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A Publication for the Radio-Amateur

Especially Covering VHF, UHF and Microwaves

VHF communications

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Four YUØB-Antennas

VHF communications

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Colin J. Brock (Assistant)

Translator: Colin J. Brock, G 3 ISB / DJ Ø OK

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Australia
W.I.A. P.O Box 300, South Caulfield, 3162 VIC, Phone 5285962

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USA
UV COMMS, K3BRS
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Contents

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Dragoslav Dobričić, YU 1 AW

The YUØB Yagi Antenna

Since its conception in 1926, the Yagi antenna has always been a center of interest due to its very favourable ratio between its electrical performance and its mechanical properties (dimensions, weight, windload). Both the theoretical analysis and practical experiments have produced many successful designs. Actual Yagi-antenna designs are not limited to the original rod-type element design: discs, loops, quads and many other forms, including combinations of the above, are also being successfully used. On the other hand an excellent agreement between computer predictions and actual experimental results has been obtained, right down to a fraction of a dB for the Yagi-antenna gain (2).

While the design of a single Yagi antenna has been developed almost to the physical limits of gain, bandwidth and side lobe attenuation, the problem of parallel operation of several identical antennas (stacking), and in particular the interactions between the single antennas connected together in the system, has received considerably less attention from researchers and experimenters.

In this article, a careful analysis of the problem of stacking Yagi antennas is made first. The results are then used to produce a very successful dual-yagi design called the YUØB Yagi antenna for the 2 m amateur band. The same results are also used to compute the stacking distances for a number of YUØB dual-yagi antennas resulting in very compact high-gain arrays. Finally all the theoretical results have actually been confirmed by many different practical measurements.

1. SINGLE YAGI ANTENNA GAIN

The gain of a single Yagi antenna depends primarily upon its boom length (**fig. 1**), only minor deviations being possible because of different reflectors, feeding systems and element length and spacing tapers. Each doubling of the boom length yields some 2.3 dB of additional gain.

It should be noted that the gain curves of **fig. 1** only show the maximum obtainable gain, for a given boom length, with a carefully optimized antenna. Many popular designs are unfortunately far below these figures!

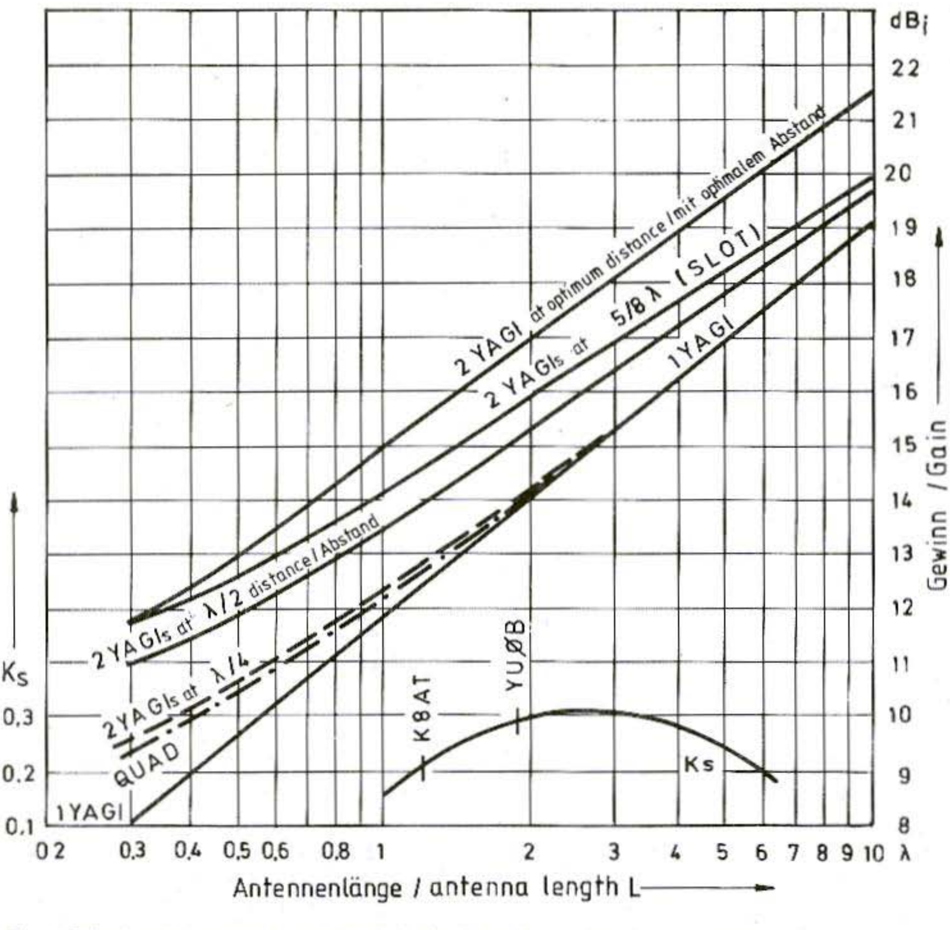


Fig. 1: Gain of optimized antennas, and optimization factor K_s, as a function of the boom length

2. STACKING YAGI ANTENNAS INTO A SYSTEM

Above a certain gain value, a single Yagi antenna becomes inconveniently long for practical construction and installation. It is therefore necessary to stack a number of shorter Yagis to form an antenna system of the required gain.

It is a widely held opinion, that the interactions between the individual antennas in an array have a destructive influence on the characteristics of the array: decreased gain and increased side lobes. Therefore, almost all the authors (and manufacturers) of Yagi antennas just specify the minimum stacking distance between the single antennas in both planes. The straight-forward parallel connection of two judiciously spaced antennas can really provide almost 3 dB of stacking gain, however, at the expense of increased side

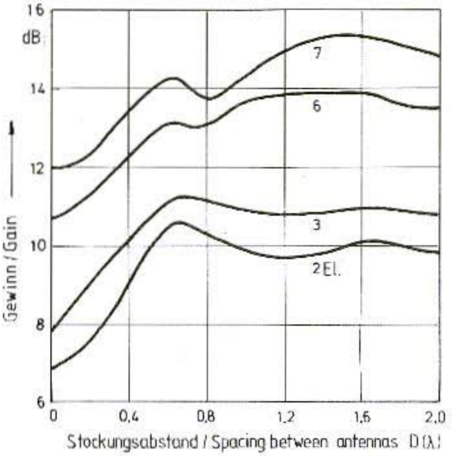


Fig. 2: Gain of two stacked Yagis with a different number of elements as a function of the spacing

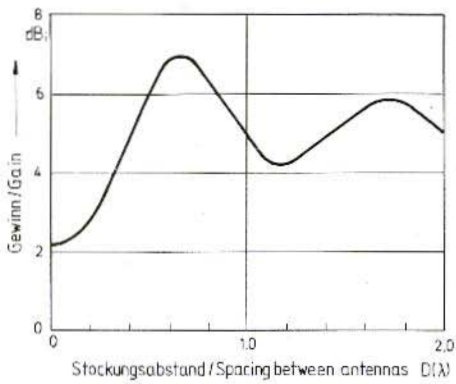


Fig. 3: Gain of two stacked half-wave dipoles dependent on the spacing

lobes, unnecessarily narrow main lobe and also requiring a lot of precious space! An in-depth analysis of the mutual influence between the individual antennas is therefore required.

It can be immediately seen from **fig. 2**, drawn for two stacked 2, 3, 6 and 7 element Yagis (1), that the system gain is not a simple function of the spacing between the antennas. The interaction between two stacked half-wave dipoles shows an even more noticeable gain oscillations (**fig. 3**). Interestingly, all these curves show a clear maximum at a spacing of 0.6 λ. Of course if the stacking distance approaches zero the gain drops to the value of a single antenna.

Besides the integration of the radiation patterns, the gain of two identical stacked antennas can also be calculated from their self and mutual impedances, provided that the gain of the individual antennas is known. This method has the side advantage that the mutual impedances can be measured much faster and to a higher accuracy than the radiation patterns!

The mutual impedance between two λ / 2 dipoles as a function of their spacing is shown on **fig. 4** (1 and 3).

Since to calculate the system gain we actually have to compute the input power to each antenna, only the real parts of R_z, the self and mutual impedances, will play a role in the calculation. It's easy to predict that the maximum gain will occur when the mutual resistance reaches a minimum.

In the general case, the stacking gain of two identical antennas connected in a system, expressed as a field intensity ratio over a single (reference) antenna, is given by the following equations (3):

in E plane:

$$G_r(\varnothing) = \sqrt{\frac{2\,R_{oo}}{R_{11} + R_{12}}} \cdot \left| \cos\left(\frac{d_r \cos \varnothing}{2}\right) \right|$$

in H plane:

$$G_r(\Theta) = \sqrt{\frac{2\,R_{oo}}{R_{11} + R_{12}}}$$

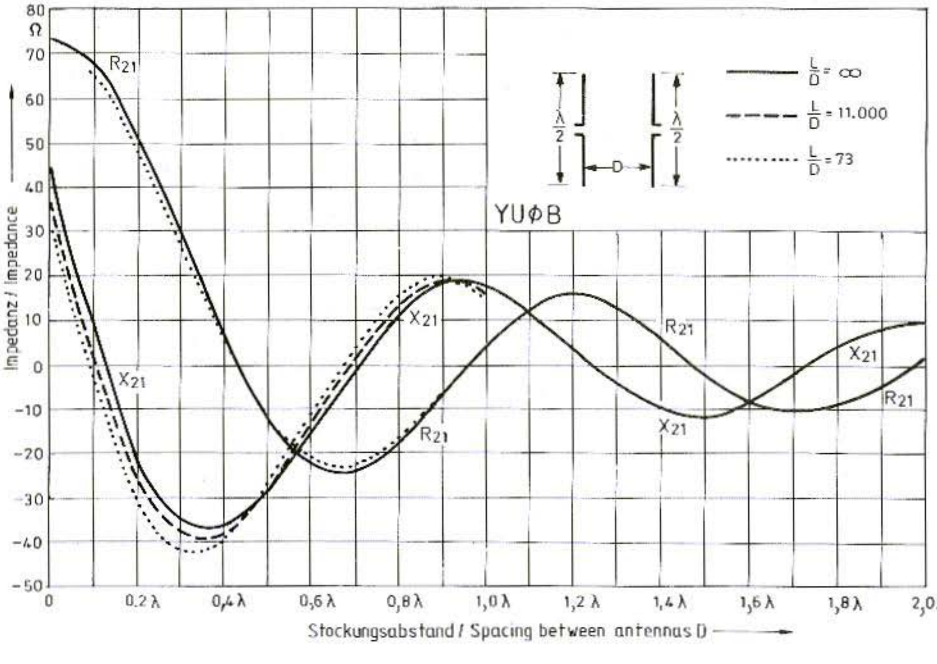


Fig. 4: The mutual impedance and mutual resistance of two half-wave dipoles as a function of their spacing D

where:

R_{oo} = self-resistance of a single (reference) antenna (73 Ω for dipole)

R_{11} = self-resistance of each antenna in system $= R_{22}$

R_{12} = mutual resistance $= R_{21}$

d_r = distance between antennas expressed in radians

With the above equations, it is very easy to explain **fig. 2 and 3**, showing the stacking gain, for λ / 2 dipoles and for short Yagis respectively, as a function of the stacking distance.

The stacking gain for λ / 2 dipoles reaches the first maximum of 3.8 dB at a distance of about 0.65 λ. The next maximum of about 3.3 dB is reached at a distance of 1.7 λ. Both maxima could have been predicted from **fig. 4**, observing the curve for R_{12}. Here it is important to notice, that a favourable interaction between stacked antennas can increase the stacking gain even above the 3 dB value!

However, the effective antenna aperture increases proportionally with the antenna gain. Of course, the stacking gain of two antennas will be decreased if their effective apertures overlap. **Fig. 5** shows the optimal stacking distance to avoid effective aperture overlapping. Aperture overlapping also explains **fig. 2**. In the case of 2 element Yagis, the respective apertures are so small that the stacking gain peak at 0.65 λ is still evident. In the case of 3 element Yagis, this maximum is already flattened and shifted to a slightly higher distance. 6 Element Yagis show an attenuated peak at 0.65 λ whilst another maximum starts at about 1 λ stacking distance, when the effective apertures no longer overlap. Finally,

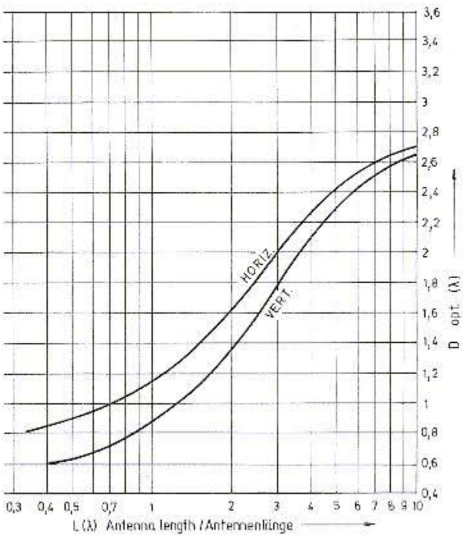


Fig. 5: **Optimum stacking distance avoiding aperture overlapping**

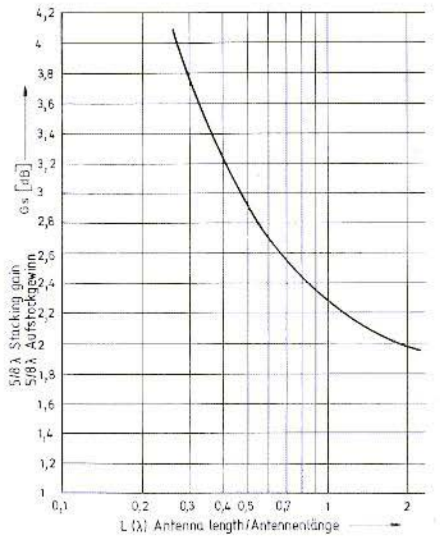


Fig. 6: Stacking gain at 5 / 8 λ distance as a function of the antenna length

for 7 element Yagis, the attenuation, caused by aperture overlapping, ends just in front of the 1.7 λ stacking distance maximum. The 0.65 λ maximum is of course noticeably attenuated.

From the above we may conclude that we have to deal with two different effects when stacking antennas:

1. A gain rise caused by a constructive interaction between the antennas;
2. A gain fall caused by effective aperture overlapping

The sum of both these effects is the actual stacking gain!

The above conclusions have, however, an even more important consequence. It is a very bad practice to arbitrary choose the length of the single antennas and then try to find an optimum stacking distance. To design an optimum system, it is better to choose a suitable stacking distance where stacking gain maxima are known to occur (0.65 λ, 1.7 λ, ...). Then the optimum single antenna gain (= Yagi antenna boom length) is chosen to avoid excessive aperture overlapping, by the use of fig. 5.

When antennas are stacked at about 0.65 λ (slot arrays), the optimum antenna length lies between 0.5 and 0.8 λ, which means quite a short antenna. Longer Yagis will exhibit a lower stacking gain at this distance, decreasing to 2 dB for 2 λ long Yagis (**fig. 6**).

If a stacking distance of 1.7 λ is chosen, the optimum Yagi length is around 2.5 λ corresponding to an antenna having a vertical radiation angle between 35° and 40°.

In the case of bayed Yagis (Yagis stacked in the horizontal (E) plane), the interaction effects show

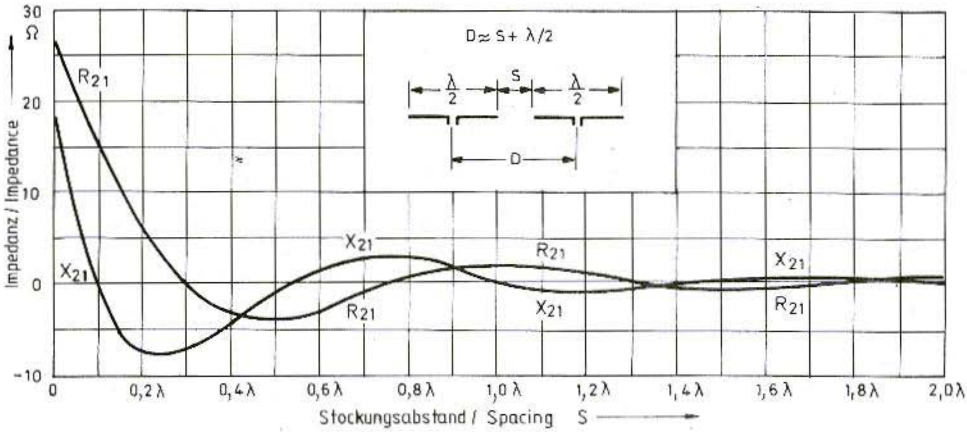


Fig. 7: Mutual impedance and mutual resistance between two half-wave dipoles as a function of distance S

a similar behavour although their magnitude is actually smaller than in the case of vertically (H plane) stacked Yagis. From the mutual impedance plot for two dipoles (shown in fig. 7) we can predict a stacking gain maximum at around 0.5 λ and another one at around 1.55 λ.

2.1. Quad and Loop Yagis

The above theory can also be used to describe the operation of a quad Yagi, since the quad Yagi can be considered as two Yagis stacked at 0.25 λ, their elements bent and joined. It can be imme-

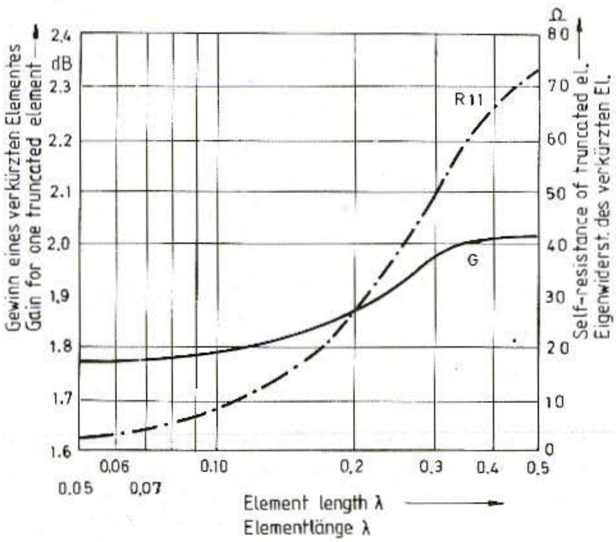


Fig. 8:
Self-resistance and gain of truncated elements dependent on their lengths

diately noticed from **fig. 2 and 3** that the 0.25 λ stacking distance is far from optimum. The gain of a quad is decreased by an additional 0.23 dB since the elements are bent (**fig. 8**).

Circular loop Yagis have about the same gain as two Yagis stacked at 0.25 λ. Other forms of loops show only smaller differences (**see table 1**) with respect to the quad and circular loops.

Equilateral loop type	Gain dBi	Driving point resistance (Ohms)
Triangle	2.70	104
Square	2.99	120
Pentagon	3.10	126
Hexagon	3.16	129
Octagon	3.22	132
Circle	3.28	135

Table 1: Estimated equilateral loop properties

Very long loop Yagis have almost the same gain as the basic rod-element Yagi, since due to the very small equivalent stacking distance, aperture overlapping is almost total. This was also confirmed by practical measurements (7).

3. OPTIMUM ANTENNAS FOR THE 0.6 λ STACKING DISTANCE

The stacking gain for 0.6 λ (about 5 / 8 λ) is shown in fig. 6. Of course the stacking gain decreases with increasing the antenna length, since aperture overlapping also increases. If our design goal were the maximum antenna system gain, it is obvious that this optimum gain (G_{opt}) can only be obtained at distances greater than 0.6 λ for antennas longer than about 0.8 λ . The actual gain difference between the optimum stacking distance and 0.6 λ can be written as dG:

$$dG = G_{opt} - G_{0,6\lambda}$$

On the other hand, in amateur conditions the space occupied is usually the limiting parameter. Widely spaced Yagis will also present a much higher wind induced torque to the rotator system. It is therefore necessary to consider also, the relative space-saving (dl), which is defined as the difference between the optimum distance l_{opt} and 0.6 λ , divided by the boom length L:

$$dl = \frac{l_{opt} - 0{,}6\,\lambda}{L}$$

It may be deduced that there exists an optimum with the best ratio between relative distance saving (dl) and gain loss (dG). We shall therefore introduce an optimization factor K_s as the ratio of the above quantities:

$$K_s = \frac{dl}{dG} = \frac{l_{opt} - 0{,}6\,\lambda}{(G_{opt} - G_{0,6\,\lambda}) \cdot L}$$

A plot of K_s as a function of the antenna boom length L, $K_s = f(L)$, is also shown in fig. 1. A rather flat maximum of K_s exists between antenna boom lengths of 2 to 3 λ suggesting the optimum antenna dimensions.

3.1. Feeding 0.6 λ Stacked Yagis

From fig. 4 it can be seen, that the mutual impedances between closely spaced antennas have considerable values. If each antenna were individually matched to the generator, a large mismatch would ensue if the two antennas are employed in close proximity to each other.

The few dB, obtained from the stacking gain, can easily be lost if the two antennas are not fed with precisely the same amplitude and phase. Careful attention must be given to constructional tolerances of the antennas and (more probably) of feed-cables. Naturally a

low Q feed-network is highly desirable, since it allows wider constructional tolerances.

A very simple and efficient method of feeding two Yagis stacked at 0.6 λ is to use the skeleton slot radiator. This is actually made of two 0.6 λ spaced half-wave dipoles with the open ends bent and joined together. The impedance at these points is very high (around 600 Ω) and can be transformed down to a convenient value of 200 Ω by a delta-match. A half-wave coaxial balun can be used to match this value to the standard 50 Ω coax.

Due to its low Q (in comparison with other feeding systems), the skeleton slot radiator is less sensitive to constructional tolerances and influences from other antennas in the array. This is especially important in large antenna arrays where a number of very similar antennas have to be built first and then connected to form an array.

