The line circuits L 1 to L 6 should have spacings of 3.5 to 4 mm from the board; if this is not the case, they cannot be resonated with the trimmers.

The coil sets for L 7 to L 9 and L 12 to L 15 seem to be slightly different than that used by the authors. This means that it may be necessary to vary the number of turns. During construction in the laboratory, it was found that one turn more was advisable when winding the above mentioned coils.

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MORE GAIN WITH YAGI ANTENNAS

by Günter Hoch, DL 6 WU

In an earlier article (1) the author discussed basic Yagi antenna principles. The present article mainly concentrates itself to the question: What gain can be obtained from a single Yagi antenna and stacked arrays, and how can it be achieved in practice ?

1. MAXIMUM OBTAINABLE GAIN

Much of the confusion about the gain obtainable from Yagi antennas has been caused by exaggerated claims of some antenna manufacturers. This has gone so far that the ARRL has banned advertisements containing antenna gain figures from its publications.

The difficulties encountered in the measurements of absolute gain figures have led to considerable discrepancies between the curves published in different amateur radio handbooks, and the values measured in practice.

In (1) the author based his findings on the only generally accepted gain data then available. They are results of measurements by Ehrenspeck and Poehler (2) carried out on uniform Yagi arrays (constant director length and spacing).

It has already been pointed out in (1) that an increase in gain can be obtained by individually optimizing both element spacings and lengths. Recently publications substantiate this increase in gain (3, 4). The results of this are shown in **Figure 1**. It should be noted that both gain and array length are in logarithmic scale which means a proportional increase in gain with length would result in a straight line sloped 45° , or 3 dB per doubling of antenna boom length (dashed line). Curve A shows results from uniform arrays according to Ehrenspeck and Poehler (2), and the US National Bureau of Standards (4). They clearly show a tendency towards a levelling off of the gain beyond about 4λ length.

Curve B shows gains of arrays having constant spacings and individually optimized element lengths. These were also taken from (4), a belated publication of measurements made in the late 1950s. (A study of this report is highly recommended to anyone interested in the construction and optimization of Yagi antennas). The 4.2λ long NBS array has been made popular by W 0 EYE.

A computer-aided approach at the Danish Institute of Technology at Lyngby (3) has shown that a further small improvement can be made by suitably tapering the spacing. The results have been verified by anechoic-chamber measurements. They are shown as curve C in Fig. 1. The extrapolation of this curve conforms well with the values found by other designers using two-way optimization. The »WCVHFC« values represent the mean of several measurements at the annual West Coast VHF Conference as cited by Overbeck (5). The »DL 6 WU« data has been obtained by anechoic-chamber measurements of arrays experimentally optimized by the author.

There is not one serious report regarding practical or even theoretical gain of a rod-type Yagi antenna that would surpass the gain of curve C. This means that it can quite safely be taken as the upper gain limit. Structures with multiple elements, loops, quads etc. may have higher gains if they have considerable extension perpendicular to the longitudinal axis.

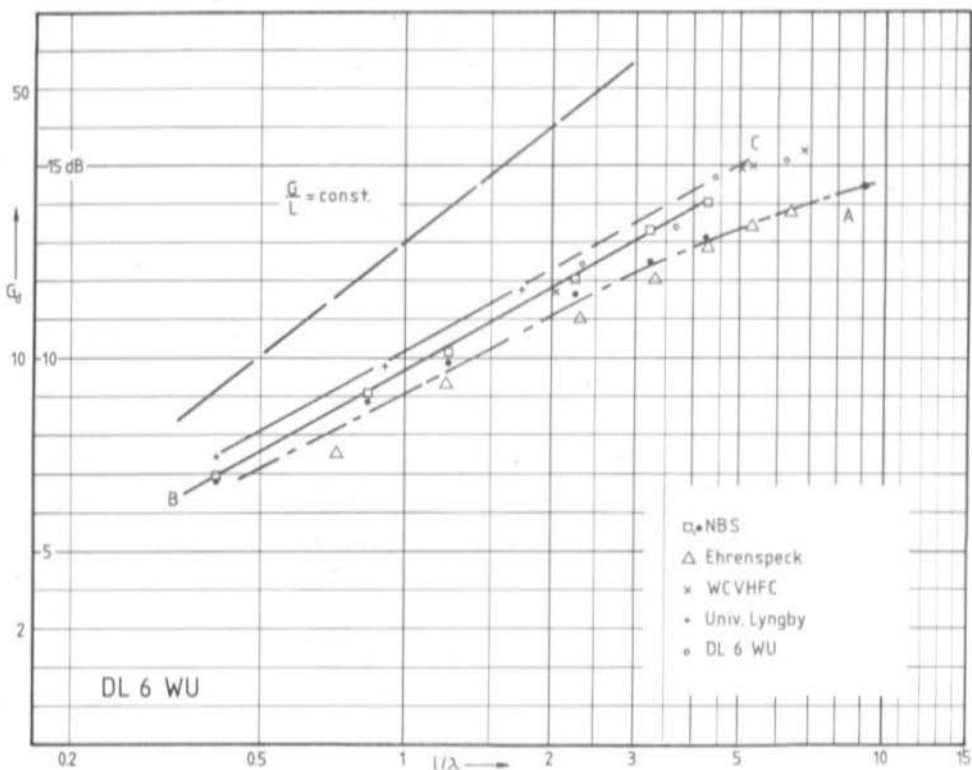


Fig. 1: Gain of Yagi arrays as a function of length
A uniform, B length-optimized, C length-and-spacing-optimized

All claims of higher gains given for Yagi antennas must be taken with extreme caution. One example is to be given to illustrate this: The gain of a very popular 2-m-Yagi of 3.05λ length is catalog-rated at 16 dBd while curve C would limit its gain to 13.6 dBd. The E and H plane -3dB angles of 32° and 34° (the former being correctly stated in the catalog!) indicate a maximum of 13.6 dBd when entered into the Kraus formula (6). Comparison measurements yield 13 to 13.5 dBd which confirm the validity of the above considerations.

No reliable data is available on Yagi antennas longer than about 7λ so the question of ultimate gain limitation cannot be answered yet. This limit, if existent at all, seems to be higher than experts have assumed to date. On the other hand the slope of all curves in Figure 1 is 2.2 dB per length doubling or less which is noticeably lower than the 3 dB commonly expected. This less-than-proportional gain increase quickly leads to very unhandy array lengths for higher gain values at lower frequencies.

Two further factors should not be overlooked which limit practical antenna length: Bandwidth and precision. Bandwidth tends to decrease with increasing array length with values around 1% to be expected at 10λ . The precision required for optimum results is stated in (4) to be 0.003λ . At 1296 MHz this would call for an element length tolerance of less than 1 mm.

2. WHICH TYPE OF ANTENNA ?

With the previously mentioned limits in mind, a graph was drawn to compare the performance of different antenna types and to determine the bands most favorable for the use of Yagi antennas in high-gain arrays.

In order to make at least a rough comparison possible, **Figure 2** assumes a maximum antenna height or width of about 8 m which seems to be somewhat of a constructional barrier. Yagis are considered only up to 10λ length for bandwidth reasons. Curtain arrays are considered in 16 element increments as this is a standard building block. It is quite evident that Yagi antennas can be used to greatest advantage on 70 cm. (The same is valid for the 220 MHz-band in the USA). On 2 m there is a challenge from phased arrays. These, however, have a disadvantage of the large number of feed connections which become the limiting factor at higher frequencies.

On 23 cm parabolic antennas can provide more gain (they are still rather clumsy on 70 cm) whereas the precision limit together with feed complications makes Yagi antennas somewhat questionable at least for high-gain arrays on this band.

The use of aperture antennas is a must on all higher GHz bands.

It is the author's opinion that work on Yagi antennas should be done especially with the 70 cm band in mind; they can then be easily recalculated to the 2 m and 23 cm bands due to the harmonic relationship.

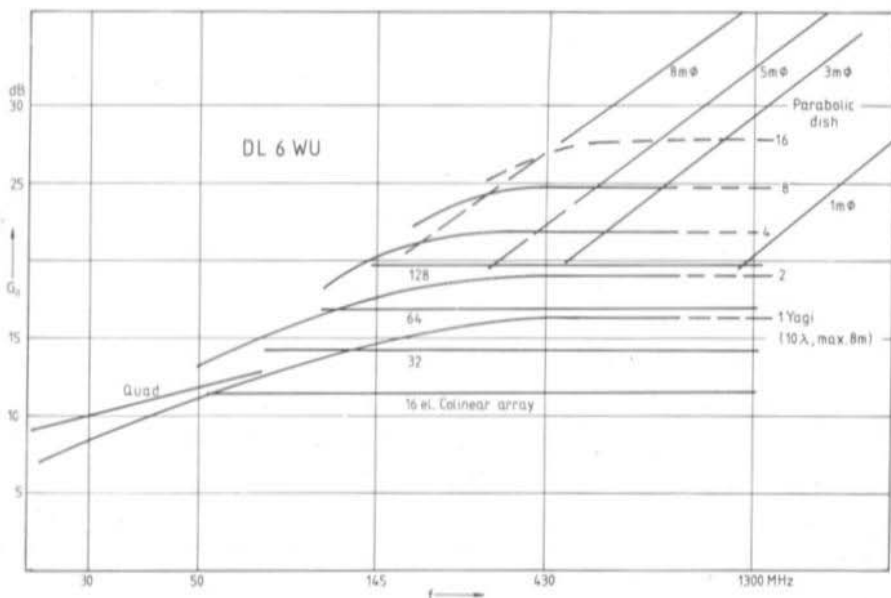


Fig. 2: Gain of different types of antenna arrays limited by maximum array dimensions of 8 x 8 x 8 m

3. PRACTICAL DESIGN

The question is now how the gains given in the curves can be obtained in practice ?

Excellent design criteria is given in (4), and the minute difference to the maximum gain theoretically possible is insignificant in practice. The «Quagi» antennas described in (5) also work very well although they are disliked by some because of the type of feed used. Still, it was found complicated that a completely different plan had to be followed for each desired array length. In order to find a more general approach, the author has collected data on two-way optimized long Yagis from numerous sources, together with data from his own experiments, and has standardized them with respect to frequency and element diameter. This was done with the aid of the equal-reactance graph given in (1). It turned out that the optimum element lengths agree so well that an average curve can be drawn with less than 0.01λ departure from any individual value (Figure 3).

This is not true to the same extent with the spacing values found to be optimum by different designers. However, the tendency is clearly that of an asymptotical approach up to about 0.4λ with the initial value strongly influenced by the type of feed.

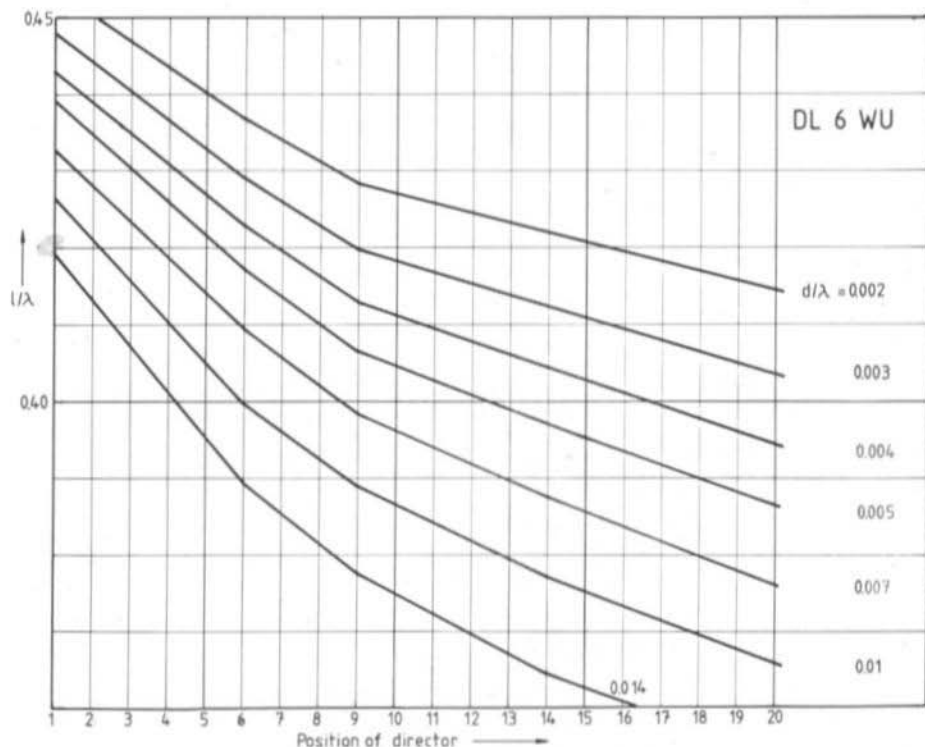


Fig. 3: Length of directors as a function of director position in long Yagi arrays (parameter: element diameter)

As many experiments have shown, a non-reactive feed using a self-resonant dipole element calls for a closely spaced first director – often called a launcher element. It takes about the position the radiator would take in designs with an adjustable impedance feed such as a delta or gamma match. For the sake of mechanical and electrical stability it was decided to concentrate on the equal-diameter folded dipole with halfwave (4:1) balun as driven element. Tests have shown that it can always be replaced with unaltered results by an open dipole with 1:1 balun transformer or, of course, any other desired matching device similar to those described in (7).

Measurements made with designs using both average element lengths and spacings showed such good results that it seems permissible to risk a »universal recipe« for long Yagi design. If the values given in Figure 3 and Table 1 are followed closely, a Yagi will result that has near-optimum gain for the length, and can be cut off anywhere beyond about 2λ length with an SWR of less than 1:1.2. Data for short antennas follow a slightly different pattern and should be taken from (4).

Antennas constructed from the data presented here will perform markedly better than optimum uniform antennas of the same length. Radiation patterns will be slightly wider and a lot cleaner, the additional gain being taken from the sidelobes. It is not claimed, however, that there is no room for further improvement. Namely the backward wave reflected from the open end of the antenna causes a more or less pronounced distortion of the amplitude and phase profile depending on the point of cut-off. This means that the skilled experimenter with the necessary measuring equipment described in (8) may be able to squeeze out a few more tenths of a dB (this is about the maximum) by further slight modification of element lengths and positions. This can be a very time-consuming and frustrating process governed by the law of diminishing returns. Only a few hints can be given here. The director farthest from the radiator has a great influence on the reflected wave. Optimizing its length and position seems the most rewarding adjustment with a slight variation of director positions near the center of the array being the next best bet. After any adjustment, the match must be checked and restored because even a slight mismatch might mask the minute gain increase that has been obtained. Never attempt to optimize without a reference antenna of known performance, and without equipment which meets the high reproducibility standards required.

The following procedure is given for the average ham who does not possess these means. All he needs is a stable power source for the design frequency and an accurate (!) SWR meter.

- a) Determine the desired antenna lengths in wavelengths and the corresponding number of elements. Take element positions from **Table 1**.
- b) Determine element diameter in wavelengths and read off or interpolate director lengths from **Figure 3** respectively. Read off the lengths of radiator and reflector from **Table 1** or from Figure 3 in (1).
- c) If the elements are to be mounted through a metallic boom, take the correction factor from Figure 4 in (1) into consideration or use **Table 2**.
- d) Cut and mount elements leaving the driven element variable, if possible. Mount antenna facing towards the sky and check SWR. If it exceeds the expected value and all measures have been determined correctly (!), the fault will most certainly lie in the driven element. Adjust its position and length not forgetting the balun which might be too long or too short. Do not change other elements' length or position at this point !

Element	Spacing/ λ	Spacing/mm	
		432 MHz	432 MHz, 4 mm dia.
Reflector	0.240	160	338
Dipole	—	—	322
Director 1	0.075	55	302
Director 2	0.180	125	299
Director 3	0.215	150	296
Director 4	0.250	175	293
Director 5	0.280	195	290
Director 6	0.300	210	287
Director 7	0.315	220	284
Director 8	0.330	230	282
Director 9	0.345	240	280
Director 10	0.360	250	278
Director 11	0.375	260	277
Director 12	0.385	265	276
Director 13	0.390	270	275
Director 14	0.395	275	274
Director 15	0.400	280	273
Director 16	0.400	280	272
Director 17	0.400	280	271
Director 18	0.400	280	270
Director 19	0.400	280	269
Director 20	0.400	280	268

Table 1: Dimensions of long Yagi arrays
 (Dimensions of a model antenna for 432 MHz with insulated 4 mm diameter elements)

Boom dia./ λ	$\Delta l / \lambda$	Boom dia./mm 432 MHz	$\Delta l / \text{mm}$ 432 MHz
0.01	0.003	7	2
0.015	0.005	10	3.5
0.02	0.008	14	6
0.03	0.016	21	11.5
0.04	0.026	28	18
0.05	0.038	35	26

Table 2: Length correction Δl to be added to length of elements passing through metallic boom.
 Influence of insulated elements falls quickly with rising distance, and can be neglected at about 1 boom radius from surface

4. STACKING

In general, directional antennas should be stacked at the distance where their effective apertures just touch (9). Unfortunately there is no definite border of the effective aperture so this distance is not well-defined. When two equal arrays are stacked, the stacking gain will rise with increasing distance reaching a first peak of about 2.5 dB and then fluctuate around this value (4, 10). These fluctuations can be explained either as varying mutual coupling or by addition and cancellation of minor lobes. A good rule of thumb is to use the distance at which the first null of the group pattern falls on the -3 dB point of the individual pattern in this plane - approximately halving the beamwidth. By simple geometry this distance is found to be $D = \lambda/2 \sin(\varphi/2)$ where φ is the -3 dB aperture angle of the arrays to be stacked. Note that distances differ in E and H planes if the patterns are not rotationally symmetrical. Greater spacings may result in slightly higher gain, 2.95 dB are stated in (10), but at the penalty of a sharper main beam and more and stronger sidelobes. Stacking distances below the above-mentioned value will quickly reduce the increase in gain obtained by stacking. Use of the compromise formula will result in first sidelobes 10 - 12 dB below the main lobe, and an increase in gain of around 2.5 dB for a bayed array, as tests have shown in practice.

