K. Weiner
– DJ9 HO –

THE UHF-COMPENDIUM

Part 1 and 2

DC9 NI DL4 NA DJ6 PI DB6 NT DL8 KH
DC9 NL DL7 YC DC9 RK DD9 NT DL9 KR

DJ9HO, Karl

The UHF compendium part 1 & 2

Why a scan version of this handbook?

When I was young, during university studies, I found in a local club in Turin this text that it becomes for me as a pragmatic guide for my new radio interests.

Now, after so many years, still I believe that the content of this text is still valid for what concern the methodology to attach and to solve a technical problem, meanwhile, for sure, the technology of the components here described is almost completely superseded!

For this last reason, the DARC, the German Union of Amateur Radio decided not to reprint this book for economical and reasonable aspects.

By the way, for me, it would have been a really lost for our community to “forget” this very important and milestone book.

Therefore, I decided, for the benefits of the all OM community, to spend some time to scan more than 350 pages, to grant the heritage in easy way to everybody.

I learned so much from these pages that I highly recommend to buy the paper version of this handbook, if you find still somewhere available!

Thanks Karl from all of us, as due reconnaissance.

Gian M. Canaparo, IW1AU

PREFACE TO GERMAN EDITION

The idea of compiling a UHF-compendium was born in 1975. That year I formulated a table of contents covering all the subjects which I regarded as being of interest to an active and technically minded UHF-amateur. Even though this book does - to a large extent - represent what I had in mind, and establishes a sound basis, it is by no means complete.

This book provides both building instructions to the home constructor and reference material for technical presentations or discussions in the course of ham - meetings. Hopefully, it will thus help to revive the technical aspect of our hobby.

I am particularly pleased by the support given to me by three other UHF district managers as well as communications engineers and capable self-taught people whose designs are published in this book under their own names. With the complete edition of the UHF-Compendium I hope to have contributed my share of idealism to amateur radio. If the endless hours of studying, designing, constructing and testing had to enter the financial considerations, then the price of this book would be considerably higher.

I wish to express my thanks to the following who have contributed major activities:

Engineering drawings:	DD9NT, DL4NA, DC9NL, DC9NI, DL9RM, DK5RW
Mechanics:	DL8KH, DC9NL, DC9NI, DL7YC, DJ6PI, DB8NP DK1FI, DK5RR
Translations:	DK5EG, DJ7YD, DC9RK
Typing:	Mrs. I. Reinel

Finally, I would like to thank my family and in particular my wife for their patience and understanding. To her I humbly dedicate this book. Only someone who has spent five years of spare time to accomplish and to continuously update such a project will understand why.

8670 Hof/Germany

55, 73 cuagn on UHF

de Karl H.

DJ 9 HO

PREFACE TO ENGLISH EDITION

The favorable reception of the German edition and the interest as shown by our English speaking friends indicated a demand for an English edition. Now that DK7LF has finished the translation I hope for the same positive response. May this English edition improve the links with our fellow amateurs in the English-speaking world!

D-8670 Hof, January 1982

73 and best DX
de Karl H.
DJ 9 HO

INTRODUCTION

This compendium has been divided into sections to cover both the theoretical background and the building instructions.
To further clarify the relations the text contains references to other sections that deal with the particular subject in greater detail. This avoids the necessity of frequently repeating remarks or explanations and results in better understanding and the saving of several pages.

The book contains sufficient theoretical material to provide the relevant background information and to explain the operation of the equipment described in the building section. Special emphasis has been placed upon the construction of GaAs FET preamplifiers and converters as well as power amplifiers and antenna systems for the 70cm and 23 cm bands. Great value has been attached to diagrams of measuring equipment and tuning aids. Consequently, there are suggestions for the construction of simple alignment tools, power measuring equipment, a fixed frequency generator for receiver alignment, a UHF dipper, a panoramic receiver and a swept frequency generator.

Practically all building projects have been verified and optimized using present standard test equipment such as signal generators and spectrum analyzers. This is followed by charts, tables, diagrams and general remarks.

THE UHF - COMPENDIUM

Table of contents

A.1 Properties of components and linear circuits

A.1.1 General remarks

At UHF construction and application of active and passive components is governed by other considerations than, for instance, at VHF and below. Up to about 300 MHz resonant circuits can be constructed of inductors and capacitors. The resonant frequency (f_{res}) is clearly defined by the capacitance of the capacitor and the inductance of the inductor.

This situation changes completely at UHF (300 to 3000 MHz) where inductance (L) and capacitance (C) will have to have extremely small values. In particular the length of valve pins and connecting leads of transistors constitute a major part of the total inductance,and the capacitance of these components greatly influences the tuned circuit. At 432MHz a tuned circuit containing - for instance - a capacitance of 5pF requires an inductance of only 27nH. Input or output capacitance of a valve is usually quite sufficient whereas an inductance of 27 nH is already represented by a piece of wire (1 mm diam.) of about 4.5 cm length. At 23 cm and 12 cm this is obviously even more of a problem.

At UHF tuned circuits composed of lumped components (as commonly used in short-wave equipment) are therefore no longer practical. They are replaced by circuits in the form of short conductors whose length is of the order of one half or quarter wavelength. But the geometric dimension is further shortened to allow for the component capacitances (see A.2).

Altogether at UHF the whole wiring- and construction-technique must be tailored to the particular requirements of this high frequency. The construction projects of sections C and D confirm that this can be achieved even with the limited means of an amateur.

Valves are exclusively used in grounded-grid configuration. The preferred arrangement for transistors is the common base circuit or - more recently for the specially developed UHF-transistors - the common emitter circuit. Semiconductor developments of well-known firms point towards future application of transistors far into the GHz region. Due to financial considerations valves will remain with us in UHF-power amplifiers for quite some time.

A.1.2 Inductance at UHF

As mentioned before, at UHF inductors may be realized by short pieces of conductor. The inductance (L) of a piece of conductor depends on its length (l), its diameter (d) and also on its distance (a) from a conducting surface. An approximate value of L per cm length of conductor may be read off the diagram (fig.1). To give an example, the case of a piece of wire, 2.5mm diameter, 5mm separation from a conducting surface and a length of 1cm was drawn in. This results in an inductance of 4.1nH.

Obviously this conductor will possess some small value of capacitance against the conducting surface. From short wave radio we are quite familiar with this property. Here a certain capacitance is formed between a conductor (dipole antenna) and a conducting surface (ground). Here, too, the conductor's diameter and its distance from ground are the governing factors. Considering a tuned circuit at UHF (e.g.70cm) one could start with the variable tuning capacitor set close to its maximum value. This would allow to reduce its value such as to compensate for the small value of capacitance formed by the conductor. However, the inductive component is dominant. This means, that the reduction of the spacing between the conductor and the conducting surface must be considered as the addition of some further inductance in parallel. This will shift the resonant frequency towards some higher value. Thus, the capacitance of the tuning capacitor must be increased to come back to the same resonant frequency when a conducting surface (chassis or cabinet wall) is brought into the vicinity of the tuned circuit.

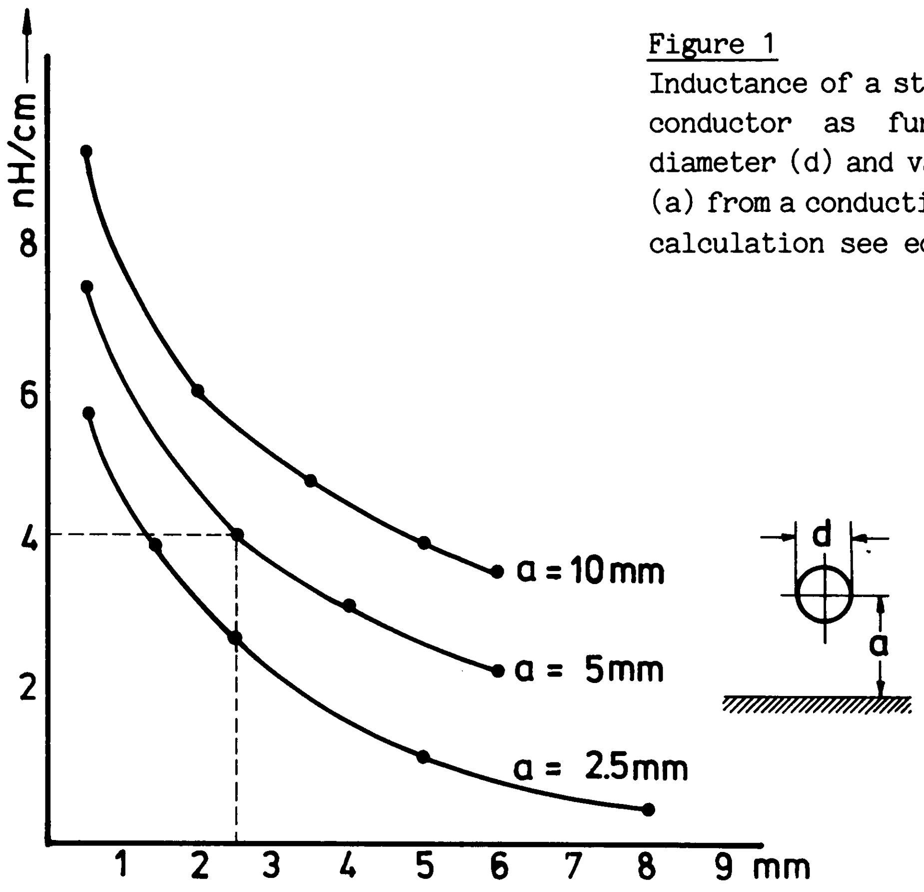
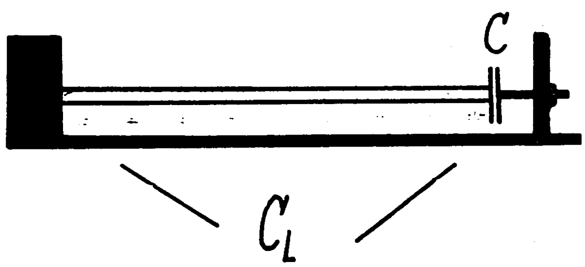


Figure 1
Inductance of a straight piece of conductor as function of its diameter (d) and various spacings (a) from a conducting surface. For calculation see equation 1.

Diameter d of conductor ---->

Figure 2
Presentation of line capacitance of UHF-circuit



Equation 1

$L = 2 \ln (4a/d)$ [nH/cm]

where

L = Inductance [nH]

a = spacing between center of conductor and conducting surface

d = diameter of conductor [cm]

The following two pages allow to conveniently obtain the inductance of a piece of conductor with known spacing and length.

Figure 3

Presentation of inductance L of a conductor with diameter (d) =1.5 mm and three different spacings to a conducting surface.

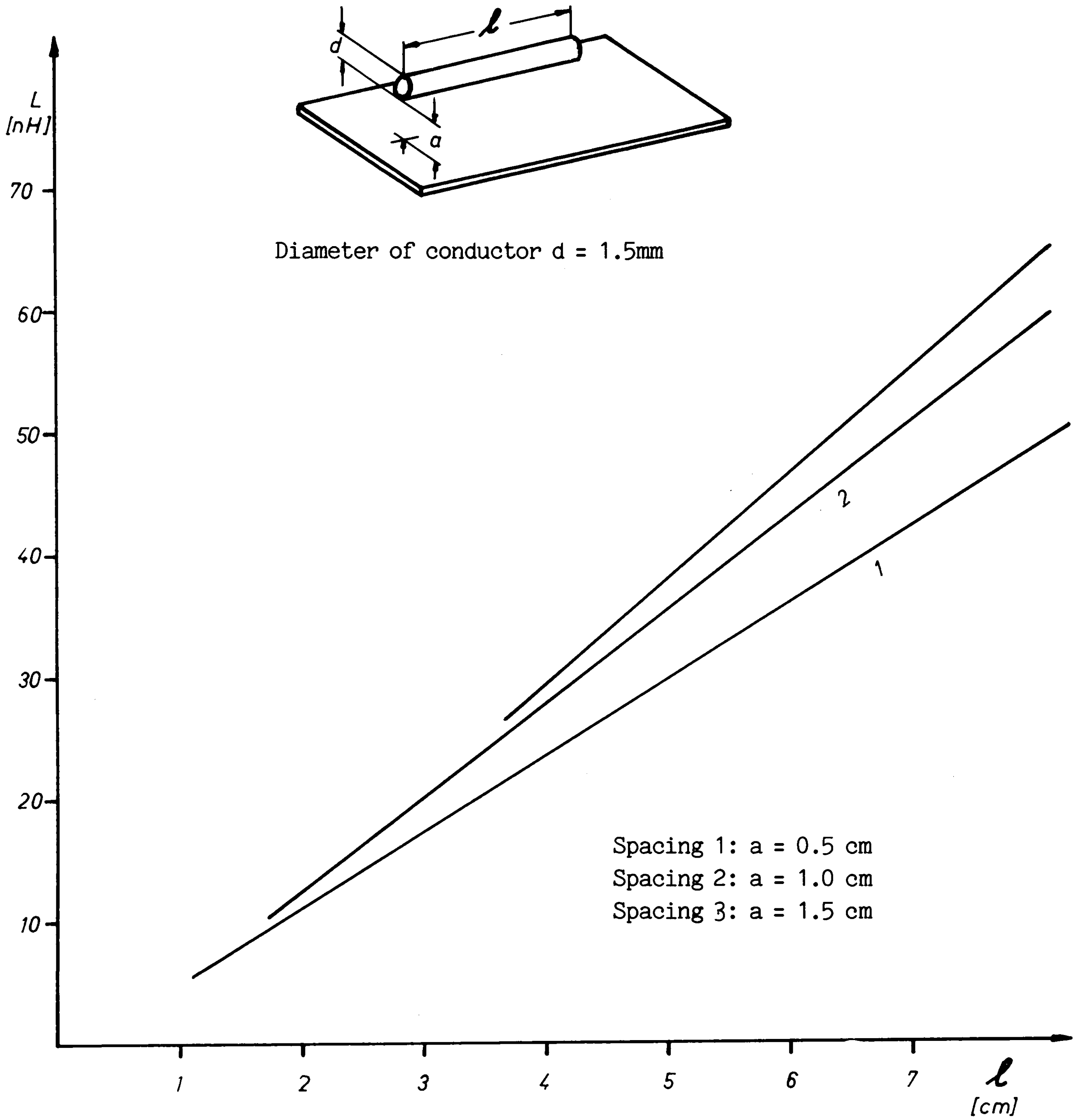
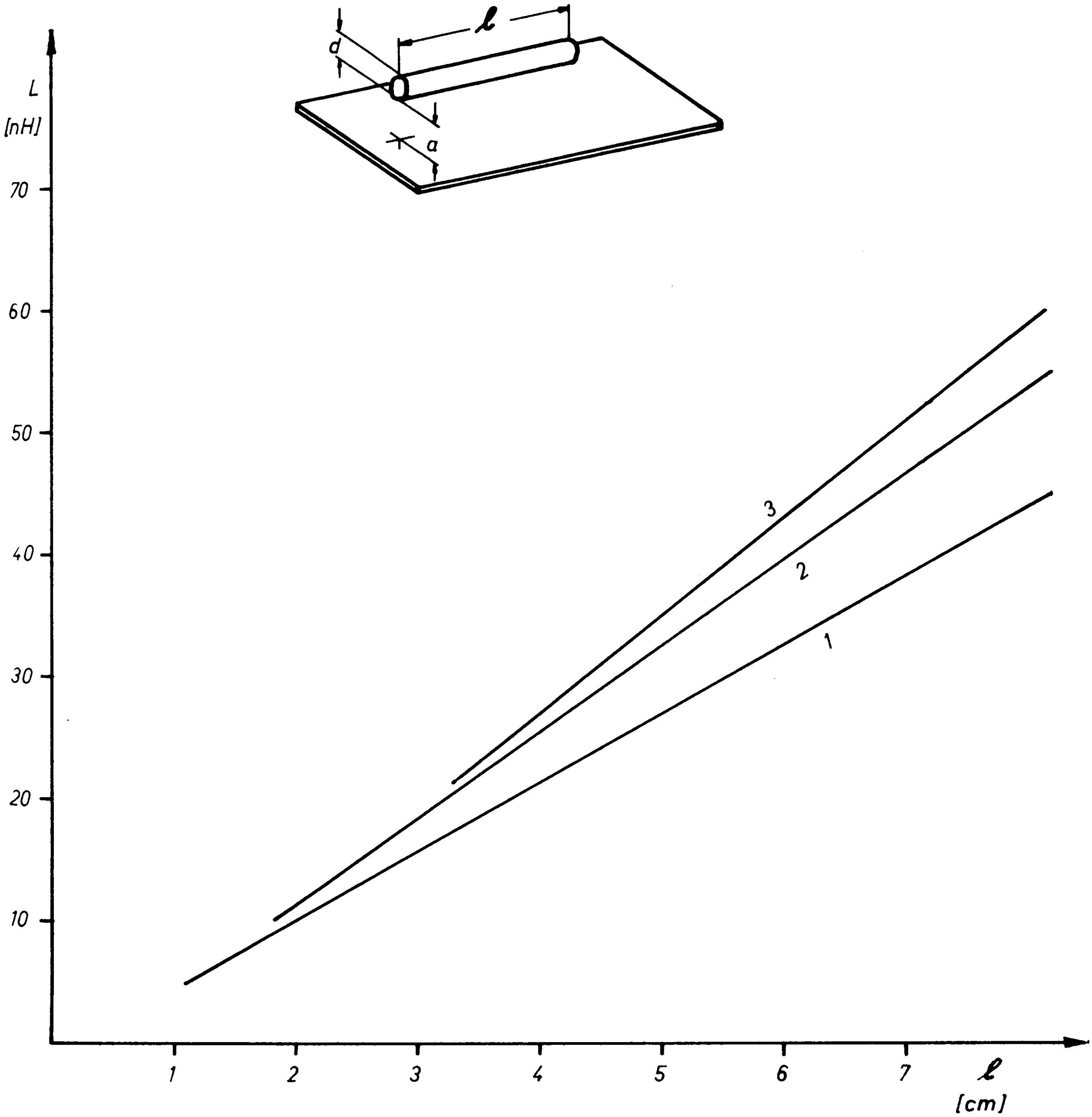


Figure 3

Presentation of inductance L of a conductor with diameter (d) = 2.0 mm and three different spacings to a conducting surface.

Diameter of conductor d = 2.0 mm

Spacing 1: a = 0.5 cm
Spacing 2: a = 1.0 cm
Spacing 3: a = 1.5 cm



A.1.3 The skin effect

We all know that a direct current utilises the total cross-section of a conductor for current conduction. At high frequencies however only a very thin layer at the surface is involved in the conduction process. Consider the conductor to be made up of a large number of thin filaments in parallel, each carrying a small fraction of the total current. Each filament is surrounded by an alternating electromagnetic field which induces secondary currents such as to oppose the primary currents. This opposing action (and the resultant weakening of the net current) is strongest towards the center of the conductor and increases further with increasing frequency. This results in HF-currents being forced out towards the surface of the conductor as frequencies go up. One could visualize this effect by considering the case of the sun and the earth: The surface of the earth is continuously exposed to the sun's radiation and the warmth enters the ground to some degree. Slow variations (summer or winter) effect a much deeper penetration of temperature changes than fast variations do (day and night). It is obvious that the material or rather its specific resistance (ρ) has an influence on the penetration depth (δ) of the HF-current.

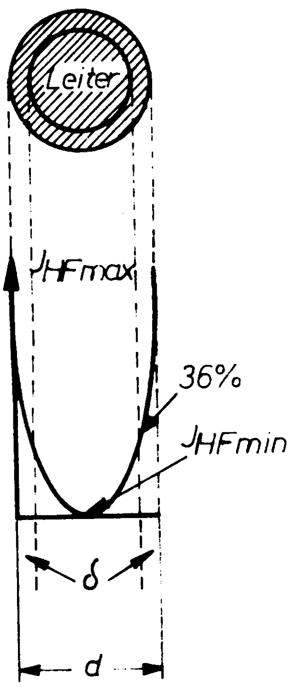


Figure 5 - HF-current distribution inside the conductor cross-section.
- Depth of penetration (equivalent thickness of conducting layer)

Equation 2

δ = 0.5 (ρ)/(u_r x f) [mm]

where

δ = penetration depth (calculated value in mm)

f = operating frequency [MHz]

ρ = specific resistance (Ohm·x mm^2) / m

μ_r = relative permeability ~ 1 for copper,gold, brass and silver

Table 1 Specific resistance (ρ) of different materials.