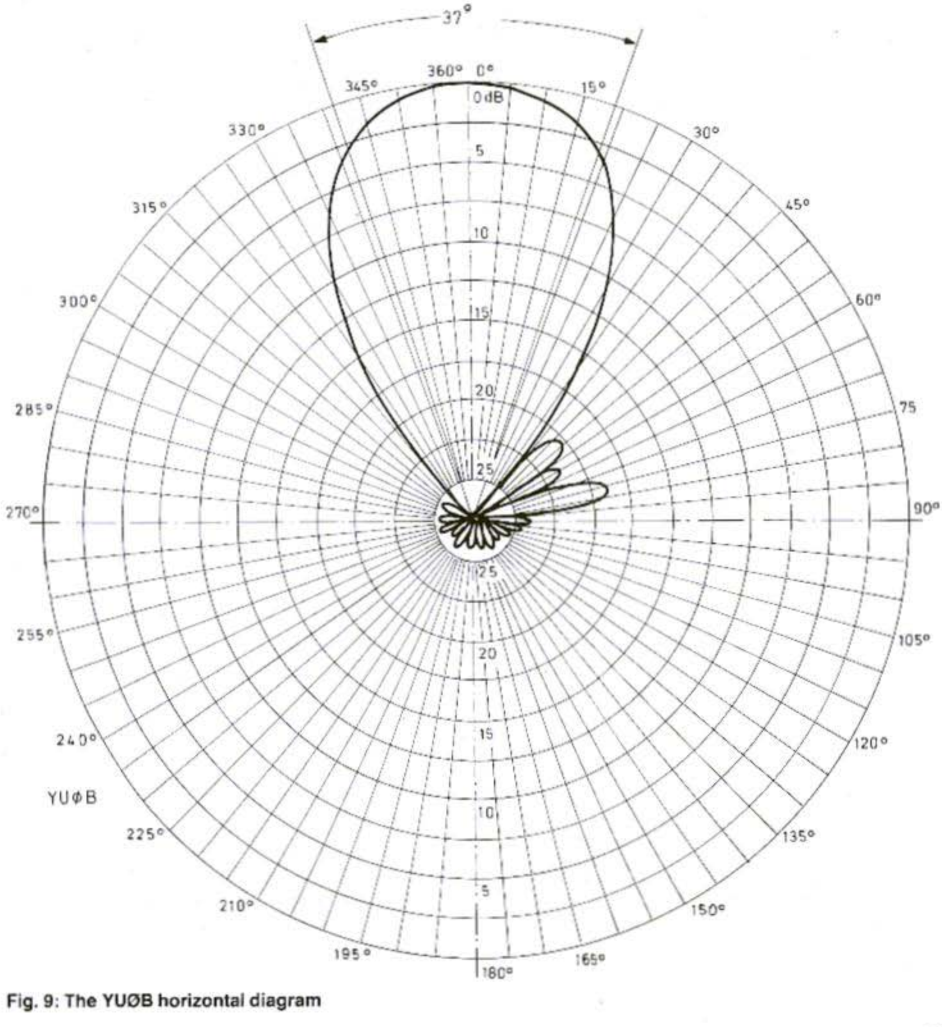


Fig. 9: The YUØB horizontal diagram

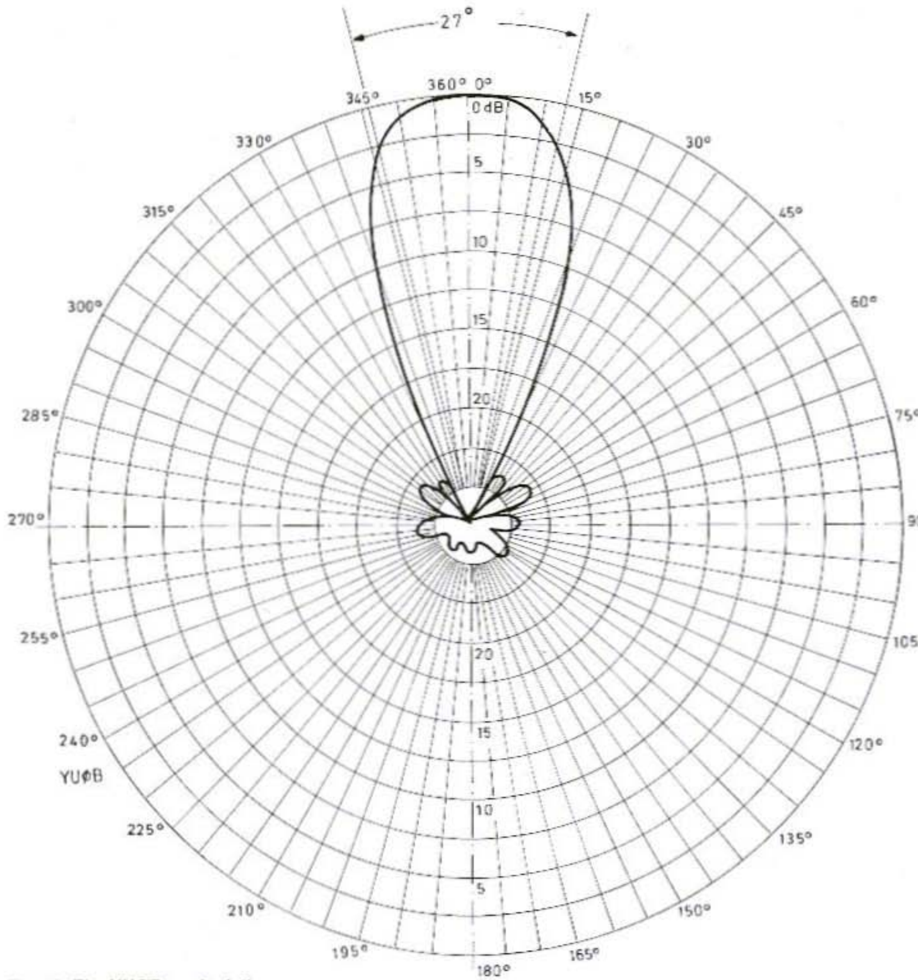


Fig. 10: The YUØB vertical diagram

4. THE YUØB ANTENNA

According to the above theoretical discussion a practical sample of a slot-fed antenna was developed by a group of amateurs.

A practical 2 λ long Yagi antenna gives about 13.8 dB_i of gain. This corresponds to a horizontal radiation angle of 37° and a vertical radiation angle of 45°. Two such antennas stacked at 0.6 λ will give an overall gain of 15.9 dB_i and the vertical radiation angle will be reduced to 27°.

The YUØB antenna derives from the K 8 AT slot Yagi developed by K 3 PGP. Acceptable dimensions and the compactness of the K 8 AT EME antenna system seemed very attractive for condi-

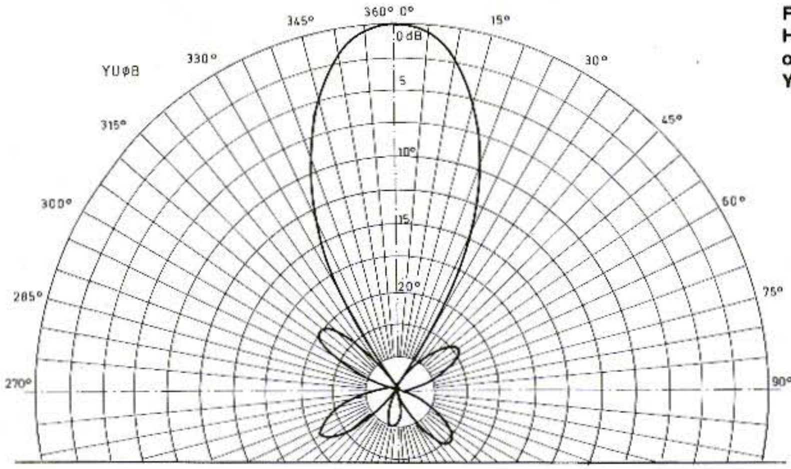


Fig. 11: Horizontal diagram of two bayed YUØB Yagis

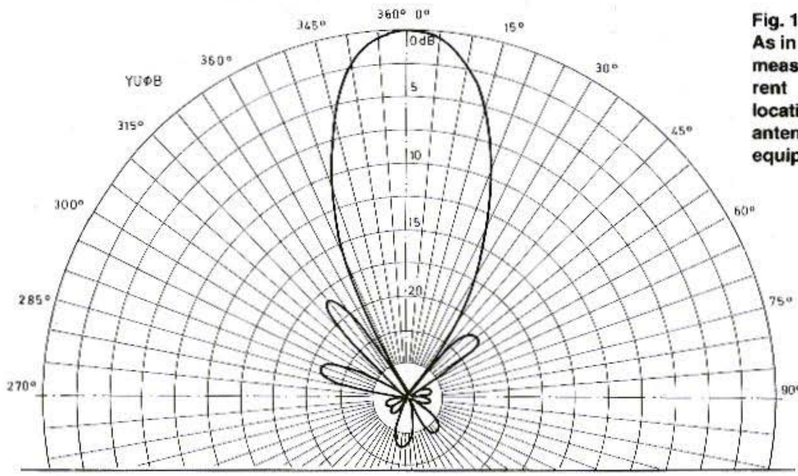


Fig. 12: As in fig. 11, but measured at a different location with different antennas and equipment

tions of restricted space. The original was modified and optimized for best results.

The first modification was the addition of two pairs of directors. This antenna seemed to have a clearer diagram and a little more gain. A system of four such antennas was built and practically tested in an EME QSO.

Later, two more directors were added and the feeding system was changed from a "sleeve" to a coaxial 1 : 4 balun. The additional directors required some more changes of element spacing, especially between the radiator and the first director. This was necessary in order to preserve a clear radition pattern.

After two years of work, including many modifications and radiation pattern measurements, we obtained an antenna of appreciable qualities:

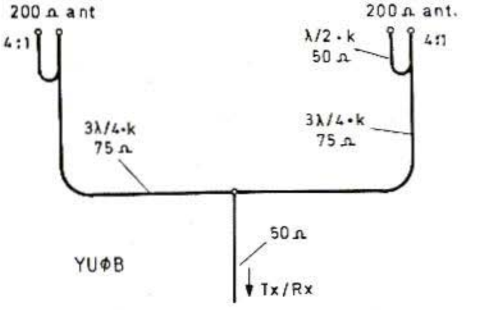


Fig. 13: Matching network for two antennas

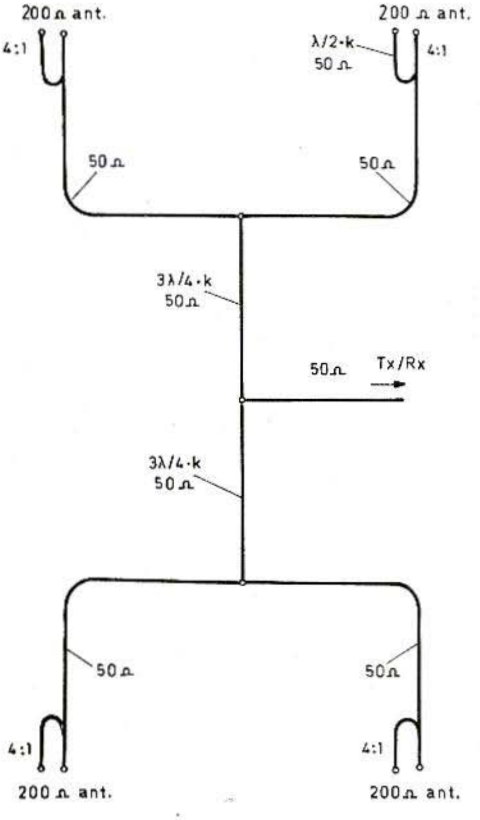


Fig. 14: Matching network for four antennas

1.) The antenna was easily reproducible and all the specimens gave a clear diagram (**fig. 9 and fig. 10**). The gain calculated from the diagram corresponded to the theoretical calculations of 16 dB.

2.) The matching was very good all over the 2 m amateur band. The SWR did not exceed 1.2 between 144 and 146 MHz.

3.) The antenna is very suitable for stacking or baying, since it exhibits a very clear radiation pattern. The results for two YUØB are shown in **fig. 11 and 12**.

To obtain accurate results, many measurements from different locations were made, using different signal levels and different equipment. Stable and repeatable results are the proof that the **measuring procedures were correct (compare fig. 11 and fig. 12).**

4.1. Stacking YUØB Antennas into Systems

As already mentioned the good features of the YUØB antenna are essential for building larger systems. Stacking YUØB antennas present two possibilities:

First, placing the antennas 0.6 λ apart; 2 dB can be expected for each system doubling.

Second, using optimum stacking distances to avoid aperture overlapping; about 3 dB can be obtained for each system doubling.

If a compact antenna system in a small space is preferred, four YUØBs can be stacked at 114 cm and bayed at 214 cm. Such a system will provide about 20 dB of gain. The dimensions are interesting: 3.5 x 3.2 x 3.8 m!

If more gain with the same number of individual YUØB antennas is required, optimum stacking distance to avoid aperture overlapping can be found from the radiation angle θ in the plane of stacking or baying (8):

$$D = \frac{\lambda}{2 \sin(\Theta / 2)}$$

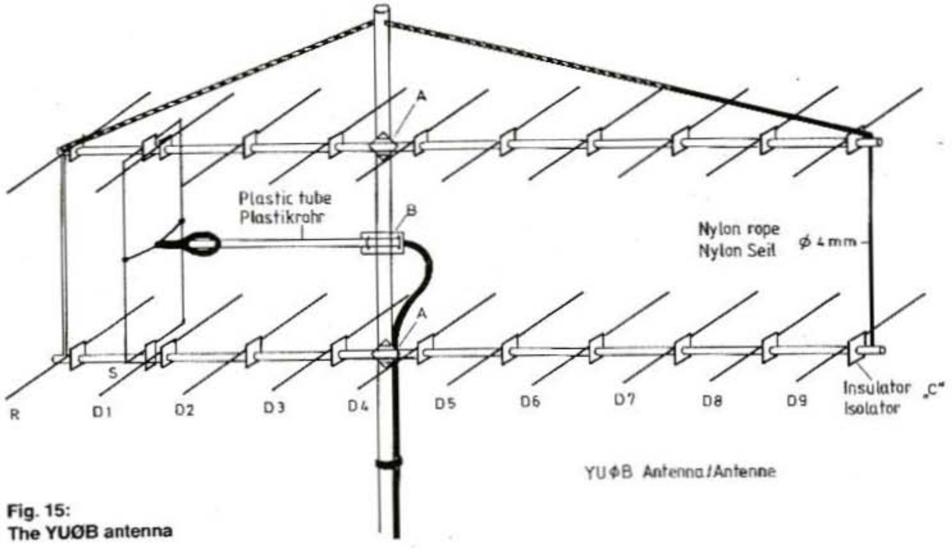


Fig. 15: The YUØB antenna

For the YUØB with $\theta_E = 37°$ and $\theta_H = 27°$, optimum spacings are 2.14 λ (4.45 m) in the vertical, and 1.6 λ (3.3 m) in the horizontal plane. Stacking in this way will give a system gain of 22 dB_i, but the dimensions are much larger, especially in the vertical plane.

From the theoretical discussion, it can be seen, that the mutual influences and corresponding additional gain are smaller when the antennas are bayed in the horizontal plane than when they are stacked in the vertical plane. It is therefore better, to use the optimum horizontal stacking distance for YUØB, since with a small increment of the system dimension (from 3.2 to 4.4 m), the maximum stacking gain can be obtained.

In the vertical plane we have the opposite situation. Due to the narrow vertical radiation angle of the YUØB, the optimum spacing is large. **So 1 dB can be sacrificed in order to retain the 0.6 λ stacking in the vertical plane, while the occupied volume is actually halved!**

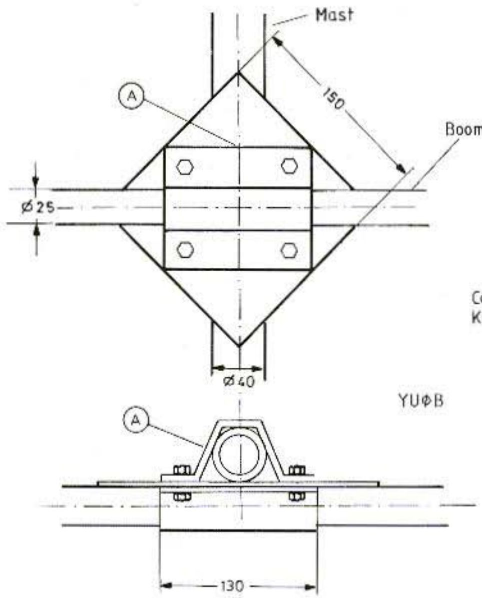
In this case the distances are 1.2 m in the vertical and 3.3 m in the horizontal plane (measured from boom to boom). System gain is 21 dB_i and the polarization is horizontal.

The matching network for two antennas is shown in **fig. 13** and for 4 antennas in **fig. 14**. Larger arrays, groups of two or four antennas, should be considered as an individual antenna.

4.2. Mechanical Construction of the YUØB

An overall view of the YUØB antenna is shown in **fig. 15**. The two booms are made of 25 mm diameter aluminium tube. The elements are installed on insulating supports (**fig. 16 c**). These insulators can be made from practically any insulating material, such as PVC, plexiglass, teflon, etc. (**fig. 16 b**).

The slot radiator is manufactured from copper, brass wire or tube of 5 mm external diameter. The various pieces are soft or hard soldered together. The most suitable material is 5 mm copper tube since the coax center conductor can simply be soldered into the hollow tube, the balun being soldered directly to the ends of the delta match transformer (**fig. 16 d**).



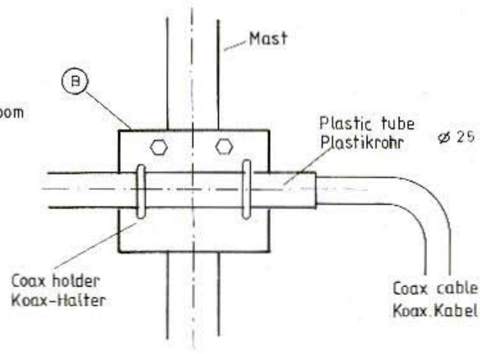


Fig. 16 a:
Mast installation and leading the coax cable

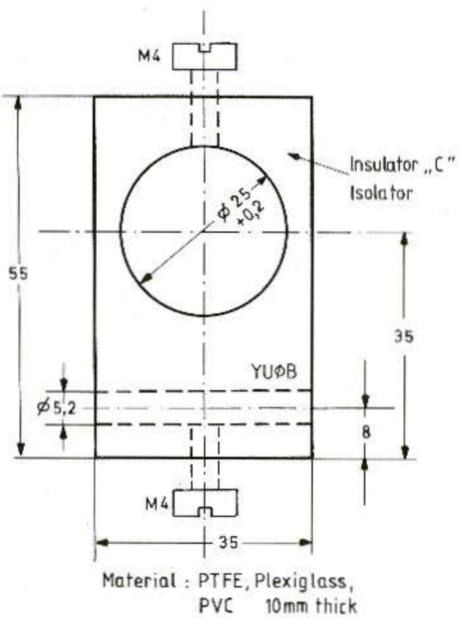


Fig. 16 b:
The insulating supports

All the directors and reflectors are made of 5 mm diameter Al welding electrodes or Al wire (dimensions and spacings - **see table 2**)

Element	Length (mm)	Spacing (mm)
Reflector	1020	Slot director 1 : 18
Director 1	933	Refl. director 1 : 380
Dierctor 2	914	Spacing between all
Director 3	908	directors is : 406
Director 4	898	
Director 5	889	All elements are
Director 6	879	made of Al 5 mm Ø
Director 7	870	
Dierctor 8	861	
Director 9	851	

Table 2: Dimensions and spacings of elements for the YUØB antenna

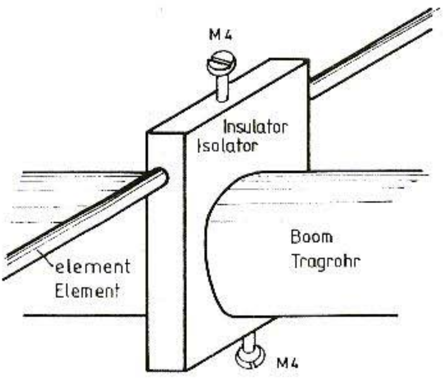


Fig. 16 c:
Boom, insulating support and element

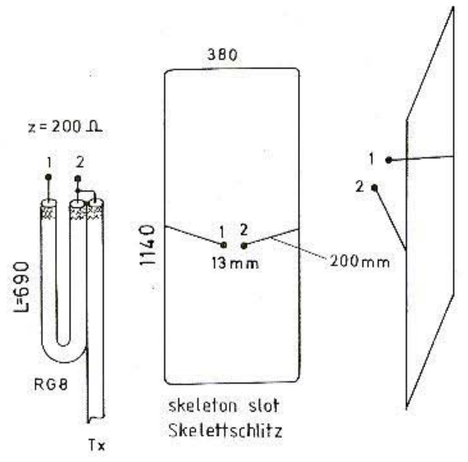


Fig. 16 d:
Feeding and matching the YUØB Yagi antenna

A 25 diameter PVC tube is used to support the cable between the balun and the mast. The nylon rope improves the mechanical strength of the antenna.
Other details are visible in **fig. 16 a**.

5. THE ANTENNA'S NAME AND THE RESULTS OBTAINED

Many YU amateurs contributed to the development, practical construction and measurements on this antenna, so it was decided to choose YUØB as its name. The letter "B" stands for "YU VHF-UHF BULLETIN" in which it was published in 1980 for the first time. Unfortunately, this call-sign could not be obtained for the activities promoted by the publisher. We got the call sign YZØB and used it in several MS and tropo expeditions to the LE, KC and KA QTH squares. In all the expeditions we used YUØB antennas.

The following amateurs contributed to the realization of the YUØB: YU 1 BB, EU, EV, NRV, MS, NZV, WA, OAM, OLO, OJP, MM, YU 2 RKY, OO, RTU, YU 7 ACO, BCD and the author of this article: YU 1 AW.

YU 7 AR, YU 1 EU, YU 3 USB and YU 3 ZV used this antenna for their first EME QSO's.
It helped to make more EME contacts and obtained good results in various contests. The antenna performed very well in MS work and was used with excellent results by YU 4 GJK, YU 7 BCD, YU 1 AWW and many others.

Finally, I wish to thank all those amateurs who participated in the development of this antenna and who used it, notwithstanding the fact that its outstanding qualities had not been fully proved.

Text revision of this article was carried out by Matjaž Vidmar, YU 3 UMV.

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Dieter Schwarzenau, Bernhard Kokot

More about the DLØ HV HF Millivoltmeter

One year following the publication of the constructional details of the HF Millivoltmeter (1) it appears to be time to review the project. The instrument has, in the meantime, been constructed by many people and quite a lot of additional experience is now to hand. The opportunity is also taken to introduce practical uses for this versatile instrument.

1. CONSTRUCTION TIPS AND CORRECTIONS

The only known actual mistake, so far, concerns the rectifier diode D 401 of the through-line probe. It is shown in **fig. 5 b** wired in the reversed polarity. The polarity shown on the component plan, (**fig. 13**) is, however, correct (1).

On the component plan for the electronic board with electronic switching (**fig. 21**) and in the circuit diagram (**fig. 7**) the polarity for all electrolytic capacitors has been omitted. As in all cases, one leg is connected to the supply rail, it would'nt have been very difficult to work out the correct way round to wire them in. With the exception of the block capacitor C 615, near the voltage regulator I 612 (**fig. 21**), all the plus leads should point upwards.

The components parts list for this module also omitted C 605, R 642, R 643 and P 601. The resistance of preset P 601 is identical with that of the corresponding potentiometer in the mechanically operated module and is 100 Ω. Capacitor C 605 is a tantalum capacitor of 2.2 µF and 25 V working voltage. The resistors R 642 and R 643 serve only as bridges and therefore, are not included in the circuit diagram. They should not exceed a value of 10 Ω. The capacitor C 607 should be removed from the circuit as it caused the regulator to oscillate instead of reducing ripple as intended.

Apart from the probe, the HF millivoltmeter may be regarded as a purely audio frequency instrument and was given the appropriate design measures with which this merits.