5. EXAMPLE

A 21 element Yagi antenna for 432 MHz was derived from the design data that has been given. Aluminium rod of 4 mm diameter (welding electrodes) was chosen because of its availability. The boom is 15 by 20 mm dry wood available in do-it-yourself shops (the central part must be reinforced by a support). A double reflector spaced 0.3λ vertically was used instead of one, because on long arrays the gain increase of about 0.3 dB is higher than obtainable by a corresponding length increase of the boom.

Gain was measured to be about 15.5 dBd in substitution measurements. A separate measurement using the rear section terminating at director no. 8 showed 12.5 dBd. **Table 1** gives the dimensions of the antenna.

Correct stacking distance for the long Yagi antenna would be 1.6 m in both planes, whereas the 11 element array would need a stacking distance of 1.2 m in the E plane and 1.1 m in the H plane.

6. CONCLUDING REMARKS

There are a multitude of good VHF and UHF antennas available on the market. The author does not mean to preach an all-out do-it-yourself philosophy. There are, however, many occasions where a mass-produced antenna would not serve the purpose or would just be too expensive - e.g. for receiving weather satellite signals, testing a temporary location, monitoring a dx tv channel, to name just a few. A cheap and quickly-made Yagi array can fill the gap. Besides, in our time of highly complicated factory-made ham equipment a good self-made antenna can impart to the builder a feeling of achievement that is hard to reach otherwise.

Acknowledgement:

The author wishes to thank Dick Knadle, K 2 RIW, for the fruitful discussion.

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A LINEAR TRANSVERTER FOR 28 MHz - 1296 MHz WITH PUSH-PULL MIXER

by U. Beckmann, DF 8 QK

Up to now, linear transverters for the 23 cm band mainly used power mixers equipped with tubes (1), (2), or varactor diodes (3). This was mainly due to the fact that low-level mixers using diodes or transistors were only able to provide a very low output, and because the transistors required for linear amplification were difficult to obtain and were expensive. This article is to describe a transmit converter equipped with a transistor push-pull mixer and three-stage linear amplifier, whose transistor complement of five transistors costs approximately 100 DM at the present time. The transmit converter is built up on a double-coated epoxy PC-board and provides an output power of 300 mW at an operating voltage of 12 V. This output power is usually sufficient for portable operation from good sites.

A special feature of this circuit is the push-pull mixer which is able to suppress the local oscillator frequency by at least 40 dB when correctly balanced. This means that only one bandpass filter similar to that described in (4) or (5) is required in order to suppress the image frequency. In the most unfavorable case when using the 10 m band as intermediate frequency, it will be 56 MHz below the required frequency and will be attenuated by approximately 20 dB even when using a simple filter as described in (4). In conjunction with the selectivity of the linear amplifier of approximately 15 dB, the total attenuation will be in the order of 35 dB. If a power amplifier equipped with a 2 C 39 tube in a cavity resonator is driven (in the case of fixed station operation), the image frequency filter will not be required since such power amplifiers provide sufficient selectivity.

1. BLOCK DIAGRAM

The basic principle of the transverter and the transistors that are used in it are given in the block diagram given in **Figure 1**. A drive power of approximately 5 mW at 28 MHz is required for driving the module. A local oscillator frequency of 1268 MHz at a power of 5 mW is also required. This can be provided using suitable crystals in module DJ 4 LB 003 (6) followed by a frequency tripler. This is not to be discussed in detail, since other suitable oscillators are either available or can be slightly modified (see DC 0 DA 005 in this edition).

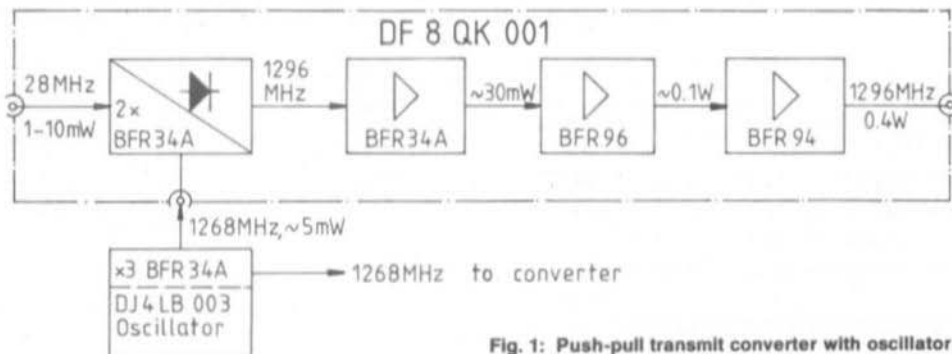


Fig. 1: Push-pull transmit converter with oscillator

2. CIRCUIT DETAILS

Figure 2 shows the circuit diagram of the transmit converter. It will be seen that the push-pull mixer with its extensive circuits represents the heart of the module. The modulated, 28 MHz signal is fed in push-pull, whereas the local oscillator signal is in push-push. In order to ensure that the 28 MHz inductance L 1 is balanced, it is wound in a bifilar manner. Since usually more than enough drive power is available, potentiometer P 1 has been provided to select the most favorable drive level. The oscillator signal is fed in approximately 8 mm from the cold end of the shortened $\lambda/4$ circuit comprising L 2 / C 6, which is then tuned to 1268 MHz. The DC-voltage operating points of the two transistors can be adjusted to the same values using the trimmer resistors P 2 and P 3. The micro-stripline L 3 is provided for the output frequency of 1296 MHz and is tuned with the aid of trimmer capacitors C 12 and C 13. This line possesses an electrical length of $\lambda/2$, the trimmers are to be found at the hot ends (max. voltage), and the voltages at the ends are in antiphase to another. The cold position (without RF voltage) is to be found at the center to which the collector voltage is fed. Any difference in the output capacitance of the two transistors is compensated for by suitable adjustment of the trimmers. This means that an exact balancing of the output circuit can be made so that the maximum suppression of the local oscillator frequency is obtained.

The loading of the output circuit of the mixer is also balanced using a similar output coupling circuit. The first linear amplifier transistor (T 3) is coupled in in the vicinity of cold position (center of L 4). Inductances L 3 and L 4 form a bandpass filter.

The collector of transistor T 3 is coupled via a short matching inductance to a capacitively shortened $\lambda/4$ circuit whose cold end is bypassed to ground with the aid of a disk capacitor. The base of the subsequent transistor T 4 is coupled via a matching network comprising C 20, C 21 and a printed inductance. The collector of T 4 also works into a capacitively shortened $\lambda/4$ circuit (L 6).

The output transistor T 5 is also matched in a similar manner. The output is transformed to 50 Ω using inductance L 7 and trimmer capacitors C 35 and C 36. All inductances with the exception of L 1 are printed striplines to ensure a high degree of reproducibility.

All base voltages are fed in via printed $\lambda/4$ chokes which are bypassed at the ends using printed capacitances. The low capacitance values required at 1300 MHz allow this simple construction.

Special attention has been paid to the base voltage dividers and the quiescent current adjustment. The following demands should be fulfilled:

- Minimum variation of the quiescent current on altering the operating voltage
- Good stability with respect to ambient temperature fluctuations
- No increase of the quiescent currents during continuous operation (endangering the expensive transistors)
- Uncritical alignment of the quiescent current

Point 3 above was the most difficult to achieve in practice, and it was necessary for diodes D 4 and D 5 to be mounted in thermal contact to the associated transistor. These diodes are to be found on the lower side of the board. D 4 is placed onto the plastic case of transistor T 4 with heat-conductive paste; diode D 5 is glued to the threaded bolt of T 5 also using heat-conductive paste.

DF 8 QK 001

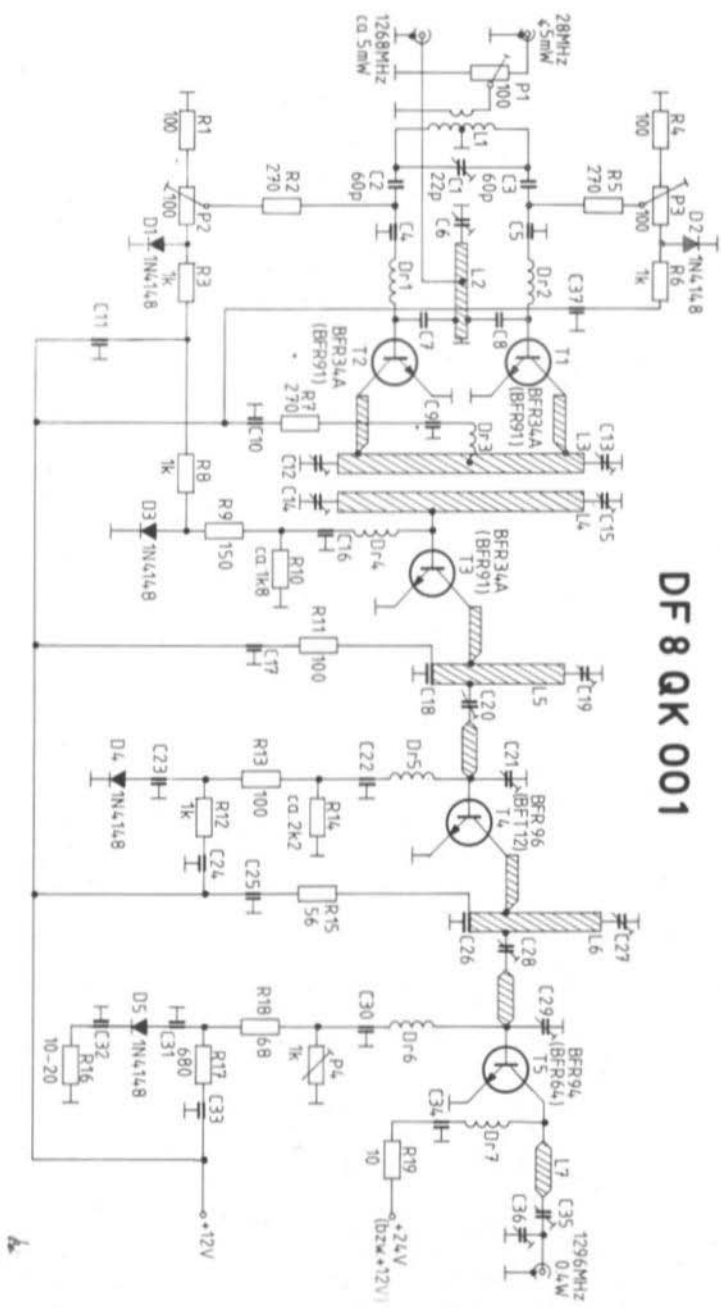


Fig. 2: 400 mW transmit converter for 144 MHz / 1296 MHz

It is also necessary for T 5 to be provided with a heat sink. A 2 mm thick aluminium strip, approximately 50 mm in length and 10 mm wide, was found to be suitable. It was bent up approximately 15 mm from both ends and screwed to the tapped bolt of T 5.

The operating voltage connection of the output stage is provided separately so that it may, if required, be connected to 24 V. In this case, the output power will then amount to approximately 400 mW.

3. CONSTRUCTION

The construction of the transmit converter should not cause any difficulty since there is only a minimum of mechanical work to be done. The 170 mm x 75 mm double-coated epoxy PC-board DF 8 QK 001 is shown in **Figure 3**.

After drilling the holes for the components, it is necessary for slots to be sawn in the PC-board with the aid of a fret saw for capacitors C 18 and C 26. A further slot is necessary at the ground end of the oscillator circuit comprising L 2 in order to be able to provide a through-contact to ground with the aid of a copper foil strip. Two further slots of approximately 5 mm in length are also required for the emitter connections of transistor T 5. After this, it is possible for the components to be mounted on the board. Transistors T 1 to T 4 are mounted into 5 mm holes. The emitter connection is bent down by 90° (bent up in the case of T 1) and is placed into the hole for the transistor. This means that the hole must be slightly widened at one side for this. A hole of 7 mm dia. is required for transistor T 5.

One of the rotary plate connections of C 29 is pushed through to the ground side; the second and the stationary plate connections are bent down by 90° and carefully soldered to the base and emitter connections of T 5.

Resistors R 7, R 11, R 15, R 16, R 19, as well as disk capacitors C 10, C 17, C 23, C 25, C 31, C 32, C 38 which are used as support, are to be found on the lower side of the board.

The local oscillator signal and the output coupling from the transmit converter can be connected by soldering the connection cable directly to the board from the lower side. The required hole is made by removing approximately 1 mm around the hole on the lower, ground surface of the PC-board. The inner conductor is placed through the hole, soldered to the appropriate conductor lane and the outer conductor soldered to the ground surface. Of course, it would also be possible for a BNC socket for single-hole mounting to be used and then modified by removing the protruding insulation so that it directly touches the ground and conductor surfaces.

The module can be enclosed in a case constructed from single-coated PC-board material. The height of the side panels should amount to approximately 40 mm and the spacing of the board to the base plate should be 15 mm.

Figure 4 shows a photograph of an author's prototype without case and using trimmers instead of resistors R 10 and R 14.

3.1. Special Components

- T 1, T 2, T 3: BFR 34 A (Siemens) or BFR 91 (Philips) with some limitations
T 4: BFR 96 (Philips), or BFT 12 for lower output power levels
T 5: BFR 94 (Philips), or BFR 64 for lower power levels
D 1 - D 5: 1 N 4148
L 1: 2 x 9 turns bifilar wound on a 6 mm dia. coil former without core coupling: 1 to 2 turns at the center
Ch 3: 1.5 turns of enamelled copper wire wound through a ferrite bead
Ch 7: 2 turns of silver-plated copper wire wound on a 3 mm former, self-supporting

All trimmers are plastic foil types of 7 mm diameter:

- C 1: 22 pF (green)
C 6, C 12 - C 15, C 19 - C 21, C 27 - C 29, C 35, C 36: 6 pF (grey)
C 2, C 3: 56 pF ceramic disk capacitor
C 7, C 8: approx. 10 pF ceramic disk capacitor of less than 3 mm diameter
C 9, C 11, C 24, C 33, C 34, C 37: 1 nF feedthrough capacitor, 3 mm dia.
C 10, C 17, C 23, C 25, C 31, C 32 :
between 220 p and 1 n, disk capacitor without connections
C 18, C 26: 1 nF disk capacitor without connections, 8 mm dia.
C 4, C 5, C 16, C 22, C 30: printed capacitances

All trimmer potentiometers are for horizontal mounting with 10/5 mm spacing:

- P 1 - P 3: 100 Ω
P 4: 1 k Ω

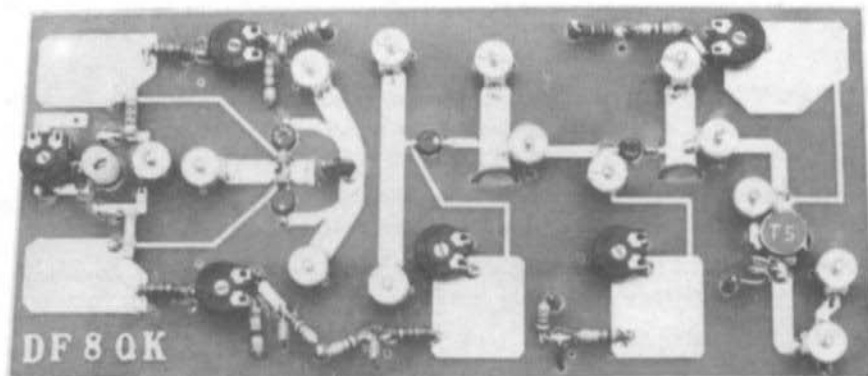


Fig. 4: Author's prototype of the transmit converter

4. ALIGNMENT

The alignment can be made with the following measuring equipment:

- Multimeter
- Absorption wavemeter for up to 1.3 GHz (as described in VHF COMMUNICATIONS)
- 50 Ω load and power meter (need not be calibrated)

The quiescent currents of the transistors should be firstly aligned to the following values:

- T 3: 2 mA \triangleq 0.20 V across R 11
- T 4: 2 mA \triangleq 0.11 V across R 15
- T 5: 60 mA \triangleq 0.60 V across R 19

Set P 1 fully anticlockwise, P 2 and P 3 to a central position (approximately 0.52 V base voltage); adjust C 1 and C 20 for approximately half capacitance, C 29 and C 36 for approximately a quarter, and all other trimmers to minimum capacitance.

The local oscillator signal should now be fed to the module and adjusted so that approximately 7 V can be measured across R 7. Switch off the oscillator again and feed in the IF signal to be converted. Align the IF circuit with the aid of C 1. Adjust the level with P 1 so that approximately 0.1 V can be measured across R 7. Switch on the oscillator again.