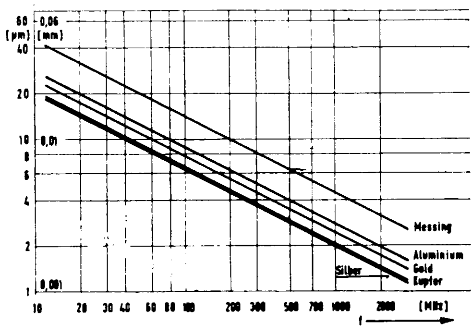
Material	(ρ) [Ohm x mm^2/m]
Silver	0.0164
Copper	0.0178
Gold	0.0223
Brass	0.077

As an example the value for copper was entered into equation 2. This gives the following results for the UHF-bands:

144 MHz	$\delta = 0.5 \sqrt{(0.0178)/(144)} = 0.0054$ mm	$= 5.4$ [µm]
432 MHz		$= 3.1$ [µm]
1296 MHz		$= 1.8$ [µm]
2304 MHz		$= 1.35$ [µm]

The obvious choice for the UHF-amateur is copper. Silver or even silver plating is generally too expensive or too difficult to achieve. The majority of building projects of this book is thus based on cheap copper clad material.

Figure 6
Equivalent thickness of conducting layer (δ) (penetration depth) as function of operating frequency (f).



Any inductance opposes the flow of HF-current to some extend. At UHF this opposing effect is composed of the reactive part (X_L) and increasingly of the ohmic part caused by the skin effect. This ohmic component is a known quantity when using copper or silver plated material. These values are still negligible at UHF so there is no need to bother the reader with equations and calculations. (For X_L see equation 3).

A.1.4 The capacitor

The use of capacitors at VHF/UHF does require some knowledge of the various types and can give rise to problems. Now, what does this mean? Let us consider the basic construction as shown in figure 7.

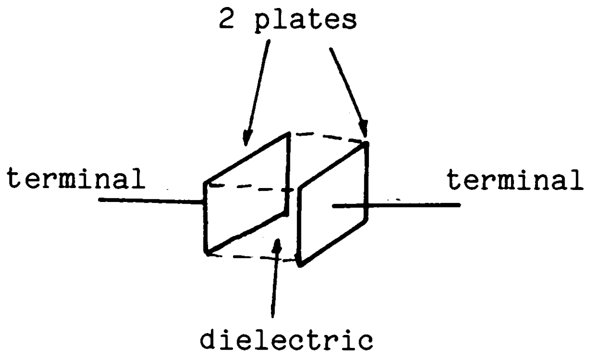


Figure 7
Basic construction of a capacitor.

As indicated in section A.1.2 every piece of wire (conductor) possesses some inductance (L). So the connecting leads of a capacitor represent a certain amount of inductance, too. The dielectric does not have infinite resistance either and can be imagined as a resistor (R_D) connected in parallel with the capacitor. Additional ohmic losses (R_S) are caused by the surface properties of the connecting leads and the capacitor plates (skin effect) and may be assumed to be series connected with the capacitor. Allowing for all these influences, the following diagram represents the equivalent circuit of a capacitor at high frequencies.

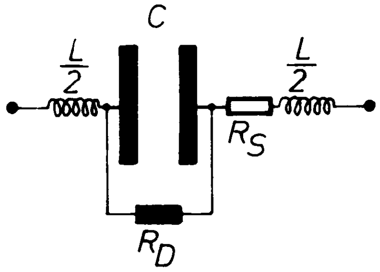


Figure 8
Equivalent circuit of a capacitor at high frequencies

L = Inductance of connecting leads
R_S = Ohmic losses of connecting leads plus skin effect losses
R_D = Dielectric losses

The reader should continue to page 12 if he does not intend to study this subject more deeply.

The equivalent circuit of a capacitor at high frequencies (figure 8) could also be represented by figure 9. Choosing a capacitor with air dielectric allows to neglect R_D. We will come back to this later.

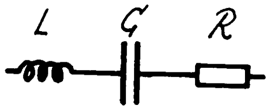


Figure 9 Series conection L,C and R

Due to the series connection all three components will carry the same HF-current I_{HF}. Resistance R_S is to be assumed as being purely ohmic whereas the imaginary components inductance (X_L) and capacitance (X_C) are frequency dependent. With increasing frequency the value of X_L will go up whilst the value of X_C will go down. The corresponding values may be calculated by equations 3 and 4.

Equation 3 (simplified)

X_L = o.oo628 x f x L [ohm]

f in [MHz]
L in [nH]

Equation 4 (simplified)

X_C = (1.59 x 1o^5)/(f x C) in [ohm]

X_C = 159ooo/(f x C) [ohm]

f in [MHz]
C in [pF]

A further representation of the series connection of L,C and R with the corresponding resistances and voltages is shown in figure 1o.

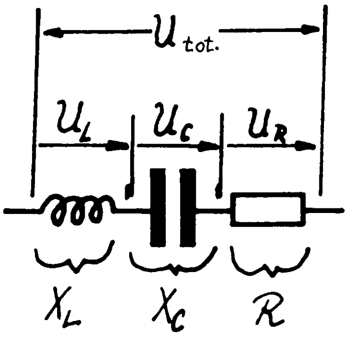


Figure 1o
Resistances and voltages of a capacitor, containing inductive connecting leads, in the UHF-alternating current circuit. (Series connection of L,C and R)

In such a circuit composed of inductive reactance (X_L), capacitive reactance (X_C) and ohmic resistance (R) the phase relationships between these components have to be considered. As may be remembered from the times before the amateurs' licensing examination, the current lags the voltage by 9o degrees in an inductor as part of an alternating-current circuit (figure 11).

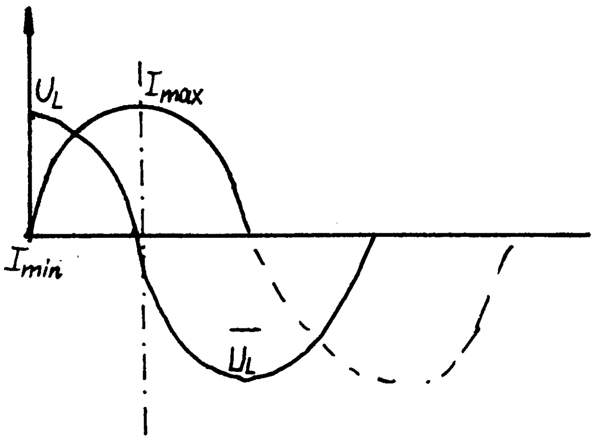


Figure 11
Current and voltage relations in an inductor as part of an ac-circuit.

- max.voltage U_L
- min.current I

In a capacitance subjected to ac the current leads the voltage by 9o degrees.

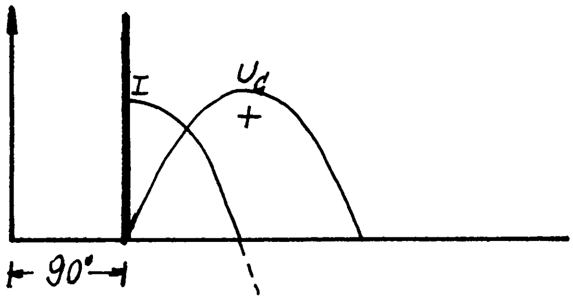


Figure 12
Current and voltage relations in a capacitor as part of an ac-circuit.

It can be seen that the voltages oppose one another and would cancel completely if they were of equal amplitude. We know this effect of cancellation by opposing phases from wave propagation, too (fading).

If both voltages have the same value (apart from the phase difference) then the two reactances X_L and X_C must have the same value, too ($|X_L| = |X_C|$), as all components carry the same current in a series circuit. This particular case $|X_L| = |X_C|$ is called "resonance". The current I_{HF} will reach its maximum value and is only limited by R. To summarize, in a series circuit (fig.10) a particular impedance Z - composed of the two reactances and the resistance - corresponds to every frequency and may be calculated by equation 5.

$$Z = \sqrt{R^2 + (X_L - X_C)^2}$$

Equation 5
Impedance Z of a series circuit [ohm]

(X_L and X_C may be calculated from equations 3 and 4).

It is important to remember that the capacitor will completely lose its character and become inductive above this frequency. Figure 13 displays the individual reactances (X_L and X_C) and impedance Z as function of frequency for a capacitor of 10 pF and connecting leads of 5mm length and 0.5 mm diameter (= 5 nH).

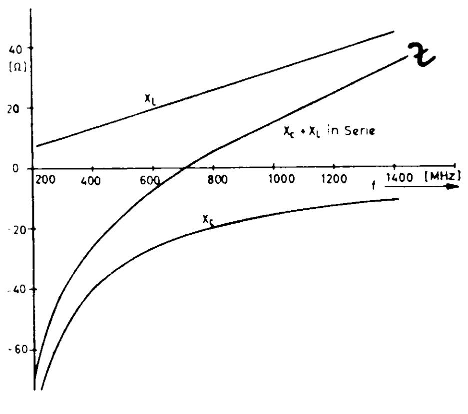


Figure 13
Impedance of a capacitor containing inductances as function of frequency. (R_S neglected)

The resultant impedance $X_L - X_C$ indicates a resonant frequency f_{res} of approximately 700MHz ($|X_L| = |X_C|$).

There are further problems of which the reader should be aware. To avoid unpleasant surprises the following types of capacitors suitable for UHF should be used. The types of capacitors shown in figure 14 are recommended as blocking- or coupling capacitors.

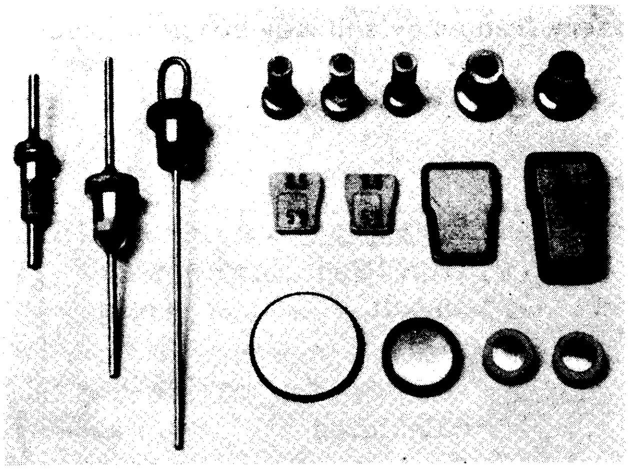


Figure 14
Preferred types of capacitors with minimal inductance for use at UHF in trapezoid- and disc- form

For adjustments of tuned circuits air dielectric variables of the Tronser type, foil dielectric variables of the Valvo type and variables with ceramic dielectric are suitable.

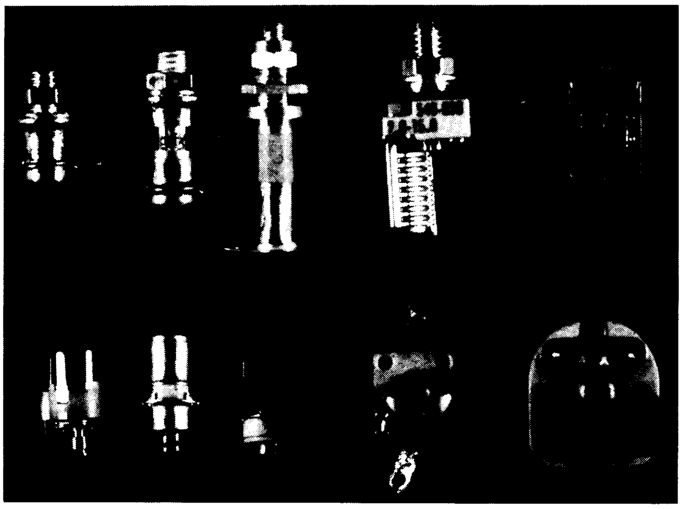


Figure 15
Suitable capacitors for resonant circuits.

A.1.5 Ohmic resistors

At higher frequencies resistors give rise to the same sort of problems that were met with capacitors. The following example indicates inductance and capacitance of a resistor.

Construction: Equivalent circuit:

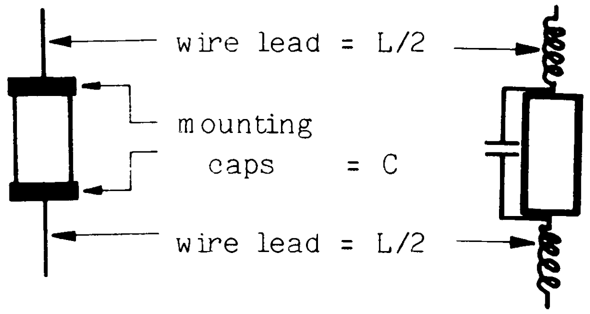


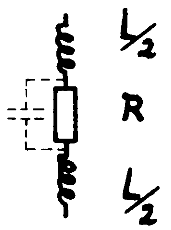
Figure 16
Composition of a resistor and development of equivalent circuit

As indicated by the equivalent circuit the above shown resistor possesses some inductance L (X_L) and capacitance C (X_C). Its impedance is thus frequency dependent. For small values of resistance the inductive component X_L will dominate as the low resistance will shunt the capacitive properties of the mounting capacitance.

With increasing values of resistance (~beyond 1000 ohm) the influence of the resistors' capacitance will increase. With increasing frequency the (real) component of the impedance will decrease as the resistance is shunted by the decreasing value of X_C. If such a resistor is - for instance - used to damp a resonant circuit then an increasing frequency will result in lowering the Q-factor of this circuit and additional damping. A low value resistor will therefore be predominantly inductive whereas a high value resistor will be predominantly capacitive. We shall now consider the two cases seperately even though the transition is not really that distinct but rather more overlapping.

The low resistance resistor

A current flowing through a resistor of low resistance gives rise to a voltage drop U_R across R and U_L across the inductive reactance X_L. From section A.1.4 we know, that in an inductor the voltage will lead the current by 90 degrees. We may therefore represent the voltage drops in a phasor diagram (figure 17). The current is identical in all components of this series connected circuit and is therefore taken as reference phasor in phasor diagrams of ac-circuits.



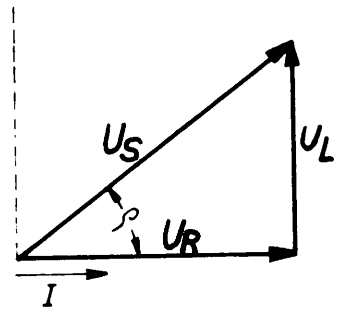


Figure 17

Voltage phasor diagram of the low resistance resistor at UHF.

U_L = voltage drop across the inductance

U_R = voltage drop across the resistance

U_S = complex voltage drop $= \sqrt{U_R^2 + U_L^2}$

Applying Ohm's law and dividing all voltages by current I leads to an impedance triangle which is similar to the voltage diagram (figure 18).

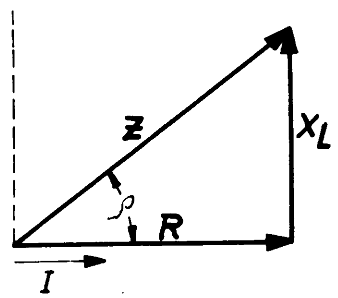


Figure 18

Impedance triangle of the described resistor at UHF.

The ratio of the voltages (figure 17) is identical to that of the resistances according to figure 18. Thus the phase shift angle φ remains the same. The linear resistance $R = U_R : I$
and the inductive reactance $X_L = U_L : I$ constitute the resultant complex resistance Z, called **"impedance"**.

Equation 6 $$Z^2 = R^2 + X_L^2 \quad [\text{ohm}]$$

$$Z = \sqrt{R^2 + X_L^2} \quad [\text{ohm}]$$

An example will help to clarify the problem:

Calculate the impedance Z of a 20 ohm resistor at a frequency of 10 MHz. The conneting leads are 10 mm long (which corresponds to an inductance of about 10 nH).

Complete calculation:

$$Z = \sqrt{R^2 + X_L^2}$$

$$Z = \sqrt{20^2 + X_L^2}$$

$$X_L = 2 \times pi \times f \times L$$
$$= 6.28 \times 432 \times 10^6 \times 10 \times 10^{-9}$$
$$= 6.28 \times 4.32 \times 10^9 \times 10^{-9}$$
$$= 6.28 \times 4.32$$
$$X_L = \underline{27.13\ [\text{ohm}]}$$

$$Z = \sqrt{20^2 + 27.13^2}$$

$$= \sqrt{400 + 736.01}$$

$$= \sqrt{1136.01}$$

$$\underline{\underline{Z = 33.70\ [\text{ohm}]}}$$

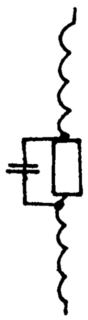
The total effective resistance at 432 MHz is hence not 20 ohm but about 33.7ohm. Ohm's law for alternating current is therefore not I = U/R, as is correct for direct current but: I = U/Z.

The same sort of calculation could be carried out for inductors. But in that case R will be a small quantity which at UHF is determined by the surface properties of the windings (skin effect). X_L will have a very high value and the low value of R will define the quality factor Q of the inductor.

Equation 7: $$Q = X_L/R = \omega L/R = (2\pi \times f \times L)/R$$

where f in [Hertz], L in [Henry], R in [ohm]

The high resistance resistor.



In this case the capacitive component will dominate. This will lead to a reduction of the effective resistance as the frequency goes up. The percentage reduction will be larger for a 100 K ohm resistor than for a resistor of 1 K ohm.

Conclusion

Under no circumstances will only the capacitive or the inductive component be of influence. Depending on the construction of the resistor and the length of its connecting leads there will be a certain amount of compensation. Wire wound resistors should not be used at UHF even when the winding is less than a quarter wave length long. Particularly well suited are **"carbon layer resistors"**. In their most modern form they can have connecting leads with an inductance of considerably less than 10 nH.

When attempting any in the building projects of this book all problems are avoided by adhering strictly to the hints and recommendations by the individual authors.

Recommended reading:
Empfangstechnik im UHF-Bereich
by
Ing. Moehring (Loewe Opta)

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A.1.6 Noise

At UHF the received signals will be accompanied by an interfering signal in the form of noise. This noise is made up of various components originating from space (entering the system through the antenna), from the antenna or generated inside the receiving system. The major component is, however, generated by the input stage of the system where it mixes with the desired signal, gets amplified and eventually appears as an audio signal at the loudspeaker. Thus the UHF - amateur's dream is a low noise input stage. It will essentially define the system sensitivity. The effort of semiconductor firms has made this dream come true even considering the limited financial means of the amateur.
But apart from noise generated by active components additional noise is contributed by passive components as well. This leads to the question about the nature of noise.

Noise - in general.

Due to thermal movements of the atoms inside a conducting material the free electrons are stimulated to move in a random manner. The intensity (energy) of this movement increases with increasing temperature. As this movement is not at a specific frequency but rather in a random fashion, a noise voltage covering a wide frequency spectrum will appear across the terminals of this conductor or component. At -273 degrees centigrade (corresponding to 0 degrees Kelvin (0^o K) or absolute zero-temperature) this movement of free electrons, which to a certain degree opposes the current flow, will cease. The conductor will reach its maximum conductivity. The resistance of lead - for instance - will drop to R = 0 ohm. This property is referred to as "supra conductivity".

At a temperature of more than 0^o Kelvin ($T > 0^o$ K) any ohmic resistor must be seen as a noise generator. The noise voltage (U_r) is relatively large across the terminals of a high value resistor and will drop to small values in the case of low value resistors. The current behaves inversely to this, so the noise power in an ohmic resistor is constant for a given temperature. This noise (professionally referred to as "white noise") appears as "grass" on the scope of a panoramic receiver (figure 19). It is easy to see that the noise energy (power density) in sections a and b, corresponding to equivalent band sections Δ f (bandwidth), is congruent and therefore identical. We may therefore conclude that over a very wide range every frequency is present with equal amplitude.