The constructor should therefore adhere strictly to the instructions concerning groundplane through connections, in order that earth loops are avoided. The old HF addage that "the more earth connections the better" does not apply here. The use of a ring-core transformer in the power supply was not found to be justified in practice. Because of the small leakage inductance of this transformer, mains interference was hardly reduced at all. If possible, intending constructors should use a normal PCB transformer or ones with an M-core, whose primary and secondary windings do not lie above each other but in juxtaposition. Under certain conditions it can occur that currents flow through the cabinet and into the mains ground line and interfere with the measurements. In this case the mains earth connection may be removed for the measurement but it must be stressed that the earth lead should never be disconnected permanently.

2. USING THE HF MILLIVOLTMETER

The measurement of frequency, voltage and power are the most important measurements undertaken by the experimenting amateur. It can also be interesting sometimes to be able to determine the reflection coefficient. The measurement of frequency presents no problems nowadays. Quite a large number of articles, describing the construction of low-cost frequency meters, exist and for those who do not wish to build there are many suitable proprietry instruments to choose from. For the measurement of the other quantities however, the picture is quite a bit different. Commercial instruments are, almost without exception, expensive and useable constructional articles are very few. In the past, many amateurs had to be content with the "go" / "no go" type of voltage measurement obtained with a rectifier and capacitor combination stuck in front of a multimeter. The following examples will serve to emphasize the considerable improvement which the described HF millivoltmeter represents.

2.1. Measuring HF Voltages

As the name suggests, the measurement of small high-frequency voltages is the main purpose of an HF millivoltmeter. Before the measurement is undertaken, some consideration should be given as to the loading the measuring device imposes upon the item-under-test (IUT). It is best when a 50 Ω termination can be used to terminate the item-under-test via the through-line connector probe. In the simplest case, a 50 Ω terminating resistor may be used, connected directly to the line output terminal of the through-line probe. The probe can, however, be part of a 50 Ω coaxial line system with say, a frequency counter on the other end. In this case the coaxial cable should be as short as possible, otherwise, the less than perfect termination may cause errors in the voltage measurement due to reflections back along the line. These errors can be considerable.

If the item-under-test cannot be made part of a 50 Ω measurement system, the high impedance test probe must be used. Although this probe has a very high impedance and low capacitance, it is always good measurement practice to select a very low-impedance test point at which to carry out the measurement. This should be practicularly observed when measuring tuned circuits and oscillators where detuning of such circuits by the probe may make the measurement meaningless.

Sometimes, when using the probe, a difficult problem occurs as to where earth connection should be taken. For measurements over 100 mV it is sufficient merely to clip the ground lead to the nearest available ground point. This, however, is not good enough for low-level measurements. This is particularly the case when the measurements are to be made in equipment where large levels are also present. To check their effect, simply connect the probe ground lead to its tip and see if there is any voltage indication on the meter range being used for the test. The extremely high sensitivity of the meter and the high-impedance probe will sense the high-level interference signals in the vicinity of the area.

If the circuit to be tested is enclosed in a tin-plate housing, it may be possible to earth the probe with a convenient point on the metal mass near the test point. If that is not possible, a short piece of wire may be used from probe ground to the nearest ground to the test point.

When making the measurement, a range should be selected which causes the meter to indicate 2 / 3 FSD or greater starting with the 10 V range. The reading can then be read off in volts or in mV. If only a normal multimeter (digital or analogue) is connected to the AF module and a dBm measurement result is required, the following equation has to be used: -

$$a\,(\text{dBm}) = 20 \log V / 0.2236 \text{ (in mV)}$$

This is only meaningful, however, when the measurement is taken across a known resistance of 50 Ω.

With the lowest scale of the indication, described in the constructional article, the value in dBm may be read off directly. It is even more simple if the logarithmic amplifier is available. The input selector switch is set to the appropriate input V 1 or V 2. The pointer indication must be read against the upper scale and added to the measurement range setting (scale value negative).

2.2. Power Measurement

As the use of the through-line probe is intended for use in a 50 Ω measurement system, a voltage measurement is always also a power measurement. The power is determined by including the measured voltage in the expression: -

$$P = V^2 / 50\,\Omega$$

It is still simpler if the voltage can be read off in dBm / 50 Ω, as 0 dBm represents a power of 1 mW / 50 Ω. With the well-known fact that doubling the power is + 3 dB and increasing it by a factor of ten is + 10 dB, it is quite easy to deduce any measured power read on the dB scale (examples: + 23 dBm = 10 dBm + 10 dBm + 3 dBm => P = 1 mW × 10 × 10 × 2 = 200 mW or − 16 dBm = − 10 dBm − 3 dBm − 3 dBm => P = 1 mW × 0.1 × 0.5 × 0.5 = 1 / 40 mW = 25 μW).

The only condition for the power measurement is that the output of the through-line probe must be terminated in a 50 Ω resistance at the frequency of measurement. It is recommended that a proprietry 50 Ω termination resistor is used to terminate the probe. The HF millivoltmeter is capable of measuring directly powers between 100 nW and 1 W.

2.3. Attenuation Measurements

The measurement of attenuation or amplification in a 50 Ω system is certainly the most interesting and, for the experimenting radio amateur, the most useful application of the millivoltmeter. It must, however, be provided with two through-line probes and the logarithmic amplifier must be used. One probe is connected between the test generator and item-under-test input and the other is connected between the IUT output and the 50 Ω termination. This arrangement may be seen in **fig. 1**.

According to which of the two voltages is greater, the logarithmic amplifier is switched to "V 1 > V 2" or "V 2 > V 1". The four LEDs will indicate whether a higher or lower range for V 1 or V 2 should be selected. The working range has been selected correctly when all LEDs are extinguished. The attenuation / amplification in dB is given

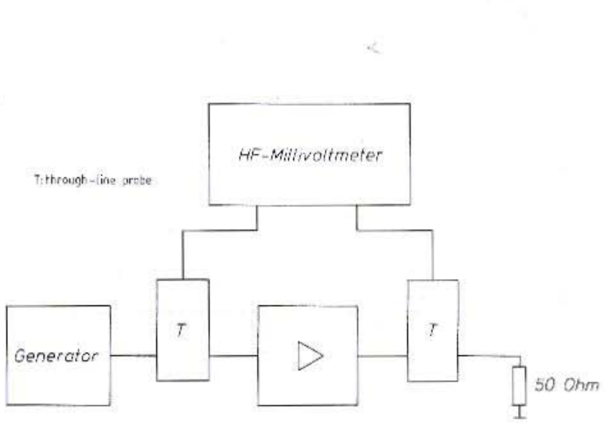

Fig. 1: Test arrangement for the measurement of insertion loss / gain

from the difference in the ranges plus the reading on the upper scale (negative).

Until now, these measurements do not mean much. The could have been obtained by using a simple HF millivoltmeter to take two separate measurements. The big advantage of this arrangement is only apparent when the test item has to be optimized and adjusted for a particular performance objective. Every alteration of the attenuation / amplification can be seen immediately by the meter deflection. It is therefore very easy to test the dynamic range or do a frequency run on the test object. Using a normal simple millivoltmeter there would be a cumbersome procedure to follow. Take for example the determination of the 3 dB-bandwidth points of a wideband amplifier. The signal-generator tuning is adjusted until the reading falls by 3 dB. The output from the signal generator would then have to be checked to ensure that it had remained constant at this frequency. The output variations of the signal generator with frequency are of no importance with the DLØHV MVM because it is the amplification, i. e. the relationship between input and output that is being displayed on the meter.

2.4. Reflection-Coefficient Measurements

The designer of HF circuits is always faced with the problem of matching circuit inputs and outputs to the nominal impedance of 50 Ω real. Only in the matched condition, i. e. when the characteristic impedance of the output, the transmission line and the input are the same, the maximum power can be transmitted. In all other cases, standing waves are developed by reflections from mismatched impedances. A measurement of the mismatch is either the reflection coefficient or the voltage standing wave ratio VSWR.

An ubiquitous instrument for measuring the reflection coefficient is known as the reflectometer. It consists of a directional coupler with two rectifiers whose output voltages can be read off on a meter. Normal reflectometers contain only a single diode rectifier with a filter capacitor. They are therefore insensitive and only suitable for the measurement of large voltages. Also the measurement is liable to errors owing to the diode barrier voltage making the characteristic non-linear. Both these disadvantages are avoided when using this HF millivoltmeter. A directional coupler, however, must be used. This can be purchased or constructed (2). If only the more simple version of the millivoltmeter is available, the incident and return voltages must be measured separately. The reflection coefficient and the voltage standing wave ratio can then be calculated.

Using two through-line probes and the logarithmic amplifier, the return loss can be directly read off the upper dB scale. Under these conditions, adjustments and tuning may be very easily carried out.
The same may be achieved with a simple HF millivoltmeter and the use of a directional bridge (3).

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Jochen Jirmann, DB 1 NV

A Miniature 70 cm Handheld FM Transceiver

In VHF COMMUNICATIONS 1 / 86 a miniature handheld transceiver for the 2 m band was introduced by DL 5 NP. This present article describes a similar unit for the 70 cm band. It is somewhat larger than the 2 m version (PCB dimensioned 70 × 130 mm), delivers the same power output of 500 mW and possesses a receiver sensitivity of better than 0.5 µV for a 20 dB signal to noise ratio. The current consumption is, at about 25 mA, somewhat higher than the 2 metre variant.

1. CONCEPT

For the realization of a UHF handheld transceiver, several variants may be considered whose advantages and disadvantages merit a short discussion.

1) The smallest constructional effort would entail using the same circuit as that for the 2 metre band. This would involve principally the synthesizer scaler being replaced by one designed for UHF and all tuned circuits re-dimensioned. There was, unfortunately, only one suitable scaler module available at the start of the design, in mid 1985, and that was the MC 12018 by Motorola. This has, however, the very awkward division ratio of 128 / 129 which would necessitate a very complicated encoding circuit between the BCD switches for the frequency selection and the synthesizer.

2) This problem may be circumvented if the VHF signals are translated in frequency by the mixing process. This would require a further crystal oscillator, a multiplier, a mixer and a filter, all costing space and current consumption. There is also the inherent risk of spurious emissions on send and birdies when switched to receive. The extensive screening measures, required for this approach, are not very practical in such a confined construction.

3) The VCO working at VHF and the working frequency achieved by multiplication. This method requires only the additional elements of multiplier and filter. The disadvantage of this technique is that the PLL phase comparator frequency must also be divided by the same factor tending to increase the lock time of the synthesizer. Of particular importance is the careful filtering of the VCO tuning potential in order that sufficient suppression of the adjacent channels is achieved.

Only a multiplication factor of two is feasible, as a factor of three would place a spurious signal in the two metre band which could interfere with cross-band working. The necessary doubling of the VCO frequency from a maximum of 220 MHz can easily be done by the same module as that used in the 2 m handheld, namely the Plessey scaler SP 8793.

The author decided upon the variant 3, as the components are easier to obtain and the demands upon components and current consump-

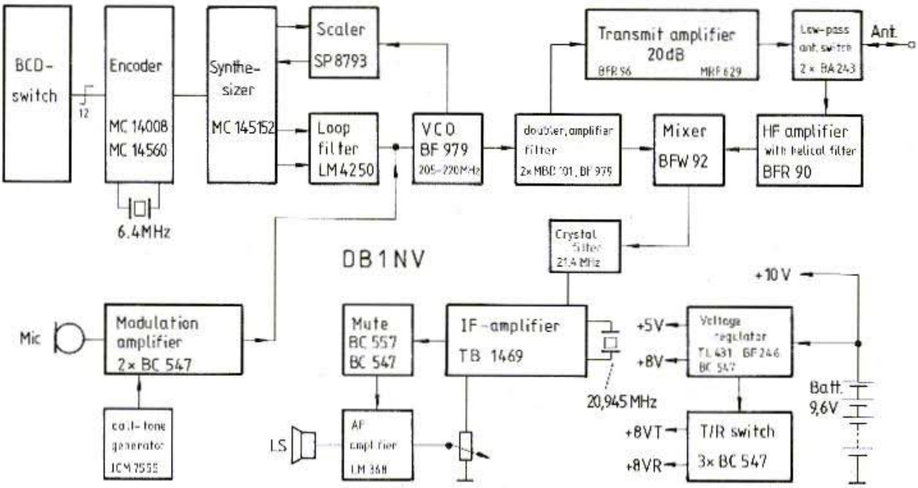


Fig. 1: 70 cm Handheld TR block diagram

tion are held within bounds. An additional advantage of doubling the VCO frequency is a better isolation between the oscillator and the send frequency.

The rest of the handheld's circuit can be simply adopted from the 2 m version. In the receiver however, an IF of 21.4 MHz was chosen in order to achieve a usuable image frequency rejection.

A proprietry helical filter was chosen for the receiver preselector.

As the channel selection of 400 channels with a frequency separation of 25 kHz is a little awkward, it was made selectable in steps of 1 MHz, 100 kHz and 25 kHz. The block diagram of the handheld may be seen in **fig. 1** and its principle specifications are tabulated in **table 1**.

Frequency range:	430 - 440 MHz
Modulation:	FM
Channel steps:	25 kHz
Relay OFF-SET:	7.6 MHz, receiver higher than transmitter to allow easier monitoring of relay traffic
Output power:	$>$ 300 mW, typically 500 mW
Mixing products:	Suppressed $<$ - 60 dB
Harmonics:	Suppressed $<$ - 50 dB
Adjacent channel output:	60 dB below carrier
Current consumption (send):	200 mA at 9.6 V
Receiver sensitivity:	Better than 0.5 μV at 20 dB S / N
Bandwidth:	Approx. 12 kHz depending on crystal filter
Image suppression:	50 dB
Current consumption (muted):	25 mA
PCB dimensions:	68 mm $\times$ 130 mm, height with components 20 mm

Table 1: Miniature handheld TR data

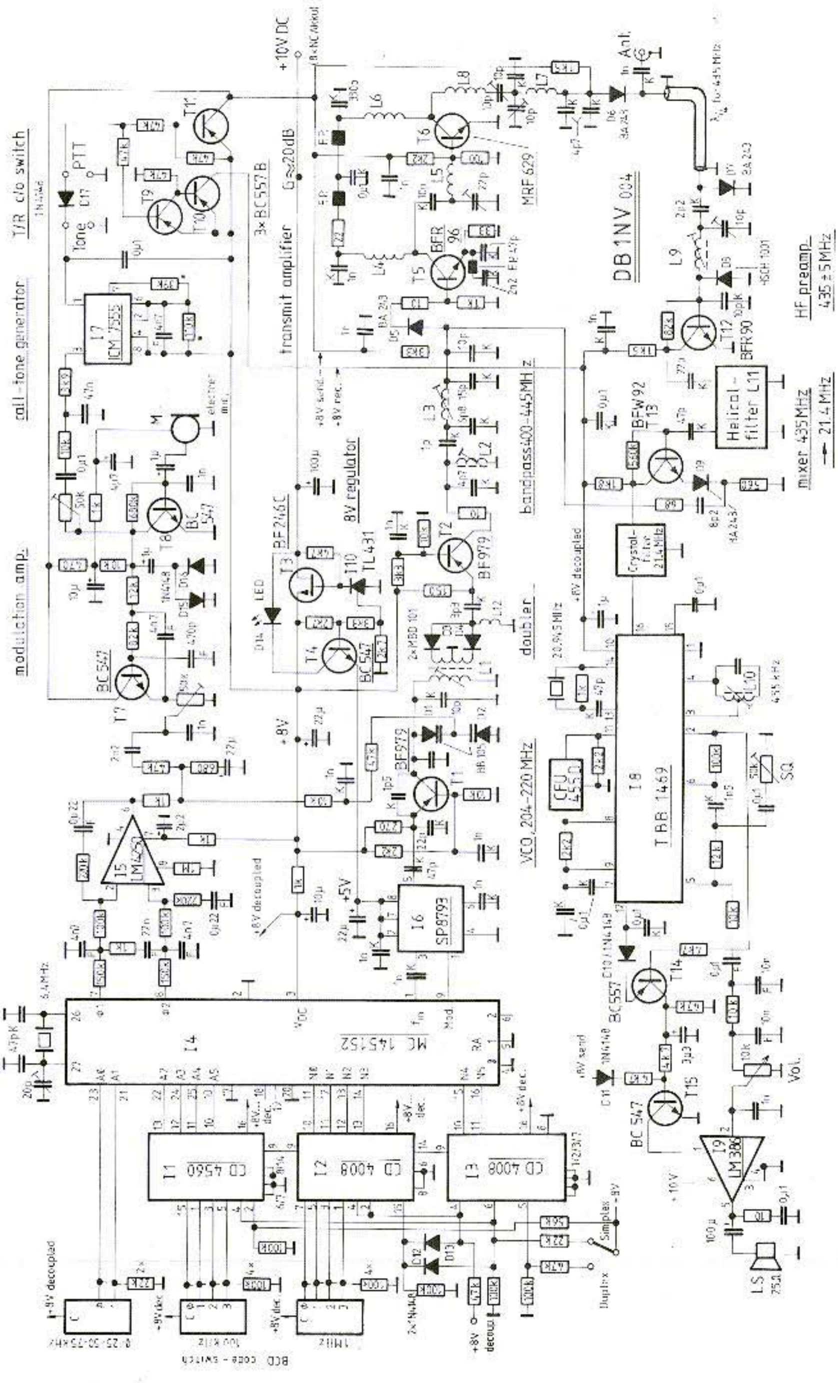


Fig. 2: Complete circuit diagram of 70 cm Handheld TR

2. CIRCUIT DETAILS

Now that the basic concepts of the handheld's design have been introduced, details of both changes of circuit from the 2 metre model and completely new circuits will now be undertaken. The complete circuit diagram is shown in **fig. 2**. A start will be made with the HF part of the synthesizer.

2.1. The Synthesizer

It has already been mentioned that the synthesizer works at half the output frequency, that is between 204.3 and 220 MHz. The circuit used by DL 5 NP was therefore directly adopted for this project. It was only necessary to programme a basic 12.5 kHz-step frequency into the MC 145 152 synthesizer IC and to dimension the loop filter LM 4250 I 5 to suit the new phase-comparator frequency. It may be seen from **fig. 2** that the dimensioning of the VCO has also been slightly altered. A UHF oscillator transistor BF 979 in a T plastic package is now employed.

The frequency selection logic I 1 to I 3 was adjusted for a direct frequency selection in 25 kHz steps and for the other relay offset of 7.6 MHz. The VCO delivers approx. 3 mW to the directly coupled frequency doubler which uses two Schottky diodes. Following the doubler there is a power of about 200 μW available, the fundamental being suppressed some 10 dB principally by a careful construction of the coupling inductor. A high-pass filter takes care of the matching to the amplifier stage T 2, a BF 979.

In the collector circuit is a two-circuit, overcritically coupled filter which guarantees the freedom of the output signal from mixing products of more than 60 dB. The twin peaks of the pass characteristic lie at 415 MHz (receive mixer frequency) and 435 MHz (sender frequency). A power of 3 to 5 mW is available at the output to drive the receive mixer and the transmit amplifier. Owing to the very lage tuning range of the 70 cm version, using only one VCO for send and receive, the phase noise of the oscillator is relatively poor. This must, however, be put up with in extremely miniaturized equipment, expecially since many mature Japanese fixed station equipments are scarcely any better. It may be just as well to point out that the signal from this handheld should not be amplified to many 100 Watts!

2.2. The Transmit Amplifier

It may be seen from the overall diagram that the transmit amplifier has hardly been altered at all from that of the 2 metre version. Only the final transistor T 6 was changed to a UHF type MRF 629 and set to work in class B in order to obtain more amplification. The MRF 629 is eminently suitable for mounting on a printed circuit board as it has a grounded emitter, delivers 2 W at 12 V and is therefore highly overrated for this application. This is just as well because of the unknown matching conditions of the various rod antennas which could be used with this handheld. The overdimensioning of the transistor does not, in any case, involve much additional expense and therefore represents a good premium against failure. The low-pass filter and antenna changeover are modelled after the 2 metre version.

2.3. Modulator and Send / Receive Switching Logic

The complete operational transmitter still requires a modulator and switching logic for the supply rail. Here also, nothing has been changed respect to the 2 metre version. Additionally, a CMOS timer I 7, ICM 7555, serves as a call-tone generator. The somewhat suspect circuitry of this generator may be explained by the fact that ground for the circuit is the positive supply rail.

2.4. The Receiver

As previously mentioned, the receiver uses an IF of 21.4 MHz in order to achieve a reasonable image rejection. Because 21.4 MHz monolithic crystal filters have become reasonably priced, an IF at this frequency is a natural solution. The complete IF / AF circuitry is taken directly from DL 5 NP.

The input signal is taken to a preselector via a

PIN-diode switch. The preselector matches the input impedance to that of the pre-amplifier transistor T 12. The BFT 66 was replaced by a BFR 90 (or BFR 34). The attenuator diode between the base of T 12 and earth has a special purpose. It is necessary because the efficacy of the PIN-diode switch is not as great at 70 cm as at 2 m and on "transmit" transistor T 12, without supply voltage, acts as a frequency multiplier sending spurious multiplication products back through the PIN-diode attenuator and reaching the antenna at levels approaching - 40 dB rel. 70 cm carrier. The additional diode simply detunes the input tuned circuit thereby reducing the RF level at T 12's base.

The pre-amplifier transistor works into a two-stage helical filter having a 10 MHz bandwidth. The following mixer is provided with a bipolar transistor BFW 92 because a FET does not provide the required amplification at this frequency. The local-oscillator signal is fed via a PIN diode, on "receive", from a doubler to the mixer's emitter. The mixer is terminated in a monolythic 21.4 MHz crystal filter. The constructor can choose here between a minimum selectivity with a two-pole filter ranging to an eight-pole filter, the only consideration being that imposed by the pocket. The author selected a six-pole filter from NKD. The rest of the IF remained as for the 2 m version except, of course, that the second oscillator works at a crystal frequency of 20.945 MHz.

Following this cursory review of the circuitry, the next paragraph will deal with the loading of circuit components on to the board.

3. CONSTRUCTION OF THE HANDHELD TR DB 1 NV 004

Note 1: The PCB DB 1 NV 004 is so densely packed with components that only experienced constructors should consider attempting this project.

Note 2: In a few places on the underside of the PCB a ground lead is taken for through-contacting of a few components. This measure enables the components around the synthesizer to be mounted more easily.