Observing the voltage drop across R 11, align C 6 and C 12 - C 15 for maximum reading. Switch off the IF signal after each adjustment of the trimmers to check that the voltage falls, otherwise the alignment will have to be made to 1268 MHz.

Measure the voltage drop across R 15 and align capacitors C 19 - C 21 for maximum. The 1296 MHz signal at the output should now be clearly seen. Align capacitors C 27 - C 29, C 35 and C 36 for maximum output power. This means that the coarse alignment has been completed.

This is followed by realigning all trimmers carefully for maximum output power. After this has been completed, it should be checked once again that this maximum is at 1296 MHz and not at 1268 MHz.

This is followed by balancing the mixer:

Switch off the IF signal, place an absorption wavemeter in the vicinity of the output circuit and align it to 1268 MHz. At this point, the oscillator signal will no doubt produce a clear indication.

Align P 2 and P 3, followed by C 12 or C 13 respectively for minimum output (the trimmers are virtually at minimum capacitance). Switch on the IF signal once more and align C 13, or C 12, as well as C 14 and C 15 for maximum output power at 1296 MHz.

These two alignment processes should be repeated several times, at the same time increasing the coupling of the absorption wavemeter until no improvement is possible.

The following voltages should be present across the given resistors after completing the alignments at an operating voltage of $U_B = 13 V$:

28 MHz	1268 MHz	1296 MHz	U (R 7)	U (R 11)	U (R 15)	U (R 19)
-	-	-	$\sim 0 V$	0.2 V	0.1 V	0.6 V
-	x	-	7 V	0.2 V	0.1 V	0.6 V
x	-	-	$\leq 0.1 V$	0.2 V	0.1 V	0.6 V
x	x	x	$\sim 6.5 V$	$\sim 3 V$	$\sim 2.7 V$	$\sim 0.6 V$

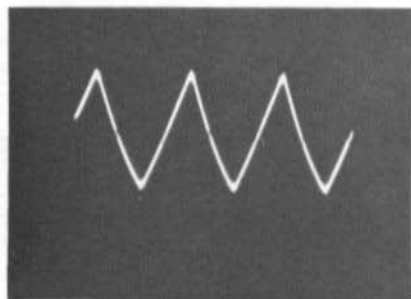


Fig. 5: Demodulated triangular wave signal during ATV test
 P_{out} : approx. 160 mW

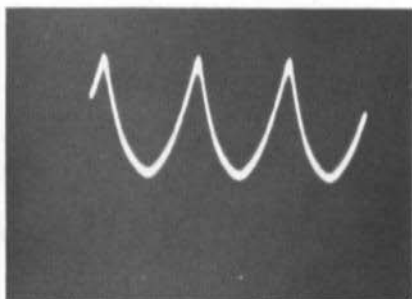


Fig. 6: Output power: 300 mW

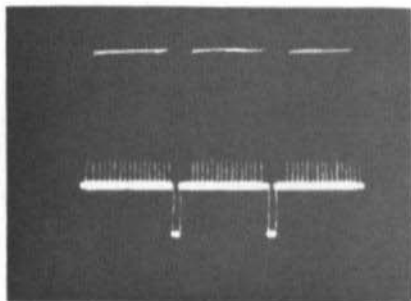


Fig. 7

Fig. 7: 160 mW composite video signal from a delta-wave generator.
 The standardized synchronizing level of 25 % is maintained.

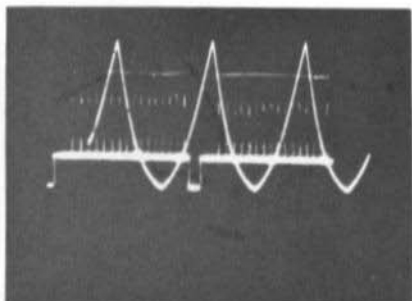


Fig. 8

Fig. 8: 250 mW composite video signal of a pattern generator, and a delta-wave voltage for comparison.
 The synchronizing signal is already limited.

5. USING THE TRANSMIT CONVERTER FOR OTHER INTERMEDIATE FREQUENCIES

An intermediate frequency of 144 MHz can also be used. It is only necessary for inductance L 1 to be changed: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, with center tap and 1 turn coupling.

The transmit converter has also been used successfully for ATV on 1252.5 MHz when fed with an input frequency of 62 MHz. An input voltage of 0.2 V (\triangleq 1 mW) is sufficient drive. In this case, inductance L 1 is reduced to 2 x 5 turns, and the required oscillator frequency is 1190 MHz.

The transmit converter operates linearly up to an output power of 200 mW composite video signal at 1252 MHz. This can be seen in **Figures 5 and 6** where the video carrier is modulated with a triangular-wave signal and the demodulated output signal fed to an oscilloscope.

When using a pattern generator, no limitation of the synchronizing pulses was observed. The UHF demodulator module DJ 4 LB 005 has been found suitable for demodulation when coupled to the output via 1 to 2 pF. The resonant circuit inductance on the demodulator board comprises a small bent wire.

The base chokes and the base voltage dividers are of such low-impedance for the modulating frequencies that no additional electrolytic capacitors are required for blocking.

6. PRACTICAL EXPERIENCE - MEASURING RESULTS

It was found that the adjustment of the mixer level was not very critical. The transmit converter operates satisfactorily with oscillator power levels of between 1 and 10 mW. The adjustment of the operating points can also be varied within wide ranges as long as attention was paid to balanced conditions. It was only in the ATV-mode that a variation could be observed with the aid of the triangular-wave signal, this was not audible in the SSB mode.

No tendency to oscillation or feedback could be observed during any of the alignment processes. On-the-air reports of the modulation show that it is just as good as the available linear converter described in (2), even in the case of local stations.

Unfortunately, it was not possible to measure the intermodulation ratio. However, the oscillator frequency is suppressed by more than 40 dB and the image in the order of 15 dB. Any spurious signals are at least 40 dB below the required signal.

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THREE-STAGE PREAMPLIFIER FOR THE 23 cm BAND

by J. Dahms, DC 0 DA

Hybrid ring mixers in printed technology with two Schottky diodes are often used on the 23 cm amateur band due to the simplicity of the circuit and high reproducibility. Samples of this are to be found in (1) and (2). These converter systems are not able to provide a single sideband noise figure of less than 10 dB under the most favorable conditions since the image-frequency noise is fully effective in the total noise figure. This problem was discussed in detail in (3). Also, not every amateur has the possibility of using a low-loss cable between the antenna and the converter. When using a cable length of 20 m of RG-213, the losses between the receive antenna and the converter will amount to at least 6 dB, which means that the receive sensitivity of the system will not be better than 16 dB. A three-stage preamplifier using the economical stripline transistor BFR 34 A (Siemens) is to be described that can reduce the overall noise figure to the order of 4 dB. However, it is necessary for the preamplifier to be placed in the direct vicinity of the antenna, as was recommended in (4).

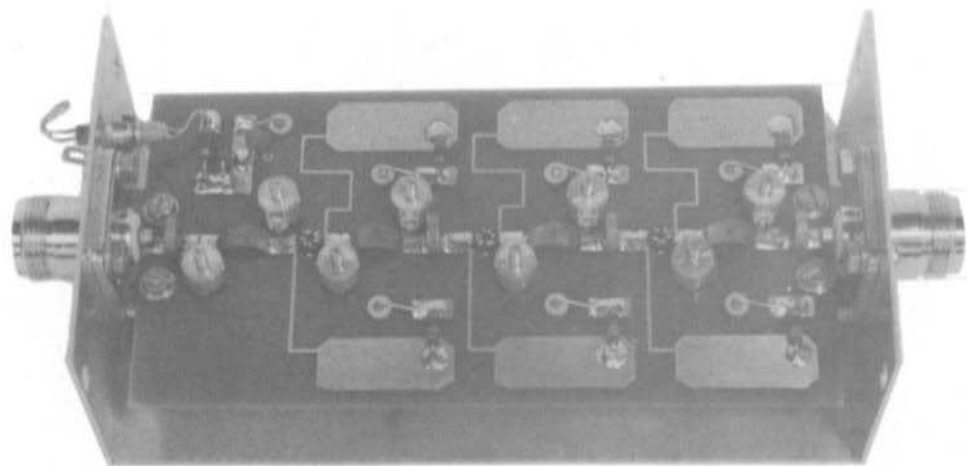


Fig. 1: Photograph of the author's prototype

Figure 1 shows a photograph of the author's prototype. This preamplifier fulfills the following conditions:

- optimum noise matching of the input stage to the antenna
- sufficient overall gain to compensate for cable losses and conversion loss of the diode mixer
- low-loss connection to the feeder cable using N-connectors, and installation in a readily available case.

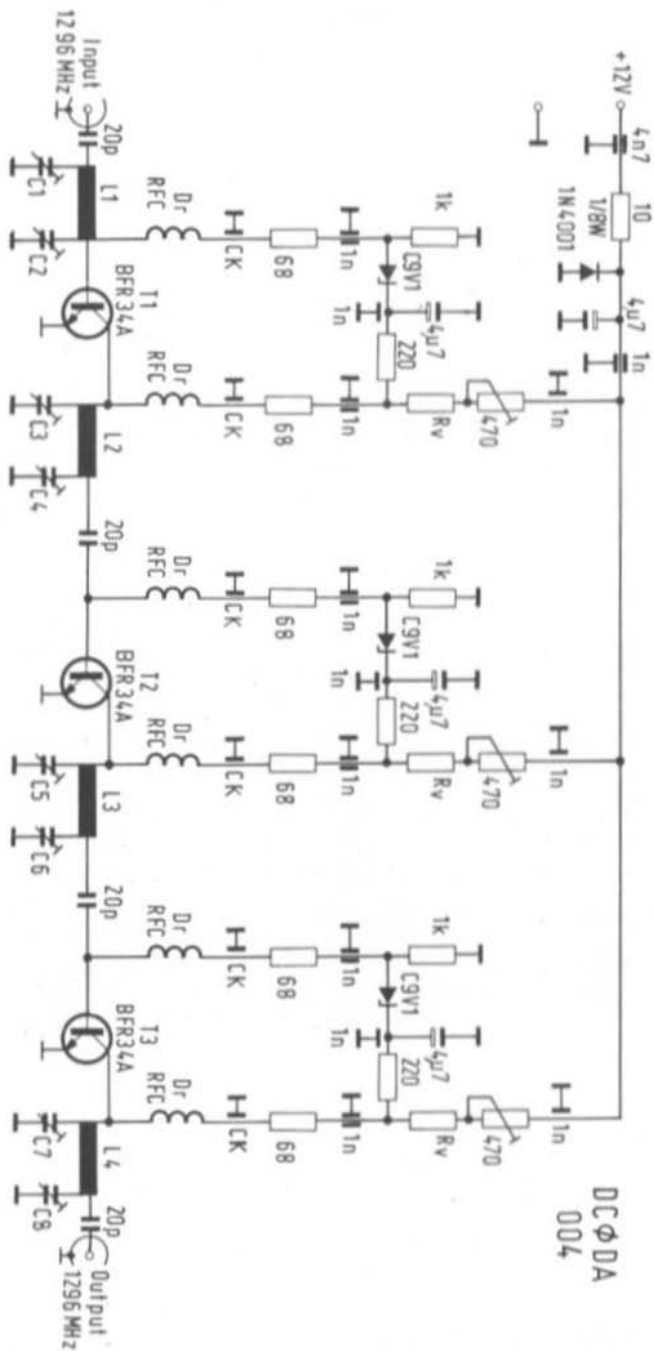


Fig. 2: Three-stage preamplifier for 23 cm

Practical experimentation showed that a bandpass filter for 1296 MHz should be provided between the preamplifier and a wideband receive mixer in order to reduce any cross modulation effects in the mixer, and to ensure a sufficient image rejection and selectivity. A suitable filter for this is to be described in section 4.

1. CIRCUIT DESCRIPTION

The three preamplifier transistors are operated with directly grounded emitters. The matching between the antenna and the first transistor, between the transistors, and from the last transistor to the output is made with the aid of Pi-filters with greatly shortened $\lambda/2$ striplines. In order to obtain a sufficiently high Q, these are not printed onto the board, but are made from bent metal plate. Plastic foil trimmers of 6 pF are used to provide the required Pi capacitance, and are often used in this frequency range. This type of matching is very suitable for alignment using amateur means. The circuit diagram of the preamplifier is shown in **Figure 2**.

Now to the alignment of the operating point of the individual transistor stages. According to the manufacturer's data the following values are most favorable for the transistor type BFR 34 A at a frequency of 1300 MHz:

Minimum noise:	$U_{CE} = 8 \text{ V}; I_C = 3 \text{ mA}$
Maximum gain:	$U_{CE} = 10 \text{ V}; I_C = 16 \text{ mA}$

Attention should be paid that the collector current is not more than 8 to 10 mA, otherwise a considerable deterioration of the noise figure will occur. The first stage is aligned for minimum noise, and the two subsequent stages are aligned for maximum gain.

So that it is only necessary for one trimmer potentiometer to be used for the stage, the collector-emitter voltage is stabilized to approximately 9 V with the aid of zener diodes. This amounts to the mean voltage. This circuit was designed according to (5) and modified as required. By use of suitable dropper resistors in series with the trimmer potentiometers, it is possible for the current variation to be made in a limited range. This ensures that the transistors are not damaged by too high a collector current.

The individual variation ranges are given in the following table. They can differ according to the spread of the transistors and components. If other transistors are to be used on the board such as MRF 901, BFR 91, or BFR 90, it is necessary for the resistance values to be modified as required.

T 1: $R_{dr} = 390 \Omega$	$I_C = 2.8 - 6.4 \text{ mA}$
T 2: $R_{dr} = 200 \Omega$	$I_C = 4 - 13 \text{ mA}$
T 3: $R_{dr} = 200 \Omega$	$I_C = 4 - 13 \text{ mA}$

2. CONSTRUCTION DETAILS

The three-stage preamplifier is built up on a 1.5 mm thick, double-coated PC-board as shown in **Figure 3**. This board is designated DC 0 DA 004 and its dimensions are 129 mm x 66 mm. The complete preamplifier can be mounted in a TEK0 box 4 B.

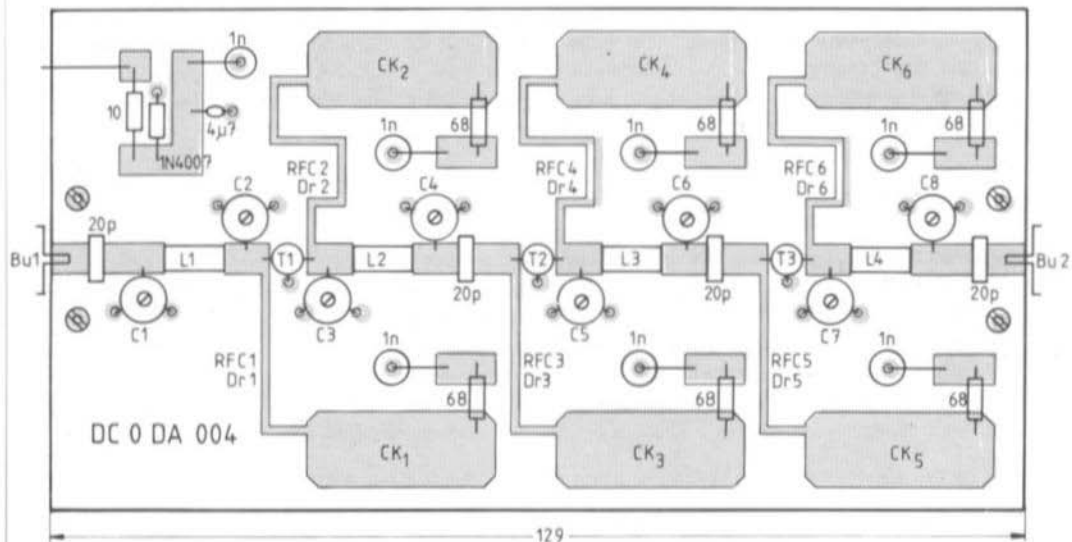


Fig. 3: Component locations on PC board DC 0 DA 004

The chokes and bypass capacitors are provided on the upper side of the board in printed technology. This considerably simplifies the construction. The completely coated lower side of the board is used as ground surface. The $\lambda/2$ Pi-circuits comprise two plastic foil trimmer capacitors and a semi-circular strip made from thin copper plate (see Figure 4). The ground pins of the trimmer capacitors are placed through 1 mm holes in the board and soldered to the ground surface on the lower side of the board. The stator pins are bent at right angles and soldered to the required conductor lane.