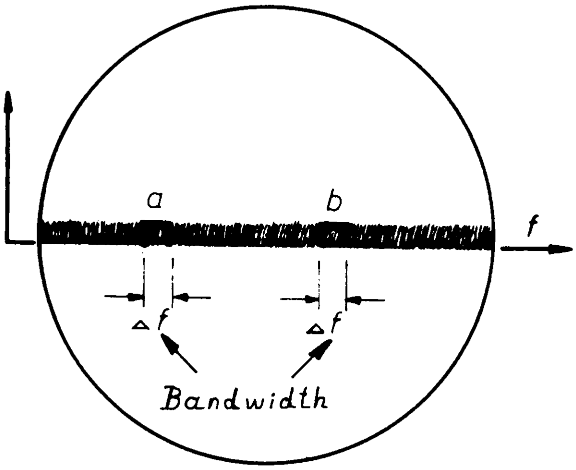


Figure 19
Presentation of noise on the display of a panoramic receiver.

As mentioned before noise power is a function of temperature (T_o) and bandwidth ^f.
Assuming a temperature of 20° C (room temperature) and a normalized bandwidth of 1 Hz the noise power (n_r) in a resistor may be calculated:

Equation 8

$$n_r = 4 \times k \times T_o \quad [W/Hz]$$

$k = 1.37 \times 10^{-23}$
(Boltzmann-constant)

T_o = absolute temperature of the resistor in degrees Kelvin. (20° C correspond to 293° K)

$$n_r = 4 \times 1.37 \times 10^{-23} \times 293$$

$$n_r = 1605.65 \times 10^{-23}$$

$$n_r = 16 \times 10^{-21} \quad [W/Hz]$$

This value of 4 kT_o may be seen as the specific noise power of a resistor. Under the condition of power matching (load resistance equals the internal resistance of the generator) a generator (in our case represented by the resistor) will pass 1/4 of the total noise power to the load.

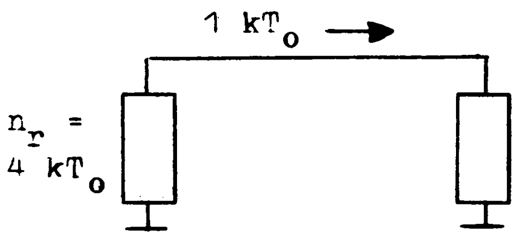


Figure 20
Noise power transfer from 50 ohm source (aerial) into 50 ohm load (receiver input)

Solar and cosmic noise (from space) which is picked up by the antenna and appears at the receiver input as noise voltage is still rather noticeable in the 2m band and must be added to the noise power generated by the internal resistance of the antenna. An increase of 10 to 20dB as indicated by a calibrated S-meter of a 2m-RX is no rare occasion. With increasing radio frequency this type of noise falls off rapidly. The sensitivity is then essentially governed by the input stage.

To calculate the antenna noise voltage (n_{rA}) one assumes the sum of these noise voltages to be generated by the internal resistance (R_A) of the antenna and enters an increased temperature into the equation (at UHF R_A is approximately the radiation resistance of the antenna). The attentive UHF-amateur will have noticed the variations in noise level depending on time of the day, frequency and antenna direction. This may be partly due to the above mentioned reasons and partly due to other active noise sources. So the reception may be impaired by interference or noise clusters around the house, its surroundings or by thermal noise from the earth's surface.
For obvious reasons this problem can not be solved by radiating vertically upwards where this type of noise is lowest. The cure lies in the antenna height above the house and in a narrow vertical and horizontal beam width of the antenna.

An additional source of noise is the tuned circuit. Here the loss resistances of inductors and capacitors give rise to noise voltages of a wide frequency spectrum. However those frequencies which fall into the pass band of the tuned circuit will be enhanced with respect to the remainder of the spectrum by the resonance effect. It is therefore logical to enter the resonant impedance rather than the loss resistance. A short comparison with the HF bands will illustrate the problems that occur at UHF. The resonant impedance of the tuned circuit at HF is high (around 100 Kohm) with respect to the equivalent noise resistances (r_e) of a valve or a transistor. The resonant impedances (of the antenna etc.) will supply a much higher noise voltage than valve or transistor do as the noise contribution is proportional to R. So with respect to the total noise the contribution of valves or transistors is small.

The situation changes completely at UHF. Due to physical reasons (A.1.9) valves and transistors will deliver a much higher noise component. Furthermore at UHF the resonant impedances are only of the order of 1 to 5K ohm. In spite of this, turning the tuning capacitor of the input tuned circuit of a functioning converter will reveal a noise maximum which is partly due to the previously described phenomenon. The maximum sensitivity does, however, rarely coincide with this maximum noise condition. So one should tune for the maximum signal level. Fine adjustment will then lead to optimum noise properties (minimum noise) with maximum sensitivity.

A.1.7 The Noise Figure (F)

A large number of publications treat the noise figure in great detail. The following summary will offer the amateur all he needs to know in simple terms.

The previous section dealt with the noise power of a resistor (an antenna etc.) and it was stated that a certain amount of power (n_r) will be delivered into the load (RX). If additional signal power (n_s) is fed to the receiver a certain signal to noise ratio (A_1) will be established.

Equation 9

$$A_1 = \frac{n_s}{n_r}$$

n_s = signal power

n_r = noise power

A receiver without internal noise generation would merely amplify this combined input signal which would then appear at the output terminals with the same signal to noise ratio A_1. Any UHF-receiver will - predominantly by valves or transistors of the input stage - add some further noise power (n_{rz}) of a level $n_{rz} > n_r$. The combined input power is going to be amplified by the receiver but the signal to noise ratio will deteriorate due to the internally generated noise. The resultant output signal to noise ratio is described by equation 10:

Equation 10

$$A_2 = \frac{V_L \times n_s}{V_L \times (n_r + n_{rz})}$$

V_L = total power amplification factor

The ratio of signal to noise ratio measured at the input (A_1) divided by that measured at the output (A_2) is called noise figure (F).

Equation 11

$$A_2 = \frac{A_1}{A_2} = \frac{n_s \times V_L \times (n_r + n_{rz})}{n_r \times V_L \times n_s}$$

A diagram will illustrate the problem. Let us assume three panoramic receivers which allow to display the various signal levels:

Figure 21 a External noise power (n_r) and received signal power (n_s)

Figure 21 b Situation inside the input stage due to the addition of the noise (n_{rz}) generated internally by the input stage to the external noise n_s

Figure 21 c Situation at input to demodulator after noise free amplification (i.e. times V_L)

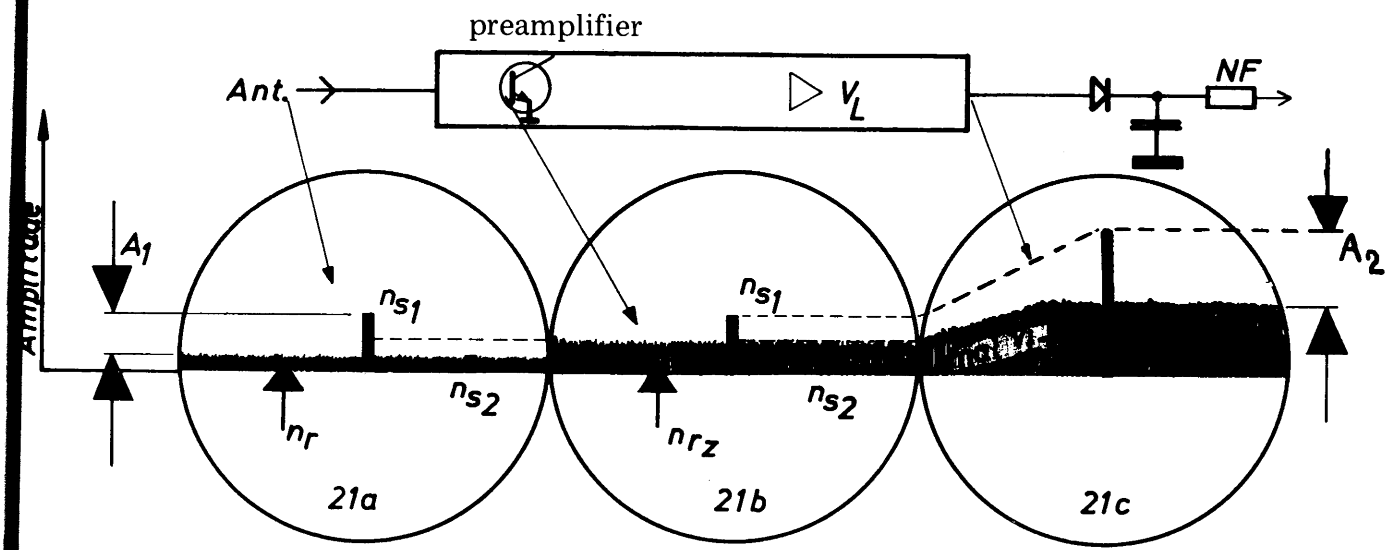


Figure 21 Presentation of signal to noise ratios at various points of the receiving system

As clearly demonstrated by figure 21 only the lowest possible value of n_{rz} will lead to near-ideal receiver sensitivity.
A_2 is essentially governed by n_{rz}. In an ideal receiver there is no additional internal noise power and $n_{rz} = 0$. There will be no deterioration of the signal to noise ratio ($A_1 = A_2$). The noise figure F of such a receiver would thus be (substituting $n_{rz} = 0$):

$$F = \frac{A_1}{A_2} = 1$$

The ideal receiver has yet to be invented and figure 21 b indicates that the wanted signal (n_s) must be equal to or larger than the internal noise power n_{rz} (assuming that n_s can be neglected) to appear at the receiver output. **The lowest threshold of detectabilty** is therefore set by the condition of $n_s \geqq n_{rz}$ in figure 21.
In Germany usually the kT_0-figure is specified to describe the power ratio whereas english-speaking countries prefer the logarithmic dB value. In transistor data sheet of foreign manufacturers the noise properties are described under the heading of **"noise figure".** Frequently the kT_0-value is designated as noise figure (resp. noise factor). So equation 12 and figure 22 permit to convert from one system into the other.

Equation 12 $$F\,[dB] = 10\,\lg F\,[kT_0]$$

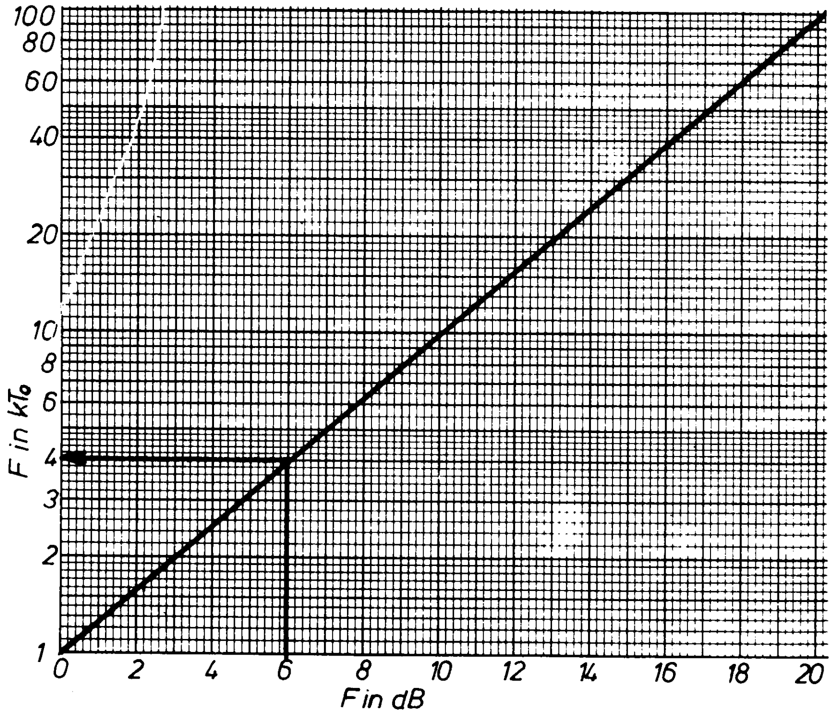


Figure 22
Relation between F in $[kT_0]$ and F in [dB]

Example:

6 dB = 4 $[kT_0]$

Frequently there are passionate discussions about the question whether the "ufb" converter should be equipped with one or two stages of preamplification. The following remarks concerning this subject must be made:

1. **The best preamplifier is the antenna.** It is the only amplifier that will give an improved signal to noise ratio (A_1) as the gain (G) goes up with only a slight increase in n_r. Any normal preamplifier will give poorer results.

2. The required number of preamplifier stages is dictated by the noise contribution (n_{rm}) of the mixing stage. We have to differentiate between active mixers (valve,transistor) and passive diode mixers. An active mixer will frequently produce more noise power than a preamplifier equipped with -for instance- the same transistor. This is caused by the mixing process. The conditions of figure 21 (but with 21a representing the preamplifier and 21b being the mixer) could again be used for demonstration. In contrast to this the diode mixer: Instead of amplification there will be mixing loss (-6 to -10 dB). So the noise level (transfer gain) will be lower than in the previous example. The following diagram (figure 23) will demonstrate the conditions that justify the use of a second preamplifier stage. A representative value of preamplification is 15 dB plus mixing losses (hence abt. 20-25dB). [See section D.1].

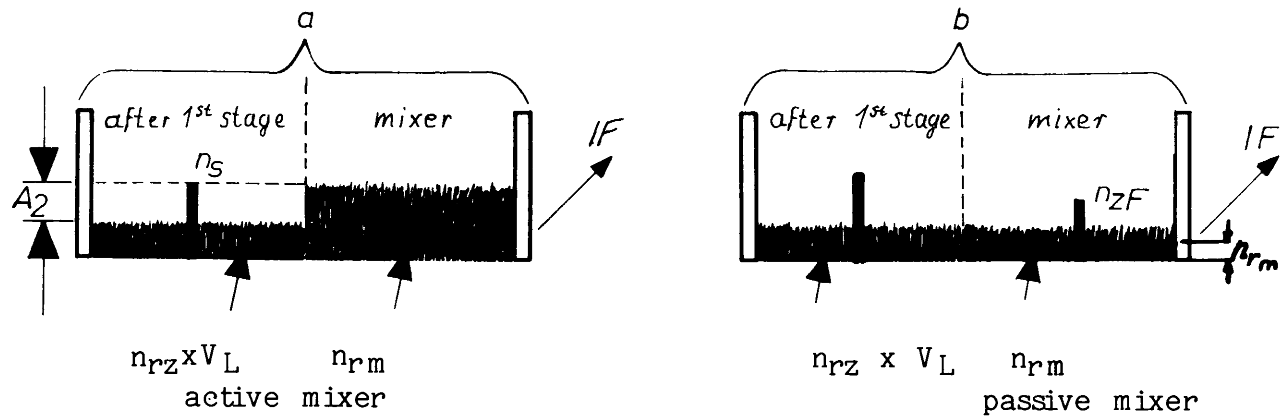


Figure 23 Pictorial U_r comparison of active and passive mixers
a) Insufficient preamplification; IF-signal = 0 dB; second preamplifier absolutely necessary.
b) Situation of well-designed diode-mixer; second preamplifier not required.

Good results are to be expected from one-high gain low-noise preamplifier stage working into a diode mixer and followed by a low-noise high-gain first IF-stage. As far as overloading is concerned this concept will satisfy more exacting demands than a concept employing two stages of preamplification followed by an active mixer.

Suitable diodes (like HP 2800 or equiv.) with low forward starting voltages are available at reasonable prices. Various technical publications mention the successful development of diodes with forward starting voltages of nearly 0 Volts. This guarantees optimum sensitivity, minimum mixing losses and a low power level for the injection (oscillator-) frequency: $n_o \leqslant 5$ mW. A following high-gain low-noise IF-stage allows the matching of the weak IF-signal to the required input level of the following main receiver.
To summarize it may be stated that the noise properties (sensitivity) are not only influenced by the input stage but that the first IF-stage contributes significantly to the overall noise level when passive mixers are being employed. This is particlarly serious as the input signal (n_s) which required great effort to amplify by 10 to 15 dB will be attenuated by about -6 dB due to the mixing losses only to be masked by the noise level of the first IF-stage.

It seems necessary to point out that for the above mentioned reasons it is impossible to mount an antenna below the roof and to simultaneously expect good sensitivity and noise properties. The essentials of a station even at UHF are the antenna, narrow vertical beam width, the shortest possible low loss coax cable (G.1) and a high gain low noise preamplifier. A lengthy antenna transmission line will not only attenuate n_s but will also contribute significantly to the overall noise in a sensitive system due to ohmic losses (skin effect). Of great importance for the reduction of noise is the band width Δf. Equation 13 allows us to calculate the theoretical noise voltage of a resistor when the band width has been specified.

Equation 13

$$U_r = 4 \times kT_o \times R \times B \quad [\text{volt}]$$

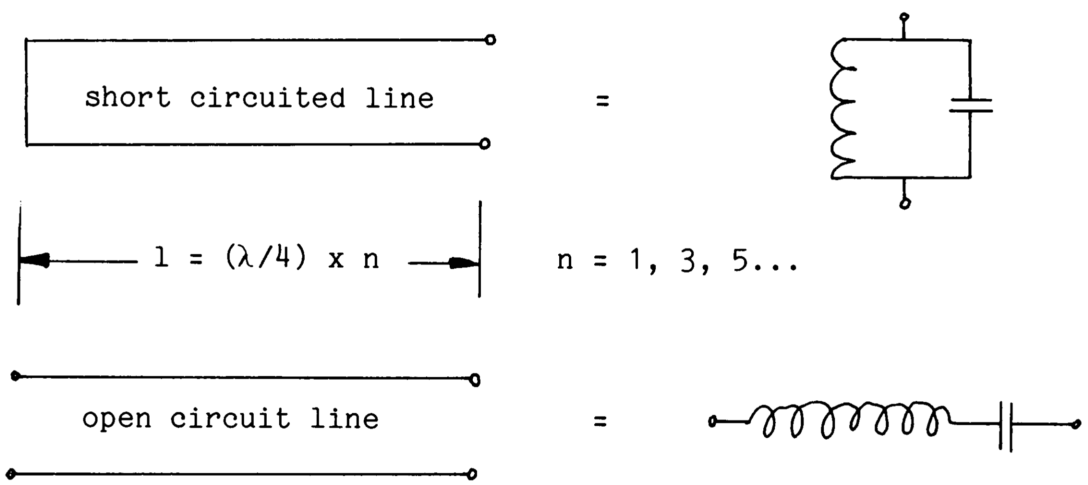
where

R = Resistance [ohm]
T = Temperature of resistor [degr.Kelvin]
B = Bandwidth [Hz]
k = 1.38×10^{-23} [Ws/°K]

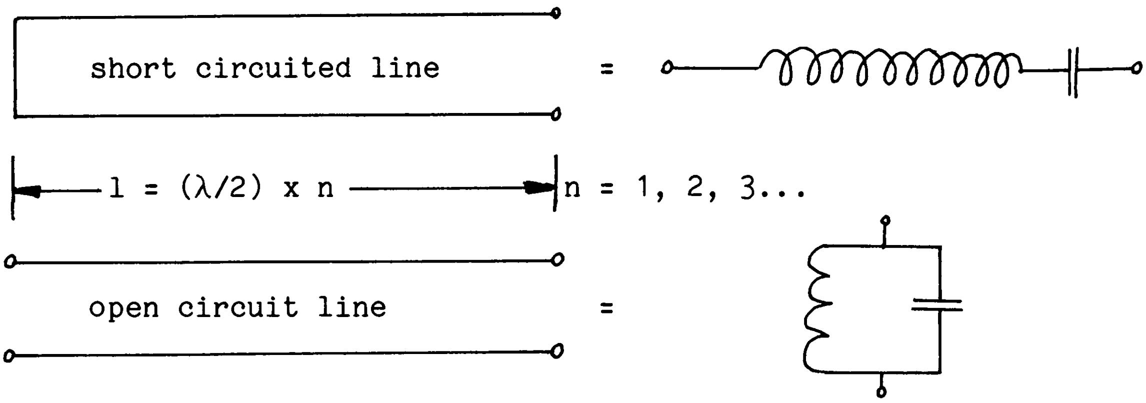
Types of UHF linear circuits.
(R.Ruell,D14NA)

As mentioned briefly in section A.1 it makes little sense to build up tuned circuits from lumped components (i.e. inductors and capacitors) at frequencies beyond approximately 350 MHz. Any active component - be it valve or transistor - has some input capacitance and some output capacitance. If such a device is coupled to a tuned circuit then this capacitance will enter the tuned circuit in whole (when coupled to the hot end) or in part (when coupled to a tap). Taking into acount the unavoidable stray capacitances and providing a variable tuning capacitor with reasonable tuning range one would end up with an inductor of very small diameter (less than 5 mm) and perhaps one or two windings or possibly just a wire loop.
Because of the unfavorable L - C ratio and large skin effect losses in the inductor such a circuit will have a low Q-factor even under no-load condition.
Compared with the resonant circuits for HF and VHF the tuned circuits designed for UHF will therefore be different.
One predominantly utilizes RF-transmission lines which are electrically a quarter wave length long (or an integer multiple thereof).They will show resonant properties when operated with open or really short circuited output terminals.