3.1. Component List for Complete TR

ICs

I 1:	CD 4560, MC 14560; RCA, Motorola
I 2, I 3:	CD 4008, MC 14008; RCA, Motorola
I 4:	MC 145152, Motorola
I 5:	LM 4250, National
I 6:	SP 8793, Plessey
I 7:	ICM 7555, Intersil TLC 555, Texas Instruments
I 8:	S 1469, TBB 1469, Siemens
I 9:	LM 386, National
I 10:	TL 431 C, Texas

Transistors:

T 1, T 2:	BF 979; Valvo, Siemens
T 3:	BF 246 C; Valvo, Texas, Siemens
T 4, T 7, T 8, T 15:	BC 547
T 5:	BFR 96; Siemens, Valvo, SGS
T 6:	MRF 629; Motorola
T 9, T 10, T 11, T 14:	BC 557
T 12:	BFR 90, BFR 34; Valvo, Siemens
T 13:	BFW 92; Valvo, Siemens

Diodes:

D 1, D 2:	BB 105, BB 505; Valvo, Siemens
D 3, D 4:	MBD 101: Motorola, HSCH 1001; Hewlett-Packard
D 5, D 6, D 7, D 9:	BA 182, BA 243; Siemens, Valvo
D 8:	HSCH 1001; HP
D 9, D 10, D 11, D 12, D 13, D 15, D 16, D 17:	1 N 4148
D 14:	LED red

Crystals, Filter:

6.4 MHz, 30 pF parallel resonance
10 ppm accuracy
20.945 MHz, series resonance
Crystal filter 21.4 MHz, 21 M 15 – series from Nikko Denshi
Ceramic filter CFU 455 D, Murata

Inductors:

L 1:	(VCO coil) Neosid coil kit 7.1 S with F 100 B core; 2 turns + 2 × 3/4 turn 0.5 Cul, fixed with paint or glue
L 2:	Neosid coil 00 514 831 brass core with turns reduced to 1.5
L 3:	As L 2 but turns reduced to 1.75
L 4, L 6:	6 turns 0.5 Cul on 3 mm former
L 5:	1/2 turn 1 mm silvered wire loop 5 mm dia, 7 mm high
L 7:	2.5 turns silvered on 2.5 mm former
L 8:	2.5 turns silvered on 3.5 mm former
L 9:	Neosid coil 00 514 831
L 10:	Toko LMC 1202
L 11:	Neosid Helical filter 005 119 651
L 12:	2.5 turns 0.5 Cul, 3mm former

Capacitors:

Ceramic 2.5 mm spacing
foil 5 mm spacing
Miniature electrolytics 16 V upright or tantalum, spacing 2.5 mm
Miniature foil trimmer 5 mm spacing

Resistors:

Type 0207 or 0204 carbon film
Miniature presets, upright, spacing 2.5 × 5 mm

Various Components:

3 Miniature BCD coder switches
Single pole change-over switch (relay Preset)
Miniature loudspeaker, e. g. Valvo AD 0580
Miniature electret microphone
10 kΩ variable pot. with switch (on / off-vol.)
50 kΩ variable pot. (muting)
3 ferrite beads
Switches for call tone and PTT
BNC antenna socket
RG-174 coaxial cable (antenna switch wiring)
Batteries e. g. 8 × 452 RS Varta (1.2 V / 460 mAh)
Charging socket for batteries
Housing: form optional

3.1.2. Printed Circuit Board

The construction is best undertaken in a step by step manner, completing whole circuits and then testing them before proceeding to the next part of the circuit. First chamfer all the holes, not indicated in **fig. 3**, with a dot, using a 3 mm drill. The component mounting is best started with the voltage regulator I 10, T 3 and then the VCO T 1. The supply voltage of 10 V is then applied and the 8 V and 5 V regulators checked for operation. They should function until the input voltage drops to as low as 8.5 V.

A frequency counter is then connected to the output of the VCO, in place of the frequency scaler I 6, and the VCO frequency is checked as the tuning voltage is varied. There should be a frequency shift of 200 to 225 MHz as the tuning voltage varies from 0 to 8 V. The centre frequency is adjusted by L 1 and the tuning range adjusted with the parallel 10 pF capacitor. If the frequency is too low, a brass 3 mm split screw can be used as a core for L 1.

When the VCO is working properly the doubler components can be mounted. The blocking capacitor of the BF 979 is critical, some examples can cause the doubler to selfoscillate. The coils used in the filter following the doubler are UHF types and their tuning range is limited, don't forget to modify them in accordance with the parts list. The tuning range also is made somewhat unpredictable owing to the ± 0.5 pF tolerance of the tuning capacitor which could lead to quite large frequency errors. If the trick with the 3 mm brass screw slug does'nt work, the next norm value of tuning capacitor will have to be tried. For this adjustment terminate the filter output (anode of PIN diode D 5) with a power indicator and tune the filter in order that both the receive frequency (415 MHz) and the transmit frequency (435 MHz) are at the peaks of the pass band humps. The output power should be at least 3 mW at both frequencies, the author obtained 5 mW. It may be possible, by bending the high-pass coil in T 2's emitter, to obtain even more power. If the two humps of the response are too far apart, the coupling capacitor must be reduced. (0.8 to 0.5 pF).

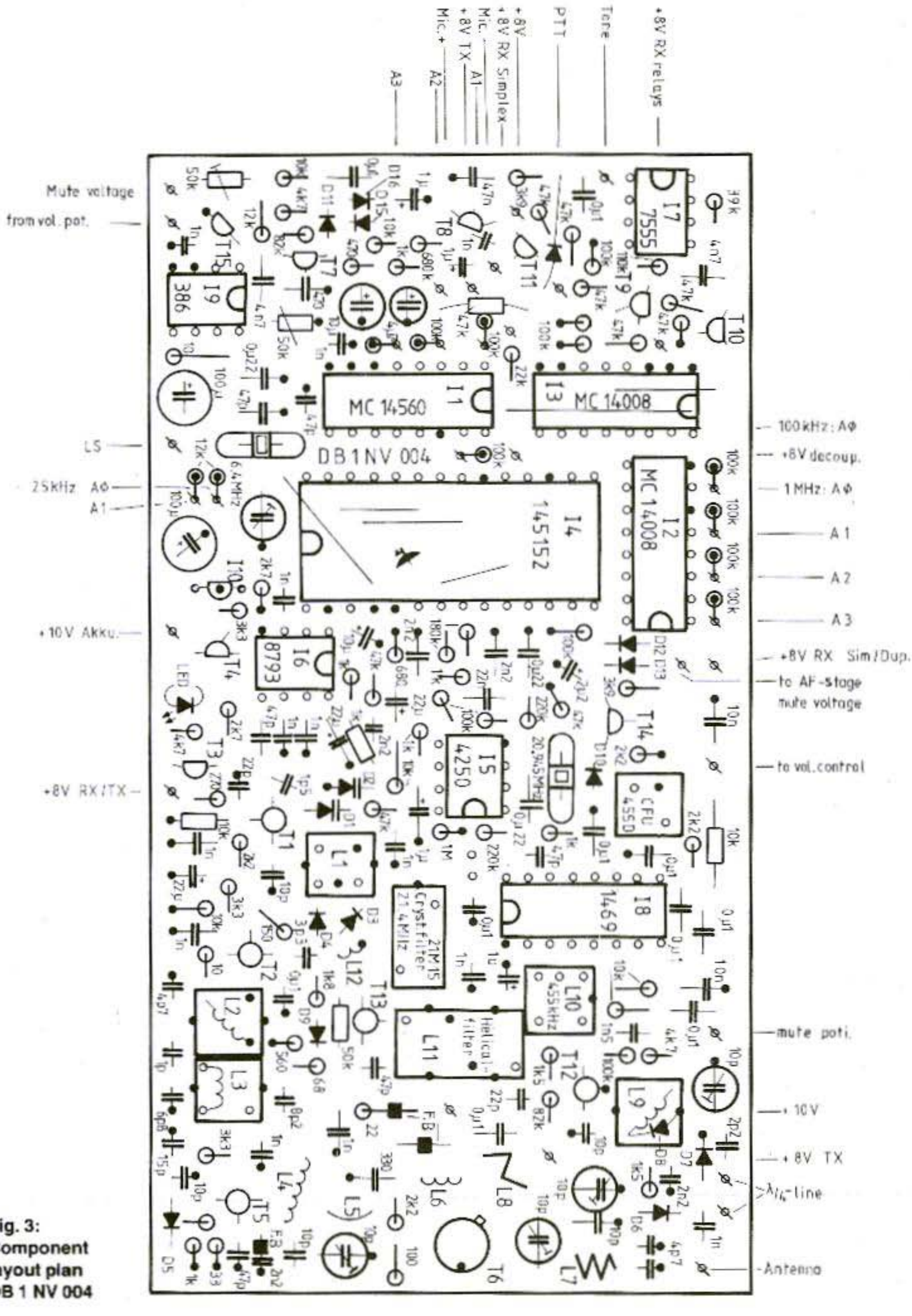


Fig. 3:
Component layout plan
DB 1 NV 004

Assuming that everthing has been tuned and adjusted satisfactorily, the scaler SP 8793, the synthesizer IC and the rest of the associated DC circuits I 1 to I 3 and I 5 may be inserted. Do not forget the four wire bridges before soldering in I 4 and the two bridges under I 3.

In order to achieve a good adjacent channel suppression, the external circuitry of the regulating amplifier I 5 should be of close tolerance, 5 % or even 1 % resistors and the foil capacitors should be of 5 % tolerance. With the coder switch in operation, together with the tuning voltages for both send and receive, the synthesizer should now operate and the exact frequency adjusted by means of the crystal trimmer.

If, after switching the unit on and off a few times, a false frequency appears, or the frequency jumps when touching the encoder ICs, it is most probable that an input to a CMOS IC is floating – perhaps a forgotten high-resistance solder connection. The electrolytic capacitors on the MC 145 152 pin 3, ECL scaler SP 8793 pin 2 and on the control amplifier LM 4250 pin 7, should not be either dispensed with, or a smaller value employed, as this will affect the spurious product output level.

When the synthesizer is working satisfactorily, the transmit amplifier can then be completed. The only point to watch here is that the ferrite bead in the emitter decoupling of T 5 (BFR 96) does not cause an oscillatory circuit to be set up with the 2.2 nF capacitor. This capacitor is required as a low-frequency decoupler. Before the transmit section can be tested, an RF choke must be connected from the cathode of PIN diode D 6 to earth. This choke consists of 10 turns 0.5 CuI on a 3 mm former soldered to ground. The choke ensures that, in the absence of the receiver portion, RF is fed through the diode to the antenna because of the flow of diode bias current. The send amplifier can now be tuned for maximum output power over the whole of the band. 500 mW should be obtainable in the centre of the band falling to 300 mW at the band limits. A final tweak of the output matching of the output transistor T 6 may help to reduce the current consumption consistent with the specified output power.

When the modulator (T 7, T 8), the transmit / receive switch (T 9, T 10, T 11) and call-tone generator I 7, are completed, the whole of the transmit side is ready. The frequency of the call-tone generator is fixed, for reasons of space saving, and is determined by a resistor (marked with an $*$) which must be selected. The timing capacitor 4.7 nF should be a high-quality foil type. It must be ensured that the ratio of the time-constant resistors must always exceed 2.5 to 3 in order to obtain sure-fire oscillator start behaviour.

The main point about the receiver construction is that the grounding connections of the helical filter are of crucial importance in obtaining a well-defined filter response curve. The receiver tuning is best carried out using a sweep generator applied to the antenna input, the sweep detector is connected to the emitter of T 13. The input is tuned for a clean, 10 MHz wide bandpass. During this adjustment, the VCO must be switched off with, for example, a temporary short between base and emitter of T 1. If a sweep generator is not available, it is possible to adjust the tuning using strong received signals at the band limits. Following the receiver input tuning the demodulator inductor L 10 is adjusted for maximum AF output.

If the audio output is distorted badly the fault may lie in a poorly adjusted second mixer crystal oscillator frequency. It should be tuned to 20.945 MHz ± 1 kHz. The loading of components onto the PCB is now completed and a few ideas concerning the construction of the handheld TR will be presented.

3.3. Housing Construction and Wiring

For the final commissioning of the PCB a few external components and connecting leads must be provided (**see fig. 4**). To itemize them, they are: –

- 10 V supply voltage from source to the point "+ 10 V from battery" and "10 V" near the Tx output transistor. The supply to the output

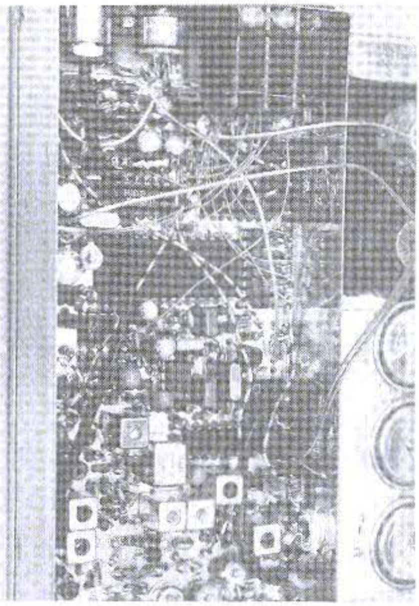


Fig. 4: 70 cm Handheld TR prototype

stage is separate in order to provide it with a power limiting resistor.

- Antenna
- Cabling of transmit / receive switch. 11 cm of RG 174 coaxial cable are required.
- Switch voltage from "+ 8 VT" at the output stage
- Stabilized voltage "+ 8 V" to "+ 8 V" in transmit / receive switch
- Loudspeaker
- Microphone
- Code switch; the wiper is connected to "+ 8 V decoupled"
- Shift switch to "+ 8 VR", "+ 8 VR relay" and "+ 8 VR Dir"
- Volume pot'meter
- Muting; There are two possibilities here, either strap the points "Mute switch voltage" and "AF switch voltage" together and make the mute level pot'meter variable from the TR exterior or solder a preset onto the PCB, and switch the preset mute noise level via an on / off switch in the above line.
- Call-tone button switch
- PTT switch

It is absolutely necessary that the housing be made of metal in order to prevent hand-capacity effects on the synthesizer. Using a few tricks, it is possible to use the same tin-plate housing as that used by the two-metre version. It might, however, be easier (and lighter) to use a larger aluminium housing.

It must be ensured that a good earthing system for the PCB is achieved as otherwise, there will be crackling noises as the set is moved around in operation. The author accomplished a solid PCB grounding by soldering a 5 × 5 mm brass angle section from 0.5 mm stock on both ends of the PCB and screwed them both directly on to the housing.

3.4. Postscript

Since early 1986 the Plessey distributors have had a new UHF scaler, with a smaller current consumption, available. For example the SP 8719, an 80 / 81 scaler for max. 520 MHz or the SP 8716 a division ratio of 40 / 41 for the same frequency limits. Using one of these scalers, a direct synthesizer at 435 MHz is easily achieved.

He who is not afraid of a little experimentation, can alter the circuit of the VCO so that it works directly at 70 cm, i. e. without the doubler, and the MC 145 152 modify to suit the new conditions i. e. 25 kHz phase-comparator frequency etc. The whole PCB could then be redesigned to make it even smaller.

4. DUAL-BAND VERSION

Right at the start of the design of the set, it was planned to construct a 2 metre and a 70 cm board in one housing thus realizing a dual-band hand-

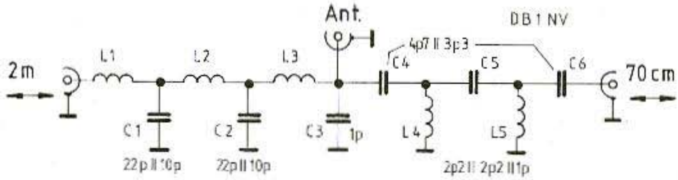


Fig. 5: Diplexer circuit diagram

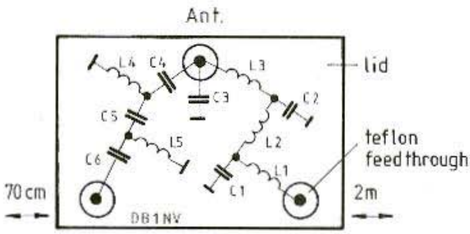


Fig. 6: Construction of diplexer

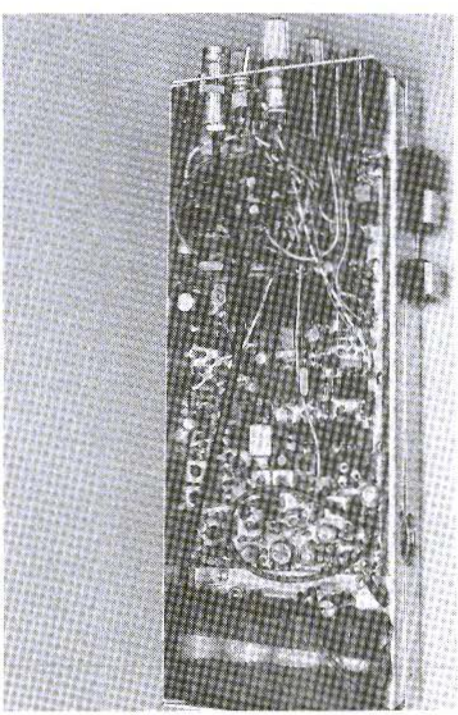


Fig. 7: Internal view of dual-band Handheld TR (70 cm part)

held TR. The concept is more interesting if it could be used to provide duplex working or the monitoring of two frequencies. This is the reason why a careful choice of mixing / multiplication frequencies was necessary to minimize detectable spurious signals caused by mixing products between the two TRs. The key element in this concept is a very small antenna diplexer which will be looked at in detail.

As can be seen in the circuit diagram of **fig. 5**, the diplexer comprises two cascaded high-and low-pass filters in a T circuit. The capacitors in every case, consist of two miniature ceramic capacitors (leads 2.5 mm) in order to minimize the self inductance. Tuning elements are, in general, not required.

The winding details are: –

L 1 = L 3:	4.5 turns int. dia 3 mm, length 6 mm
L 2:	8 turns int. dia 3 mm, length 9 mm
L 4 = L 5:	2 turns int. dia 3 mm, length 3 mm

All 0.8 mm-dia silvered wire.

The housing used was one which happened to be available, size 47 mm × 28 mm × 12 mm, and provided with a lid. Constructors could use the TEKO container 433 (50.5 mm × 31.5 mm × 13 mm). All the diplexer components are soldered into the lid as shown in **fig. 6**. The HF is connected via three teflon feedthrough points. No screening walls are necessary as the isolation between elements is achieved by physical orientation.

Following the soldering of the diplexer into the lid the following data was obtained: –

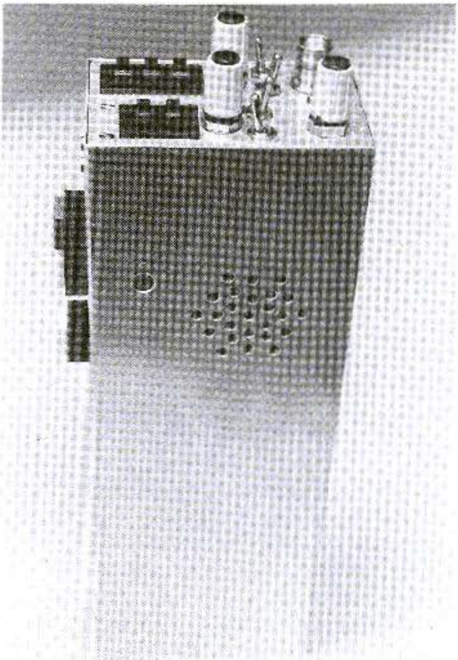


Fig. 8: Dual-band Handheld TR

Insertion loss 2 m:	$<$ 0.2 dB
Insertion loss 70 cm:	0.5 dB
Stop-band loss 2m to 70 cm:	45 dB
Stop-band loss 70 cm to 2 m:	48 dB

The cross-band isolation was considered to be more than sufficient. To give some idea, two mast-mounted antennas for the two different frequencies have a cross-coupling loss of between 20 and 40 dB.

The diplexer is capable of handling powers above 10 W making it suitable for fixed-station use. A similar construction, using three coaxial sockets, serves the author for the connection of 2 m and 70 cm FM transceivers to a discone antenna. Using this diplexer, perfect duplex operation is assured.

The **figs. 7 and 8** give an impression of the construction of the 22 cm × 7.5 cm × 4.7 cm dual-band handheld TR. A shortened Lambda / 4 antenna (40 cm long radiator) fitted to a base inductor can be used. Telescopic antennas driven in a 1/4 or 3/4 wave mode may also be employed.

Matjaž Vidmar, YU 3 UMV

Microstrip Transverters for 23 and 13 cm Part 1

1. INTRODUCTION

The microstrip technology is certainly well known to anybody working in the RF or microwave field, since most of both professional and amateur equipment is, at least, partially built using it. Due to its widespread use, the name "microstrip", itself does not really say much about its application in equipment, ranging from simple transmission lines and matching transformers built in microstrip and combined with components manufactured in other technologies (cavity, coaxial, waveguide) to complete circuits including single resonators, complex filters, power splitters, couplers, chokes, matching transformers, tuning stubs, small capacitors and even antennas! In the production of professional equipment, the microstrip technology brings a significant reduction of the manufacturing time and, consequently, a reduction of the overall cost. In addition to this, a number of theoretical tools, measuring instruments and computer programs were developed to reduce both the design time and the production line tuning of single circuits.

Unfortunately, most amateurs do not have access to the expensive professional instrumentation or computer-aided-design (CAD) tools. On the other hand, many very common design problems cannot easily be solved by theoretical tools or CAD programs like (real world) lossy laminates having an anisotropic dielectric constant ε, or semiconductor devices operating in their non-linear region (mixers, varactor and transistor multipliers, power amplifiers). Of course there are other even less predictable factors, such as, the influence of various shields and / or the resonances of the metal case actually containing the microstrip circuit. Practical experiments are therefore necessary in any case, even with the best CAD program.

Fortunately, we amateurs only have moderate requirements such as narrow-band operation or gain tolerances. Since most of components used in our designs are usually not sufficiently characterized at microwave frequencies in data sheets, such as cheap plastic case transistors or conventional glassfiber-epoxy laminate, the logical design procedure is to roughly calculate or estimate the circuit parameters and then practically optimize the circuit performance.

Microstrip circuits are usually built as a double-sided printed circuit board. The transmission lines and other microstrip components are all etched on one side of the PCB. The other side is not etched, since it acts as a ground plane for the transmission lines and other components. Since the distance between the transmission lines and the ground plane (thickness of the laminate) and the widths of the lines are small, compared to the wavelength and to other circuit dimensions, it is assumed that most of the electric and magnetic field is constrained to the close proximity of the transmission line. Since the magnetic and electric field intensities decrease rapidly with distance, microstrip circuits usually do not require any shields or additional ground planes. Additional metal planes or even closed metal boxes generally only have a very small influence on the circuit. Unfortunately, closed metal boxes have self resonances with very high Q-factors. At these particular frequencies they can introduce considerable unwanted couplings even between physically distant microstrip transmission lines. There are many efficient solutions for such problems, and some of them will be shown later in this article. Actually it is necessary to understand, that improper shielding may even introduce new problems at microwave frequencies!

Selective circuits at microwave frequencies may be implemented using λ / 4 or λ / 2 microstrip resonators as stand-alone resonators or arranged in more or less complex filters. In the amateur literature two basically different designs are described. The first uses full-size fixed-tuned resonators and therefore requires very close tolerances of the PCB laminate and the circuit pattern etched onto it. Due to the manufacturing tolerances, the loaded Q-factor of the single resonators has to be kept low and a large number of resonators are required to obtain the desired spurious frequency rejection. A large number of resonators in series calls for a low-loss, expensive teflon laminate as a substrate material. Practical experimenting is difficult and costly.

The other design approach employs significantly shorted λ / 4 resonators (acting practically as coils) by capacitive trimmers. Suitable trimmers allow a very broad tuning range. Unfortunately, this also means that the circuit may be easily tuned on the wrong mixer sideband, harmonic or other spurious frequency! Since the trimmer does not act only as a tuning element but it also provides a significant part of the capacity required in the circuit, the tuning may become very sharp and critical and the mechanical stability (post tuning drift) may not be sufficient. This is especially true when using cheap trimmers, not originally designed for microwave frequencies, close to their minimum capacity. Suitable microwave trimmers are, at least, an order of magnitude more expensive and are not easily available. In any case the circuit contains an unpredictable variable, the parasitic reactances of the trimmers used, making the duplication in amateur conditions considerably more difficult.