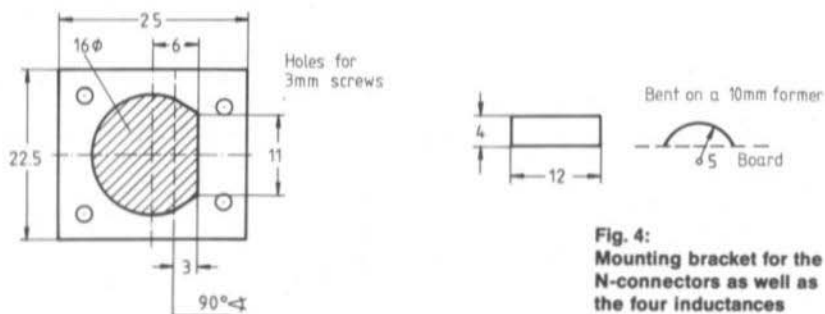


Fig. 4: Mounting bracket for the N-connectors as well as the four inductances

The three transistors are placed into holes in the PC-board so that base and collector connections are directly soldered to the conductor lanes. The emitter connection is bent down by 90° and placed through the hole in the PC-board together with the transistor and soldered into place on the lower side of the board. The collector and base of the subsequent transistors are galvanically isolated using chip capacitors between two parts of the conductor lanes. The input and output connections are also made via isolating capacitors.

The networks for the supply voltages of the three transistors are to be found below the PC-board where they are wired together. The author used bypass capacitors, which were soldered to the ground surface of the board, for mounting these components. This also ensured that the power supply voltage was bypassed. The voltage of the individual networks is fed to the transistor connections on the upper side of the board via feedthrough capacitors of approximately 1 nF and filter resistors of 56 to 82 Ω (without caps).

Special care must be taken when mounting and connecting the input and output sockets. N-connectors type UG-58/U are used.

Two brackets should be cut from thin copper plate and drilled and finished as shown in **Figure 4**. When mounting the board into the case, the two N-connectors are mounted from the inside of the side panels and are screwed into place with the previously mentioned brackets and the case. A spacing of approximately 1.5 mm should result between the plate and the connection pin. It is now possible for the board to be placed through this spacing from the side and screwed to the bracket with the aid of two 3 mm dia. screws. The connection pin of the N-connector should now be directly touching the conductor lane of the board. This somewhat complicated mounting allows the board to be easily installed and removed, and for the input and output connectors to be connected to the board in a low-reflective manner.

After completing the alignment, it is possible for the cover to be mounted into place. The result is a compact UHF preamplifier which is suitable for both portable, mobile and fixed station operation.

2.1. Components

- T 1 - T 3: BFR 34 A (Siemens); MFR 901 (Motorola); BFR 91, BFR 90 (Philips)
- L 1 - L 4: 4 mm wide copper strip, max. 0.3 mm thick bent and cut as shown in Fig. 4
- C 1 - C 8: Plastic foil trimmer, 7 mm dia. (Philips)
- 2 N-flange connectors UG-58/U
- 1 TEKO box 4 B
- 4 chip capacitors of 10 to 20 pF
- 7 feedthrough capacitors of 100 to 1000 pF; value uncritical;
as small as possible for solder fitting
- 1 feedthrough capacitor of approx. 5 to 10 nF; value uncritical; for screw mounting
- 6 carbon resistors of 56 to 82 Ω without caps; max. 10 mm long
- 6 bypass capacitors of 100 to 1000 pF, value uncritical;
(for the resistors on the ground side of the board)
- 1 resistor of approx. 10 Ω and 1/8 to 1/4 W rating (for polarity protection)
- 4 tantalum electrolytic capacitors of 4.7 to 10 μ F for at least 12 V
- 3 resistors of 1 k Ω
- 3 resistors of 200 Ω
- 3 zener diodes of 9.1 V; miniature version
- 1 silicon rectifier diode 1 N 4007 or similar
- 3 trimmer potentiometer of 470 to 500 Ω , vertical mounting, spacing uncritical
- 2 dropper resistors 200 Ω R_{dr} for T 2 and T 3
- 1 dropper resistor 390 Ω R_{dr} for T 1

3. ALIGNMENT OF THE PREAMPLIFIER

The first preamplifier stage is aligned for a current of approximately 3 mA. The subsequent stages are aligned to a current of approximately 8 mA. The alignment of the line circuits on the PC-board can only be made with the aid of an external signal. The amplifier is aligned for the best signal-to-noise ratio and not for maximum gain. The various stages should be aligned alternately until the most favorable alignment has been found.

4. BANDPASS FILTER FOR 1296 MHz

A small bandpass filter comprising $\lambda/4$ coaxial lines was developed by the author. The filter is constructed from single or double-coated PC-board material (**Figure 5**). The length of the chambers is dependent on the length of the alignment capacitors used and their commencement capacitance. If possible, this should be less than 0.5 pF. The construction of the filter can be seen in **Figures 5 and 6** and need not be described in more detail here.

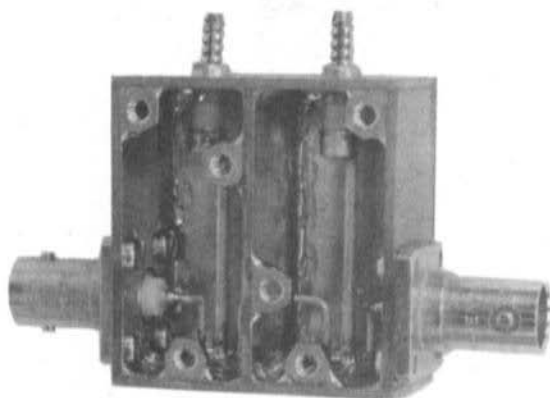


Fig. 5:
Bandpass filter for 1296 MHz

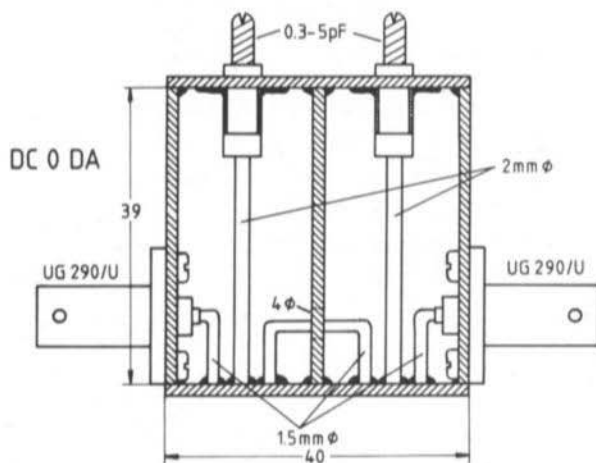


Fig. 6:
Line circuits and coupling links
spaced approximately 4 mm above
the base plate.
Height of the chamber: 20 mm;
intermediate panel made from
double-coated PC-board material

Several filters have been constructed. The insertion loss at a frequency of 1296 MHz was always in the order of 0.5 and 1.5 dB which is negligible.

The suppression of unwanted frequencies can be seen on the selectivity curve given in **Figure 7**.

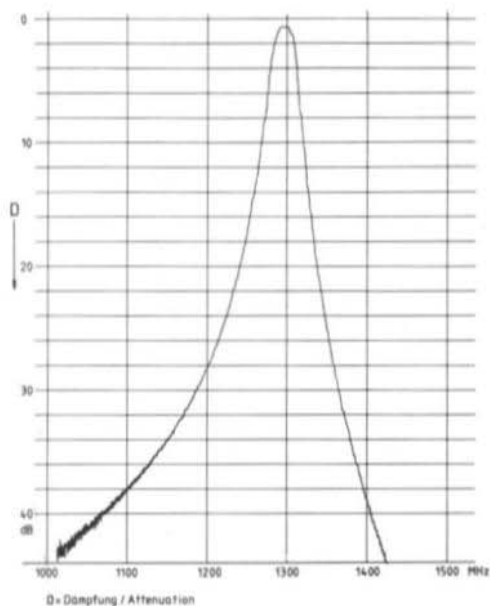


Fig. 7: Selectivity curve of the filter

The described filter should be connected between the output of the preamplifier and the input of the converter. This filter will improve the signal-to-noise ratio of converters with wideband hybrid ring mixers (1), (2), even without preamplifier. Furthermore, all other frequencies outside of the 23 cm amateur band will be suppressed.

5. RESULTS

The described preamplifier has been constructed numerous times. All versions could be aligned without difficulty. No tendency to oscillation was observed. The choke circuits for the base and collector were designed by DL 3 NQ; the author used wound chokes and chip capacitors in the original prototypes. The following measured values were obtained on an automatic noise measuring system manufactured by Hewlett-Packard:

Single-sideband noise figure:	2.5 dB
Overall gain:	21.0 dB
Image rejection at 28 MHz IF:	7.0 dB

The excellent noise figure seems to come from a very good transistor, since the manufacturer guarantees a maximum of 3.5 dB. The passband curve of the filter was measured using a network analyzer (100 MHz to 12 GHz).

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- (2) 1296 MHz Converter MMC 1296/28 Manufactured by Microwave Modules
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VHF COMMUNICATIONS 7, Edition 4/1975, Pages 217 - 235
- (5) H.P. Shuch: A Low-Cost 1296 MHz Preamplifier
Ham Radio Magazine, October 1975, Pages 42 - 46

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A NEW CONCEPT FOR 2 m to 70 cm TRANSVERTERS

by E. Berberich, DL 8 ZX

The simplest method of converting a 2 m transceiver to the 70 cm band is to use a transverter in which the transmit or receive frequency is mixed with a local oscillator frequency of 288 MHz. The required transmit signal is then obtained from the sum of the 144 MHz and 288 MHz signals which amounts to 432 MHz. In the receive mode, the difference of the 432 MHz and the 288 MHz local oscillator frequency is used which amounts to 144 MHz. The difficulty when using this direct method is that the transmit mixer does not only mix but also triples, which means that the tripled frequency of 144 MHz is also present as an unreadable (SSB) signal. Since this unwanted third harmonic cannot be filtered out because it falls into the required signal bandwidth, early 70 cm transverters used a local oscillator frequency not directly coinciding with 288 MHz – e.g. 287.5 MHz in the case of (1). This meant that the mixed and tripled frequency were no longer present at the same frequency in the 70 cm band and that the transmission was no longer interfered with. However, the transverter continued to radiate two frequencies, and interference to other stations was therefore to be expected. For this reason, this method can no longer be recommended.

For this reason, more modern equipment used a double conversion system using a frequency plan of, for instance, 144 MHz – 336 MHz – 432 MHz (2), or 144 MHz – 28 MHz – 432 MHz (3). This allowed the unwanted harmonics to be filtered out in the transverter during the frequency processing so that they were suppressed by more than 30 dB. However, the circuitry required for this is rather excessive and additional spurious signals can result in the case of non-ideal construction due to the additional conversion process. For this reason, a new, simple concept is to be discussed that is especially attractive when a 2 m transverter is to be constructed. As far as the author knows, this concept has not been published previously.

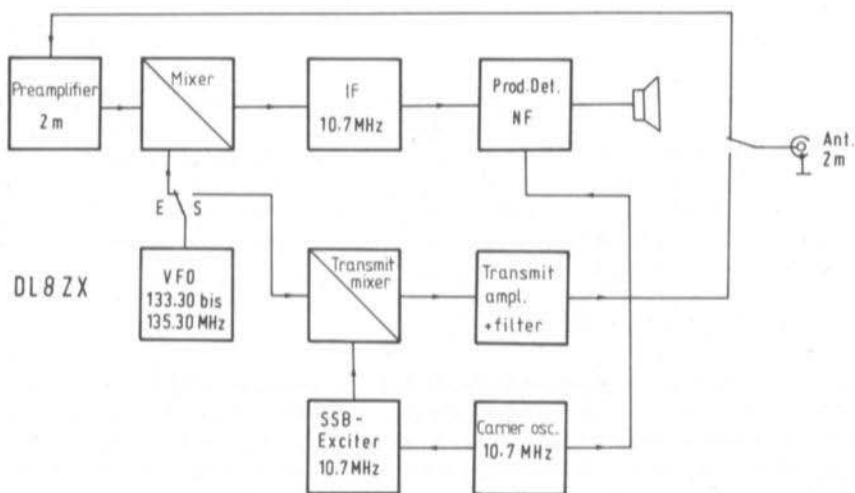


Fig. 1: Block diagram of a 2 m transceiver

1. A DIFFERENT FREQUENCY PLAN

The new concept is to be discussed with the aid of the block diagram of a 2 m transceiver given in **Figure 1**: Two different frequencies are generated in the transmit mixer: the required frequency of $133.3 \text{ MHz} + 10.7 \text{ MHz} = 144 \text{ MHz}$, as well as the image frequency of $133.3 \text{ MHz} - 10.7 \text{ MHz} = 122.6 \text{ MHz}$. This frequency range of $122.6 - 124.6 \text{ MHz}$ in the case of a 10.7 MHz IF can be filtered out, amplified to approximately 5 mW and fed to a (ring) mixer where it is mixed with a local oscillator frequency of 309.4 MHz . This results in a sum frequency of $432 \text{ MHz} - 434 \text{ MHz}$ on the 70 cm band. When using other frequencies, the same principle can be used for other intermediate frequencies, e.g. for 9 MHz . The frequency range of 122.6 to 124.6 MHz can also be used for the 23 cm band.

However, since this frequency range falls into the aeronautical communications band, special attention must be paid to the construction and filtering so that no spurious radiations are made in this band. It is possible to avoid this by placing the frequency of the local oscillator above the required 2 m frequency instead of below. This means that the local oscillator should operate at 155.7 MHz at the centre of the 2 m band (145 MHz). The image frequency used to obtain the 70 cm signal is now located at 166.4 MHz . The required local oscillator frequency should then correspond to 266.6 MHz . The previously mentioned frequency plan is shown more clearly in the form of a scale in **Figure 2**. The first concept is shown above the line and the second concept below.

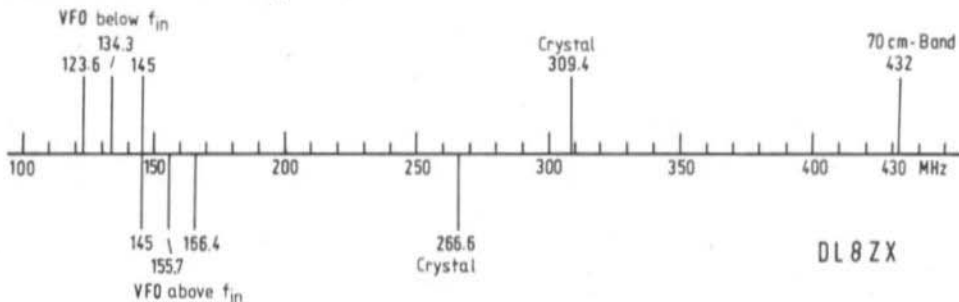


Fig. 2: Frequency plan for the described concept

In order to ensure that exact transceive conditions exist, the receive signal should be converted using the same oscillators as the transmit signal. This means that the intermediate frequency of $122.6 - 124.6$, or approximately 166.4 MHz respectively, are also used for reception. The great advantage of this frequency plan is that the 2 m band is completely avoided and that no strong 2 m stations can break through.

2. RECOMMENDED CONSTRUCTIONS

The block diagram given in **Figure 3** shows how the suggested frequency concept can be realized in practice. The mixers are equipped with Schottky ring mixers (IE-500) of 50Ω impedance which are switched with cheap miniature relays (e.g. RH-12) to the various amplifiers and filters. The frequencies resulting when the VFO is operative above the 2 m band are given in brackets. This recommendation can be realized relatively simply during the construction of a $2 \text{ m} / 70 \text{ cm}$ transceiver, and the details given are no doubt sufficient for those amateurs intending to carry out such a project.

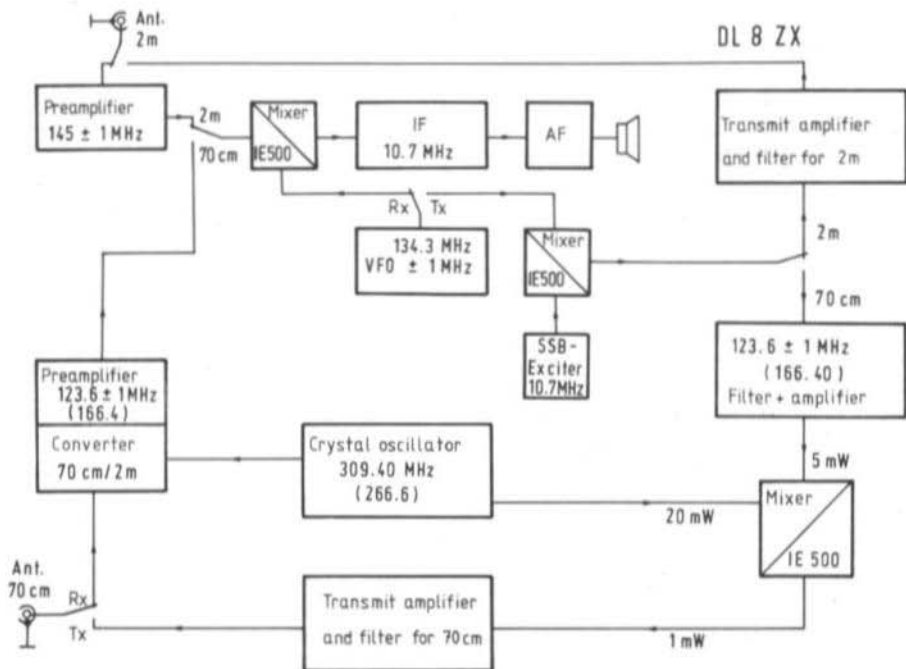


Fig. 3: Block diagram of a 2 m / 70 cm transceiver

It is not simple to give details for modifying existing 2 m transceivers since the frequency plan differs greatly. However, it seems possible that the receiver input stages and the transmit mixer for 70 cm to be retuned to 123.6 MHz with the aid of varactor diodes. However, sufficient room should be available in the transceiver for the required filtering, and measuring instruments are required that are not usually available to the radio amateur. It is probably more favorable to leave the 2 m input stage in its original condition and to provide the 123.6 MHz preamplifier stage in the 70 cm converter. **Figure 4** shows a typical circuit of a preamplifier and mixer circuit of a 2 m transceiver. **Figure 5** shows how this circuit can be modified and extended to accommodate the described double function.