Example: Quarter wave length transmission line. Figure 24



Example: Half wave length line   Figure 25

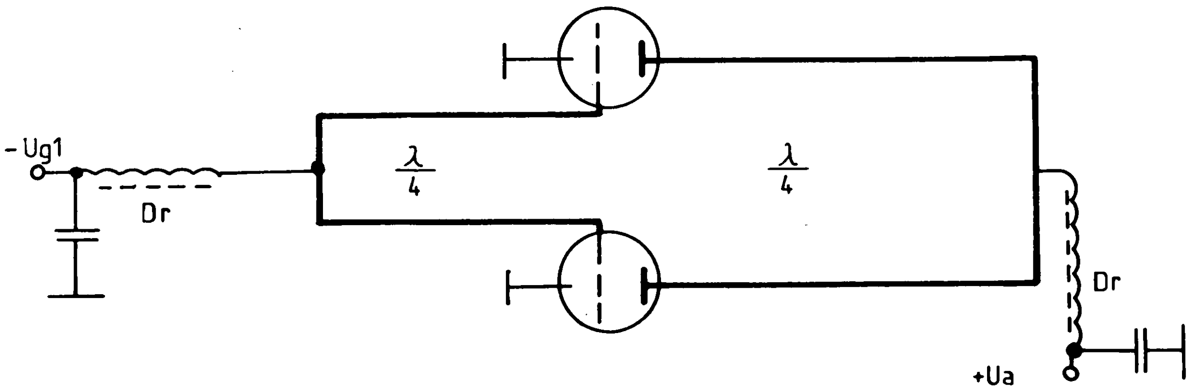


Suitable RF-transmission lines are the symmetrical twin-lead as shown in the examples and all other commonly used cables such as coaxial cables and micro strip lines. Cavities are favoured at even higher frequencies.

A.2.1 The quarter wave length linear circuit (Lecher line)

As mentioned before this is a symmetrical RF-line and thus particularly suited for symmetrical circuits.

Figure 26
Example of application of
Lecher line

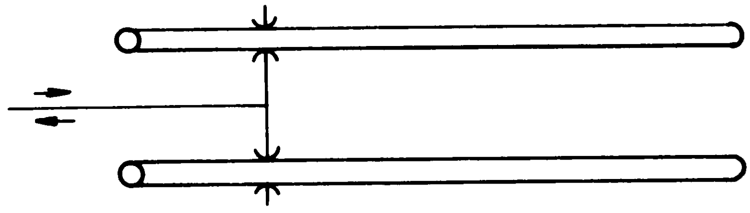


Due to the capacitive loading at the hot end of the line the physical length of it will be less than $\lambda/4$.

Tuning of this arrangement can be achieved in two different ways:

1. The effective length of the line is varied by sliding an adjustable RF short along the line.

Figure 27



2. The physical length of the line is arranged so as to shift its resonant frequency above the desired frequency. A variable tuning capacitor connected across the hot end of the line will allow tuning of the line towards lower frequencies (additional capacitive loading).

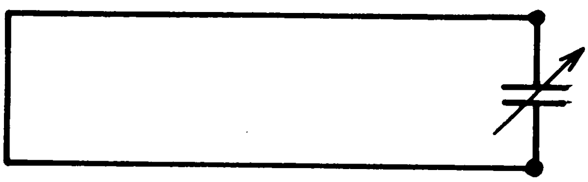
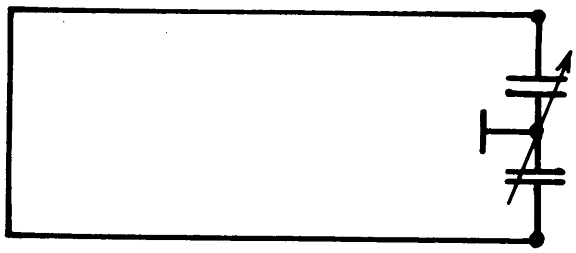


Figure 28

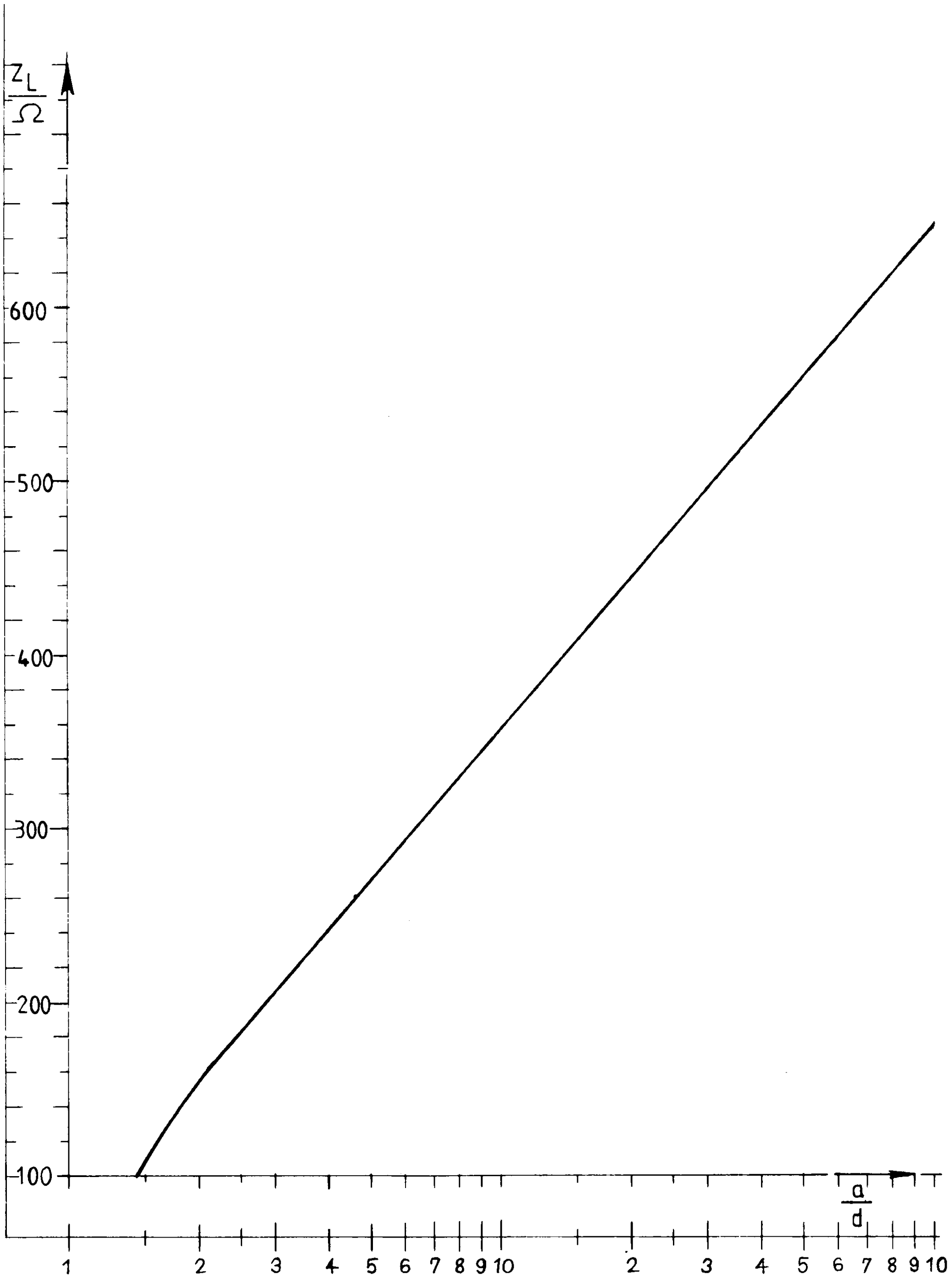
In this case the axle of the tuning capacitor must be insulated.

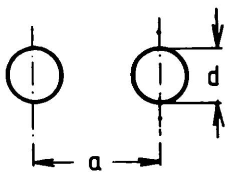


When using a butterfly- type tuning capacitor its axle should be connected to ground.

The main field of application for the Lecher line as a tuned circuit is that of power amplifiers equipped with valves (such as QQE 06/40, QQE 04/20, QQE o3/20, QQE 03/12) and in symmetrical balanced mixers with diodes or transistors.
In power applications the line material should be silver plated tubing with an external diameter of 6 to 10 mm to reduce skin effect losses (A.1.3) as far as possible. The characteristic impedance (Z_L) of the Lecher line should be chosen to lie between 160 and 300 ohm.

Figure 29





$$\frac{Z_L}{\Omega} = \frac{120}{\sqrt{\varepsilon_r}} \cdot \operatorname{arcosh} \frac{a}{d}$$

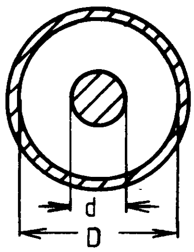
A.2.3 Coaxial lines.

For all other cases the use of coaxial lines will pose least problems to the amateur.

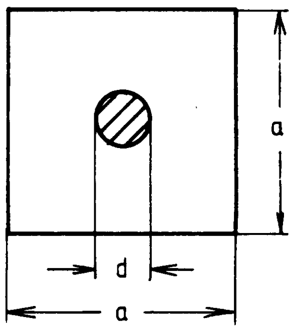
The coaxial line is an asymmetrical RF line which makes it particularly suitable for all single phase (unbalanced) circuits.

There are different types of coaxial lines:

Figure 30 Presentation of different coaxial lines and computation of their characteristic impedance Z_L



$$Z_L = \frac{138}{\sqrt{\varepsilon_r}} \lg \frac{D}{d}$$

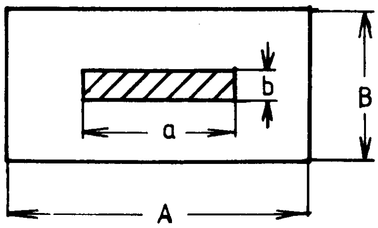


$$Z_L = \frac{138}{\sqrt{\varepsilon_r}} \lg 1{,}08 \frac{a}{d}$$

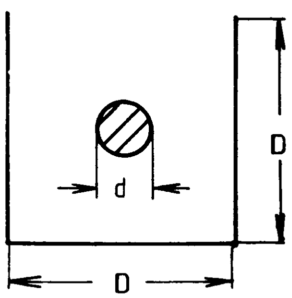
$$d = \frac{a \cdot 1{,}08}{10^{c}}$$

con $c = \frac{Z_L \sqrt{\varepsilon_r}}{138}$

(for a/d > 2)



$$Z_L = \frac{138}{\sqrt{\varepsilon_r}} \lg \left(\frac{A + B}{a + b}\right)$$



$$Z_L = \frac{138}{\sqrt{\varepsilon_r}} \cdot 1{,}17 \cdot \lg \frac{D}{d}$$

The most commonly used and easiest to implement versions are the combination of a circular inner with a rectangular outer conductor and rectangular inner and outer conductors.

The dielectric between inner and outer conductor defines the factor by which the line has to be shortened.

The value of factor $\frac{1}{\sqrt{\varepsilon_r}}$ equals 1 in the case of air being the dielectric. In this case there is no shortening of the line.

If the space between inner and outer conductor is filled with an insulating material then its dielectric constant ε_r must be entered. This value may be obtained from section G.3.2 for printed circuit materials and from section G.1 for insulating materials inside coaxial cables.

The characteristic impedance (Z_L) of coaxial lines should be of the order of 50 - 150 Ohm to achieve high no-load Q-factors. Outside these values the resonant impedance will drop off sharply. High resonant impedances are indicative of high quality resonant circuits. Furthermore there should be the least amount of capacitive shortening since this, too, will lead to a significant reduction of the resonant impedance (R_{res}).

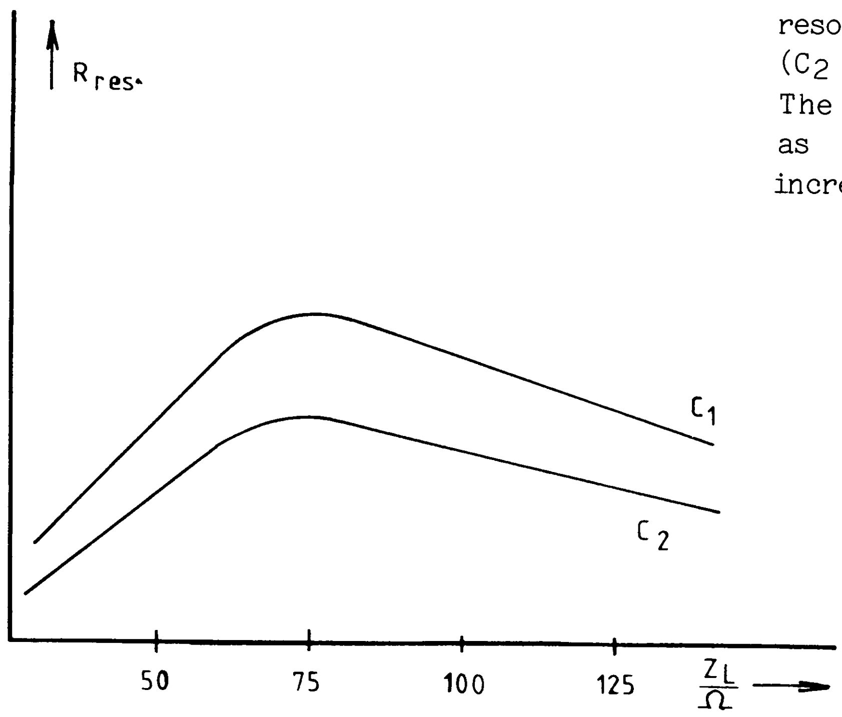


Figure 31
Diagram of relative resonant impedance of a resonant circuit.
($C_2 > C_1$)
The Q-factor is reduced as the capacitance is increased.

Another subject of great importance in coaxial circuits is the spot where the inner conductor is fixed to the base of the cavity. Good contact must be guaranteed since this is the point of maximum current flow. It is recommended to bolt the inner conductor to the cavity base (trough wall) particularly in all power amplifiers and in all equipment for the 70 cm and higher bands. In all other cases the inner conductor should have a tight fit and should be soldered from the outside. If possible the innner conductor should be shaped such that only its circumference will sit on the base surface (counter-sink or expand conically). By this measure very high contact pressures can be achieved which will result in low transfer resistance.

A.2.3.1 Calculation of tuning capacitance for shortened parallel wire lines and coaxial circuits.

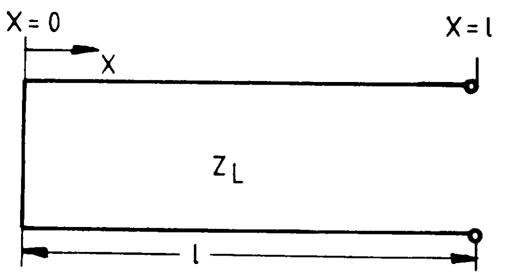


Figure 32
Diagram of circuit under consideration

At this stage a short theoretical derivation is required. For practical purposes the results may be read from the following diagrams. Equations from the transmission line theory allow the following expressions for current and voltage at line input:

$$\underline{U}_e = \underline{U}_o \times \cos\beta l + j\ \underline{I}_o \times Z_L \times \sin\beta l$$

$$\underline{I}_e = \underline{I}_o \times \cos\beta l + j\ (\underline{U}_o / Z_L) \times \sin\beta l$$

$\underline{U}_e$ resp. $\underline{I}_e$ voltage resp.current at X = 1
$\underline{U}_0$ resp. $\underline{I}_0$ voltage resp current at X = 0

$\beta = \frac{2\pi}{\lambda}$ 1 = physical length of line
λ = wave length on line

Since we are considering a short circuited line $\underline{U}_0$ is equal to zero.

From this the input admittance $\underline{Y}_e$ can be calculated:

$$\underline{Y}_e = \frac{\underline{I}_e}{\underline{U}_e} = \frac{1}{jZ_L} \times \cot\beta l$$

$$\underline{Y}_e = -j\frac{1}{Z_L} \times \cot\beta l \qquad \text{(inductive character of circuit)}$$

To fulfil the condition of resonance the input conductance must be made zero (or real). As the circuit is inductive a shunt capacitance is required for compensation.

$$\underline{Y}_e = 0 = -j\frac{1}{Z_L} \times \cot\beta l + j\omega_r \times C$$

$$\frac{1}{Z_L} \times \cot\beta l = \omega_r \times C$$

$$C = \frac{1}{\omega_r \times Z_L} \times \cot\beta l$$

$\omega_r = 2\pi \times f_r$ [Hz]

C in [F]

Z_L in [ohm]

$= \frac{2\pi}{\lambda}$

<u>Example:</u>

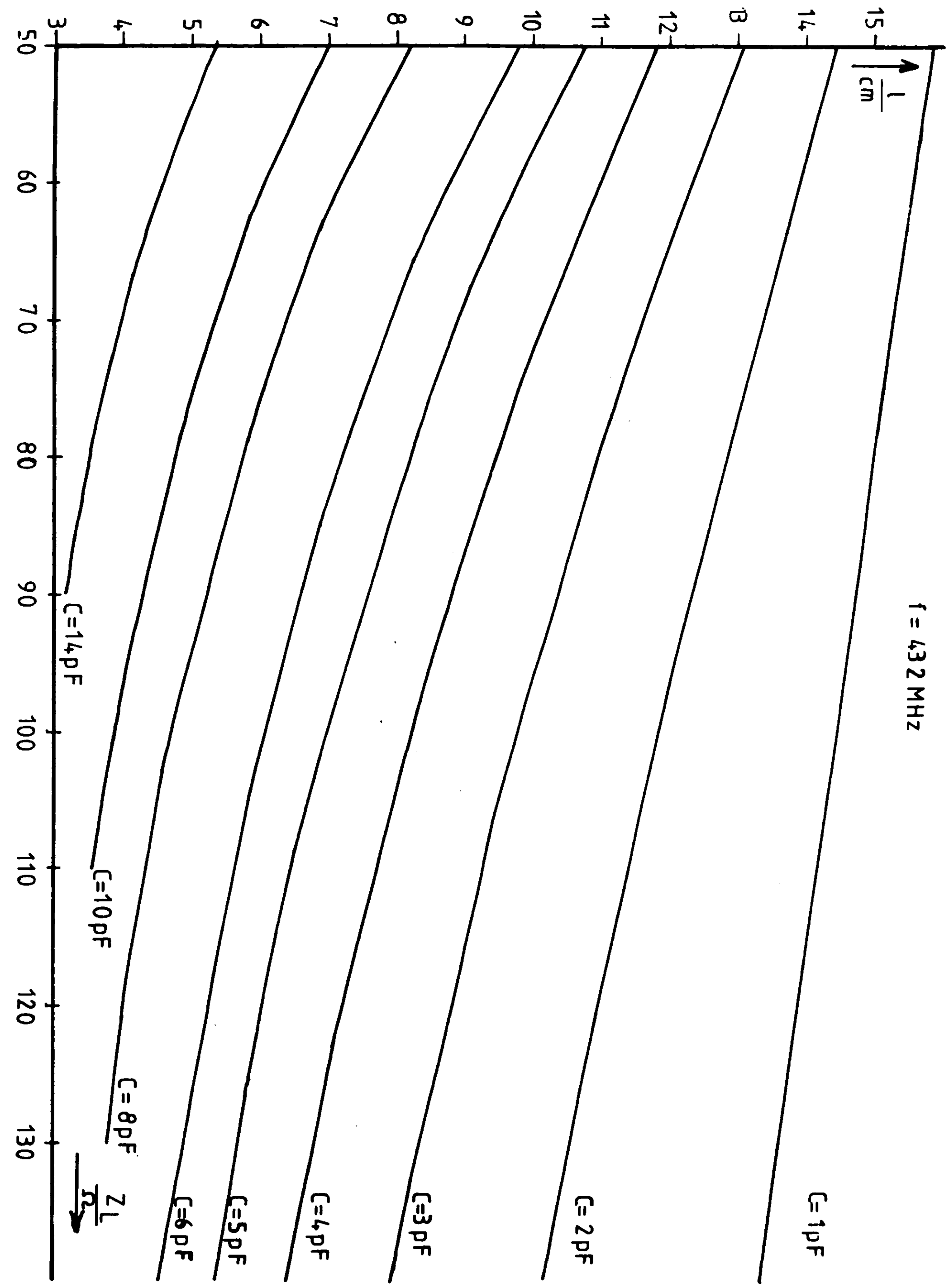
$f_{res} = 432 \times 10^6$ [Hz]
Z_L = 70 [ohm]
l = 7 [cm]
λ = 70 [cm]

$$C = \frac{1}{2 \times \pi \times 432 \times 10^6\text{Hz} \times 70} \times \cot\left[\frac{2\pi}{70\text{ cm}} \times 7\text{ cm}\right]$$

C = 7.2 [pF]
============

If transistors or valves with input and output capacitances are coupled to the circuit then these capacitances are transformed into the circuit partly or in full (depending on the type of coupling) and have to be taken into consideration.