In the transverters described in this article, a different solution was sucessfully tested. The filters in the transverters are made of single or coupled, full-length λ / 4 resonators. The λ / 4 microstrips are etched very close to the final dimensions and the tuning is performed by adjusting the length of the resonating strips at the hot end (see also fig. 1). Cutting the hot end of the strip, produces large frequency variations and is of course an irreversible operation. A fine frequency adjustment can be obtained by soldering a short length of 1 mm Ø silver-plated copper wire at the hot end of a 2 or 2.5 mm wide strip (characteristic impedance 60 or 50 Ω respectively on an 1.6 mm thick glassfiber epoxy laminate).

Of course a reliable method has to be used to detect the actual resonant frequencies of the microstrips. A very simple method is to use a small dielectric rod (plastic screwdriver for RF ferrite cores) and approach it to the hot end of the microstrip. The presence of the dielectric rod causes a decrease of the resonant frequency of the microstrip. Monitoring the output of the circuit it can be immediately discovered whether the microstrip resonator is too short – the presence of the dielectric rod increases the output signal, or whether it is too long – the presence of the dielectric rod decreases the output signal.

Where larger capacity variations are required due to the lower loaded Q (as in the 1296 MHz power amplifier), a small piece of thin copper plate is used in place of the silver-plated wire to tune the circuit.

In both transverters for 1296 MHz and 2304 / 2320 MHz the required RF selectivity is not concentrated in a single multi resonator filter but it is distributed among the RF amplifier stages, both in the receive signal path and in the transmit signal path, mainly in the form ct two resonator filters (see **fig. 2**) which are used, at the same time, as matching devices between two amplifier stages.

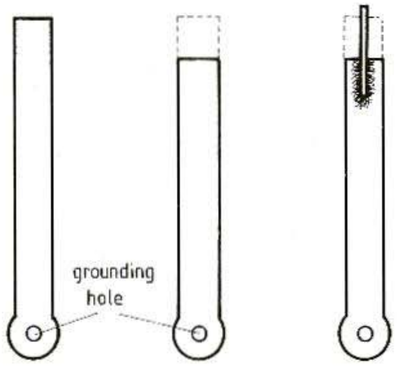


Fig. 1: Tuning a λ / 4 microstrip resonator

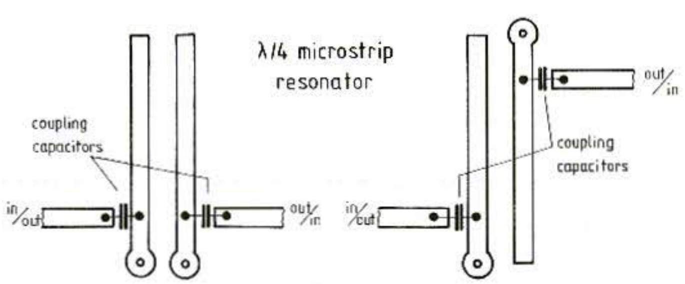
Left: Original λ / 4 microstrip resonator as etched on the PCB

Middle: The resonant frequency is increased by removing part of the microstrip with the aid of a sharp knife

Right: The fine frequency tuning is accomplished by soldering a short length of wire at the resonator's "hot end"

To obtain a usable value of coupling between two λ / 4 resonators there are two basic arrangements: both microstrips parallel and oriented in the same direction (**fig. 2 a**) and both microstrips parallel but oriented in opposite directions (**fig. 2 b**). The coupling capacitors allow a more convenient selection of the taps on the microstrips and provide also DC decoupling of the amplifier stages.

Designing the transverters, particular care was taken to use exclusively cheap and easily available materials and components without degrading the overall performance or the reproducibility. Both transverters are built on low-cost glass-fiber epoxy FR 4 laminate which has noticeable losses at 2304 MHz (this is probably its frequency limit for high Q selective circuits). Except for the RF power amplifiers all the transistors are packaged in low-cost plastic cases. **The reproducibility can only be enhanced by designing out the needs for critical components like chip capacitors or microwave trimmers.** All the critical RF grounds are therefore directly connected to the ground plane on the other side of the PCB or to "printed" capacitors. The remaining capacitors are conventional ceramic disc (max. diameter 5 mm) or pearl types with wire leads, even those used to couple the microstrip resonators, since no difference could be measured in the electrical performances when replacing them with the more expensive, and fragile, chip capacitors.



Figl. 2:
Two resonator filters in microstrip technology
Left: parallel microstrips
Right: opposed microstrips

2. BLOCK DIAGRAMS

The block diagrams of the 23 cm and 13 cm microstrip transverters are shown in **figs. 3 and 4** respectively. Both transverters are of modular construction, each "box" on the block diagrams representing a single module built on its own printed circuit board. Separate mixers are being used in the transmit and receive signal paths. Since the single - ended bipolar - transistor mixers are termination sensitive, each converter has its own last LO multiplier stage or stages.

Both transverters include a solid state, RF-antenna switch with PIN diodes to replace expensive and potentially unreliable coaxial relays. The VOX module is used to interface the transverters to any conventional 144 MHz base transceiver having a common transmit / receive antenna connector. The VOX module includes an RF detector driving a solid state DC supply switch, a receive IF preamp at 144 MHz with a base station, TX protection circuit and a power attenuator to reduce the base station TX power feeding the transmit converter. Of course the operation of the VOX module must be "transparent": it must not limit the operational performance of the transverter in any circumstances. On the other hand the VOX module simplifies the operation and increases the reliability, since a single-connection cable is used between the base RTX and the transverter, and the circuit of the transverter can not be damaged by a wrong connection or a faulty cable.

The 23 cm transverter has a single local-oscillator module, since the 1296 MHz segment is being used for narrow-band operation in most countries. The LO power splitting at 576 MHz is made with a simple capacitive divider (**fig. 16**). The 13 cm transverter has two local-oscillator modules since only 2304 MHz segment is allowed in some countries (Italy) and only the 2320 MHz segment is allowed in some other countries (Germany). Fortunately, in Yugoslavia and in many other countries, both segments are allowed and transverters covering both subbands are required to be compatible with all possible correspondents. A diode switch is required in this case to switch between the outputs of the LO modules and the LO inputs of the converters. If operation in a single subband only, is required, the LO module output may be connected as in the 23 cm transverter.

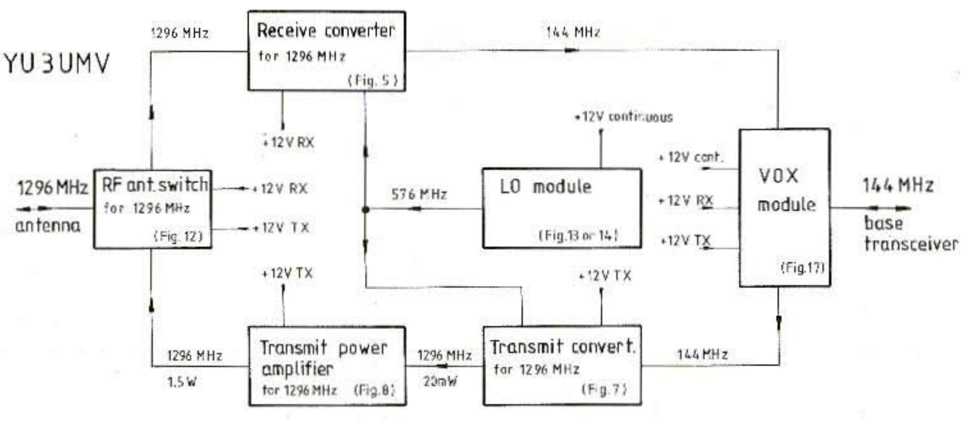


Fig. 3: Block diagram of the 1296 / 144 MHz transverter

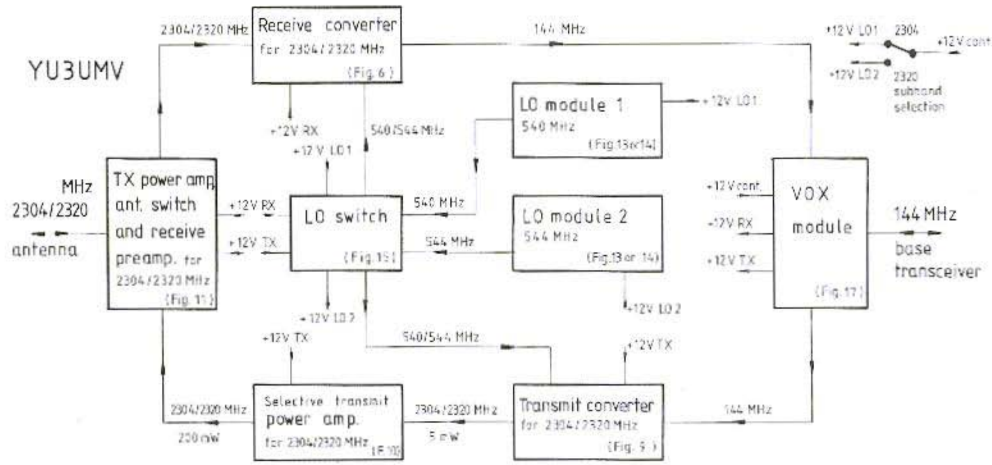


Fig. 4: Block diagram of the 2304 / 2320 / 144 MHz transverter

Finally the modular construction allows a number of variations, induding modules built in other technologies, and / or transmit only, and receive only converters primarily for the satellite uplink band around 1270 MHz and downlink band around 2400 MHz respectively.

3. RECEIVE CONVERTERS

The receive converters for 23 cm (**fig.5**) and 13 cm (**fig. 6**) are basically of identical design except for the obvious changes due to the almost 2 : 1 frequency ratio. Both converters have two RF amplifier stages (T_1 and T_2), the main RF selectivity provided by the two resonator interstage filters. The input resonator only provides a broad selectivity to reject far removed interference, since its insertion loss has to be kept low to avoid noise-figure degradation.

The receive mixers employ a single bipolar transistor. Both LO and RF signals are applied to the base of the mixer transistor through $\lambda / 4$ lines (L_{10} and L_{16} in **fig. 5**, L_{10} and L_{22} in **fig. 6**). The purpose of these lines is to transform the relatively low out-of-passband impedance of the filters into a high impedance at the base of the mixer transistor. In this way the RF filter does not attenuate the LO signal and the LO filter does not attenuate the RF signal. To increase the conversion gain of a transistor mixer, the base must be efficiently grounded for the output frequency and the collector must be efficiently grounded for the input frequencies. The base of the receive mixer transistor is virtually grounded for 144 MHz through the 10 nF capacitor. The collector is connected to a low-pass π filter / impedance matching network tuned to 144 MHz. The first capacitor of the filter is printed on the PCB to minimize its parasitic inductance.

The 23 cm converter includes a single frequency doubler stage (T_3) to obtain the 1152 MHz signal from 576 MHz. The 13 cm transverter needs two frequency doubler stages (T_3 and T_4) to obtain first 1080 (1088) MHz and finally 2160 (2176) MHz from the original 540 (544) MHz signal. Transistor multiplier stages have a similar requirement as mixer stages, concerning the input and output impedances: the base should see a low impedance for the output frequency and the collector should see a low impedance for the input frequency. The function of the two stubs L_{16} and L_{17} on **fig. 6** (λ / 4 at 2160 MHz including the parasitic inductivity of the transistor package) is to provide a short circuit for the output frequency of the multiplier stage. At lower frequencies, a capacitor between base and emitter is usually sufficient (low-impedance lines L_{11} on **fig. 5** and L_{11} on **fig.6**).

The 23 cm converter reaches an overall noise figure of around 3 dB. Since the performances of the transistors used fall off rapidly with increasing frequency, the performance of the 13 cm converter is considerably worse, the overall noise figure being around 7 dB. This performance can also be attained by a far simplier interdigital-cavity diode converter, however the manufacture of the interdigital cavity requires a considerable amount of work and the mixer and multiplier diodes are not easily available, they actually cost more than all the plastic-case transistors used in the microstrip converter. Of course it is possible to use better transistors, since the tuning elements, already present in the circuit, enable a correct matching for almost any bipolar microwave transistor.

4. TRANSMIT CONVERTER AND POWER AMPLIFIER FOR 23 cm

The transmit converter for 1296 (1270) MHz is shown in **fig. 7**. The frequency doubler (T_1) from 576 MHz to 1152 MHz is very similar to that in the receive converter. The transmit mixer (T_2) is a single-ended configuration using a single bipolar

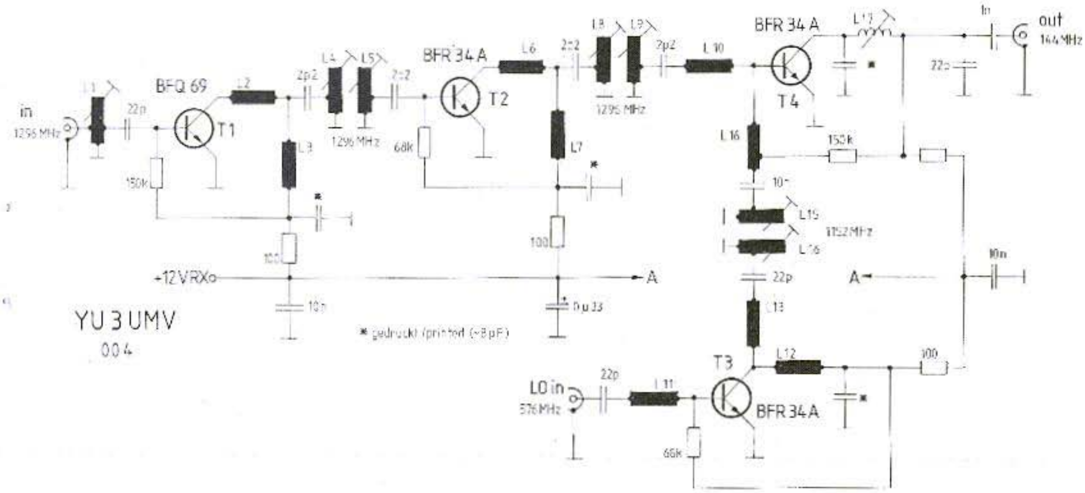


Fig. 5: Receive converter for 1296 MHz. All transmission lines L_1 to L_{16} are printed on the PCB. L_{17} self supporting, i. d. = 5 mm, wire = 0.7 CuL, L_{17} 8 turns, variable spacing!

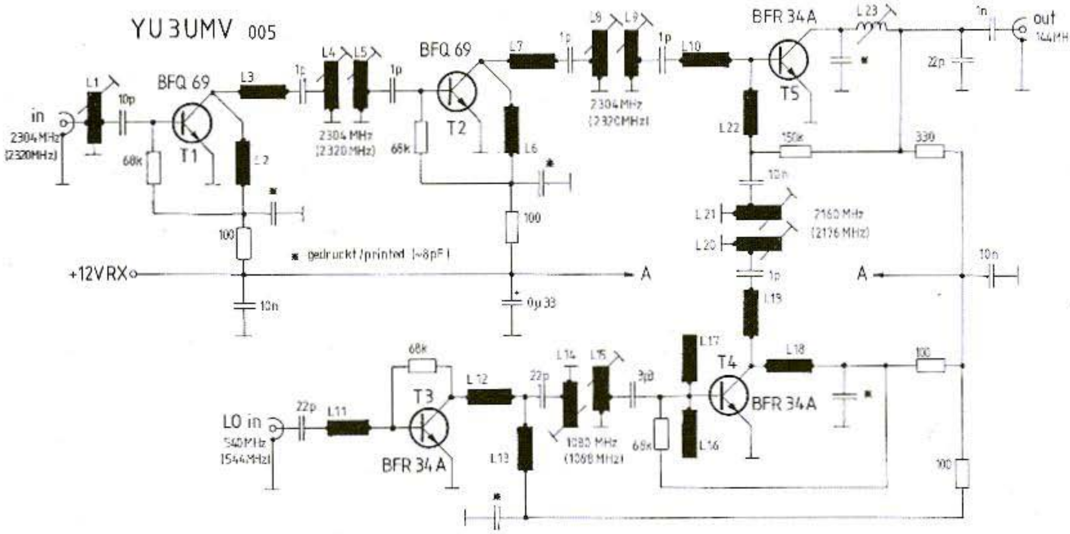


Fig. 6: Receive converter for 2304 / 2320 MHz. All transmission lines L_1 to L_{22} are printed on the PCB. L_{23} self supporting; i. d. = 5 mm, wire = 0.7 CuL, L_{23} 8 turns, variable spacing!

transistor. Both LO and 144 MHz IF signals are applied to the base of the mixer transistor. An additional 10 dB attenuator is placed in the IF signal path as it is more convenient to perform the base station TX-signal attenuation in two consecutive steps thus avoiding some otherwise, critical connections.

The transmit mixer is followed by two selective RF amplifier stages (T_3 and T_4) at 1296 MHz. The five in total λ / 4 resonators are completely sufficient to attenuate all unwanted signals such as the LO at 1152 MHz and other unwanted products generated in the mixer stage. The second amplifier stage supplies about 20 mW of power at 1296 MHz and the transmit converter can already be used as a low-power transmitter in the 23 cm band.

The transmit power amplifier for 1296 (1270) MHz is shown in **fig. 8**. It includes three amplifier stages to increase the output power to around 1.5 W. The main function of the microstrips is to provide interstage matching with minimal insertion loss. The first two amplifier stages use BFR 96 transistors, which can provide 6 to 7 dB power gain at 1296 MHz depending on the output power level and bias conditions. The first BFR 96 (T_1) operates in class AB supplying about 100 mW to the secound BFR 96 (T_2). This transistor increases the power level to about 400 mW. This is probably the maximum **safe** power level a plastic-case transistor, like the BFR 96, can supply. For higher power levels more expensive transmission transistors are required, packaged in metal-ceramic cases with a stud or flange for heat dissipation. The transistor used in the third amplifier stage (T_3), 2 N 5944, does not provide a very high gain (about 5 dB), but it is quite rugged since it was designed for transmitter operation. Since this transistor is internally matched for operation in the 70 cm band, its input impedance at 1296 MHz has a very high reactive component, compensated with L_8 and L_9. L_7 is an air-wound λ / 4 choke since a single printed microstrip λ / 4 choke was not sufficient.

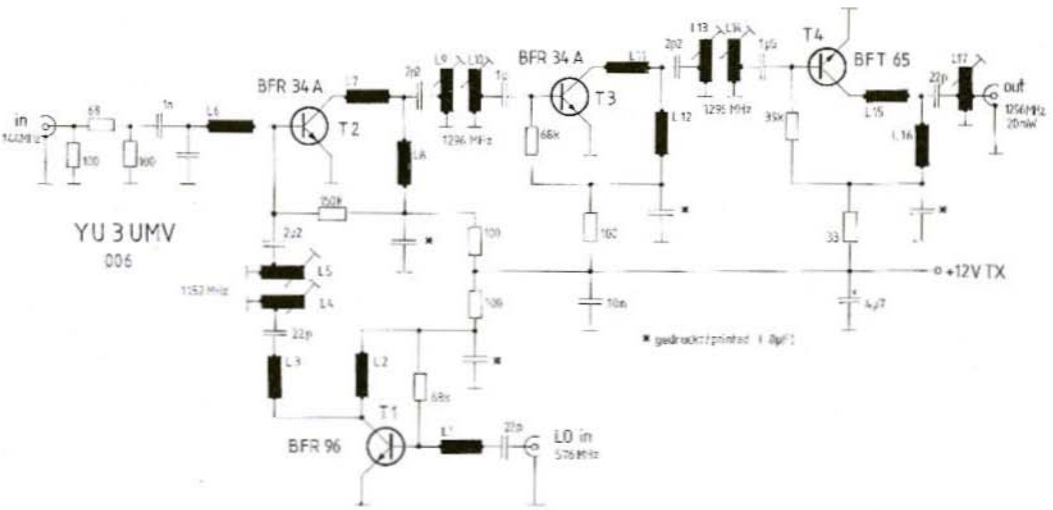


Fig. 7: Transmit converter for 1296 (1270) MHz. All transmission lines L_1 to L_{17} are printed on the PCB.

5. TRANSMIT CONVERTER AND POWER AMPLIFIERS FOR 13 cm

The transmit converter for 2304 (2320) MHz is shown on **fig. 9**. The two frequency multiplier stages (T_1 and T_2) from 540 (544) MHz to 2160 (2176) MHz are very similar to those in the receive converter. The transmit mixer (T_3) is practically identical to that for the 23 cm band, including the 144 MHz IF attenuator. However, due to the higher frequency, the transistors have a lower gain, and more amplifier stages are required. The residual LO signal and other unwanted mixing products are relatively less distant from the desired signal and therefore more filtering is required. Unfortunately, laminate losses become significant at 2.3 GHz and some gain is also necessary to overcome the losses in the micro-strip resonators.

The two selective RF amplifier stages (T_4 and T_5) following the mixer, provide about half of the selectivity required (attenuation of unwanted signals) and increase the wanted 2304 / 2320 MHz signal level to about 5 mW.

This signal feeds the selective transmit power amplifier for 2304 / 2320 MHz, shown on **fig. 10**. This amplifier consists of four amplifier stages. The first two stages (T_1 and T_2) provide the remaining selectivity and about 10 dB of gain thereby increasing the useful level to about 50 mW.

The following two stages (T_3 and T_4) employ BFR 96 transistors. With careful input matching, these can supply about 3 dB of gain per stage, and about 200 mW of power at 2304 / 2320 MHz.