The mixer circuit is provided with a varactor diode (e.g. BB 105); the original parallel capacitance may have to be decreased. The cathode of the varactor diode is not completely bypassed but via approx. 33 pF. This ensures that a low-impedance RF output is provided at this position. The 100 kΩ resistor for the operating voltage is therefore effective as a choke and should possess short connections.

2 m operation: In this mode, the varactor diode is provided with the full operating voltage so that it obtains the lowest capacitance of approximately 3 pF. The mixer circuit is now brought to resonance at 145 MHz by tuning the inductance. This means that the original conditions exist for 2 m operation, since the capacitive tap of the circuit is at very low impedance. In addition to this, the preamplifier transistor for the intermediate frequency of 123 MHz is not provided with any operating voltage in the 70 cm converter. This means that its drain circuit is short-circuited at low impedance and has no effect.

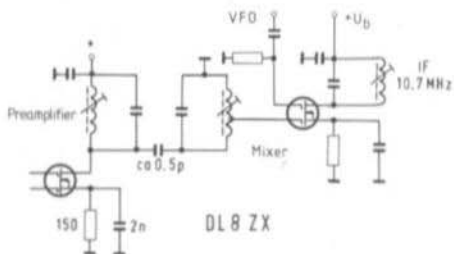


Fig. 4:
Conventional preamplifier and mixer of a 2 m receiver

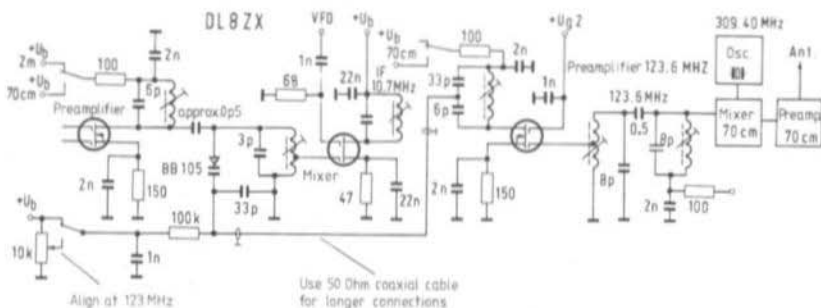


Fig. 5: Switchable preamplifier and mixer stage for 123/144 MHz

70 cm operation: In this mode, the operating voltage of the 145 MHz preamplifier is switched off and fed to the 123.6 MHz preamplifier in the 70 cm converter. The drain circuit of the 2 m preamplifier is now dampened at low impedance by the transistor so that the coupling capacitance of approximately 0.5 pF is effective as additional parallel capacitance for the mixer circuit. In addition to this, the varactor diode is fed with a lower voltage from the trimmer potentiometer so that its capacitance is increased and the circuit brought to resonance at 123.6 MHz. This means that the 70 cm converter is active via the 123.6 MHz preamplifier and is fed to the original 2 m mixer, which is now tuned to 123.6 MHz.

In order to ensure that the full 2 MHz wideband is covered, the alignment should be made, if possible, with the aid of a sweep generator. The 123.6 MHz bandpass filter between the preamplifier and mixer stage possesses a capacitive base coupling; the degree of coupling and thus the bandwidth are adjusted by varying the capacitance of the two capacitors (33 pF in the circuit).

The transmit circuit can be modified in a similar manner. Of course, it would be possible for switching diodes to be used instead of relays, as is often the case in TV tuners. It is also possible for the scale of the 2 m transceiver to be used as before.

3. REFERENCES

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- (2) W. Rahe: A Linear Transverter for 2 m / 70 cm with Double-Conversion
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A MODULAR ATV TRANSMITTER WITH VIDEO AND AUDIO MODULATION AT IF LEVEL

Description of a complete ATV transmitter from previously published, improved modules and one new module.

by G. Sattler, DJ 4 LB

The following article is to summarize previously published modules for ATV and describes the interconnection of these modules to form an ATV transmitter with an output power of approximately 1.5 W on the 70 cm band. Two modified modules, e.g. DJ 4 LB 001a and DJ 4 LB 002a, as well as the new module DJ 4 LB 007 are to be introduced. The more important measured values for these three modules are to be given.

1. PREVIOUSLY DESCRIBED MODULES

The basic modules of an ATV transmitter with video and audio modulation at IF level were described in (1): video and sound IF module, local oscillator module, and 70 cm mixer. Other modules were described later such as the video/sound coupler, and an AFC circuit for the sound IF in (2), a vestigial sideband filter in (3), as well as a linear amplifier for 435 MHz in (4). For clarity, these references and the modules described therein are as follows:

- | | |
|-------------------------|-----------------|
| (1) DJ 4 LB 001 - 005 | (3) DJ 6 PI 004 |
| (2) DJ 6 PI 002 and 003 | (4) DJ 4 LB 006 |

1.1. Modified Modules

The possibility of using the smaller coil sets D 41-2165 instead of the previously used larger types D 21-1634 was the reason for modifying PC-boards DJ 4 LB 001 and 002. This allowed the height of the components on the boards to be reduced from approximately 25 mm to 15 mm, which meant that the smaller TEKO box 4 A would be used instead of the higher TEKO box 4 B. Since some circuit modifications were also carried out to improve the electrical characteristics, the modified modules are to be described completely.

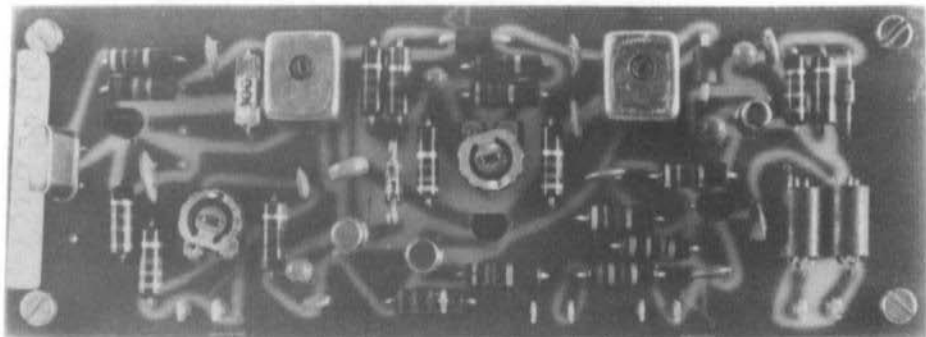


Fig. 1: Photograph of the author's prototype DJ 4 LB 001a

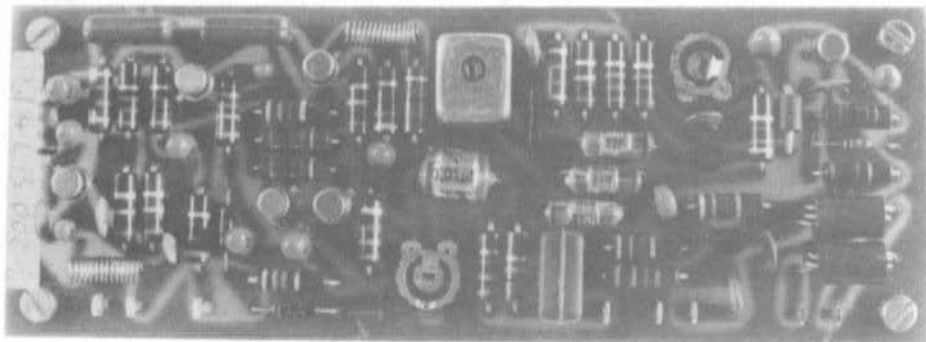


Fig. 3: Photograph of the author's prototype DJ 4 LB 002a

As can be seen in the circuit diagram of the new module DJ 4 LB 002a given in **Figure 4**, the microphone preamplifier has not been modified.

The double varactor diode type BB 104 (D 203) used in the oscillator for frequency modulation and AFC brings a number of advantages over the previously used single diode BA 124. The common cathode of the double diode is connected via the RF-decoupling resistor R 218 to a point in the circuit where it receives a stabilized DC-voltage of approximately 6 V via R 220, the audio signal via C 214 - R 217, and the AFC voltage via R 219. The standardized preemphasis of the AF signal is mainly made in the CR-link comprising C 214 / R 219. In order to ensure that it is able to fulfill its function, the other side of resistor R 219 is bridged via C 216 to the stabilized operating voltage and is thus short-circuited for audio frequencies.

The AFC input Pt 206 receives a DC-voltage of approximately 6 V via the series circuit comprising resistors R 220 and R 219. This means that it is suitable for direct connection to the AFC output of an integrated FM circuit. Normally, approximately half the operating voltage is present at such an output when the frequency coincides, e.g. 6 V for IC type CA 3089 E in module DJ 6 PI 003. When the frequency does not coincide, AFC voltages of between 3 to 9 V (CA 3089 E) are obtained, which will be superimposed on the DC-voltage of 6 V at the AFC input of the audio-IF module DJ 4 LB 002a. Due to the voltage division across resistors R 219 and R 220, approximately 90 % of the actual AFC voltage will be effective at the varactor diodes.

1.1.2.1. Alignment of Module DJ 4 LB 002a

It is very helpful if a frequency counter is available during the alignment of the sound IF oscillator since the core of inductance L 203 can cause a frequency variation of virtually 1 : 2. In contrast to the original version, the AFC input should not be grounded during alignment and operation. Its input should remain free, or connected to the AFC circuit in the AFC mode..

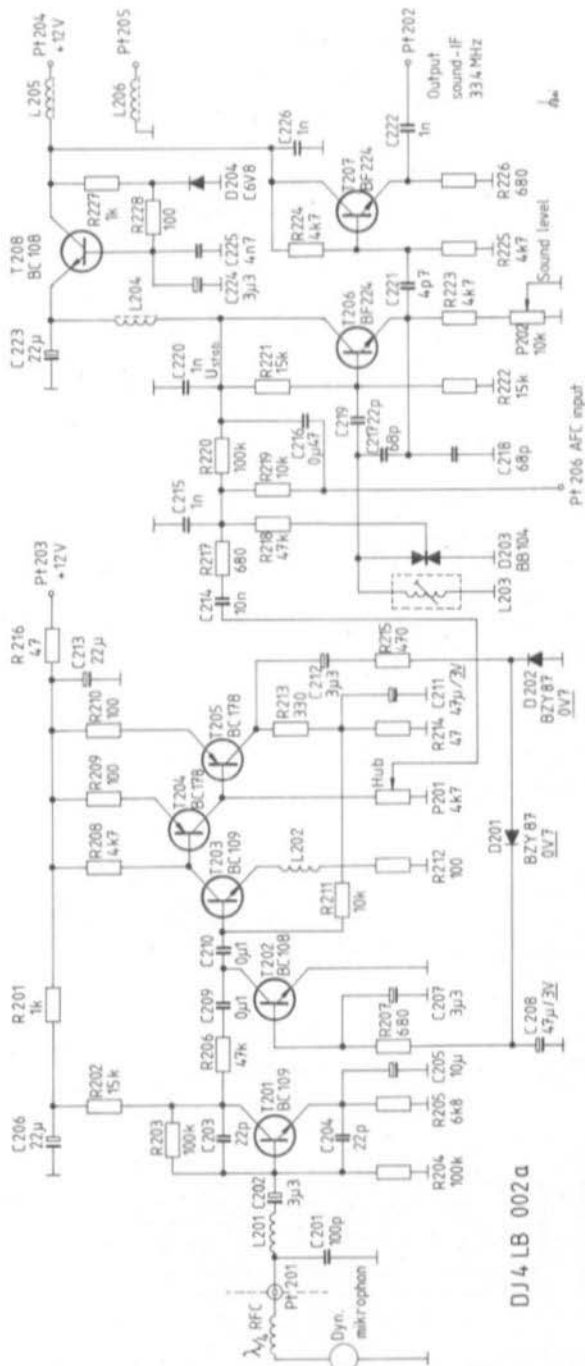
1.1.2.2. Characteristics of Module DJ 4 LB 002a

The given values represent mean values from measurements made on several prototypes.

Current drain at 12.0 - 12.5 V

Microphone amplifier: 16 mA

Sound IF oscillator: 9 mA



DJ 4 LB 002 a

Fig. 4: Circuit diagram of the sound-IF module DJ 4 LB 002a

- L 201, L 202: approx. 17 cm of 0.4 mm dia. (26 AWG), coil dia. 3 mm, self-supporting, length 10 mm
 - L 203: 8 turns of 0.2 - 0.3 dia. (32 AWG) enamelled copper wire in special coil set
 - C 217, C 218: 68 pF styroflex capacitor
 - C 219: 22 pF styroflex capacitor
- All electrolytic capacitors: tantalum drop types for 16 V when not given otherwise

Microphone input Pt 201:

for dynamic microphones with 200 - 500 Ω impedance, noise-matched

RF output voltage at Pt 202, adjustable with the aid of potentiometer P 202:

120 - 380 mV peak-to-peak = 40 - 135 mV RMS into 60 Ω

Frequency shift with above level adjustment: approx. 20 kHz

Frequency variation on adjusting the DC-voltage (6 V) at D 203: approx. 270 kHz/V

Harmonic component of the output voltage: second harmonic: - 32 dB

third harmonic: - 47 dB

2. THE NEW IF DEMODULATOR WITH VIDEO/SOUND COUPLING DJ 4 LB 007

This module is able to take over important functions in a modular ATV transmitter and is therefore to be described in detail.

2.1. Block Diagram

As can be seen in the block diagram given in **Figure 5**, module DJ 4 LB 007 has three, equal priority IF inputs: Pt 701, Pt 702 and Pt 703, each having approximately 60 Ω impedance. In the transmit mode, the video and sound IF signals for modules DJ 4 LB 001 (a) and 002 (a) are connected to two of the inputs. The third IF input can be used for ATV reception. The input signals are coupled in via a resistor network for decoupling and feed two independent amplifiers.

The amplifier with transistors T 1 and T 3 is in the signal path of the transmitter. The IF level at the output (Pt 705) can be adjusted to the most favorable value for the subsequent UHF mixer or linear amplifier with the aid of a DC-voltage, so that it can be remote controlled.

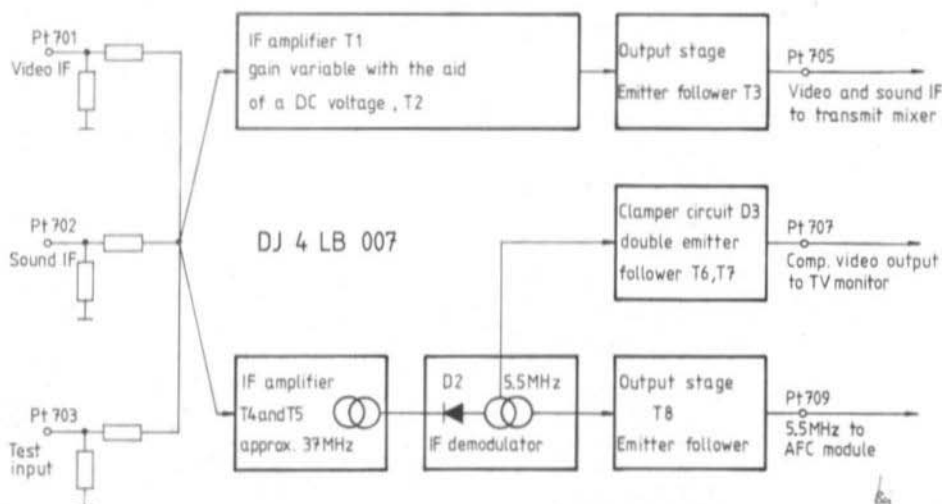


Fig. 5: Block diagram of module DJ 4 LB 007

In the second amplifier, the input transistor T 4 drives an amplifier stage equipped with transistor T 5 which roughly corresponds to the last IF stage of a television receiver. A composite video signal as well as the intercarrier frequency (difference frequency) between the video and sound IF of 5.5 MHz are generated in the IF demodulator diode D 2. The frequency modulated 5.5 MHz sound carrier is filtered out from the composite video signal in a bandpass filter. For AFC and sound demodulation, the signal is fed via the emitter follower comprising T 8 to Pt 709.

The composite video signal with suppressed 5.5 MHz component is fed via a two-stage emitter follower (T 6, T 7) with clamper circuit (D 3) and is available at low impedance at Pt 707. If a monitor is connected to this composite video output, it will be possible to monitor the transmitted signal without interference.

If the output signal from a tuner and IF amplifier of the receiver is connected to the other IF input, the video signal of the received signal will also be fed to the monitor.