Figure 33 Physical length of line in the various UHF-bands. Shunt capacitance (C) and characteristic impedance (Z_L) are given.



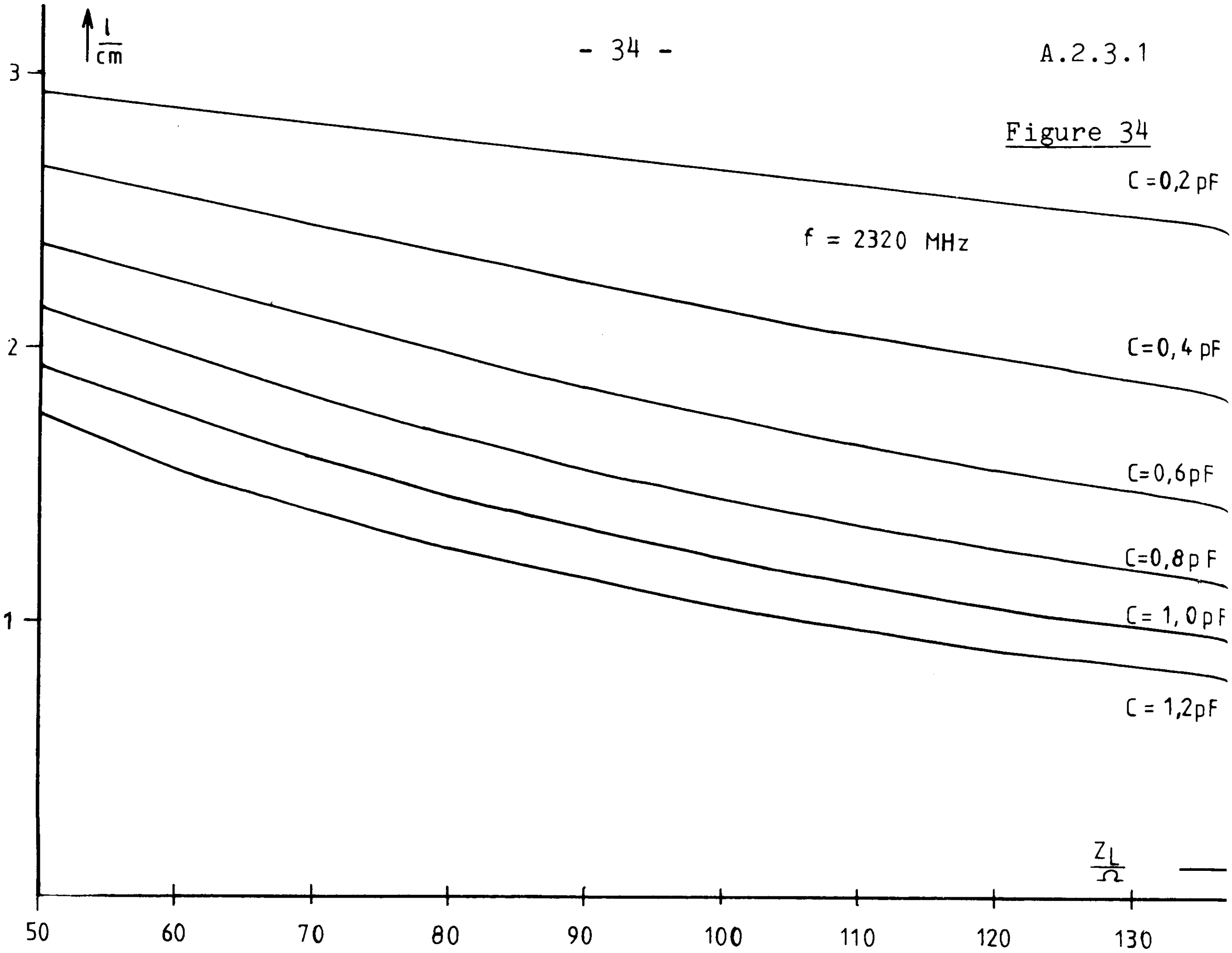


Figure 34

f = 1296 MHz

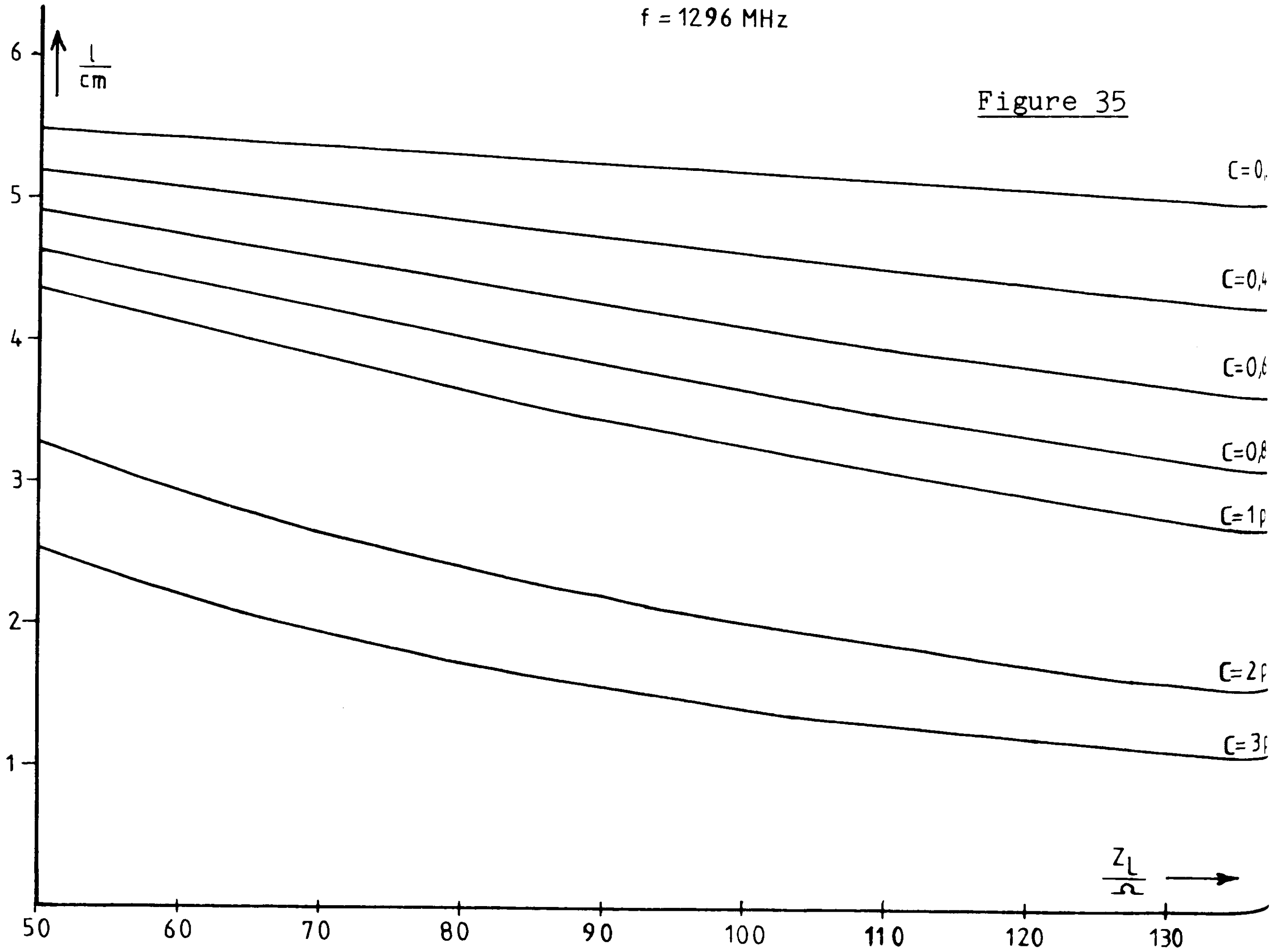


Figure 35

A.2.4 Circuits in microstripline technique.

A different approach to implement connections and resonant circuits in UHF and SHF modules is by means of the microstripline technique. Its great advantages are excellent reproducabilty and the avoidance of extensive "plumbing".

One disadvantage must, however, be stated: In 90 % of all amateur work the base material will be epoxy resin. This material will result in significantly lower no-load Q-factors than the coaxial technique! This will lead to a marked reduction in achievable stage gain and the RF-band width will increase:

- High no-load Q-factor = high quality circuit
- Low no-load Q-factor = poor quality circuit.

Commercially produced equipment makes use of AL_2O_3 or BeO as substrate and gold deposit as conductor material. This leads to imroved Q-factors but is unsuitable for the amateur due to technological reasons.
In high quality circuits double-sided copper-clad PTFE could be used. This is commercially available under the trademark Duroid.

Types of microstriplines.

Microstripline circuits are made of double-sided copper-clad base material. In this case one side (reverse) remains as a continuous ground surface. The front is etched to form the lines. The characteristic impedance of these lines depends on:

1. width "W" of the lines,
2. ratio W/h where h is the thickness of the base material,and
3. on the relative dielectric constant ε_r of the base material.

For epoxi ε_r has a value of about 4.8.
(PTFE approximately 2.2; see section G.3.2)

The characteristic impedance of the line may be computed as follows. Here we have to consider two different cases:

Case a $\frac{w}{h} < 1$

$$Z_L = \frac{60 \times \ln}{\sqrt{\varepsilon_{eff}}} \left(8 \frac{h}{w} + \frac{1}{4} \frac{w}{h} \right)$$

Case b $\frac{w}{h} \geqslant 1$

$$\frac{Z_L}{\Omega} = \frac{120\pi}{\sqrt{\varepsilon_{eff}}} \times \frac{1}{\frac{w}{h} + 1.393 + 0.667 \times \ln(\frac{w}{h} + 1.44)}$$

Into these equations for the characteristic impedance a constant E_{eff} has to be entered. This E_{eff} is the effective relative dielectric constant and depends on the ratio w/h.

$$\varepsilon_{eff} = 0.5 (\varepsilon_r + 1) + 0.5 (\varepsilon_r - 1) \times F$$

Factor F is a function of the ratio w/h.

Again two cases have to be considered:

Case a $\frac{w}{h} \leqslant 1$

$$F = (1 + 12 \times \frac{h}{w})^{-0.5} + 0.04 (1 - \frac{w}{h})^2$$

Case b $\frac{w}{h} \geqslant 1$

$$F = (1 + 12 \frac{h}{w})^{-0.5}$$

ε_{eff}:

Figure 36

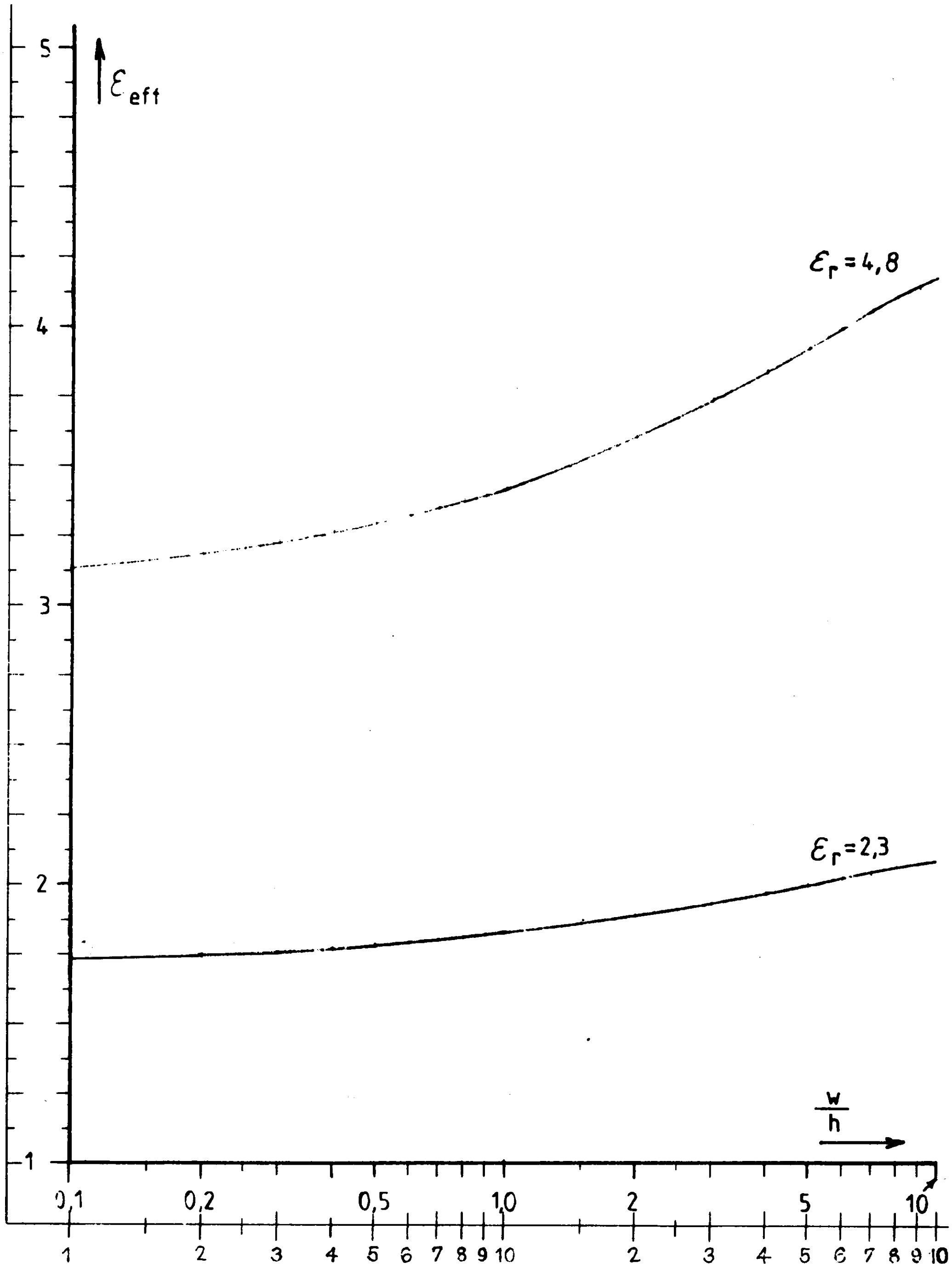
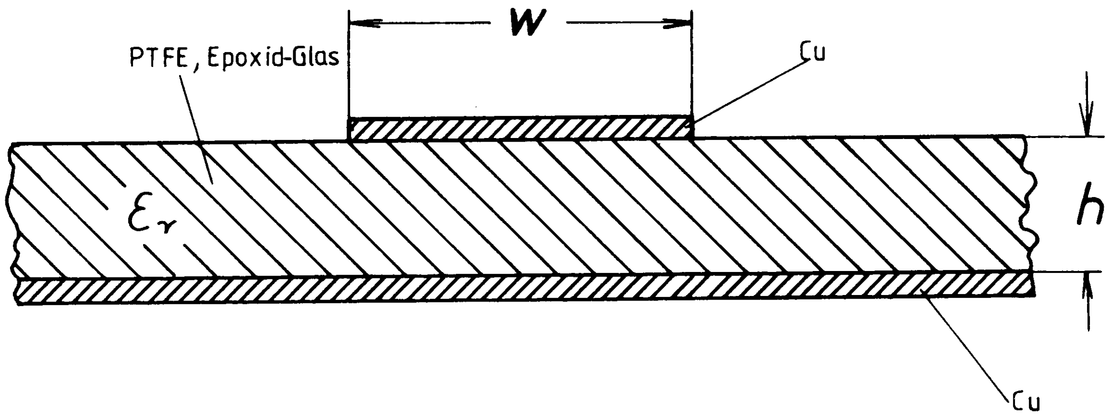


Figure 37 Construction of a microstripline.



For epoxi resin $\varepsilon_r = 4.8$

Computation of shortening factor for microstripline circuits.
The velocity of propagation of an electromagnetic wave on a microstripline depends on the dielectric constant of the material separating the two lines.

$$C = \frac{C_o}{\sqrt{\varepsilon_r}}$$

C.........velocity of propagation along the line.

C_o........Velocity of light in free space.