Note that all the amplifier transistors are biased in class A to obtain the maximum possible gain. When bipolar transistors are operated in class A,

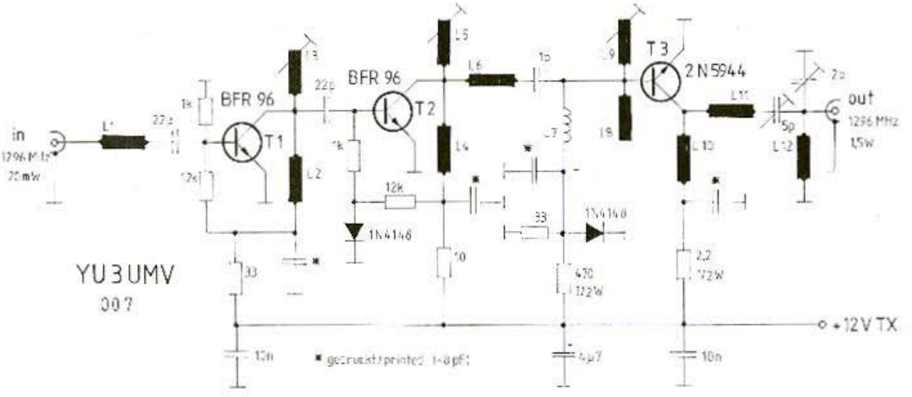


Fig. 8: Transmit power amplifier for 1296 (1270) MHz. All transmission lines L_1 to L_6 and L_8 to L_{12} are printed on the PCB. L_1 is self supporting, i. d. = 3 mm, wire = 0.7 CuL, L_7 8 turns, spaced to 12 mm

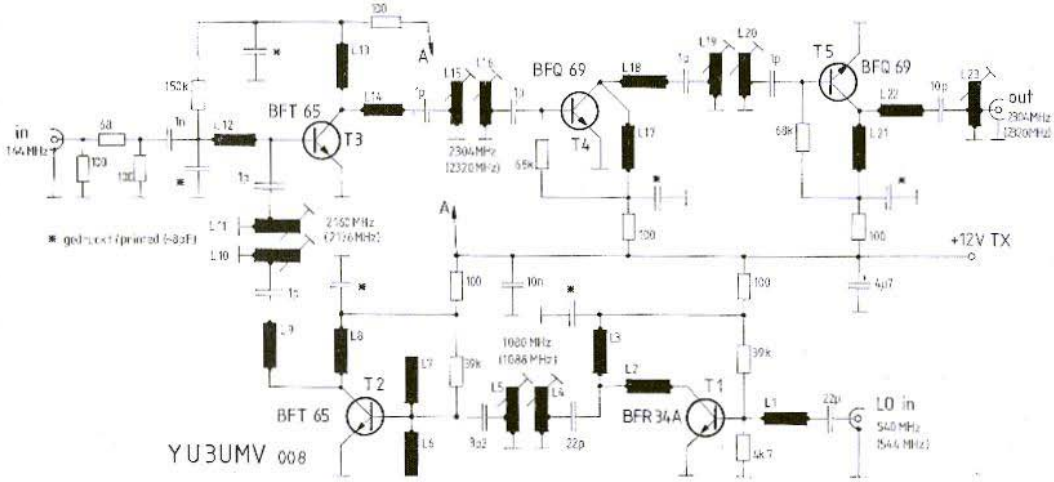


Fig. 9: Transmit converter for 2304 / 2320 MHz. All transmisssion lines L_1 to L_{23} are printed on the PCB

close to their maximum useable frequency and at high signal levels, it is very common to observe a "negative rectification" phenomenon: with the drive power applied, the collector DC current also decreases. This is actually just the opposite of what we are accustomed to when working with RF amplifiers in class AB or B at lower frequencies!

The 200 mW available from the last BFR 96 transistor are already sufficient for a low-power transverter. In this case only a RF antenna switch, such as that described in the following section, needs to be added to complete the microwave part of the transverter.

However, in the case a slightly higher output power and somewhat better receiver sensitivity are desired, a transmit power amplifier and a receive preamplifier are required. To avoid interconnection losses, both stages are integrated, together with a PIN diode RF antennas switch, onto a single printed circuit board (see **fig. 11**). The RF power transistor (T_1) BFQ 34 requires a fairly complex matching network to allow the use of similar microstrip tuning elements as in the lower level stages. T_2 BC 213 is a bias regulator for the RF power transistor. It stabilizes the operating point of the RF transistor around the optimum value of 140 to 150 mA of DC collector current in order to counter the "negative rectification" problem. The obtainable output power, subtracting the losses in the PIN antenna switch, is in the 500 mW range at the antenna connector.

The receive preamp improves the receive-converter noise figure by about 1.5 dB when equipped with the relatively cheap transistor BFQ 69.

6. RF ANTENNA SWITCHES FOR 23 cm and 13 cm

Microwave coaxial relays are still very expensive components due to the high amount of skilled mechanical work required for their construction. Since they are electromechanical components, they are also subjected to wear and are thus potentially unreliable. Fortunately, for low-power transmission only, for moderate insertion loss and cross-talk requirements, a solid state replacement is readily available. Popular PIN diodes like the BA 379 can be used to switch 5 to 10 W of RF power depending on the circuit configuration of the switch and number of diodes used. The maximum switched power is limited both by the power dissipation rating and breakdown voltage of the single diodes. The BA 379 has, however, another very interesting property: the PIN-diode structure is also very slow to turn on. Therefore RF voltages

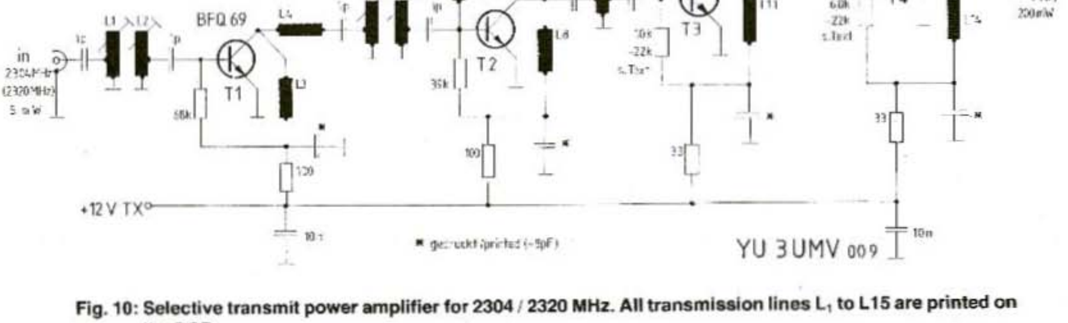


Fig. 10: Selective transmit power amplifier for 2304 / 2320 MHz. All transmission lines L_1 to L15 are printed on the PCB

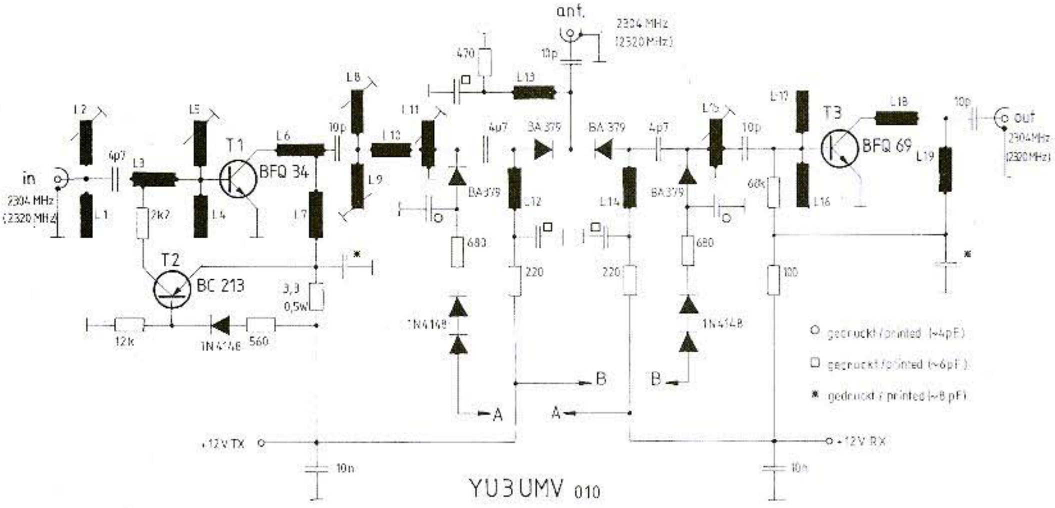


Fig. 11: TX power amp, antenna switch and receive preamp 2304 / 2320 MHz. All transmission lines L_1 to L_{19} are printed on the PCB

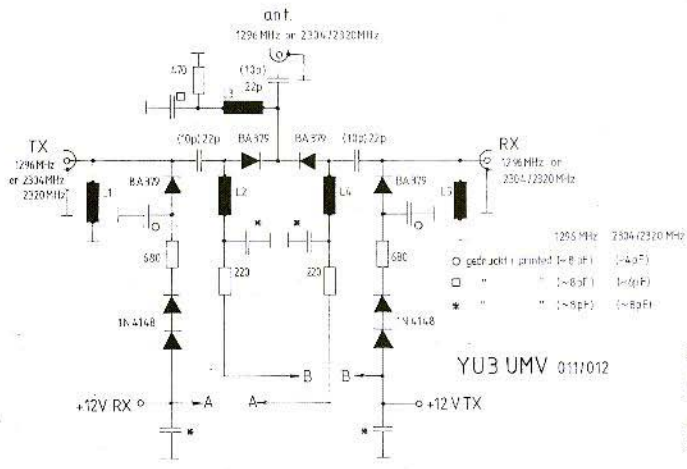


Fig. 12: RF antenna switch for 1296 MHz or 2304 / 2320 MHz. All transmission lines L_1 to L_5 are printed on the PCB

of sufficiently high frequency can not switch the diode on even if the positive halfwave amplitude greatly exceeds the diode turn-on voltage of about 0.7 V. In our particular application, this means that these diodes do not need any reserve DC bias in the non-conducting state, even if an RF voltage of more than 20 V_{PP} is applied to these diodes during transmission.

Due to the residual diode resistance in the on-state, and other prasitics, the insertion loss of the RF antenna switch shown in **fig. 12** is around 0.5 dB in the 23 cm band and around 1 dB in the 13 cm band. Accurate insertion-loss measurements are difficult due to the mismatches at coax to microstrip transition, microstrip radiation and other causes. The cross-talk attenuation is sufficient for the application shown, limiting the 13 cm TX power to below 1 W to avoid RX frontend damage.

The RF antenna switch is controlled by the two supply voltages + 12 V RX and + 12 V TX switched by the VOX module. The silicon diodes 1 N 4148 in series with the supply of the "shunt" PIN diodes are required to speed up the switching, since the supply voltages do not fall immediately to zero after a transmit / receive or receive / transmit switchover.

Concluding part in de next edition.

Jochen Jirmann, DB 1 NV and Friedrich Krug, DJ 3 RV

Microcomputer System
Part 1: Switched-Mode Power Supply (S.M.P.S.)

We would like to introduce the power supply to start the descriptions of the individual modules announced in the 4 / 85 edition of VHF COMMUNICATIONS (1). This unit is realized as a switched-mode power supply, the theoretical groundwork for this being covered extensively in (2) by DB 1 NV.

1. CONCEPT

The circuit concept of the power-supply module was motivated by the idea that, even with the largest complement of possible extension modules, all would be powered with good efficiency. In order that under partly loaded conditions the efficiency would also be favourable, the module was built as a primary switching regulator using a single-transistor choke converter.

The disadvantages of the choke converter are that the choice of suitable and sufficiently rated components is critical and that the output filtering must be quite extensive in order to suppress interference to a workable level. This will be mentioned in more detail in the circuit description. For the circuit conception, the total power requirements and working voltages were the determining factors. Working potentials of $V_1 = +5$ V, $V_2 = +12$ V and $V_3 = -12$ V at a total secondary power output of 100 W were provided under the following considerations:

For the powering of a modern microprocessor module, only a fixed voltage of + 5 V is necessary, this being standard practice since the introduction of TTL integrated circuits. The tolerance for this voltage amounted, in the most unfavourable case, to only ± 5 %, although for high-speed CMOS chips, a supply range of 2 V to 6 V is permissable.

For the operation of floppy-disc drives, monitors and a few special units, a supply of 12 V ± 5 % is necessary and a V 24 interface requires a − 12 V supply. A supply of − 5 V for older memory elements or the use of a supplementary 8080 microprocessor was not provided as, if it becomes necessary, it may be relatively easily taken from the − 12 V supply rail by means of a fixed voltage regulator.

The estimated power requirements for a maximally utilized computer system, complete with

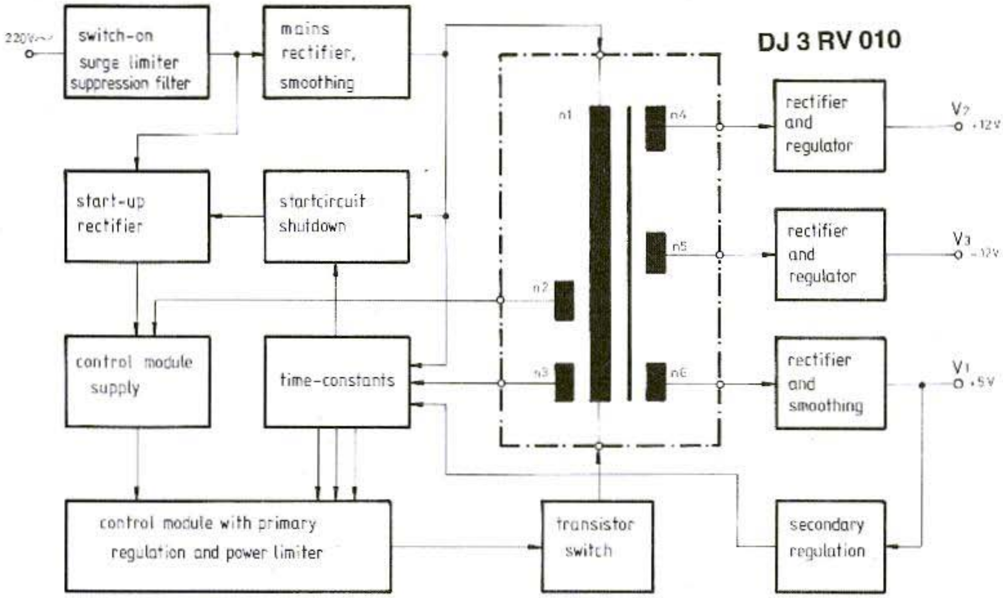


Fig. 1: 100 W Choke converter microcomputer switched-mode power supply

monitor and two floppy-disc drives, all in constant operation is:

+ 5 V / 6 A (8 A)	=	30 W
+ 12 V / 3 A (6 A)	=	36 W
− 12 V / 0.1 A	=	1.2 W
Total		67.2 W

The figures in brackets represent short-term peak values which arise from the drive motors of the floppies or from a transistory operation of a further unit such as an EPROM programmer card. It was not thought necessary to provide extra heat-sinking to cover these transistory powers.

The power loss in the rectifiers, stabilizing circuits and in the transformer must all be added to the 67.2 W given above. A figure of 100 W would cover all requirements plus a reserve for the peak demands. That is the sort of power requirement that a modern television would have, and for this application, there exists a wealth of well-tried circuits and favourably priced components. Besides this, nearly all semi-conductor manufacturers supply convenient, highly-integrated control and regulating chips for the switching transistor. At the same time, these chips provide a variety of monitor functions which greatly increases the functional integrity of the SMPS and also its sure-fire duplication. We have decided to use the Siemens control IC TDA 4600-2 or TDA 4601 for the switching transistor control module, this being comprehensively covered in (3) to (5). Both types are pin and function compatible with each other, the TDA 4601 having a low-voltage protection facility which is of no importance to this application.

Should more information be required about the basic function of switched-mode power supplies, reference is again made to (2) and also to (6), (7) and (8). The principle function of the

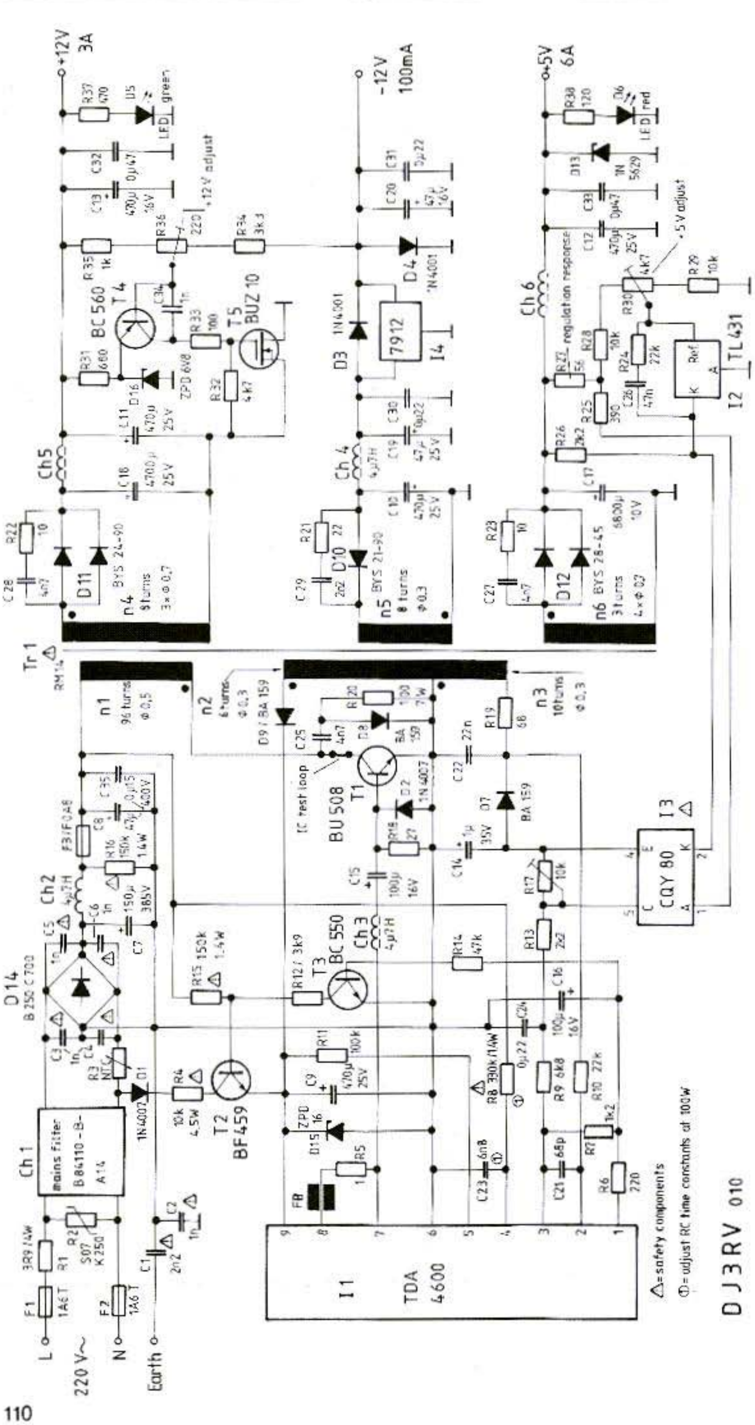


Fig. 2: Complete circuit schematic of SMPS DJ 3 RV 010

subject SMPS will be explained with the aid of the block diagram of **fig. 1**.

The choke converter stores all its energy in a "transformer" in magnetic form during the conducting phase of the switched transistor. Following the cut-off phase of the switch, the transformer winding voltage polarity changes. The rectifiers in the winding n 2 to n 6 circuits conduct and deliver energy from the transformer core into the charging capacitors. The charge time depends upon the energy stored in the core and the load on the secondaries. At the same time, the secondary voltage is heavily dependent upon the current load. In order to achieve as constant an output voltage as possible, the control circuit must control the period that the switched transistor conducts according to the quantity of energy stored in the core. A reference signal from winding n 3 assists in this purpose. The switching frequency is then a function of the load and varies between 18 kHz (full load) and 50 kHz (no load).

As the secondary load distribution can vary drastically between windings n 4 to n 6, the V_1 (+ 5 V) is regulated from a secondary regulator which also reacts upon the control-circuit regulator. The V_2 (+ 12 V) and V_3 (− 12 V) supplies are both stabilized additionally by series pass regulators. At the same time, the unavoidable magnetic coupling between windings is balanced out.

The other circuit parts in the block schematic concern the mains supply to the unit. The 220 VAC is fed into the mains rectifier via an rfi suppression filter. The rectified output is then filtered and fed to the pulsed transducer (the main "transformer" functions in this circuit electrically as a choke).

Owing to the pulsed operation of the SMPS, the required outlay, in terms of smoothing and suppression measures, is quite high.

The supply for the control circuit during operation is taken from winding n 2. Upon initial switch-on, however, this winding cannot supply any power and the control circuit must be supplied from a rectified starter voltage. After the start condition and the power supply has settled down to steady operation, the control-circuit supply is automatically switched over to the auxiliary winding n 2.

2. CIRCUIT DETAILS

Referring to the circuit schematic of **fig. 2**, the power-supply function will be described more closely and the details of the block diagram in **fig. 1** will be highlighted. The component specifications can be taken from the parts list.

2.1. Surge Limiter and Suppression Filters

The mains input voltage is taken via two cartridge fuses and R 1 to the mains interference suppression filter. This comprises a current-compensated choke of 2 × 27 mH and two X-capacitors, each of 150 nF. The varister R 2 removes dangerous voltage peaks from the mains supply. Resistor R 1 and the nTC thermistor R 3 serve to limit the switch-on surge current. They also form, together with the capacitors in the suppression filter, a low-pass filter for additional rfi suppression. This is also improved owing to capacitors C 3 to C 6 which are connected across the bridge rectifier diodes. These have the effect of Y-capacitors and are connected via C 1 and C 2 to the ground / earth of the mains supply. Capacitor C 2 is only necessary in the event that the power supply is to be fully isolated from the mains. If the conventional construction, using a metal (screening) housing, is employed, C 1 and C 2 may be dispensed with the low potential side of the mains rectifier / filter being taken directly to chassis ground.

The construction of this power supply is attended by the observation of mandatory safety measures. The capacitors C 1 to C 6 must only be those types which are expressly manufactured for this application and which have been approved by the relevant national standards authorities.

2.2. Mains Rectifier and Filters

The mains rectifier bridge D 14 provides the DC and capacitors C 7 and C 8 serve as charge capacitors. Because of the division of the capacity into two, connected via choke CH 2, they also work, in conjunction with C 35, as a high-frequency suppression filter. The fast-blow fuse F 3 is fitted for protection against fault currents. In the event of a blown fuse, R 16 acts as a bleeder resistor for C 7 – another safety measure. This resistor should, of course, be capable of withstanding the voltage across it. This goes also for resistors R 4, R 8 and R 15.

2.3. Starting Supply for Control Module and Shut-Down Circuits

Upon switching on the power supply the control circuit module I 1 is immediately supplied with power via D 1, R 4 and T 2. Capacitor C 9 charges to 12 volts in about 800 ms and the control module starts producing start pulses for the switched transistor T 1. Capacitor C 9 then discharges. The Zener diode D 15 in paralell with C 9 serves to protect the capacitor in the event of a fault. Following the operation of the transistor switch, the winding n 2 supplies a voltage, which is rectified by D 9, to the control module. Pin 1 of I 1 at the same time produces a reference voltage causing T 3 to conduct and T 2 to cut-off. Resistor R 15 has, at the same time, the function of a bleeder for C 8.

2.4. Control Module

The switching transistor T 1 is caused to conduct during the periods of the pulses supplied by the control module I 1 pin 8 via R 5, CH 3 and C 15. Resistor R 5 is a feedback element and serves to fix the base control current of T 1 with the output of I 1 pin 7. The ferrite bead FB prevents parasitic oscillation but has a minimum effect upon the form of the current pulses. Use only the ferrite bead mentioned in the parts list. During the development of this circuit, the use of the wrong ferrite material resulted in a parasitic oscillation which was difficult to locate.

The period of the base current pulse is controlled by the voltage on I 1 pin 3. It is developed across the potentiometer formed by R 6, R 7, R 9, R 13 and R 17 and supplied by a combination of the positive reference voltage at I 1 pin 1 and the rectified negative voltage from winding n 3. The secondary voltage may be adjusted by R 17 providing that the windings have a high mutual inductance. A transformer which has been hand wound usually possesses a relatively high leakage inductance and an additional control from the secondary side, via the opto-coupler I 3 and R 17, is necessary. Unfortunately, because of the long time constants in the secondary, the control is largely ineffective, C 24 prevents control-loop oscillation.