2.2. Circuit Description

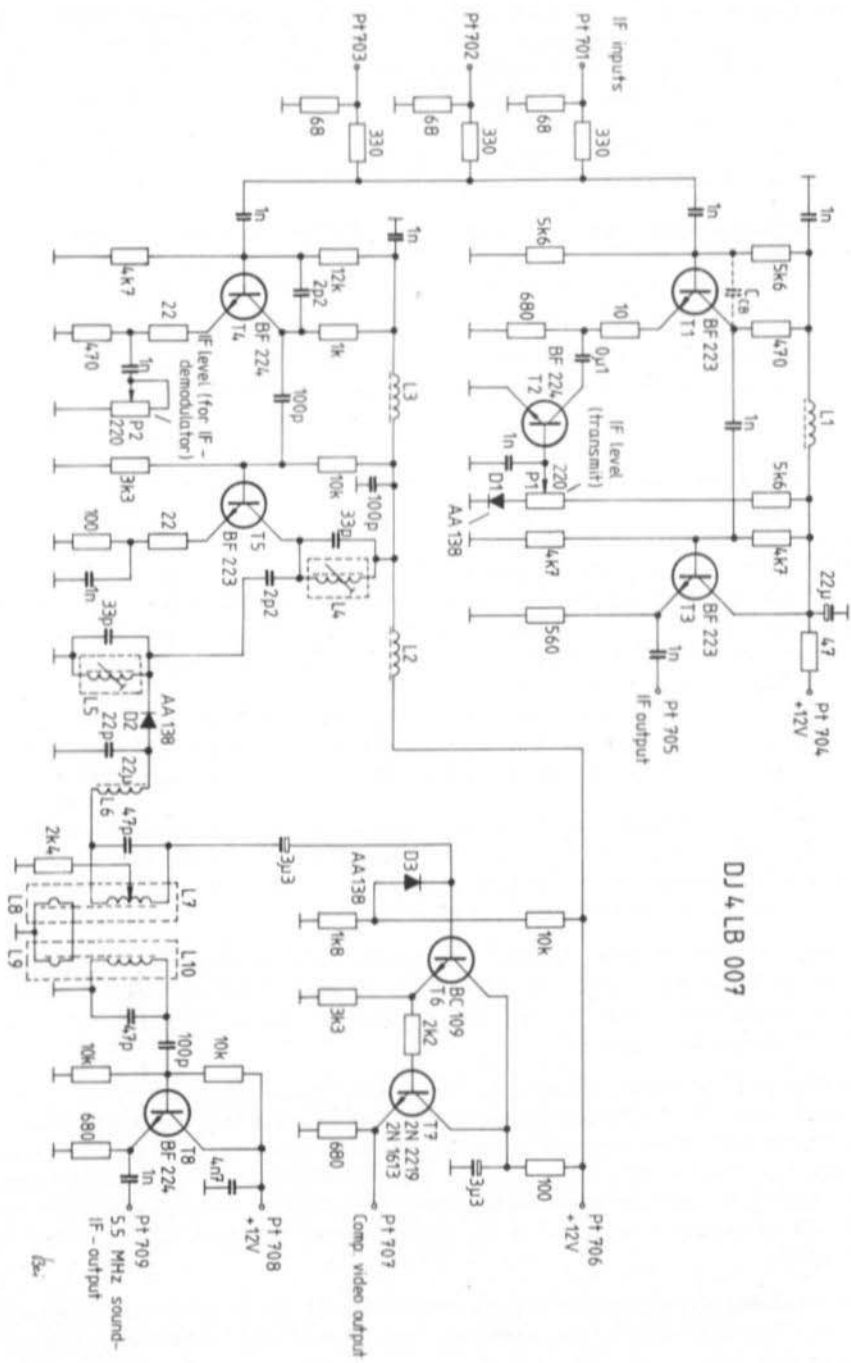
The circuit diagram of the IF demodulator with video/sound coupling (DJ 4 LB 007) is given in **Figure 6**. The amplifier stage equipped with T 1 compensates for the level loss of the interconnection of the inputs via the resistors. The $680\ \Omega$ emitter resistor of transistor T 1 is connected in parallel to the collector emitter path of transistor T 2 with respect to high frequencies. This means that the gain of T 1 can be adjusted by placing a DC-voltage at the base of T 2. The germanium diode D 1 in series with potentiometer P 1 ensures, together with a dropper resistor, a virtually linear relationship between the position on the potentiometer and the RF voltage. **Figure 7** shows the video and sound IF signals at the amplifier inputs in a ratio of 3:1, as well as the combined video/sound IF signal at the output (below).

The amplifier chain comprising transistors T 4 and T 5 works into an undercritically coupled bandpass filter comprising L 4 and L 5, whose passband curve is given in **Figure 8**. The sound carrier is approximately 20 % and the video carrier approximately 50 % of the voltage that it present at the filter at its center frequency of 37 MHz.

The high-frequency slope of the filter corresponds to the Nyquist slope in a TV receiver. The simple passband curve means that practically no pulse distortion of the composite video signal takes place, as can be seen in **Figure 9** in the care of the shape of a demodulated 250 kHz squarewave signal.

Together with the resistor at the center tap, which is also used as load resistor of the demodulator diode D 2, the resonant circuit comprising L 7 forms a bridge filter that completely suppresses the 5.5 MHz components of the composite video signal. The resonant circuit comprising L 10 is coupled in via L 8 and L 9 and extends the bridge filter to form a bandpass filter for 5.5 MHz with a 3 dB bandwidth of approximately 500 kHz. The passband curve of this filter as measured at the output of the emitter follower comprising T 8 (Pt 709) is given in **Figure 10**.

The two-stage emitter follower with clamper circuit for the composite video signal allows a $60\ \Omega$ or $75\ \Omega$ termination at the composite video output Pt 707; a decoupling capacitor is not required.



DJ 4 LB 007

Fig. 6: Circuit diagram of the IF-demodulator module

6x.

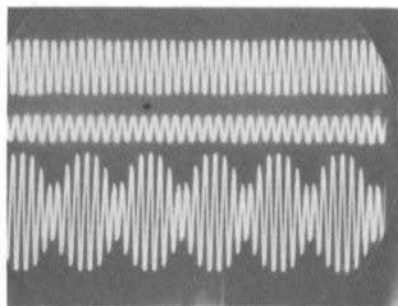


Fig. 7:
Video IF and sound IF
individually and combined

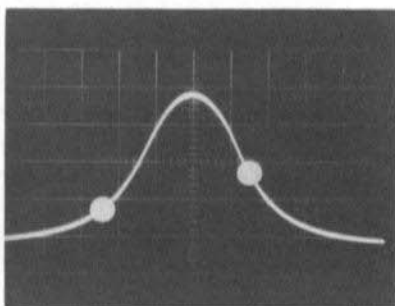


Fig. 8:
IF-bandpass curve. Markers:
Sound carrier (left),
video carrier (right)

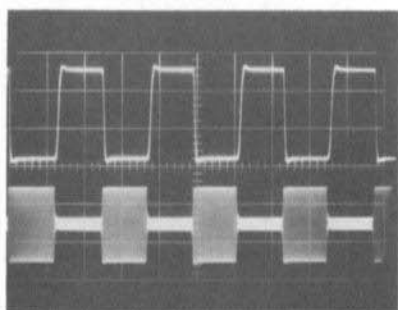


Fig. 9:
Below: Video IF carrier with
250 kHz squarewave modulation
Above: Demodulated squarewave
signal at composite video output

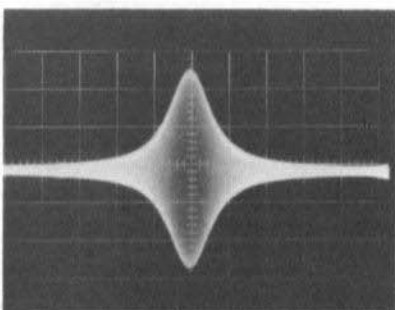


Fig. 10:
Passband curve of the
5.5 MHz filter

2.3. Construction and Alignment Details

The PC-board of module DJ 4 LB 007 also has the dimensions 135 mm x 50 mm and is single-coated (see **Figure 11**). This module can also be mounted in a TEKO box 4 A. A photograph of the author's prototype is given in **Figure 12**.

Any spread of transistor T1 can be compensated for with the capacitor C_{CB} and will, if necessary, suppress any tendency to oscillation. A value between 0 and 3.3 pF is suitable according to the IF level and the required degree of feedback.

If P1 is mounted vertically, it will be necessary for the wiper to be connected using a wire bridge (two solder connections are provided on the PC-board). These wire bridges can be connected via the break contact of a switching socket. This type of wiring allows the IF level to be either adjusted internally or with the aid of an external DC-voltage fed to the switching socket.

The bandpass filter comprising L4 and L5 can be aligned even without sweep generator by feeding in a 37 MHz signal and aligning both circuits for maximum (negative) DC-voltage at

- L 6: RF-choke, approx. 22 μ H
- L 7: 2 x 26 turns bifilar wound, 0.2 mm (32 AWG) enamelled copper wire in special coil set as for L 4, L 5, center tap
- L 8: Coupling winding on L 7; 1 turn of 0.2 mm (32 AWG) enamelled copper wire wound between L 7 and the connections of the coil former. Leave off the ferrite cap of the core, otherwise coupling will be too tight.
- L 9: Coupling winding for L 10; 1 turn wound on L 10. Use ferrite cap-
- L 10: 52 turns of 0.2 mm (32 AWG) enamelled copper wire in special coil set as for L 4, L 5, L 7

Resonant circuit capacitors parallel to L 4 and L 5: 33 pF, styroflex

Resonant circuit capacitors parallel to L 7 and L 10: 47 pF, styroflex

2.5. Characteristics of Module DJ 4 LB 007

The given values are mean values from measurements made on several prototypes.

Current drain at 12.0 - 12.5 V without signal: 45 mA

with composite video signal: 65 mA when terminated with 60 Ω at Pt 707.

RF output voltage at Pt 705 when module DJ 4 LB 001a and 002a are connected to the inputs: approx. 60 - 500 mV peak-to-peak = 20 - 80 mV RMS into 60 Ω , adjustable with P 1

Spectrum of the RF output signal (intermodulation) referred to the video IF carrier at 38.9 MHz:

5.5 MHz: - 36 dB

27.9 MHz (= 33.4 - 5.5 MHz): - 48 dB

44.4 MHz (= 38.9 + 5.5 MHz): - 40 dB

Output voltage (composite video at Pt 707):

0.3 - 1.5 V peak-to-peak into 60 Ω , adjustable with P 2

Output voltage of the sound IF (Pt 709):

approx. 10 mV peak-to-peak value = 3 mV RMS into 60 Ω .

3. BLOCK DIAGRAM OF A COMPLETE ATV TRANSMITTER

The extended block diagram given in **Figure 13** shows how the previously mentioned modules can be combined to form a complete ATV transmitter.

Figure 14 shows the front view of a ATV transmitter. The case dimensions are 305 x 105 x 260 mm. **Figure 15** shows the arrangement of the TEKO box modules in the case. The numbers on the narrow edges of the boxes correspond to the last numeral of the module.

3.1. Video/Sound Coupling

The video and sound IF coupling can be made either in the video/sound coupling stage DJ 6 PI 002 or in the IF demodulator with video/sound coupling DJ 4 LB 007. Module DJ 4 LB 007 generates the 5.5 MHz sound carrier internally, e.g. at IF level. If module DJ 6 PI 002 is used for video/sound coupling, the intercarrier frequency of 5.5 MHz can be obtained using the UHF demodulator DJ 4 LB 005.

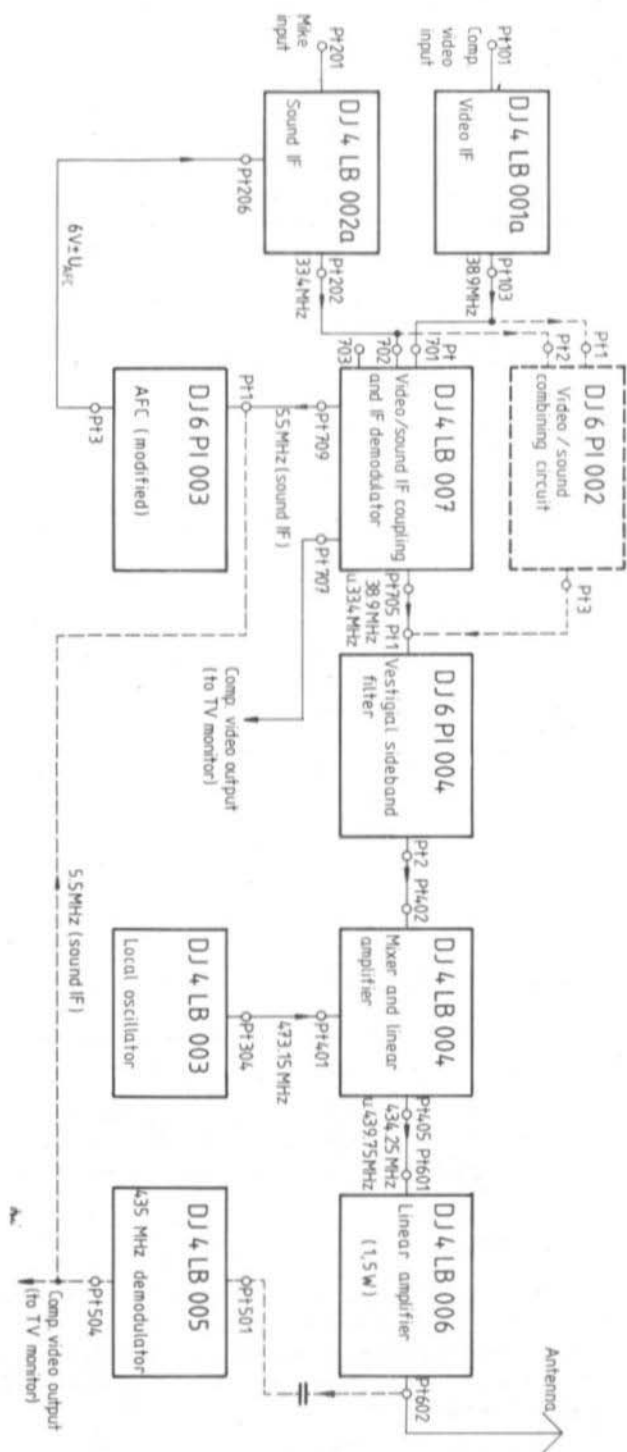


Fig. 13: Extended ATV-transmitter, block diagram

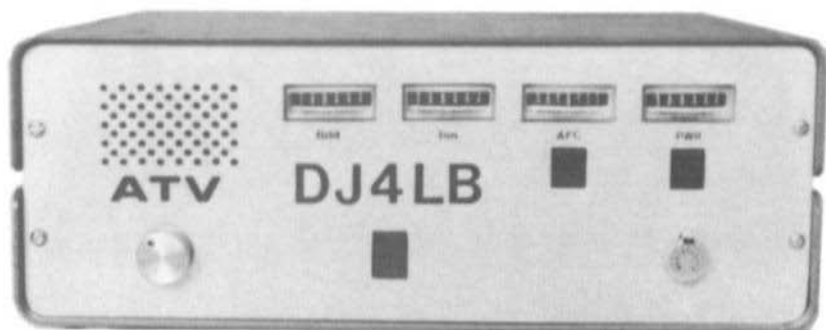


Fig. 14: Front view of the ATV-transmitter

3.2. AFC; Modified AFC Module DJ 6 PI 003

The AFC module DJ 6 PI 003 was designed as accessory for the sound IF module DJ 4 LB 002. It can also be used for controlling the frequency of the new sound IF oscillator in module DJ 4 LB 002a, but only after modification as will be seen in the following consideration:

If the sound IF oscillator is operating lower than 33.4 MHz, for instance due to some temperature effects, this means, that the DC-voltage at the varactor diode must become more positive in order to control the frequency error. In the original version of the AFC module DJ 6 PI 003, a lower DC-voltage would be generated than when the signal is at the nominal frequency. The integrated circuit CA 3089 E does not possess an inverted AFC output, but it is still possible to invert the AFC voltage in a simple manner: it is only necessary for the RF choke L 4 in module DJ 6 PI 003 to be replaced by a capacitor of 2.2 - 3.3 pF. This causes a phase shift of 180° at the reference circuit of the coincidence demodulator and causes a reversal of the AFC voltage curve. The damping resistor R 2 should not be used in this modification since this would cause the AFC control voltage to be too low.

3.3. Level Adjustment

The power ratio between video and sound carrier of a TV transmitter is standardized at 10 : 1. This can be achieved by aligning the sound IF voltage at Pt 202 to one third of the video IF voltage at Pt 103. The potentiometers in the video/sound coupling stages DJ 6 PI 002 and DJ 4 LB 007 effect the level of the coupled video/sound signal and allow the drive to the following stages such as mixer and linear amplifier to be varied without changing the power relationship between the video and sound carrier.

3.4. Vestigial Sideband Filter

The vestigial sideband filter DJ 6 PI 004 can easily be placed between the video/sound coupling stage and the UHF mixer. No level variations result since the insertion loss for the pass-band is very small. The filter does not only suppress the vestigial sideband but also harmonics and intermodulation products of the IF-signals.