$= 3 \times 10^{10}$ [cm/sec]

The wave length on the line may now be computed:

λ_0.......Wave length in free space.
λ........Wave length on line.

$$\lambda = \frac{C}{f} = \frac{C_o}{\sqrt{\varepsilon_r} \times f}$$

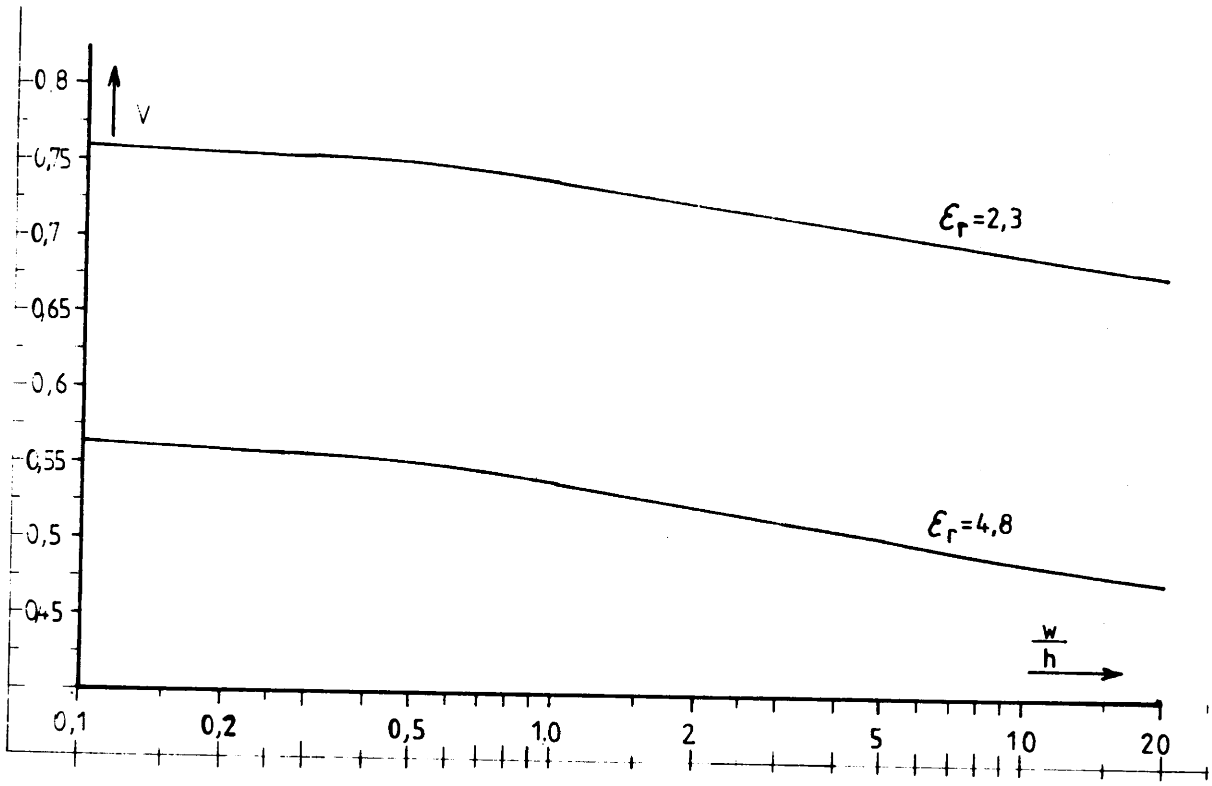
$$\lambda_0 = \frac{C_o}{f}$$

$$\lambda = \frac{1}{\sqrt{\varepsilon_r}} \times \lambda_0$$

In this case the shortening factor is $\frac{1}{\sqrt{\varepsilon_r}}$

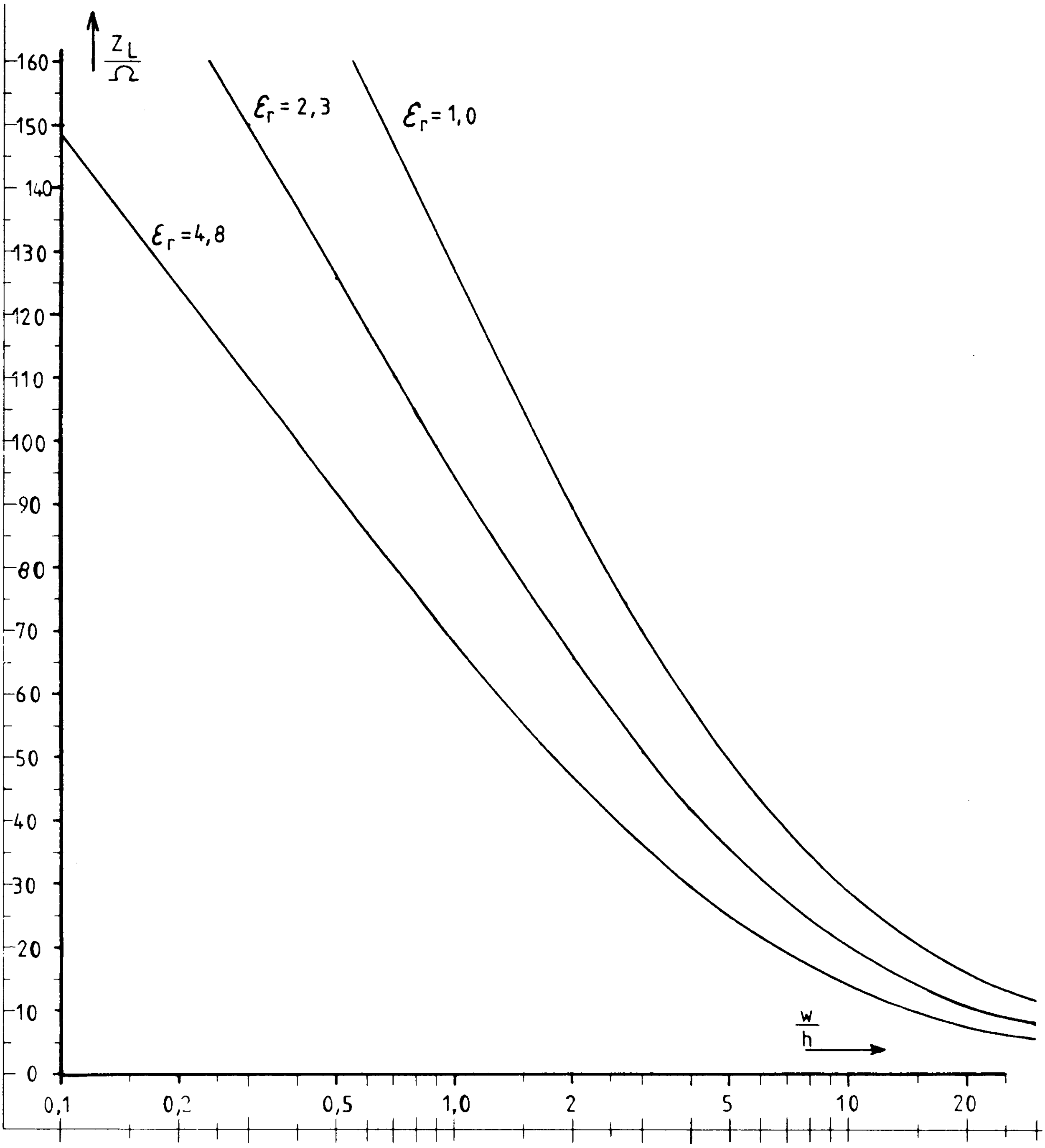
For microstriplines the change in effective dielectric constant due to the ratio w/h has to be considered (The corresponding value of ε_{eff} must be entered).
The shortening factors for epoxy and PTFE may be read from the following diagram:

Figure 38



All circuits designed for coaxial technique may also be built in stripline technique.If parallel resonant circuits in microstripline technique are required then the highest no-load Q- factors will be achieved when - similar to coaxial technique- the characteristic impedance Z_L is around 70 ohm.

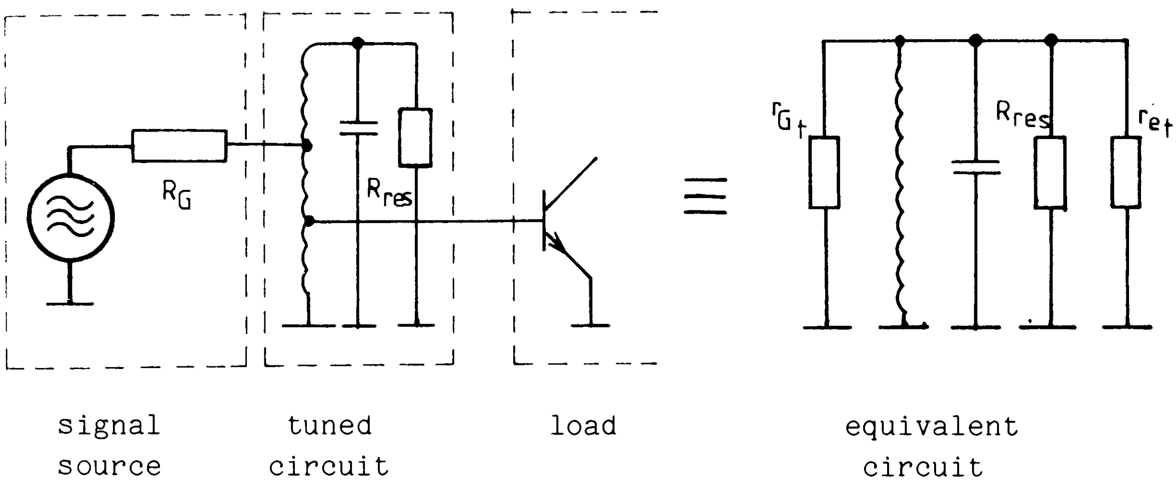
Figure 39 Characteristic impedance of a stripline as function of w/h for various base materials possessing different relative dielectric constants (Dk).



A.3 Coupling of linear circuits
(R.Ruell, DL4NA)

Tuned circuits in RF-applications can not be considered as independent modules separated from external components. They are mainly used in the signal path at the input or output of active components (valve,transistor). It is therefore necessary to couple RF-energy into and out of tuned circuits.
By coupling an antenna (respectively the output of an active device as signal source) or the input of an amplifier stage to a tuned circuit the internal resistance of the signal source or the load impedance will be transformed into the circuit and the no- load impedance will be shunted by this value.

Figure 40



R_G = internal resistance of signal source
R_{res} = no-load resonant impedance
r_{Gt} = transformed internal resistance of signal source
r_{et} = transformed load resistance

By combining these three components (no-load resonant impedance, transformed internal resistance of signal source,and transformed load resistance) we obtain the operational resonant impedance.
The operational resonant impedance is thus always smaller than the no-load resonant impedance of the tuned circuit.

There is a direct relationship between the resonant impedance and the quality factor Q according to the following equation:

$$\boxed{R_{res} = Q \times X_L}$$

R_{res}..resonant impedance
Q.....quality factor of circuit
X_L....inductive resistance of inductor at resonant frequency ($X_L = 2 \times \pi \times L \times f_{res}$)

Furthermore the following equation applies:

$$b = \frac{f_{res}}{Q}$$

b........bandwidth of resonant circuit
f_{res}...resonant frequency of circuit

Thus the operational resonant impedance has a direct influence on the bandwidth of a tuned circuit.
By suitable choice of the transformation ratio (separation of coupling point from cold end, size, form and position of coupling loops, capacitance of coupling capacitors) for signal source and load the operational resonant impedance and thus the bandwidth of the tuned circuit may be adjusted.
(It is assumed that proper construction resulted in the no-load resonant impedance being sufficiently high. Otherwise the no-load resonant impedance will have a dominating influence on the bandwidth).

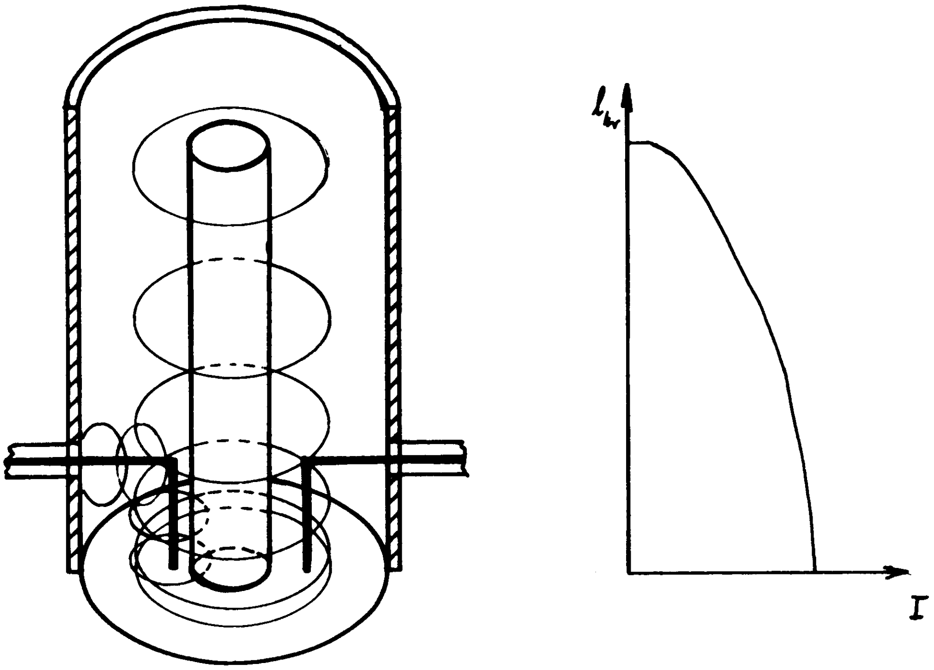
There are different ways of coupling RF-energy into and out of tuned circuits.

The various types of coupling are now demonstrated for the case of the most commonly used circuit, the coaxial circuit.
Printed parallel resonant circuits may be treated like coaxial circuits. In parallel wire linear circuits frequently just coupling loops are used to perform the coupling.

A.3.1 <u>Galvanic coupling</u> This is a special case of inductive coupling and will be covered by the following section A.3.2.

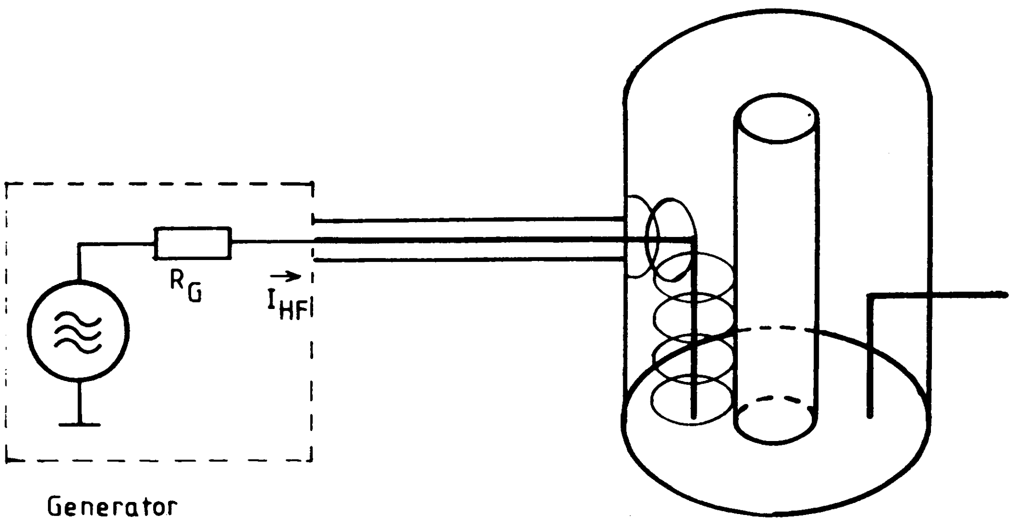
A.3.2 <u>Inductive coupling</u>
Coupling via the magnetic field at maximum current position of a coaxial circuit.

Figure 41



Coupling is performed by a wire loop. Connecting this loop to a generator will close the current path and high frequency alternating current will flow inside the circuit (The generator delivers power).

Figure 42



Current flowing in the coupling loop will set up a magnetic field around this conductor (The field lines are represented as circles around the wire). This alternating magnetic field will induce an alternating voltage inside the inner conductor of the coaxial circuit at the generator frequency, thus stimulating the circuit to oscillate.

If the resonant frequency of the tuned circuit differs from the frequency of the induced voltage, then the circuit can not follow these oscillations. Little or no power at all will be transferred to the output (attenuation). If, however, the resonant frequency equals the generator frequency, then strong oscillations will build up.

This resonant current will set up a magnetic field around the inner conductor (illustrated by the field lines around the latter). The magnetic field will be maximum (closest spacing between field lines) where the current is maximum. This magnetic field will permeate the output coupling loop and induce a high frequency voltage. High frequency power is transferred to the output and the insertion loss is low.

The spacing between the coupling loop and the centre of the inner conductor as well as the area which is enclosed by the loop defines the coupling factor and thus the load as seen by the circuit (operational resonant impedance).

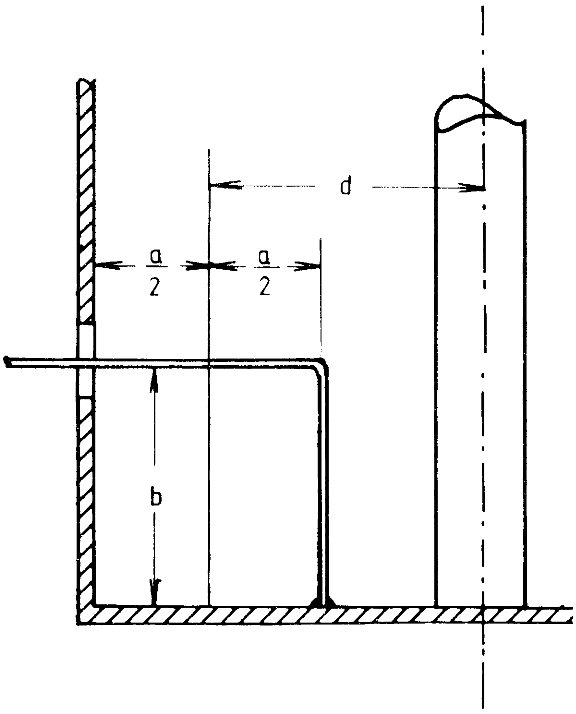


Figure 43

$F = a \times b \quad [cm^2]$

F = area of coupling loop

See reference (5) for equations concerning the calculation of a coupling loop placed at current maximum.
For amateur use the following simplified formula will suffice:

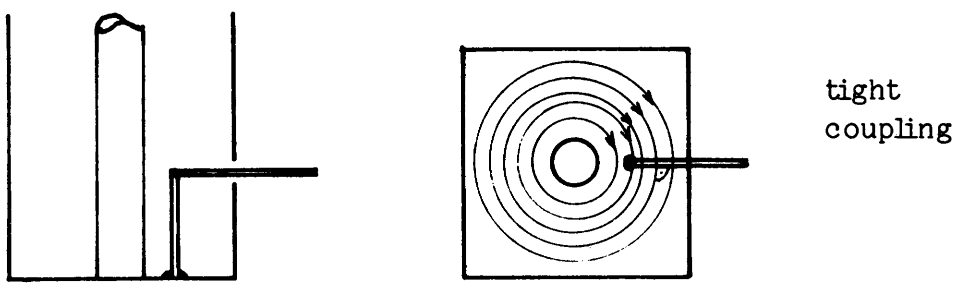
$F = 0.5 \times M \times d \times 10^9$ [cm]   where d...mean distance of enclosed area from centre of inner conductor in [cm].

$$M = \sqrt{R_A \times 0.5 \times \frac{1}{R_{res}} \times Z_{Kr}} \times \frac{\sin \frac{2 \times \pi \times l}{\lambda}}{2 \times \pi \times f}$$

R_A.....load or internal resistance to be matched to circuit
Z_{Kr}....characteristic impedance of coaxial circuit.
R_{res}...operational resonant impedance.
l......length of inner conductor of coaxial circuit [cm].
λ..[cm]

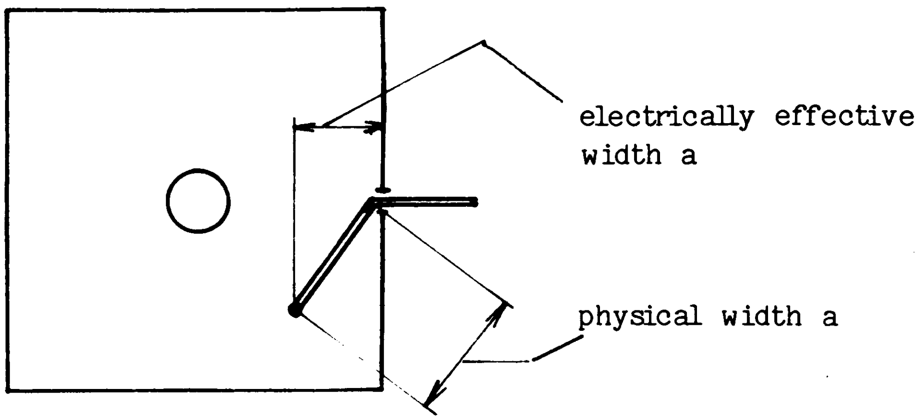
If the plane of the coupling loop is positioned normal to the penetrating magnetic lines the effective coupling area will be maximum - resulting in the tightest possible coupling.

Figure 44



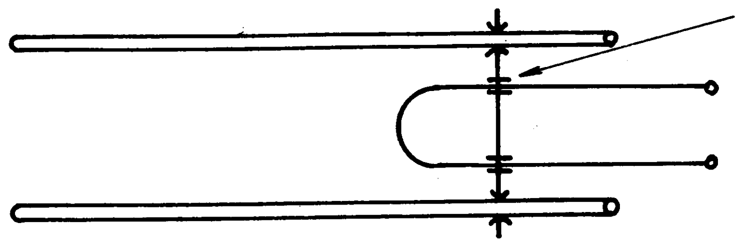
If the plane of the coupling loop is inclined such that the penetrating magnetic lines are no longer normal to it the effective coupling area will be reduced.