The voltage provided by winding n 3 via filter R 19 / C 22 and R 10 / C 21 to I 1 pin 2 serves as a through-zero transition detector. With this, the discharge of C 15 via I 1 pin 7 is initiated. The voltage present at I 1 pin 4, with a time constant of $\tau = R\,8 \times C\,23$, determines the power overload point, set here to 100 W. Making C 23 larger in value increases the overload limit.

At pin 5, the supply voltage to the control module I 1 is monitored via resistor R 11. In the event of insufficient voltage – perhaps a secondary short-circuit which brings down also the voltage across n 2 – the current pulse to T 1 is suppressed. This renders the power-supply output short-circuit proof.

2.5. Choke-Converter Transformer and Switch Transistor

When the switched transistor T 1 is conducting, magnetic energy produced in the winding n 1 is stored in the transformer. This energy must be stored without the transformer core saturating and therefore a large core with an air gap must be provided. The core has, in fact, an AL factor of 250 nH per turn^2. When working at its full output power of 100 watt, the duty cycle (T on / T off) is 1 : 2 and the switching frequency an inaudible 18 kHz. The conduction time T on = 18 μs.

The stored energy W (per pulse) produced with a linear slope pulse in the winding n1 may be given as: –

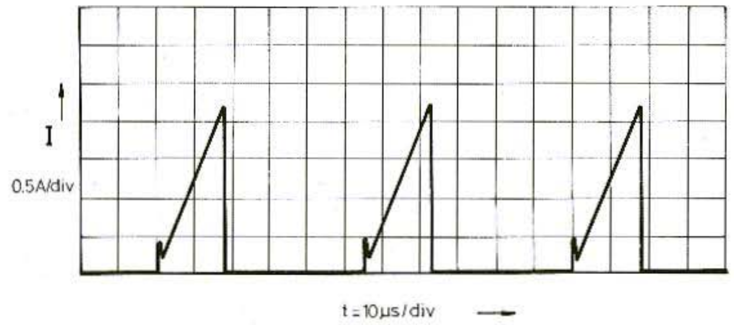
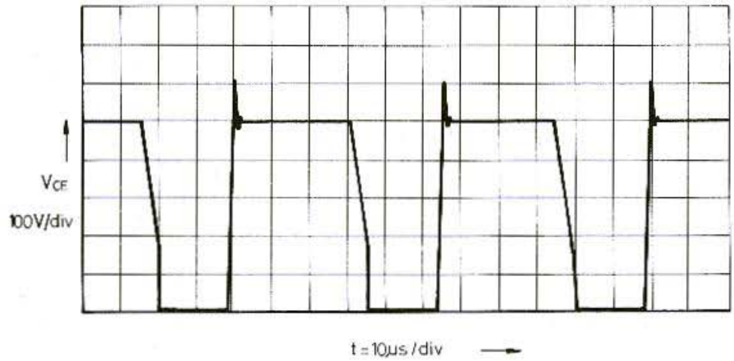


Fig. 3: Collector current I_C and collector voltage V_{CE} at T 1 at full load (100 W) output



$$W = P / f = 0.5\, L\, \hat{I}^2 = 0.5\, T_{on}^2\, V_{DC}^2 / L$$

$$L = 0.5\, f\, T_{on}^2\, V_{DC}^2 / P$$

When V_{DC} is 280 V across C 8, L = 2.3 mH

At the given AL factor of the core, the required number of primary turns is:

$$N = \sqrt{L \cdot AL} = 96 \text{ turns}$$

The current in the winding rises to $\hat{I}$.

$$\hat{I} = T_{on}\, V_{DC} / L = 2.2 \text{ A}$$

When T 1 is switched off, the T 1 collector voltage rises, diode D 8 conducts, C 25 charges and limits the rise of the collector potential. Upon switching on T 1 again, the C 25 discharge is limited by R 20. An oscilloscope monitor point has been provided on the board in order that the T 1 conduction current may be controlled. C 25 and R 20 can then be optimized for the transformer.

The current and voltage waveforms, at full load, are shown in **fig. 3** for transistor T 1.

The peak at the moment of switch-on caused by the discharge of C 25 can be clearly seen. Note the linear rise of the current to its peak value of $\hat{I}$. This linear characteristic indicates that the transformer is not saturating.

The switch-on current peak should be as small as possible as, at this instant, the collector voltage V_{CE} has not, as yet, dropped to zero, thereby developing a power loss I_C V_{CE} in the transistor. By a suitable selection of value for resistor R 20, most of the energy stored in C 25 may be diverted from the transistor into the resistor. The BU 508 A has a maximum permitted junction temperature of only 150° C and the RCD combination (slow-rise-network) of C 25, D 8 and R 20 serves to protect the transistor.

When T 1 is turned off the collector voltage rises to twice or three times the applied voltage V_{DC} and then falls, as the secondary diodes start conducting, to a steady value of about 480 V. The voltage surge is delayed by C 25 and D 8 to about 500 V / μs in order that it is limited to under that specified for the collector-emitter voltage V_{CE_0} = 700 V.

During the time that the collector potential is constant, the magnetic field is declining as the secondary delivers energy to the load. The voltage on the output winding, calculated on the data given, is about 2 V per turn thus determining the number of secondary turns necessary for the required output voltage.

indicator as well as being a bleeder for the electrolytics.

In order to hold the output within a tolerance range of ± 0.25 V during mains and load fluctuations, an additional regulation supplements that carried out via winding n3. This supplementary regulation uses the reference element I 2 and opto-coupler I 3.

The output voltage sample is taken via voltage dividers R 27 to R 30 and compared in I 2 with a 2.5 V reference voltage. The difference voltage is then caused to alter the current through the opto-coupler photo diode in the appropriate sense.

R 27 is a feedback resistor which, together with R 24 and C 26, increases the loop reaction time thereby reducing the tendency for loop instability. The combination C 26 and R 24 also serves this purpose. The photo transistor in the opto-coupler is connected in parallel to the voltage divider resistor R 17 in the regulation loop and also participates in the regulation. The time constant of R 13 and C 24 is so chosen that voltage transients, due to large load variations, remain within the specified voltage tolerance.

2.6. + 5 V Supply

Across winding n 6 a voltage of about 6 V appears during the cut-off period of T 1. The reservoir capacitor C 17 is charged up to 5 V owing to the potential difference across rectifier D 12. In order to keep the efficiency high and the ripple voltage low, all the secondary rectifiers employ Schottky diodes. The electrolytic capacitors also are special low-resistance and inductance types, which have been specifically designed for switched-mode power supplies. In spite of this, a further filtering process is necessary to keep the output voltage as free as possible from residual-interference-producing components. This is carried out by choke CH 6, C 12 and C 33. The choke is wound on a ferrite rod in order that it has a high saturation point.

The diode D 14 serves as an over-voltage protection, which, in the event a fault, limits the output voltage to + 6 V whilst LED D 6 acts as an "on"

2.7. − 12 V Supply

Winding n 5 delivers a voltage of about 16 V which is rectified by Schottky diode D 10, producing 15 V across electrolytic C 10. Following the filter elements Ch 4 and C 19, this voltage is stabilized by a series stabilizer I 4. The diodes D 3 and D 4 are protection against a polarity fault. As the − 12 V is also used as a reference voltage for the + 12 V supply, a residual load is formed by resistors R 34, R 35 and R 36. This assists the regulation and also bleeds the capacitors after switching off the supply when the load has been disconnected.

2.8. + 12 V Supply

Winding n 4 supplies a full-load voltage of approximately 15 V which is rectified, filtered and then stabilized with series-pass transistor

T 5. The reference voltage for the regulation is supplied by D 16 to the emitter of T 4, to the base of which, is applied the output sensor voltage supplied by the voltage divider R 34, R 35 and R 36. Transistor T 4 amplifies the error voltage via the gate of T 5. Resistor R 23 and capacitor C 34 serve only in the stabilization of the regulating loop and elements.

The LED D 5 serves also as a bleeder load during unloaded operation.

3. POWER SUPPLY CONSTRUCTION

Despite the rather closely packed PCB DJ 3 RV 010, the construction presents no real problems. Basically, the cooling of the power semi-conductors must be attended to. Resistors R 1 and R 20 develop a lot of heat and must be suspended above the board thus allowing a cooling air-convection current to flow around them.

Mounting the printed circuit board vertical against the inside back wall of the cabinet also helps with the convection cooling. The PCB should be so positioned that the resistor R 20 is uppermost in the cabinet. **Figure 4** gives an example of how the board is mounted, the rear cabinet wall assisting in the dissipation of heat.

The connecting leads to the module and the floopy disc drive must be of large cross-sectional area. This applies particularly to the + 5 volt output lead. A copper connecting lead of 1 mm^2 cross-sectional area has a resistance of 17 mΩ per metre, which means that with 6 A flowing, there will be a potential drop of 0.1 V per metre.

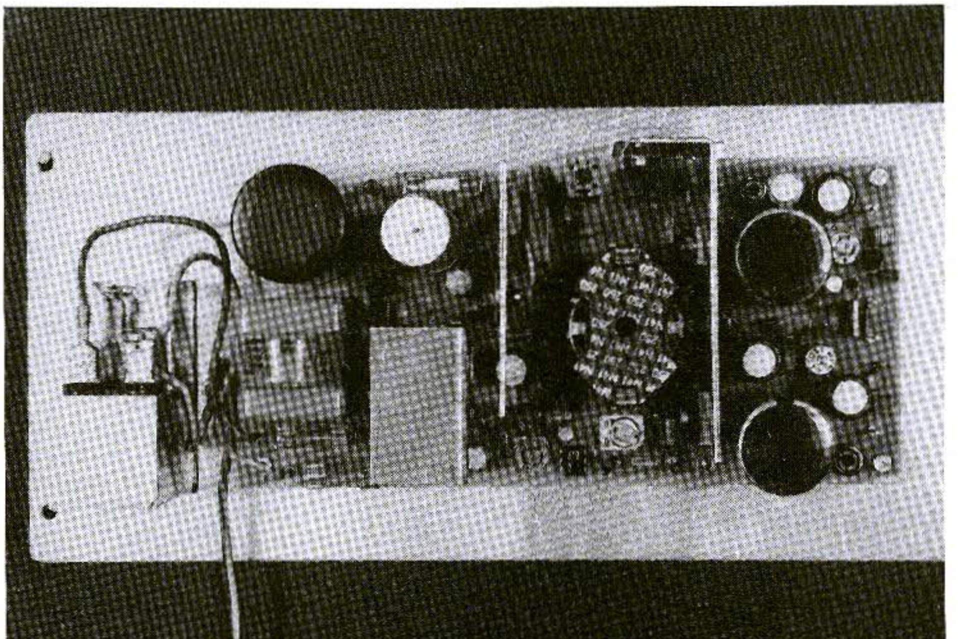


Fig. 4: The SMPS board mounted on the back wall of the cabinet

3.1. Components

For this type of SMPS with a switch frequency of between 18 kHz and 50 kHz, normal standard components from the junk box are usually unsuitable. Also, the relatively compact construction requires a few special components which are detailed in the following parts list together with manufacturers and suppliers.

Semiconductor

T 1:	BU 508 A (Valvo, Motorola etc.)
T 2:	BF 459 or similar ($V_{CE} \geq 350V$)
T 3:	BC 550 or similar
T 4:	BC 560 or similar
T 5:	BUZ 71 (Siemens) or other power FET e. g. Motorola MTP 10 N 60
D 1 - D 4:	1 N 4007
D 5, D 6:	LED ($I_D = 20$ mA)
D 7 - D 9:	BA 159 or other fast HV diode
D 10:	BYS 21 - 90 (Siemens) Schottky diode
D 11:	BYS 24 - 90 (Siemens) Schottky parallel diodes $I_D \geq 8$ A
D 12:	BYS 28 - 45 (Siemens) Schottky parallel diodes $I_D \geq 25$ A
D 13:	1 N 5629 (Thomson-CSF) TAZ Suppressor diode for 5.5 V
D 14:	B 250 C 1000 / 700 or equiv.
D 15:	ZPD 16
D 16:	ZPD 6.8
I 1:	TDA 4600-2 or TDA 4601 (Siemens)
I 2:	TL 431 CLP (TI)
I 3:	CQY 80 (Tfk), VDE proof
I 4:	7912 CT

The semi-conductors can be supplied by various firms, but Buerklin, Schillerstraße 40, 8000 Muenchen 2, supplies everything.

Capacitors

C 1:	2.2 nF, 250 V (ERO) VDE 0560 specs.
C 2 - C 6:	1 nF, 250 V (ERO) VDE 0560 specs.
C 7:	150 µF, 385 V (Siemens) B 43306
C 8:	47 µF, 350 V, upright or axial e. g. Valvo 2222 133 45479
C 9 - C 12:	470 µF, 25 V for SMPS use e. g. ROE, Type EKR
C 17:	6800 µF, 10 V (Siemens) B 41336 or 10000 µF, 10 V (Valvo) 2222 051 54103 for SMPS use
C 18:	4700 µF, 25 V (Siemens) B 41336 or (Valvo) 2222 051 56472 for SMPS use
C 19:	47 µF, 25 V for SMPS e. g. ROE, Type EKR
C 25:	4,7 nF, 1500 V (Siemens), B 32650 - K 1472 - T Pulse cap.
C 35:	0.15 µF 400 V, MKT, RM 15

All other capacitors are standard components. Values and dimensions can be seen on the diagram or component outlay.

Resistors

R 1:	3.9 Ω, 4.5 W (Vitrom) 208 - 3, wire-wound upright
R 2:	S0 7 K 250 (Siemens) Q 69 - X 3044
R 3:	K 231 / 33 Ω (Siemens) P.T.C. Q 63023 - K 1330 - N
R 4:	10 kΩ, 4.5 W metaloxyd resistor, > 300 V proof e. g. WK 8 (Resista)
R 8:	330 kΩ, 1.4 W, > 300 V proof
R 15, R 16:	150 kΩ, 1.4 W, > 300 V proof
R 17:	10 kΩ, preset. cer. horizontal
R 20:	100 Ω , 7 W (Vitron) KV 212-3, wire-wound upright
R 30:	4.7 kΩ, preset. cer. horizontal
R 36:	220 Ω preset cer. horizontal
R 37:	470 Ω, 0.5 W

All other resistors are of the type 0207.

Other Components

Ch 1:	rfi suppression filter, 2×27 mH / 1.4 A, (Siemens) B 84110 - B - A 14
Ch 2 - Ch 4:	4.7 μH Ferrite rod choke
Ch 5, Ch 6:	Ferrite rod core (Valvo) 4322 020 36810
FB:	Ferrite bead, 5 mm, mat. N 22 or FXC 3 B
Tr 1:	RM 14, ferrite core transfr. complete AL = 250 nH / turn^2 (Siemens)

3.2. Winding Details

Chokes Ch 5 and Ch 6, as well as the transformer, must be wound

Ch 5:	20 turns Cul 1 mm dia on rod core 5 mm dia
Ch 6:	12 turns Cul 1.5 mm dia on rod core 5 mm dia
Tr 1:	Start with the collector end of n 1 as reference
n 1:	96 turns 0.5 mm dia Cul in three layers with insulating foil between layers and also on top layer.
n 2:	6 turns 0.3 dia Cul
n 3:	10 turns 0.3 dia Cul Both n 2 and n 3 are laid besides each other in the same layer. The remaining space is filled with teflon tape. Use four layers of teflon tape to cover windings and isolate mains.
n 4:	8 turns $3 \times$ Cul 0.7 mm dia parallel wires, covered with insulation foil.
n 5:	8 turns Cul 0.3 mm dia
n 6:	3 turns $4 \times$ Cul 0.7 mm dia parallel wires

Both n 5 and n 6 are wound in a single layer and insulated as for n 3 and n 4 above.

note: Cul = enamelled copper wire.

The windings should be carefully wound and as tightly as possible, otherwise there will not be enough place in the available winding volume. High frequency Litz-wire can also be used in place of single conductor copper wire. As the Litz-wire is insulated with silk as well, the risk of internal arc-over due to mechanical movement of the wires in response to the current pulses is minimized. It must, however, be stressed that the winding volume is only just sufficient.

n 1:	30×0.1 Cul S
n 2, n 3, n 5:	15×0.1 Cul S
n 4:	120×0.1 Cul S
n 6:	$2 \times 120 \times 0.1$ Cul S

The Litz-wire may be obtained from the suppliers Buerklin.

3.3. Component Placement

The important thing to watch is that all the components have the specified ratings. The printed circuit board layout plan, **fig. 5**, indicates the component arrangement. The order attending the component mounting is unimportant, i. e. except for the semi-conductors. These should be left until last and after they have been mounted upon their heat sinks. This refers to T 1, T 5, I 1, D 11 and D 12, suitable heat sinks may be seen in **fig. 6** and their mounting in **fig. 4**. Take particular care with diodes D 11 and D 12 that they are insulated from their heat sinks. By means of a tin-plate wall, mounted across the board, the heat sinks are mechanically secured and thermally bonded – the wall acting as a heat sink for the PCB as well. The heat sinks of T 1 and I 1 are, on mechanical grounds, bolted together by means of nylon screws and insulating spacers – both sinks work at high (different) potentials. This latter point must be taken into account when T 1's heat sink is bolted on to the board as it has collector potential. No HV tracking must be allowed and ceramic stand-off insulators are recommended.

4. COMMISSIONING

In order to adjust the module, a (variable) isolation mains transformer and two high-wattage resistors for the output load, are required.

A digital multimeter and a two-channel oscilloscope with a current clamp and 1 kV probe are also necessary. The SMPS is switched on without load and with potentiometers R 17, R 30 and R 36 set to a mid-range position.

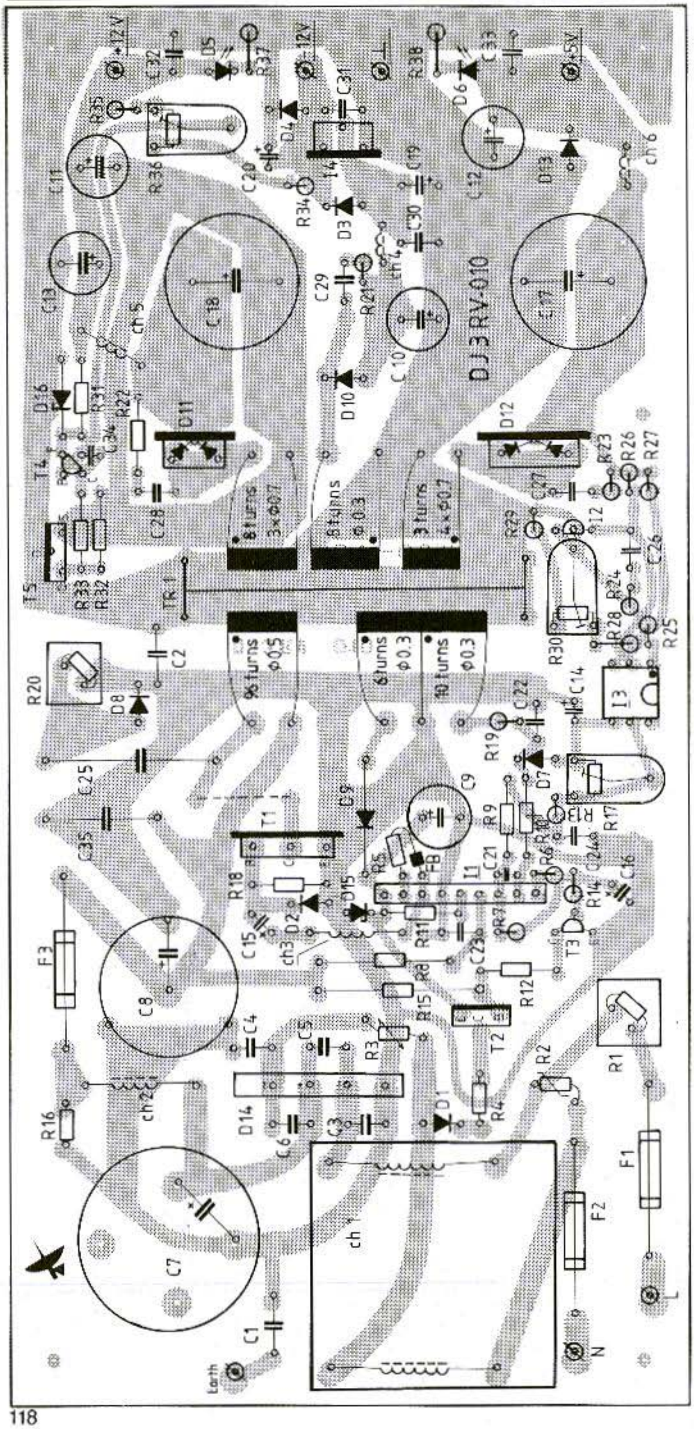


Fig. 5:
Component layout
of PCB DJ 3 RV 010

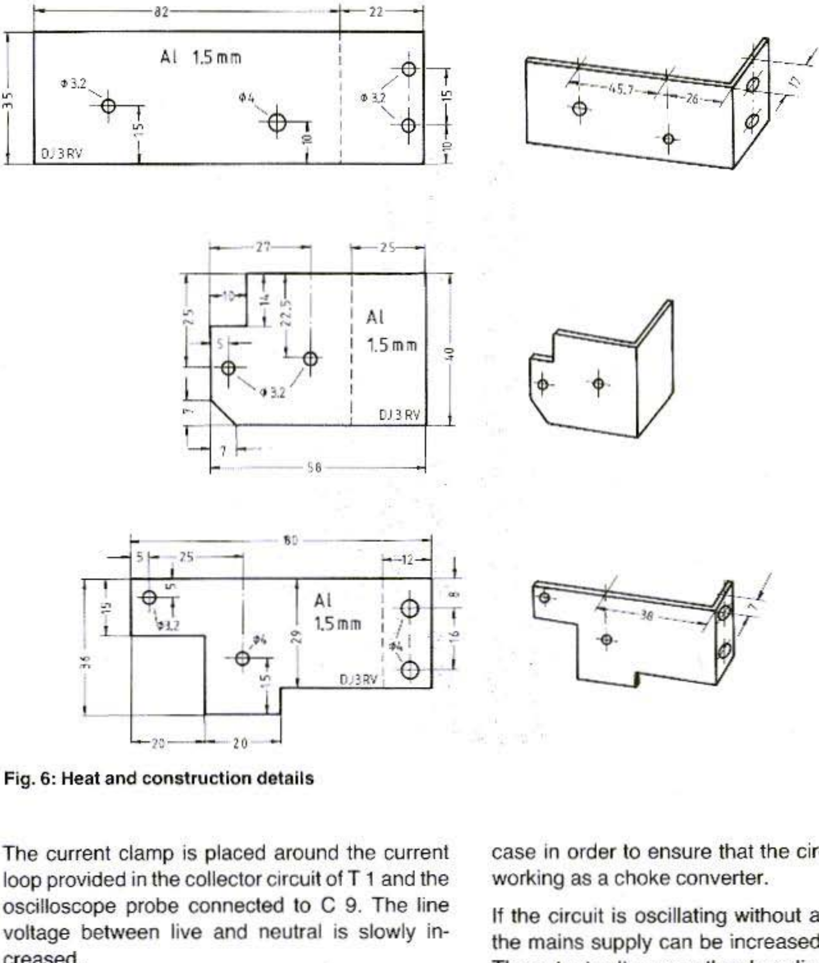


Fig. 6: Heat and construction details

The current clamp is placed around the current loop provided in the collector circuit of T 1 and the oscilloscope probe connected to C 9. The line voltage between live and neutral is slowly increased.