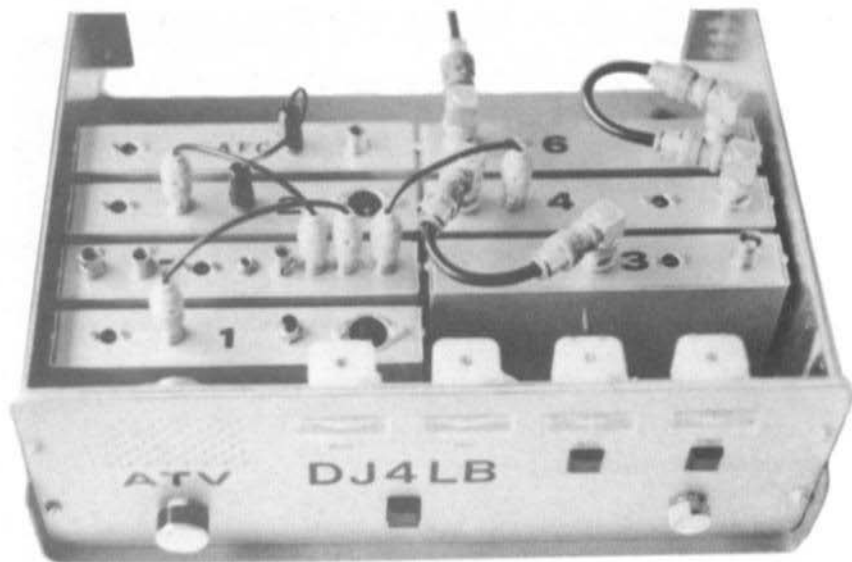


Fig. 15: ATV-transmitter from above with power supply removed

3.5. Improvements to Linear Amplifier DJ 4 LB 006

An inductance with three to four turns of 0.4 mm diameter (26 AWG) enamelled copper wire, coil diameter 3 mm, self-supporting, is sufficient for choke L 5 in the collector circuit of T 1 (originally a $\lambda/4$ -choke). If 100Ω is selected for P 2 instead of 220Ω , transistor T 2 can only be slightly shifted towards class B operation under high RF drive levels.

4. REFERENCES

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VHF COMMUNICATIONS 5, Editions 1 and 2/1973
- (2) J. Grimm: ATV Information
VHF COMMUNICATIONS 8, Edition 2/1976, Pages 90 - 95
- (3) J. Grimm: A Vestigial Sideband Filter for ATV
VHF COMMUNICATIONS 9, Edition 1/1977, Pages 14 - 18
- (4) G. Sattler: A Two-Stage ATV Linear Amplifier
VHF COMMUNICATIONS 9, Edition 1/1977, Pages 10 - 13

A TRANSCEIVER FOR 10 GHz

Part 3

by B. Heubusch, DC 5 CX · Dr. Ing. A. Hock, DC 0 MT · H. Knauf, DC 5 CY

7. CONSTRUCTION OF A TRANSCEIVER

There are various concepts for 10 GHz transceivers; one version is to be described here that was recommended by G 3 RPE, and that has proved itself for such communications. The principle is as follows: A microwave oscillator provides the transmit energy which is then radiated by the antenna. On the way to the antenna, a small portion of the oscillator power is tapped off via a directional coupler and fed simultaneously together with the receive signal from a second antenna to a mixer. The receive signal and oscillator frequency are mixed and the resulting intermediate frequency is filtered out, amplified and fed to a VHF-FM broadcast receiver. **Figure 23** shows the basic circuit.

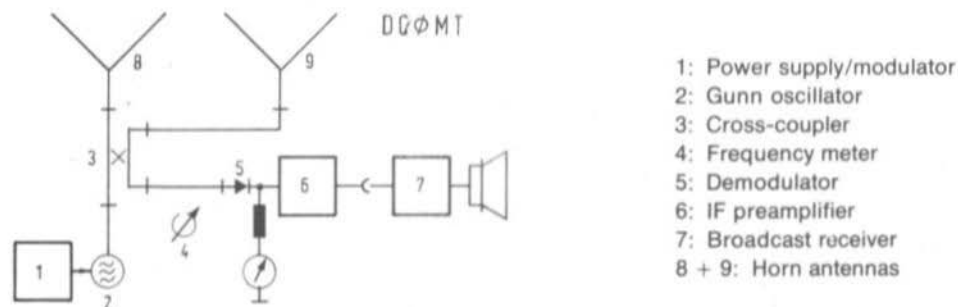
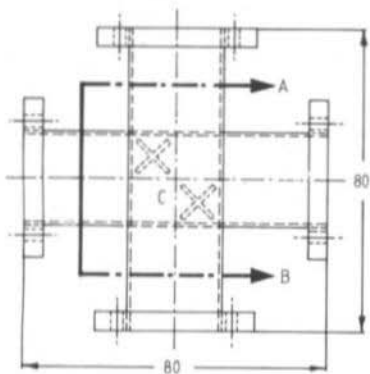


Fig. 23: Principle diagram of the 10 GHz transceiver

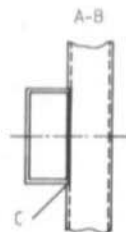
Since it is sometimes difficult to align the two antennas, a different version has been designed by the authors using only one antenna and a waveguide switch. This considerably simplifies the alignment of the antenna, especially when a high-gain parabolic is used. However, an expensive or difficult to obtain change-over switch is required. In addition to its function as change-over switch, it must also provide the lowest possible reflection for the Gunn oscillator signal in the receive mode, since large frequency shifts would otherwise be present.

7.1. The Directional Waveguide Coupler

It is advisable to use a directional coupler in order to obtain the required oscillator power of between 0.1 and 5 mW (1 N 23) or 0.25 and 10 mW (Schottky diode). Different methods of coupling out part of the oscillator power are possible, but the directional coupler provides the cleanest configuration. At transmit power levels in the order of 15 mW, a coupling loss of approximately 10 dB is required; this corresponds to approximately 1/10 of the oscillator power fed to the transmit antenna.

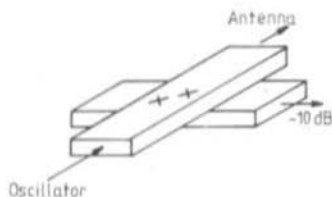
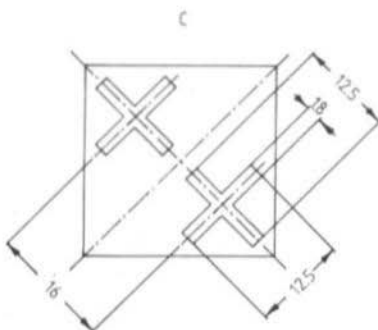


Waveguide and flange dimensions, see part 2



Mill both waveguides to a width of 25mm and 1mm deep and solder it to part C

Mat: Brass 63



Mat: Brass 63
25 x 25 x 1mm

Fig. 24: Construction details of a 10 GHz crossed-slot coupler

A coupler using crossed slots as shown in **Figure 24** is easy to construct. This results in a return loss of approximately 25 dB at a standing wave ratio of less than 1.2., which can be easily maintained in practice. This means that the frequency shift of the Gunn oscillator remains within narrow limits. Figure 24 shows the mechanical construction, and **Figure 25** the construction of this coupler. In order to avoid a large mismatch, it is important that the brass plates with the crossed slots are placed into the waveguide so that no steps are present in the waveguide path.

7.2. The Receive Mixer

When no great demands are to be placed on the mixer, the diode probe described in section 4 of part 2 can be used. The noise figure that can be obtained with an average mixer diode (1 N 23 D) is approximately 8 dB; this value can be improved to approximately 6 dB when using high-quality diodes (1 N 23 H). However, practical operation has shown that the noise figure is relatively uncritical. Considerable improvements can only be made with diode probes that can be aligned exactly for minimum standing wave ratio at the operating frequency.

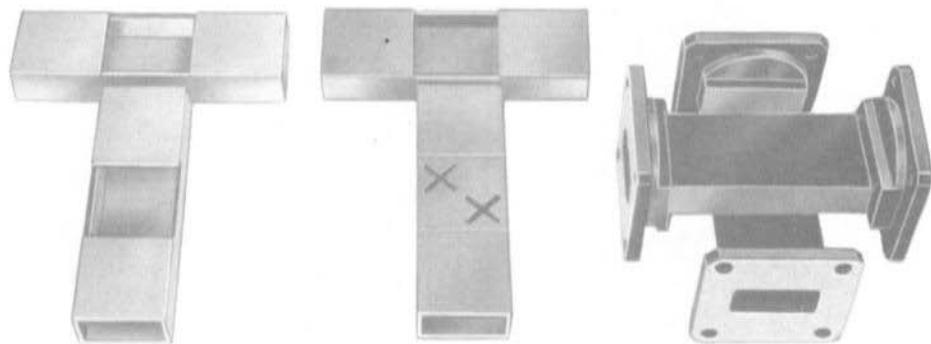


Fig. 25: Individual parts of the completed crossed-slot coupler

These components are extremely difficult for the radio amateur to construct and usually very difficult to align due to the non-existence of the required measuring equipment. This means that the described diode probe represents an acceptable compromise. In the case of a 1 N 23 diode, a diode current of approximately $600 \mu\text{A}$ should flow during operation. At this point, the oscillator power will be most favorable to obtain the lowest noise figure. However, this is not very critical as long as the current is between approximately 0.2 and 3 mA.

7.3. The IF-Preamplifier

For various reasons, it is important that the mixer diode is as well power-matched as possible on the IF side. This is a prerequisite for obtaining the most favorable noise figure and mixer slope (lowest conversion loss). Mixer diodes of type 1 N 23 require a load impedance of approximately 300Ω which must be offered by the first IF amplifier stage. This is not the case with VHF-FM broadcast receivers that are very often used as IF amplifiers, which means that an impedance converter stage will be required. Radar equipment operating in the 10 GHz range often use the field effect transistor type E 300 (Siliconix), which was specially developed for this purpose. If this FET is operated in a common gate circuit, it will offer the exact load resistance required for the mixer diode at its source input. According to the data sheets, the noise figure of this transistor at 100 MHz is far lower than that of the mixer diode which means that it will not deteriorate the overall noise figure. If a resonant circuit is provided at the input, it will be possible to attenuate some of the image noise. The overall gain of the impedance converter is not critical; approximately 10 dB can be obtained which has been found to be more than sufficient. The circuit is given in **Figure 26**, and **Figure 27** shows the PC-board of the preamplifier with the dimensions 80 mm x 48 mm.

The output coil tap is selected so that the output is matched to a conventional car radio ($Z_{in} \approx 150 \Omega$). If a different type of broadcast receiver is used as IF amplifier, it may be necessary for the coil tap to be adjusted for maximum noise. **Figure 28** shows the author's prototype of the preamplifier. Since an intermediate frequency of 104 MHz (center frequency) has been selected, it is extremely necessary for the amplifier to be screened, which is also valid for the interconnection between the diode probe and the amplifier, in order to avoid cross modulation from broadcast transmitters in the VHF-FM range.

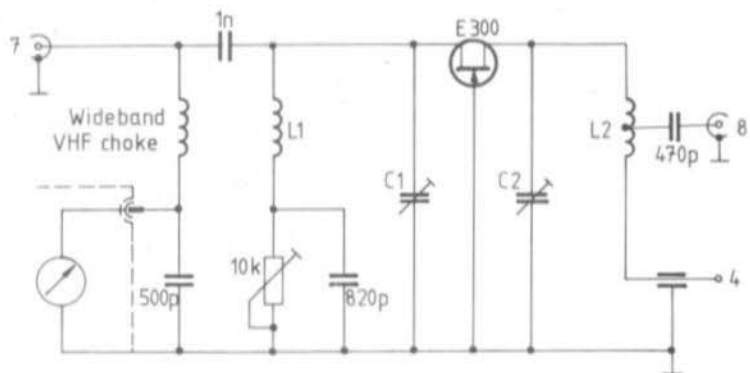


Fig. 26: Low-noise preamplifier for 100 MHz

L1,L2:
7 turns, 8mm dia.
s-p. copper wire
0.8mm dia.

C1,C2:
Air spaced
capacitors
6-21p

6a

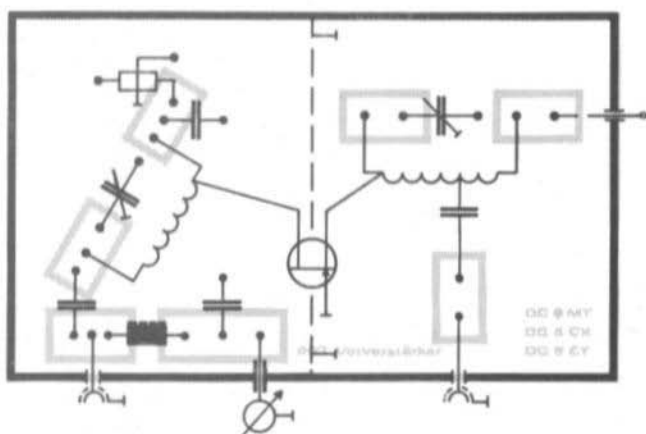


Fig. 27: PC-board DC 0 MT 003 for the 100 MHz preamplifier

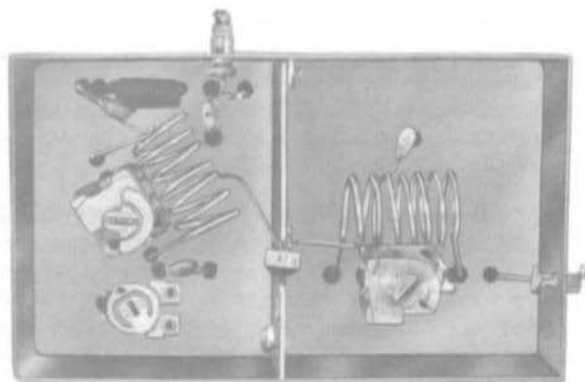


Fig. 28:
Author's prototype
of the preamplifier

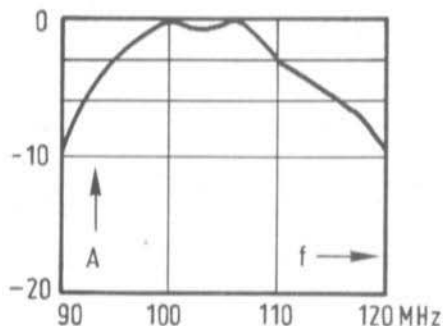


Fig. 29: Selectivity curve of the preamplifier

The preamplifier board also has a connection for measurement of the diode current. Since the directional coupler possesses a constant directivity over the whole band of 10 to 10.5 GHz, this meter can also be used as relative output meter. Conventional moving coil meters having a full-scale deflection of 1 mA can be used.

The selectivity curve of the preamplifier is given in Figure 29.

7.4. Frequency Meter

Practical operation has shown that it is very advisable when the operating frequency can be measured accurately. This can be made by placing a cavity frequency meter between directional coupler and diode probe. This consists of a half-wave resonator with variable short-circuit piston which is loosely coupled to the waveguide.

Figure 30 shows constructional details. The construction is based on normal waveguide pieces. The coupling between the two waveguides is made with the aid of a 6 mm hole. If the short-circuit piston is aligned so that a cavity wavelength of $\lambda_H/2$ results in the frequency meter (or also $3\lambda_H/2$ etc.), the cavity will be in resonance and will accept energy from the waveguide. This is visible as a noticeable dip on the diode-current meter. If a standard waveguide is used for the frequency meter, the calibration curve given in Figure 31 will result.

The accuracy that can be obtained is mainly dependent on the sharpness of the dip, which is in turn dependent on the magnitude of the coupling hole. The smaller the coupling hole, the sharper and more definite will be the dip, but it will also be shallower. The given coupling hole represents a good compromise, and accuracies of 0.001 % can be obtained which amounts to an absolute frequency accuracy of approximately 10 MHz. Prerequisites are, however, that micrometer screws having a high resolution are used. Due to the unavoidable play between the short-circuit piston and the micrometer spindle (see Figure 30), it is to be recommended that the frequency measurement is only made in one direction in order to avoid errors in the frequency indication. Of course, more expensive solutions to avoid this problem are possible. Figure 32 shows the completed frequency meter.