Figure 45



The same applies when the loop is swivelled such that the electrically effective height b is reduced.
In a parallel wire linear circuit the coupling loop is arranged as depicted in the following figure:

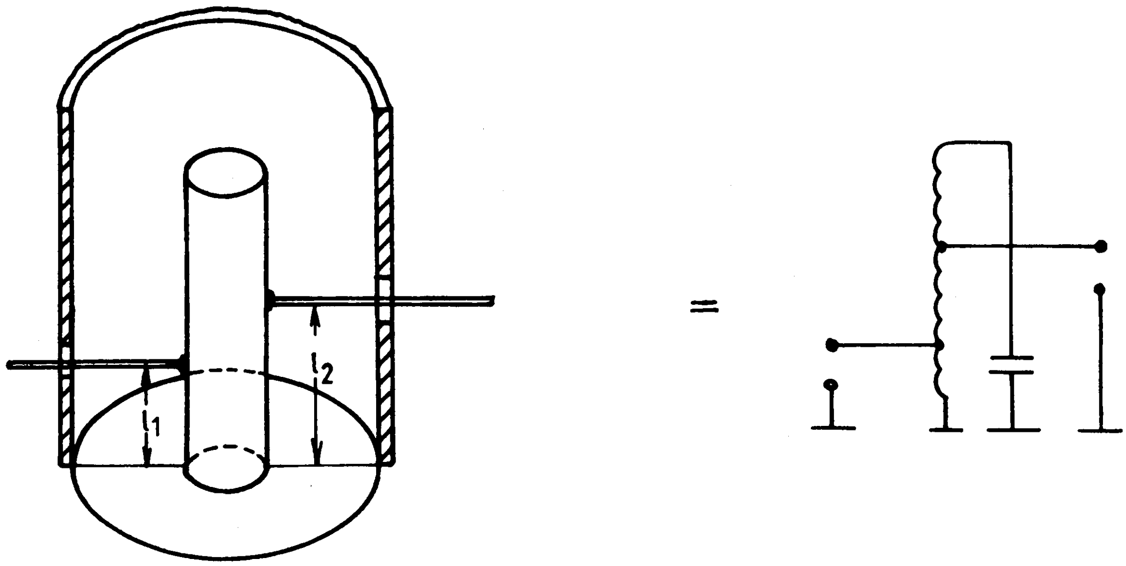
Figure 46



Great care must be taken to guarantee symmetric coupling. When connecting asymmetrical feed lines suitable measures must be taken (like inserting a balun).

The galvanic coupling

Figure 47



The galvanic coupling is a special version of coupling via a coupling loop. In this particular case part of the inner conductor is utilized for the coupling loop (Compare with auto-transformer in mains power supplies). The coupling is now defined by the spacing between coupling points l_1 resp. l_2 and the cavity base. The smaller l_1 resp. l_2 are made with respect to the total length (l_{Kr}) the lower the loading of the circuit by generator and load.

The distances l_1 and l_2 define the transformation ratio from input impedance to output impedance. Distance l for a tap point may be calculated as follows:

$$l = \frac{\arc\sin\left\{ \sqrt{\frac{R_A}{R_{res}}} \times \sin \frac{2 \times \pi \times l_{Kr}}{\lambda} \right\}}{\frac{2\pi}{\lambda}}$$

l.......spacing between tap and cavity base
R_A......impedance to be matched
R_{res}....operational resonant impedance
l_{Kr}.....length of inner conductor of coaxial circuit

A.3.3 Capacitive coupling

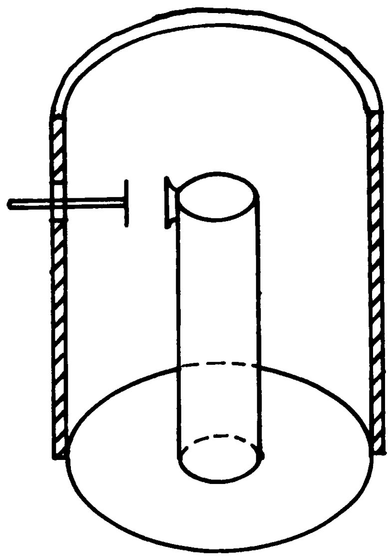
Capacitive coupling in the electric field (at voltage peak) makes use of a coupling capacitor to couple RF-energy into or out of a circuit. This coupling capacitor is placed at the position of peak voltage (coinciding with zero-current) which is the free end of the inner conductor (this is also the point to which the tuning capacitor is connected).This is the place of maximum circuit impedance (With the external components connected this equals the operational resonant impedance). The coupling must therefore be of high impedance - resulting in small values of coupling capacitance.

$$X_K = \sqrt{(R_{res} \times R_A) - R_A^2}$$

$$C_{coupling} = \frac{1}{2 \times \pi \times f \times X_K}$$

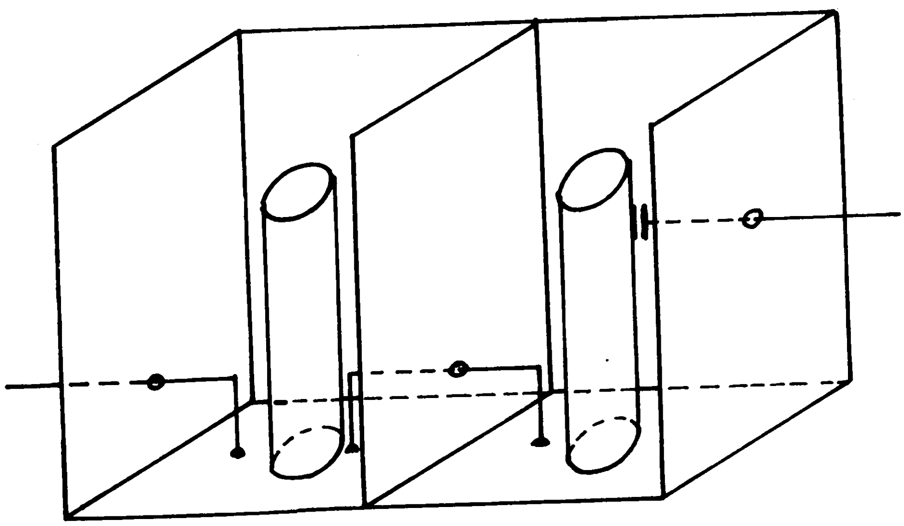
R_{res}...operational resonant impedance
R_A.....impedance to be matched

Figure 48



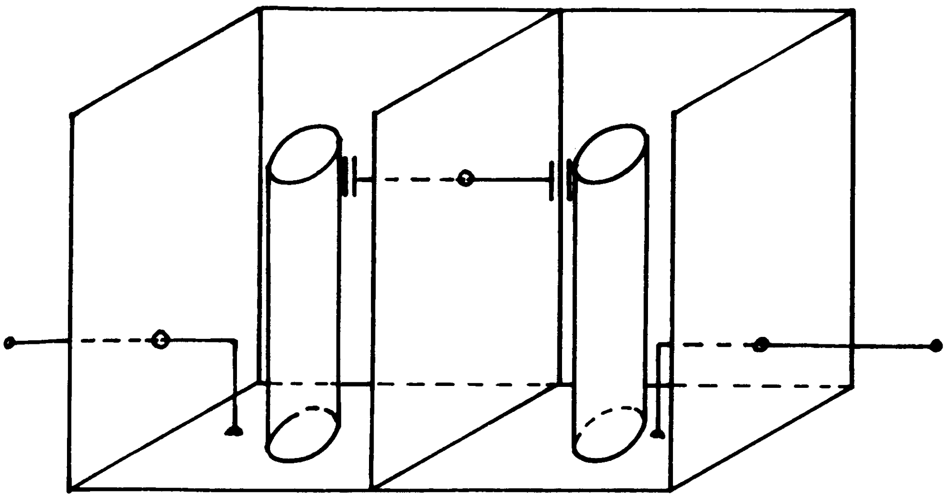
If two or more tuned circuits are to be coupled to one another then this can be achieved by using either inductive or capacitive coupling (See section C: Filters for signal selection).

Figure 49 Example of coupling



Inductive coupling between the tuned circuits.
Input: inductive coupling.
Output: capacitive coupling.

Figure 50 Example of coupling



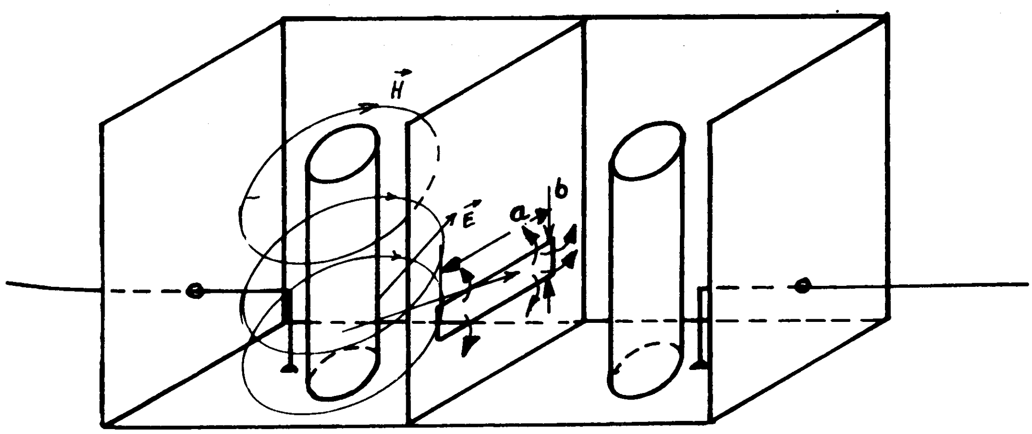
Capacitive coupling between tuned circuits.
Input and output: Inductive coupling

A further type of coupling exists for the interconnection of tuned circuits:

A.3.4 Radiation coupling

This type of coupling is achieved by slotting the wall that separates the tuned ciricuits. Electric and magnetic fields may pass through this slot from one circuit to the other.
If the resonant frequencies of both circuits coincide with the frequency of the generator then the second circuit will also be stimulated. The two circuits are coupled through their electric and magnetic fields. RF-energy can now be passed from the input of the first to the output of the second circuit.

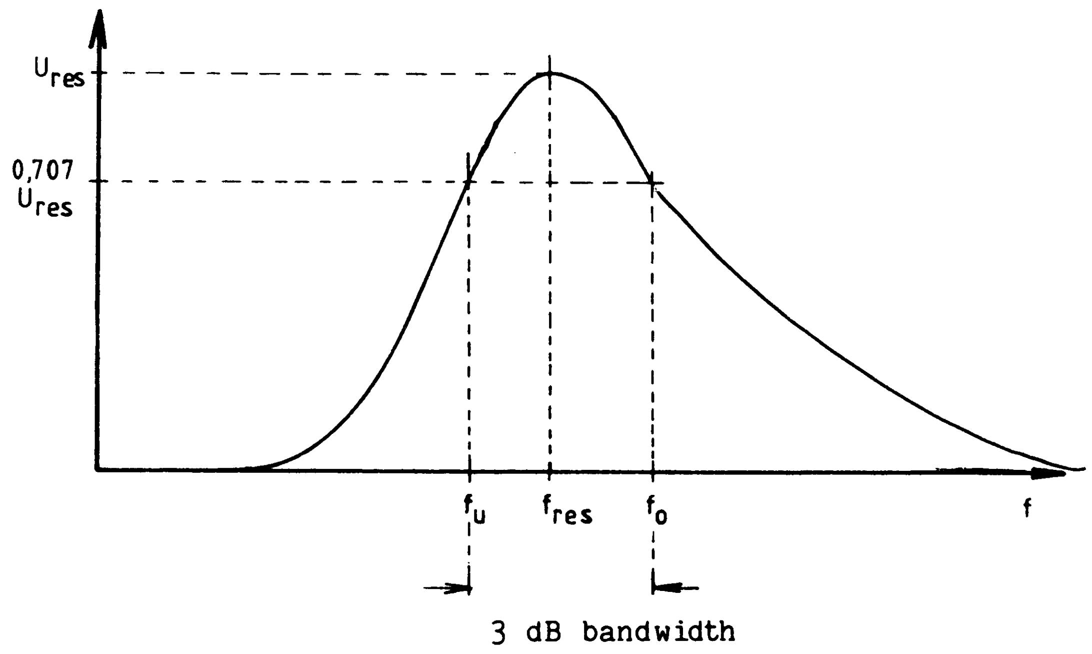
Figure 51 Example of radiation coupling.



The slot should be cut in such a way that "a" is larger than "b". Height "b" will control the degree of coupling. An increase in "b" will result in tighter coupling between the two circuits.

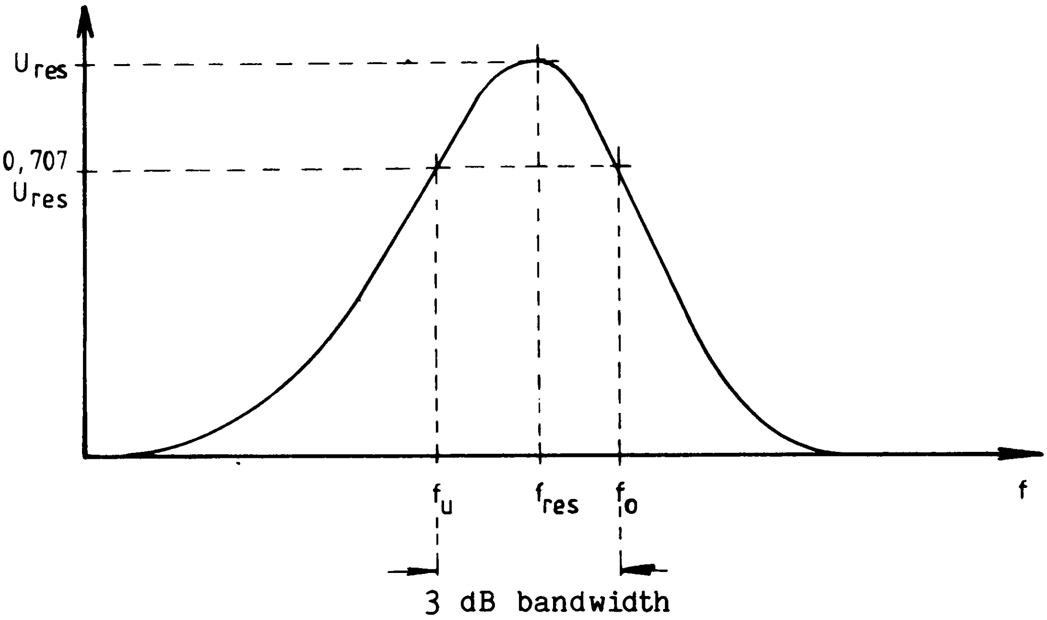
Another aspect should be born in mind when selecting the type of coupling (inductive or capacitive):
The type of coupling has an influence on the behavior above and below the resonant frequency. Capacitive coupling will result in high pass properties (the flank of the resonance curve is steeper for frequencies below the resonant frequency than it is for frequencies above resonance - See section C.3.1 and C.3.2).

Figure 52 High pass properties due to capacitive coupling.



When choosing inductive coupling, a low pass behavior will be the result (the flank of the resonance curve is not as steep for frequencies below the resonant frequency as it is for frequencies above resonance).

Figure 55 Low pass properties due to inductive coupling



If both types of coupling are used in a filter arrangement a reasonably even shaped resonance curve above and below resonance can be achieved.

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A.4 Amplification at UHF

Rather than treating the theory of amplifiers (the reader is referred to other publications) the following section will compare valve and transistor operation in amplifiers and indicate trends in transistor development. Furthermore some technical terms are discussed in their widest sense as related to amplifiers to improve the understanding of data specified in the building section of this book.

A.4.1 Valve or transistor as power amplifier.

Even in UHF-amateur equipment there is a noticeable trend towards the transistor PA. But due to financial and other reasons the valve will remain at the heart of 70 cm and 23 cm power amplifiers for years to come, in particular as the cost/performance relationship is rather favorable. But transistor technology is progressing, too. Power levels around five watts are quite possible but for the average amateur these types are still out of reach. Reasonably priced valves (2C39 and 4CX250) can produce output signals of a spectral purity and of a power level that is hard to duplicate by transistors at several times the financial outlay. When considering the low material effort of PA- arrangements (in D.5.9) the future of the relatively easy to build 2C39 PA becomes obvious.

But the low power transistor has got a chance as well. In phased array antennas several radiating elements - each connected to an individual low power amplifier - are placed in front of a reflector arrangement. In the transmit mode the sum of all power components and the antenna gain are effective. This approach to higher power has a bright future with amateurs as well since the usual stacked antenna systems require only little additional effort. Phase relations must not be neglected: This calls for identical lengths between driver and power amplifiers - respectively power amplifiers and radiating elements. The insertion of phase shifters will allow the antenna beam to be steered electronically (and very fast, too). Expensive rotors with inaccurate indicators could become things of the past. Building instructions for such antenna systems even for amateur application are certain to appear in technical publications in the coming years.

The range of power transistors is dominated by those of the GaAs- field effect variety. However, before considering their application in the near future the reader is reminded of the remarks at the beginning of this chapter!

A.4.2 Receiver sensitivity

The sensitivity of a receiver is governed by the system noise. Noise fundamentals were treated in A.1.7 and mathematical relations in D.1. At this stage only the sensitivity is going to be discussed. Even the attenuation of the antenna cable (which must be added to the overall noise) is disregarded.

The noise power depends on bandwidth (B_n) and temperature, as can be seen from equation 13 in A.1.7. It is common practice to choose the room temperature (T_o = 290°K) as reference. This limits the theoretical sensitivity ($N = K \times T_o$) to -174 dBm for 1 Hz of bandwidth. But the amateur is not likely to operate a receiver at a 1 Hz bandwidth; the sensitivity will thus be reduced as the bandwidth is increased:

B_n:	Theor.sensitivity	Factor	at mode of operation
1 Hz	- 174 dBm		
10 Hz	- 164 dBm	10 dB	
100 Hz	- 154 dBm	20 dB	--- CW, RTTY
1 KHz	- 144 dBm	30 dB	}-- SSB and NBFM are around
10 KHz	- 134 dBm	40 dB	}-- wideband FM
100 KHz	- 124 dBm	50 dB	
1 MHz	- 114 dBm	60 dB	}-- ATV
10 MHz	- 104 dBm	70 dB	

The data specified in dBm refer to 1 mW. A table of dBm versus mW is included in section G.3.1. As the receiver sensitivity is essentially governed by the preamplifier the theoretical sensitivity is decreased by that amount. With insufficient preamplifier gain (G_V) some part of the mixer noise will also have to be taken into consideration. A rough check may be performed by disconnecting the antenna. If this results in an audible noise variation then G_V is generally adequate.
The receiver sensitivity is defined as being that amount of signal power (S_{min}) at the input which will produce a signal-to-noise ratio equal to 1 at the receiver output.

Worked example:

$S_{min} = kT_o \times B_n \times F$

Bandwidth: 1 KHz,factor	=	30 dB
F of preamplifier	=	4 dB
Other losses	=	4 dB

S_{min} = - 174 dBm + 30 dB + 4 dB + 4 dB = - 136 dBm

In other words: If a 1 mW signal is attenuated by 136 dB and fed into a receiver a S/N of 1 will result at the output. This value of -136 dBm should be achieved by every good receiver or VHF- converter followed by a HF-receiver. If this is not the case then certain improvements should be considered.
The noise of the second preamplifier stage has not been taken into account since it is usually of no importance.