At about 70 VAC the control circuit starts delivevering pulses to T 1. The oscilloscope trace shows the voltage pulses and collector current is indicated in the clamp ammeter.

Above approximately 120 VAC the SMPS starts to oscillate and an output voltage appears. The switching frequency is about 45 kHz and the pulses have the form shown in **fig. 3**.

If the circuit is not forced into oscillation by the start impulses, the sense of the winding connections on the main transformer should be checked. This check should be carried out in any case in order to ensure that the circuit is in fact, working as a choke converter.

If the circuit is oscillating without any problems, the mains supply can be increased to 220 VAC. The output voltage can then be adjusted with R 36 and R 30 to a voltage of 12 V and + 5 V at partial load currents of 1 A and 2 A respectively. Under partially loaded conditions (of the 12 V supply), the 5 V supply should not be left unloaded, otherwise, the strongly regulated 5 V supply will cause the voltage across n 4 to be diminished.

The SMPS should now be tested at the load extremes in order to test the output voltage regulation tolerances. If the predominating effect of the 5 V regulation is too strong, R 17 can be adjusted, thereby decreasing the feedback from n 3 via the opto-coupler. The 5 V output must then be reset to nominal by means of R 30.

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Friedrich Krug, DJ 3 RV

Microcomputer Clock-Pulse Generator linked to DCF 77 Off-Air Time-Standard

For the correct evaluation of coherent transmitted computer signals the synchronization must be of the highest order of accuracy and stability. Together with the high input dynamic range and interference resistant DCF 77 receiver, DJ 3 RV 006 / 007 described in (1), and the subject module DJ 3 RV 009, the necessary synchronizing signals are generated and phase-locked to an off-air (German) frequency standard. The module delivers the 4 MHz CPU sync. and the 16 MHz dot clock for the terminal card of the forthcoming computer described in VHF Communications 4 / 85 (2).

1. CIRCUIT DESCRIPTION

The circuit of this module is shown in **fig. 1**. The low noise junction FET (T 1) Collpits oscillator has already been used in this form in the circuit DJ 3 RV 007. The basic mode of the crystal (Q) oscillates in parallel resonance at 8 MHz and is fine-adjusted with the 10 pF trimmer and the varicap diode D 1 in order that it is phase-locked to a radio-standard derived frequency.

The oscillator signal is then low-reactively coupled to transistor T 2 where it is amplified and passed to a filter L 1 and a level translator T 5. The latter amplifies the signal to CMOS level and controls the frequency divider in I 3. The 8 MHz signal is then divided by a factor of 40 and compared with a DCF 77 derived, 200 kHz signal. The integrated circuit I 3 contains two programmable dividers connected in series. The first with a division factor of 4 : 1 and the second with a division factor of 10 : 1 fixed. The output of the first divider is available at pin 1 and the second at pin 2. These signals at 2 MHz and 200 kHz are available to adjust the frequency of the crystal oscillator.

The output of the phase comparator is taken from the I 3 tri-state output pin 3 and controls the phase

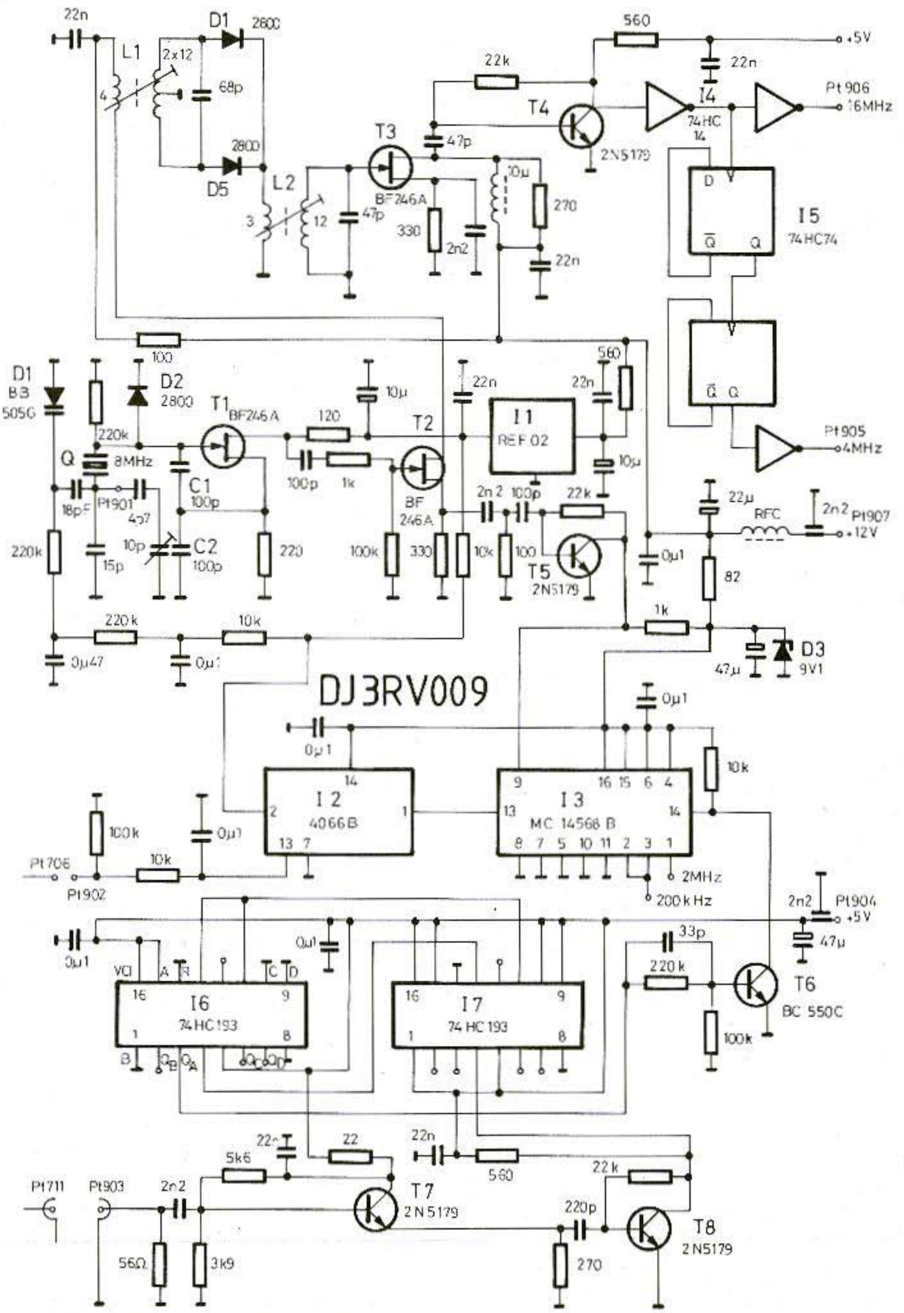

Fig. 1: Circuit of Clock-Pulse Generator Module

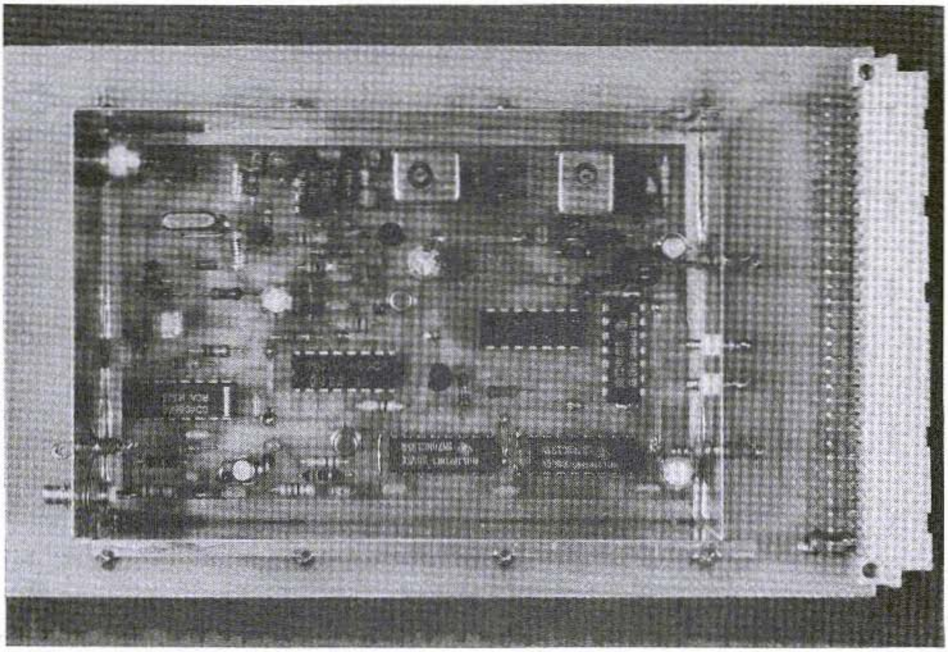

Fig. 2: The bottom cover of the tin-plate housing is soldered directly to the PCB

of the oscillator via CMOS switch I 2, the low pass filter and the varicap diode D 1.

The CMOS switch I 2 opens in the event of a radio transmitter failure, or a fault in module DJ 3 RV 007 through lack of voltage at Pt 902. The varicap (D 1) is then supplied by a fixed + 5 V from the precision regulator voltage reference I 1. The oscillator is working in an unlocked condition and therefore it has to be adjusted initially in order that its nominal frequency is exactly 8 MHz when the precision 5 V is being supplied by I 1. The 6.2 MHz reference signal from the module DJ 3 RV 007 present at Pt 903 at a power of − 6 dBm / 50Ω, is amplified by T 7 and T 8 up to TTL level. The dividers I 6 and I 7 (both 74 HC 193) have been connected to scaledown by a factor of 31, bringing the reference frequency down to the necessary 200 kHz (i. e. that of the phase comparator). Transistor T 6 amplifies the level to that of CMOS.

The 8 MHz oscillator signal, after being amplified by transistor T 2, is converted into a balanced signal by L 1 and then doubled in frequency by D 4 and D 5. L 1 is tuned to 8 MHz and L 2 to 16 MHz. The 16 MHz signal, selected by L 2, is amplified by transistors T 3 and T 4 to TTL level. The following inverter I 4, with a high-speed CMOS Schmitt-Trigger input, forms a square-wave signal which is fed to a dual D flipflop (I 5), inverted again and delivered to the computer BUS. I 5 is connected as a 4 : 1 divider which delivers the 4 MHz signal to the computer BUS via an inverter used as a driver.

The + 5 V and + 12 V supplies, needed for the operation of this module, are taken from the

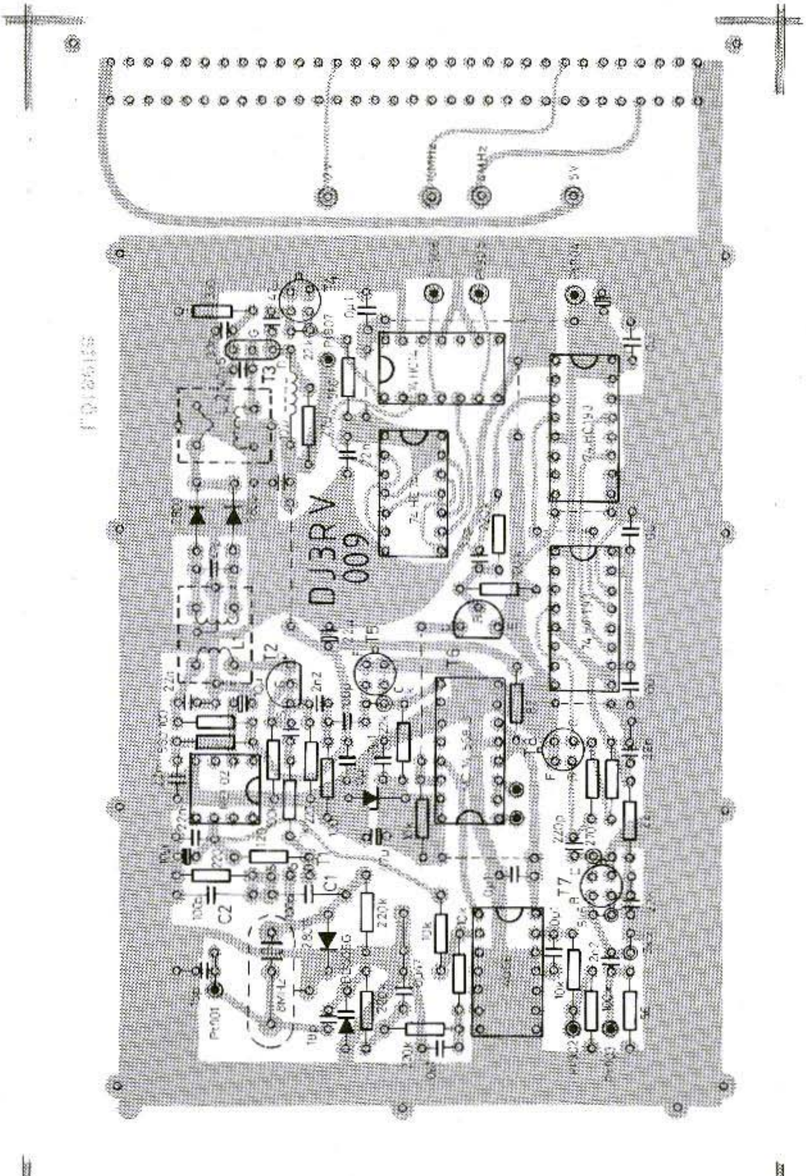

Fig. 3: Component layout of board DJ 3 RV 009; there are 9 wire bridges on the board and 11 through-ground connections to the housing

computer rails. The supplies for the CMOS chips I 2, I 3 and driver T 6 are taken from a 9.1 V zener diode which is placed across the + 12 V rail. Also a low-noise, highly-stable 5 V supply is derived from the + 12 V supply which is used for the 8 MHz VCO. The + 5 V from the computer rail is not suitable to be used as an oscillator supply as it is not stable enough and prone to interference from the computer workings.

2. CONSTRUCTION

The module is constructed using a double-sided PCB in the Euroformat of 100 × 160 mm as may be seen in **fig. 2**. It is screened with a tin-plate housing soldered on to the board – this helping to minimize radiated interference.

The construction (**fig. 3**) is unproblematic, demanding the same requirements as the associated modules DJ 3 RV 006 and 007 described in (1). The component data may be taken from the following parts list.

2.1. Parts list

T 1, T 2, T 3:	BF 246 A (TI etc.)
T 4, T 5, T 7, T 8:	2 N 5179 (RCA) or similar UHF Transistor, e. g. BFX 89; BFY 90 (Siemens etc.)
T 6:	BC 550 C (Siemens etc.)
I 1:	REF 02 (PMI, Bourns)
I 2:	4066 B (RCA etc.)
I 3:	MC 14568 B (Motorola)
I 4:	74 HC 14 (TI etc.)
I 5:	74 HC 74 (TI etc.)
I 6, I 7:	74 HC 193 (TI etc.)
D 1:	BB 505 G (Siemens etc.)
D 2, D 4, D 5:	2800 (HP etc.)
D 3:	Zener diode 9 V 2
L 1:	Filter kit Vogt D 41-2165, Colour orange 2 × 12 turns, 0.2 CuL, 7 μH, 4 turns, 0.2 CuL
L 2:	Filter kit Vogt D 41-2165, Colour orange 12 turns, 0.2 CuL, 1.7μH, 3 turns, 0.2 CuL
Q:	8.000 MHz, parallel 20 pF
Trimmer:	10 pF air trimmer, Johanson Type 5200
C 1, C 2:	100 pF, Styroflex

The components not specifically detailed can be taken as being standard.

Resistors:	Type 0207 metal film
Capacitors:	Ceramic with 2.5 or 5 mm spacing grid.

3. TUNING INSTRUCTIONS

The adjustment of the crystal oscillator is carried out with no input to Pt 902, to obtain a frequency of exactly 2,000,000 Hz at I 3 pin 1. Inductors L 1 and L 2 are then adjusted for a maximum at the drain of T 3. The module is now ready to be connected to the computer BUS. Before using this module to supply the computer's sync. requirements, the existing sync. arrangements on the CPU card and the terminal card must be rendered inoperative.

4. REFERENCES

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BRIEFLY SPEAKING...

On the 20th - 21st Sept. 1986, Borehamwood & Elstree Amateur Radio Society and B.B.C. Elstree Centre will commemorate 50 Years of High Definition Television.

On the 2nd November 1936, the B.B.C. started the first High Definition Television Service in the world. The first programmes were seen by probably no more than 2,000 people, a figure which had grown to 50,000 by the start of the war in 1939 when transmissions abruptly ceased. In 1985, the audience for the B.B.C.'s relay of the "Live Aid" concert was estimated worldwide to be 1500 million. In half a century it has become the most powerful means of communication in the world.

The Borehamwood & Elstree Amateur Radio Society will be operating S. S. B., C. W. and possibly R. T. T. Y. on 2, 10, 15, 20, 40 and 80 metres subject to propagation conditions. A special QSL Card has been designed and it is hoped that contacts will be made throughout the world.

The transmissions will begin on September 20th at 1200 hours G. M. T. and continuing until September 21st at 2000 hours G. M. T. under the call sign **GB 2 TV**.

Borehamwood & Elstree Amateur Radio Society

Editor's Note

The constructional details of a wideband IF and FM demodulator has been received in the office from Roman Polz, DC 2 CS. This module will be a suitable addition to the TV satellite frontend described by YU 3 UMV. The wideband IF / demodulator article is planned to be published in the VHF COMMUNICATIONS 3 / 86 edition.

Practical advice on Home-made PCBs

By chance, I came across a special foil made by the firm CHEMITEC with which it is possible to make PCBs easier than hitherto. How does it work?
The A 4 format foil is placed into the paper magazine of a dry-copy machine. If possible, the copier is adjusted in order to achieve an optimum blackness. Then the circuit layout is copied onto the special foil.

The foil is then laid onto the copper side of the PCB and a smoothing-iron, set to "linen" or "cotton" is used to transfer the printed circuit onto the copper surface. This has to be done with a certain "feel" since either too weak or too short, an application of the iron will result in the transfer being incomplete. If the application is too long or too hard, the tracks will be diffused. A professional way of doing it, would be, to use a heat-press or a heated roller. Following the printing, the foil is removed and the PCB placed in the etching bath. For the choice of etching material and etching techniques there are no restrictions. A mixture of 200 ml hydrochloric acid (35 %), 30 ml hydrogen peroxide (30 %) and 770 ml of water (note: the figures in brackets are the concentrations, also this mixture could be dangerous, use good ventilation).
When the etching process has finished, the PCB is washed in water and the etching resist on the tracks removed with acetone.
The special foil (TEC 200) is obtainable from CHEMITEC GmbH, Bergisch-Gladbacher Straße 977, 5000 Koeln 80, West-Germany. Five DIN A 4 sized foils cost DM 9,50.

Dr. Roland Milker, DL 2 OM

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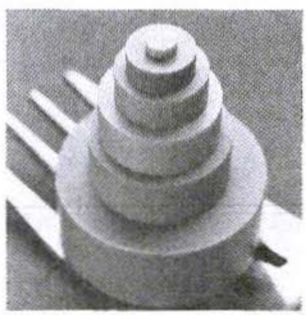

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 2 / 1986 of VHF COMMUNICATIONS

DJ 3 RV	**Microcomputer Clock-Pulse Generator linked to DCF 77 Off-Air Time-Standard**		**Art. No.**	**Ed. 2/1986**
PC-board	DJ 3 RV 009	single-sided, without component plan, drilled	6946	DM 30.—
DJ 3 RV / DB 1 NV	**Microcomputer System, Part 1: (S.M.P.S.)**			**Ed. 2/1986**
PC-board	DJ 3 RV 010	single-sided, with component plan, drilled	6947	DM 36.—
Kit	**DJ 3 RV 010**	**with all components**	**6952**	**DM 498.—**
DB 1 NV	**A Miniature 70 cm Handheld FM Transceiver**			**Ed. 2/1986**
PC-board	DB 1 NV 004	double-sided, etched on one side, without comp. plan, drilled, silvered	6957	DM 39.—
Components	DB 1 NV 004	10 ICs, 13 transistors, 17 diodes, 1 LED, 2 crystals, 1 crystal filter 21.4 MHz / 15 kHz, 1 ceramic filter, 6 ready-made coils, wire for self-supporting coils, 39 ceramic caps., 12 blocking caps., 11 foil caps., 9 electrolytics, 5 foil trimmers, 1 preset, 1 poti'meter with switch, 2 presets, 84 resistors, 1 tin-plate housing, 1 charge socket, 1 switch, 2 push-buttons, 1 BNC socket, 1 m RG - 174, 3 BCD, code switches	6958	DM 529.—
Kit	**DB 1 NV 004**	**complete as listed, except microphone and batteries**	**6959**	**DM 539.—**

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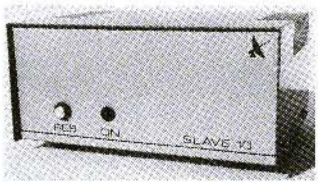

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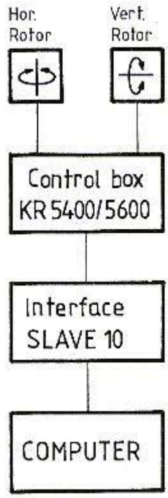
Table of commands:

command		response		function
R	CR	R	CR	rotation clockwise
L	CR	L	CR	rotation counter clockwise
U	CR	U	CR	rotation up
D	CR	D	CR	rotation down
S	CR	S	CR	all rotators stop
V	CR	V	CR	rotator stop vert.
H	CR	H	CR	rotator stop horiz.
G xxxxyyyy	CR	G	CR	preset position
F	CR	F xxxxyyyy	CR	interrogation position

xxxx: Vertical position (4 digits)
yyyy: Horizontal position (4 digits)
CR: CARRIAGE RETURN

System's block diagram



Technical data:

Data exchange:	3-wire asynchron. full duplex input and output negative or positive
Data format:	1 start bit 8 data bits 2 stop bits
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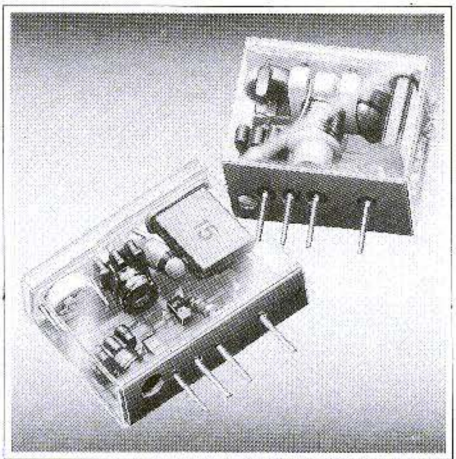

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