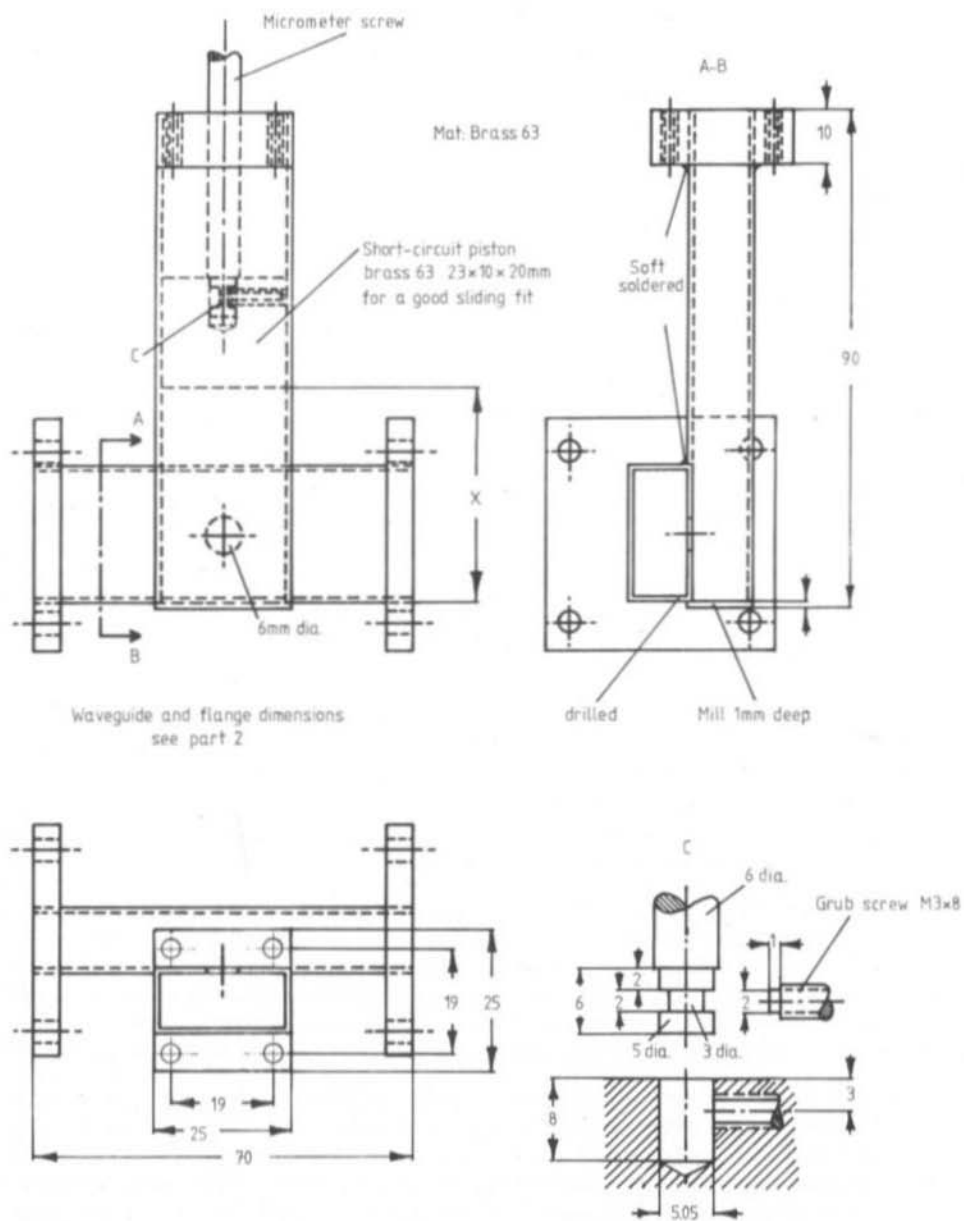


Fig. 30: Construction details of a cavity frequency meter for 10 GHz

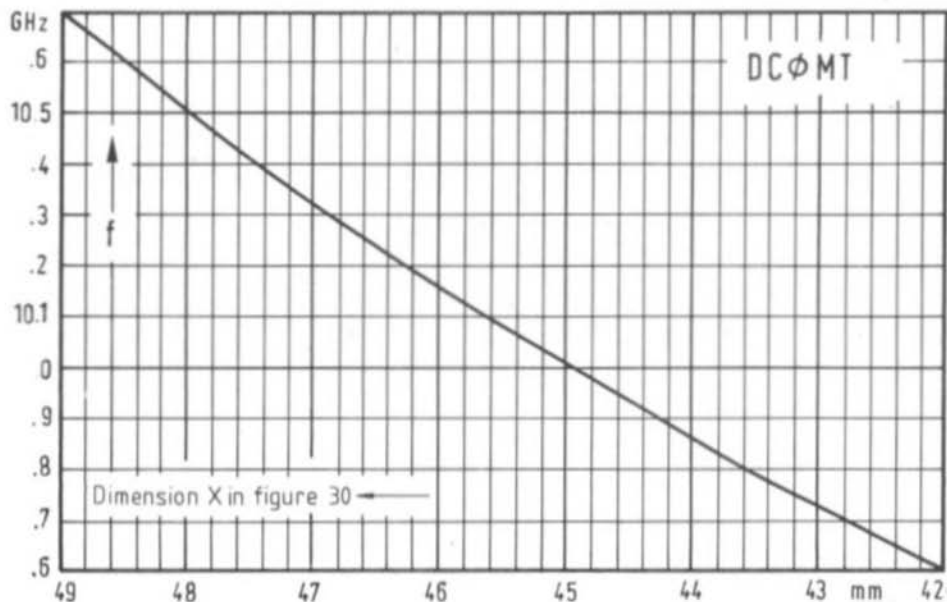


Fig. 31: Calibration curve of the frequency meter

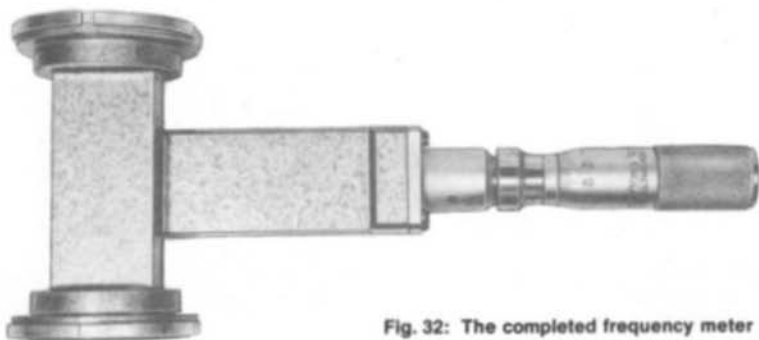


Fig. 32: The completed frequency meter

7.5. Construction

A complete transceiver as shown in Figure 23 can now be constructed from the described parts. The connection of the waveguide components to the individual boards should be made according to **Figure 33**. The relay board shown in Figure 33 is used for transmit receive switching and can be driven from a PTT switch. In principle, this switching is not necessary but allows battery current to be saved during portable operation. In the receive mode, the modulator is switched off, as are the IF preamplifier and broadcast receiver used as IF circuit, in the transmit mode.

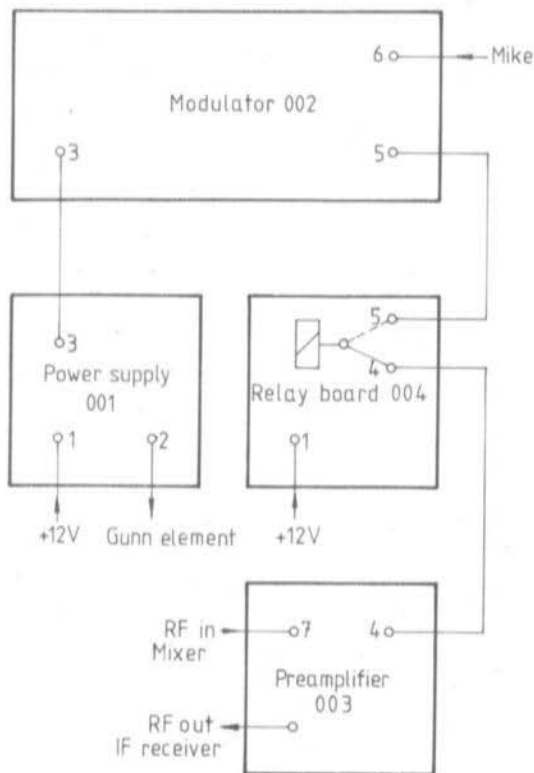


Fig. 33:
Interconnection of the individual modules

Fig. 34:
The completed waveguide components

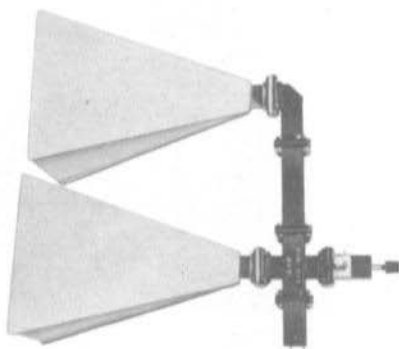


Figure 34 shows the construction of the waveguide components without frequency meter.

Principally speaking, as soon as the Gunn oscillator described in section 4 is ready for operation, the whole transceiver will be completed. The IF preamplifier is aligned for maximum noise on the subsequent broadcast receiver which is aligned for a frequency of approximately 104 MHz. If a conventional car radio is used, the matching will be correct as described; however, when using portable radios or other types of receiver, it may be necessary to shift the tap on the output coil of the preamplifier until maximum noise is obtained.

If the diode current is far too high ($> 2 \text{ mA}$), this will mean that either the Gunn element is too powerful, or the directional coupler has too low a coupling attenuation. In both cases, it will be necessary for the crossed slots to be replaced by smaller ones. If the diode current is too small, this will show that the diode is inferior; the diode should be selected to obtain the minimum value of the mixer slope of $I_D > 200 \mu\text{A}$. The transceiver is ready for operation after the preamplifier has been aligned and a diode current of 200 to 600 μA is present.

8. PRACTICAL OPERATION

It is important that the preliminary experiments on the 10 GHz band be made over shorter distances in order to obtain the necessary experience with the system. Since even medium-sized antennas produce very sharp beams at these high frequencies, it is recommended that small horn antennas should be used at first. G3RPE once said that more communication links were unsuccessful due to too high an antenna gain than with too little. It is imperative for a 2 m or 70 cm link to be used when aligning the antennas. The frequency deviation should be adjusted at short distances according to the report of the partner station. After having gained some experience, experiments can be made over greater distances, however, exact maps should be used to find paths having true line-of-sight. For instance, for communication over a distance of 50 km, the highest obstruction including trees, building etc., should not be nearer than 20 m to the signal beam in order to ensure reliable communications. The path loss in free space can be calculated as follows:

$$a(\text{dB}) = 92.4 + 20 \log d(\text{km}) + 20 \log f(\text{GHz})$$

For example, a value of 146 dB results for a distance $d = 50 \text{ km}$. If antenna gains of 30 dB are available at both ends (e.g. parabolic antennas with a diameter of approximately 40 cm), and approximately 5 mW are available from the transmitter, a receive power of approx. 10 pW will result. This is more than sufficient for conventional receivers; this corresponds to a voltage of 22 μV at an impedance of 50 Ω . In practice, the atmosphere can provide higher attenuation values due to rain, fog or snow. However, 50 km can always be bridged as long as line-of-sight conditions exist. The antennas can be aligned easily with the aid of optical sights and according to landmarks taken from the map. Of course, this gets more and more difficult the higher the antenna gain.

If both stations have selected the same intermediate frequency on the receiver, it will be possible to operate duplex as described in section 5. One is always surprised at the sound quality that can be achieved once communication has been made.

REFERENCES

- (1) Dr. D. Evans: Getting started on the 10 GHz Band
VHF COMMUNICATIONS 9, Edition 1/77, Pages 19 - 29
- (2) D. Vollhardt: Mixer and Preamplifier Noise at SHF
VHF COMMUNICATIONS 8, Edition 4/76, Pages 234 - 242

MATERIAL PRICE LIST OF EQUIPMENT

described in Edition 4/1977 of VHF COMMUNICATIONS

DC 0 DA 004	THREE-STAGE AMPLIFIER		Ed. 4/1977
PC-board	DC 0 DA 004	(double-coated, no thru-contacts)	DM 16,—
Semiconductors	DC 0 DA 004	(3 transistors, 4 diodes)	DM 33,—
Minikit	DC 0 DA 004	(19 trimmers, chip and disc caps., 1 TEKO box, 2 N conn.)	DM 40,—
Kit	DC 0 DA 004	with above parts	DM 88,—
DC 0 DA	BANDPASS FILTER FOR 1296 MHz		Ed. 4/1977
Minikit	DC 0 DA	Bandpass filter (3 ceramic tubular capacitors, 2 BNC connectors)	DM 11,—
DC 0 MT 003	IF PREAMPLIFIER		Ed. 4/1977
PC-board	DC 0 MT 003	(with printed plan)	DM 9,—
Minikit	DC 0 MT 003	(1 transistor, 1 choke, 2 feedthru-caps., 4 disc caps., 1 trimmer potentiometer)	DM 22.50
Kit	DC 0 MT 003	with above parts	DM 31,—
DF 8 QK 001	LINEAR TRANSVERTER FOR 1296 MHz		Ed. 4/1977
PC-board	DF 8 QK 001	(double-coated with printed plan)	DM 18,—
Semiconductors	DF 8 QK 001	(5 transistors, 5 diodes)	DM 98,—
Minikit	DF 8 QK 001	(14 plastic-foil trimmers, 2 chip, 8 disc, 6 feedthru caps., 1 coil-former, 1 ferrite bead, 4 trimmer potentiometers)	DM 34,—
Kit	DF 8 QK 001	with above parts	DM 148,—
DJ 0 FW 002	70 cm TRANSMITTER		Ed. 4/1977
PC-board	DJ 0 FW 002	(double-coated, thru-contacts)	DM 30,—
Semiconductors	DJ 0 FW 002	(13 transistors, 1 IC, 19 diodes)	DM 108,—
Minikit 1	DJ 0 FW 002	(4 coilsets, 1 choke, 6 trimmers)	DM 34,—
Minikit 2	DJ 0 FW 002	(40 ceramic bypass, 14 ceramic, 8 electrolytic caps., 3 trimmer pots., 54 resistors)	DM 47,—
Transmit crystals	HC-18/U	Available: R 69 - 72, R 72 - 75, R 75 - 78, R 78 - 81, R 81 - 84, R 84 - 87 - each:	DM 16,—
Set of crystals	DJ 0 FW 002	any 4 of above crystals	DM 60,—
Kit	DJ 0 FM 002	complete with above parts (incl. set of crystals)	DM 275,—
DJ 4 LB 007	ATV TRANSMITTER / IF DEMODULATOR WITH VIDEO-SOUND COUPLING		Ed. 4/1977
PC-board	DJ 4 LB 007	(with printed plan)	DM 11,—
Semiconductors	DJ 4 LB 007	(8 transistors, 3 diodes)	DM 24,—
Minikit 1	DJ 4 LB 007	(4 coilsets, 1 ferrite choke, 3 ferrite beads, 1 tinplated box, 1 feedthru capacitor)	DM 19,—
Minikit 2	DJ 4 LB 007	(25 capacitors, 35 resistors, 2 trimmer pots., 5 m enamelled copper wire 0.2 mm dia., 6 feedthroughs)	DM 32,—
Kit	DJ 4 LB 007	complete with above parts	DM 85,—

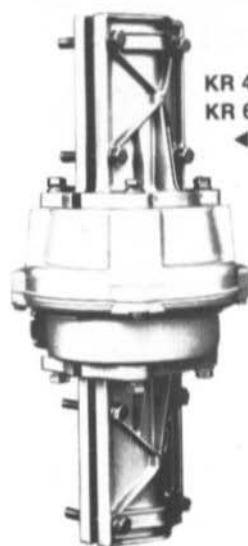
Complete set of kits for ATV TRANSMITTER DJ 4 LB

DJ 4 LB 001a, 002a, 003-007, DJ 6 PI 003, 004 (latter ready-to-operate)

Kits with all parts listed in pricelist **DM 710,—**

Set of PC-boards DJ 4 LB 001a - 007, DJ 6 PI 003, DJ 6 PI 004 **DM 85,—**

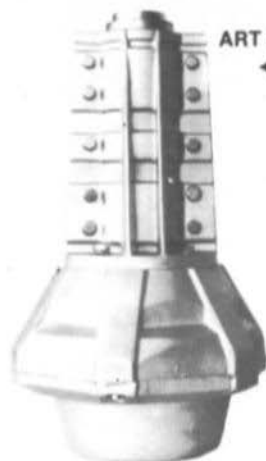
ANTENNA ROTATING SYSTEMS



KR 400
KR 600



KR 2000



ART 8000

SPECIFICATIONS

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

*) 1 kpm \triangleq 9.81 Nm

Controllers for above rotators

Our well-known rotators KR 400, KR 600 and KR 2000 are now available with large 360° compass indicators of 105 mm diameter. These models are designated by the suffix »RC«.

KR 400, KR 600, KR 2000

KR 400 RC, KR 600 RC, KR 2000 RC

ART 8000



UKWtechnik

Terry Bittan · Jahnstr.14 · Postfach 80 · D-8523 Baiersdorf

Tel. 09133/855



CRYSTAL FILTERS OSCILLATOR CRYSTALS

**SYNONYMOUS FOR QUALITY
AND ADVANCED TECHNOLOGY**

NEW STANDARD FILTERS

CW-FILTER XE-9NB see table

SWITCHABLE SSB FILTERS

for a fixed carrier frequency of 9.000 MHz

XF-9B 01

8998.5 kHz for LSB

XF-9B 02

9001.5 kHz for USB

See XF-9B for all other specifications

The carrier crystal XF 900 is provided

Filter Type	XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9NB	
Application	SSB Transmit	SSB	AM	AM	FM	CW	
Number of crystals	5	8	8	8	8	8	
3 dB bandwidth	2.4 kHz	2.3 kHz	3.6 kHz	4.8 kHz	11.5 kHz	0.4 kHz	
6 dB bandwidth	2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz	
Ripple	< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 0.5 dB	
Insertion loss	< 3 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 6.5 dB	
Termination	Z_t	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω	500 Ω
	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
Shape factor	(6:50 dB) 1.7	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 2.2	
		(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 4.0	
Ultimate rejection	> 45 dB	> 100 dB	> 100 dB	> 100 dB	> 90 dB	> 90 dB	

XF-9A and XF-9B complete with XF 901, XF 902

XF-9NB complete with XF 903

KRISTALLVERARBEITUNG NECKARBISCHOFSHHEIM GMBH

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