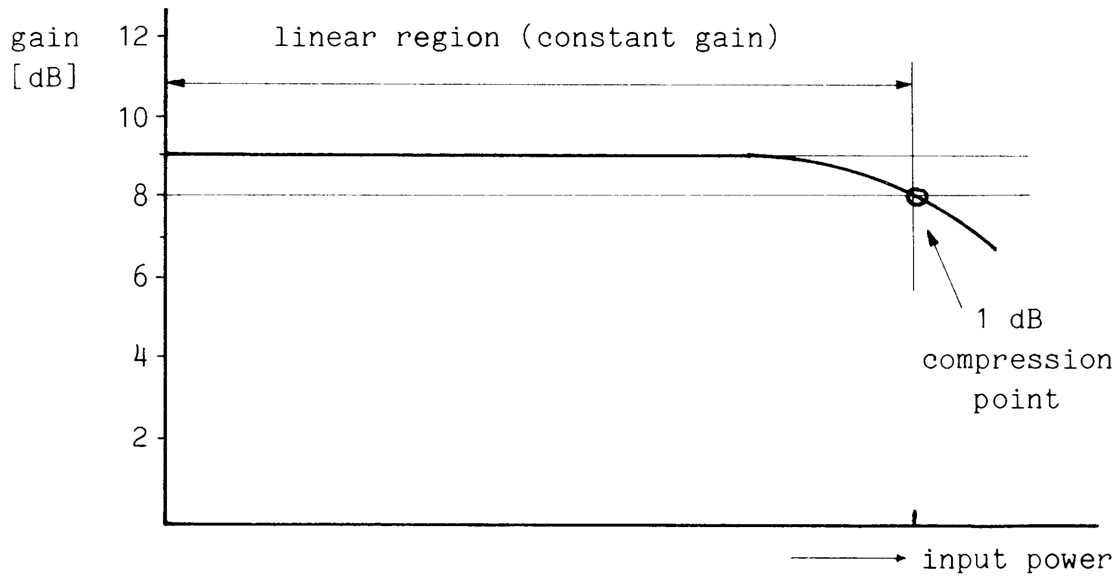
Its influence may however be calculated: $F_{ges} = F_1 + \frac{F_2 - 1}{V_1}$

It is apparent that this noise has little significance if the first-stage gain is reasonably high.

A.4.3 Compression point

For a wide range of signal levels fed to any amplifier (be it a preamplifier, driver, power amplifier or receiver) the output power will be proportional to the input signal power. The factor of proportionality is thus a constant and represents the small signal gain. Throughout this region the amplifier is said to be linear. From a certain input level onwards saturation effects will cause a reduction in gain i.e. the output signal will no longer equal the product of input signal multiplied by the small signal gain. It is convenient to define the limit of the linear range as that input signal for which the effective gain is one dB less than the small signal gain.
This is the "1 dB-compression point".

Figure 54



A.4.4 Receiver dynamic range

The dynamic range describes the power ratio of the weakest and the strongest signal which an amplifier or receiver can handle without any of its components reaching saturation. The maximum signal level is defined by the 1 dB compression point whereas the minimum signal level (S_{min}) is set by the sensitivity. (See section A.4.2). The following example will illustrate this:

Receiver: Bandwidth (B_n) ....10 KHz
Noise figure of preamplifier....2 dB
S_{min} = -174 dBm + 40 dB + 2 dB = -132 dBm

Assuming a 1 dB compresion point of -50 dBm the receiver dynamic range results as the difference between -132 dBm and -50 dBm and has a value of 82 dB.

Modern spectrum analyzers have dynamic ranges of 80 to 100 dB.

A.5 UHF-Diode Mixers

Due to the development and continuous improvement of Schottky diodes and due to the fact that they are readily available at reasonable prices mixing stages even of amateur equipment are almost entirely based on diodes. This section will provide the newcomer with a summary of advantages and disadvantages of various circuit arrangements and also with hints for the construction of diode mixers for the 70 cm and 23 cm bands in both open-air and stripline technique. Starting with the balanced mixer, the theory of operation will be outlined in section A.53. The single diode mixer is discussed just briefly in several applications.

A.5.1 Simple Mixer for 432 and 1296 MHz

This type of mixer has found wide application in converters employing so-called interdigital filters (finger filters) for the 23cm and meteosat bands. In commercial equipment the use of single diode mixers is limited to SHF and EHF. Functioning single diode mixers are easily built but have poor isolation properties and require additional measures. Converters with interdigital filters (1) are fitted with resonant circuits thus limiting their bandwidth. Generally speaking these inter- digital filters could be adopted for 70cm, too. This requires only little additional mechanical effort; but here balanced or double balanced mixers are the obvious choice.
An additional application - in particular for wideband measurements - is the directional coupler. It is, however, a severe disadvantage that the coupling factor of a directional coupler (see B.4) means a corresponding attenuation of the oscillator signal (P_{fo}). P_{fo} must therefore be such that sufficient power is delivered to the diode in spite of this attenuation. If - for instance - a P_{fo} of 0 dBm (=1 mW) is required at the diode and assuming a transfer attenuation of 10dB a P_{fo} of 10 mW must be injected. Since the transfer attenuation of directional couplers that are not specifically designed for mixer applications is likely to be much higher (16 to 30 dB) it might prove difficult to generate the required power. The useful operating range is therefore limited by the frequency dependent transfer attenuation as well as the dimensions of the directional coupler, the lengths of the cables and diode parameters.

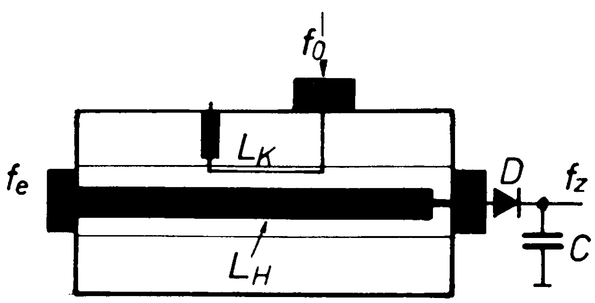


Figure 55
Example of a wideband mixer based on a directional coupler.

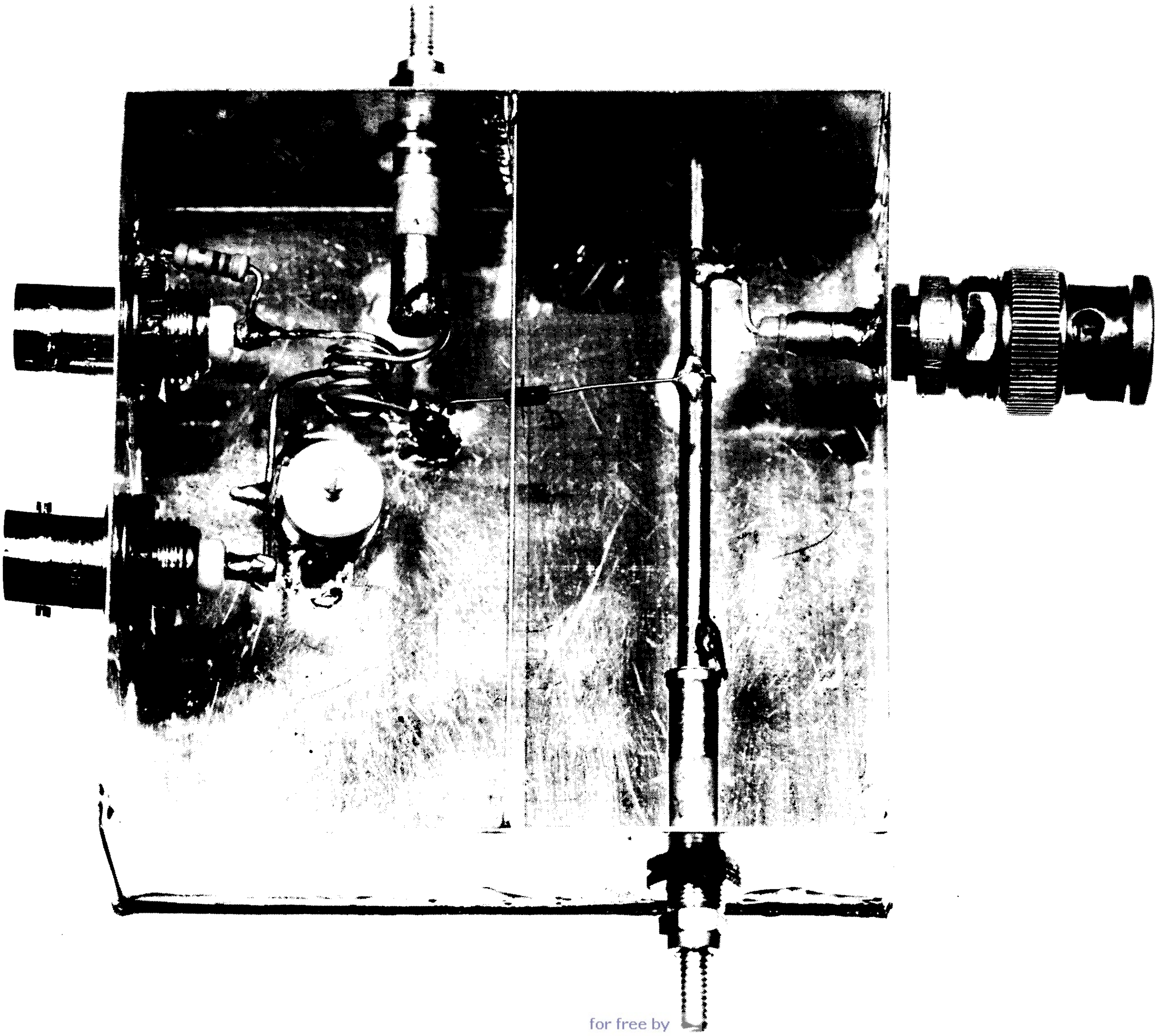
f_e = frequency of input signal
f_o = oscillator frequency
f_z = intermediate frequency
L_K = coupling line
L_H = feed-through line
C = RF-blocking

Bibliography: (1) UKW-Berichte No. 4/77

Assuming a coupling factor of only 3 dB then the input signal will also be attenuated by 3 dB due to the power splitting effect of the directional coupler (see B.4).

For demodulating a 70cm signal with the minimum amount of hardware the following arrangement could be duplicated. Since the photo gives all the relevant information a circuit diagram is not shown. The RF-signal is fed via the BNC-connector to a tap of the input linear circuit (13cm off the cold end). Its length depends on the dimensions of the tuning capacitor (0.5 - 12pF). In this case L is a piece of copper rod (diameter 3mm). One diode lead is connected to this at a point 25mm off the cold end. The other lead is soldered (together with the IF-coil) to a ceramic support as shown in the picture. The oscillator output is inductively coupled to the mixing diode. P_{fo} is terminated by a 50 Ohm resistor to improve the electrical stability. The inductors have been wound using a 5mm mandrel, the number of turns may be seen in the picture. The variable capacitor tuning the IF-circuit ranges from 4.5-70pF.

Figure 56 Extremely simple construction of single diode mixer for the 70cm band. Chassis dimensions: 72 x 72 x 30 mm.



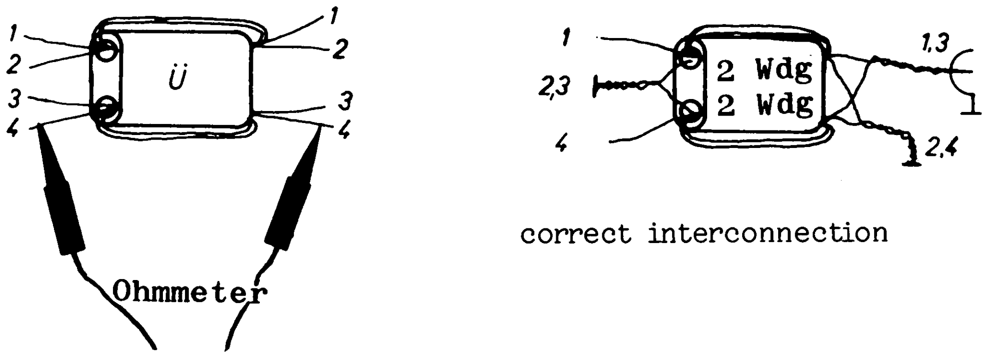
A.5.2 RX/TX-Balanced Mixers for 432 and 1296 MHz

This section presents simple, cheap balanced mixers employing Schottky-diodes. They are suited for both transmitters and receivers. Again they are designed as compact modules for universal application. In commercial equipment for millimeter and sub-millimeter waves the constantly improved Schottky diodes dominate as mixers. Furthermore the so called Mottky diodes (1) find increasing use as they require 3 dB less oscillator power for best noise figure. Extensive developments of Schottky diode mixers up to several hundred Giga Hertz will result in improved mixers for amateur applications. The basic building blocks for the 70- and 23cm bands are described here. As advanced products become available all that is required is to replace the two mixer diodes by new types to upgrade the equipment to the latest standard.

The advantages of the described mixers are simple construction, no power supply, duplex operation (RX/TX) with a saving in components, large dynamic range and significant suppression of the oscillator signal at the output terminals (-43dB). Especially in transmitting equipment these advantages over active mixers are obvious.

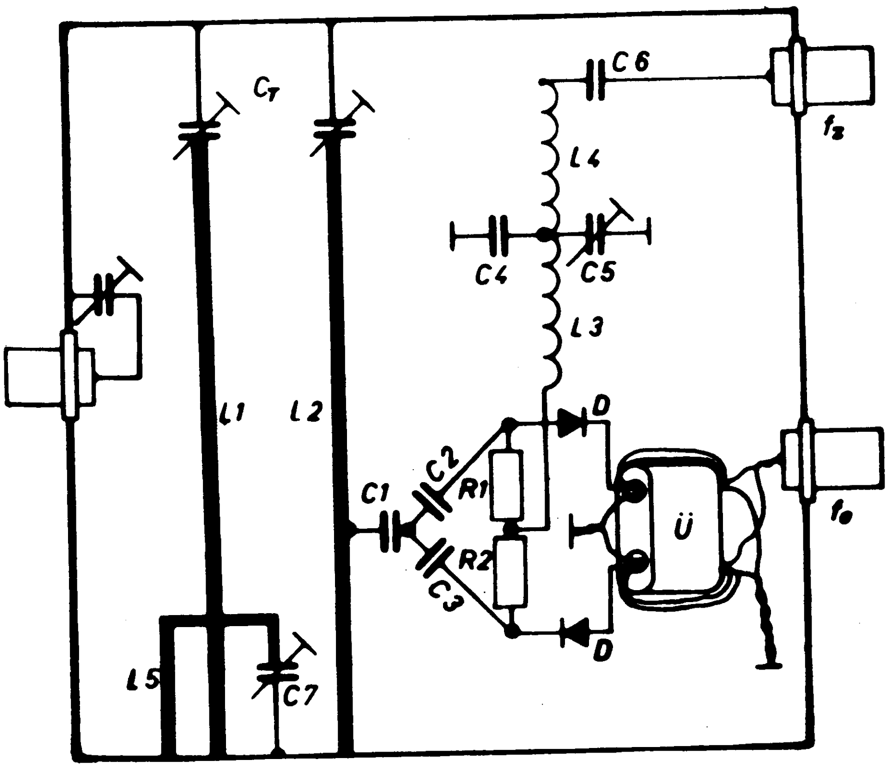
Theory of operation: The transmit IF-signal (144 or 28 MHz) is entered via a tuned low pass filter to be mixed with the oscillator frequency by the two diodes. The capacitor combination C1-C3 passes all mixing products on to the bandfilter L1/L2. L5 and C7 are tuned to f_o for the transmit mode and to the image frequency f_{sp} for the receiving mode. The choice of IF (28 or 144 MHz) will be discussed in section D.2.
The winding sense of transformer (U) may be taken from the drawing. The two ports f_z and f_o may be swapped around. But the input combination L4/C5 as well as L5/C7 must be tuned to the frequency of the signal being fed to port f_z. Further details and component values are given in section D.4.1 where this mixer is employed at 70cm.

Figure 57 Winding sense, finding the ends of the windings and interconnection of the transformer windings. The transformer is a Siemens two-hole core made of U 60 material. But the core of a balun transformer salvaged from the UHF-section of an old TV set will serve the same purpose. A piece of twin lead (2 x 0.2 mm Cu or equiv.diam.) is wound according to the drawing, verified and interconnected. For best clarity the drawing shows only one turn per winding.



correct interconnection

Figure 58 Wiring diagram of the balanced mixer for transmit and receive applications on 70 cm. Further data is given in D.4.1. As laid out the circuit will fit into a chassis of 72 x 72 mm outside dimensions. In combination with a preamplifier from section D.1 a modern converter with excellent strong signal handling capability will result (see D.1 and D.7.1).



F_Z = 28 MHz band about 10 mW input for transmitting

F_O = 404 MHz
RX = 5 - 15 mW
TX = 15 mW

Circuit for the 23cm Band

In contrast to the arrangement for the 70 cm band selectivity is achieved by circuits in $\lambda/2$ technique. These circuits allow a narrowing of the bandwidth down to 10MHz. This concept is thus suited for narrow band transmitting and receiving equipment, whereas a mixer according to A.5.4 is superior in wide-band applications (such as panoramic receivers or scanners). The balanced configuration based on Schottky diodes can handle high output levels as required in transmitter applications. Transistorized mixers are usually below optimum at P_o-levels of 0.5mW and begin to deteriorate due to saturation at 1.5mW. The response of this mixer is shown in the following diagram:

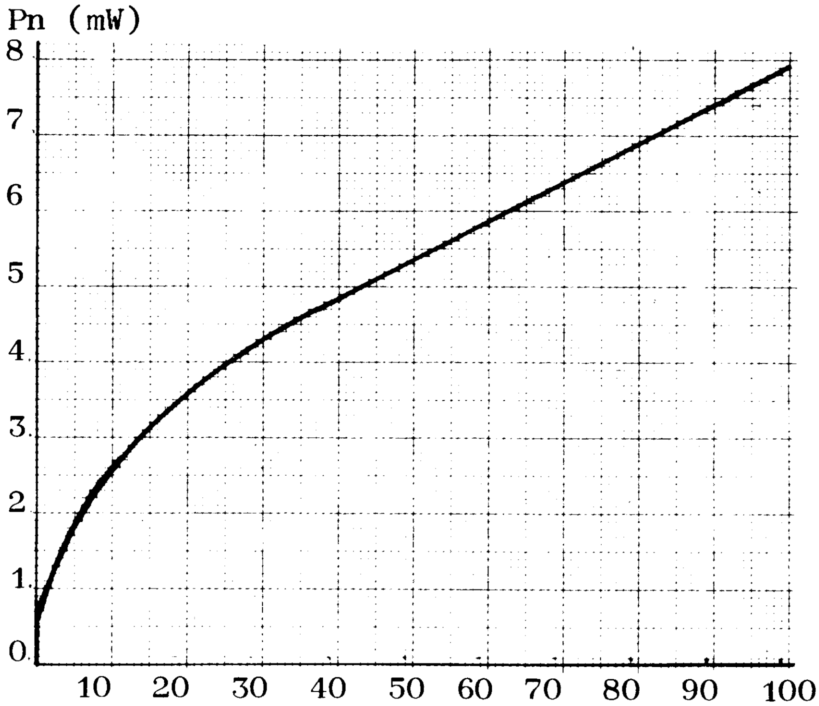


Figure 59
Measured data of TX/RX mixer in the 23-cm band

f_o = 1152 MHz
P_o = 30 mW
f_z = 144 MHz

Example:
P_{fz} = 20 mW
P_n = 3.8 mW

f_z-level [in mW]

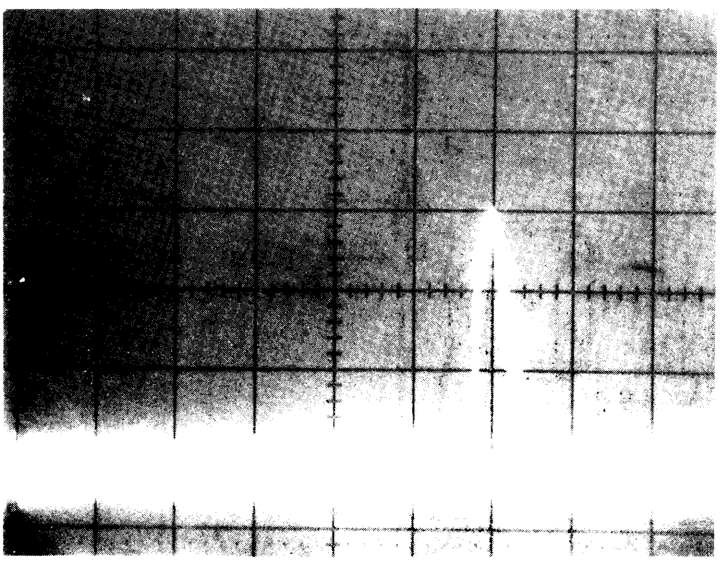


Figure 60
Passband of the mixer at the IF-output port as generated by a scanning frequency generator sweeping from 50 - 2000 MHz

Scale:
h: abt.50 MHz/line
v: 10 dB /line

Figure 59 shows that the mixer can handle a wide range of power levels. All that is going to change is the output level at port 3.

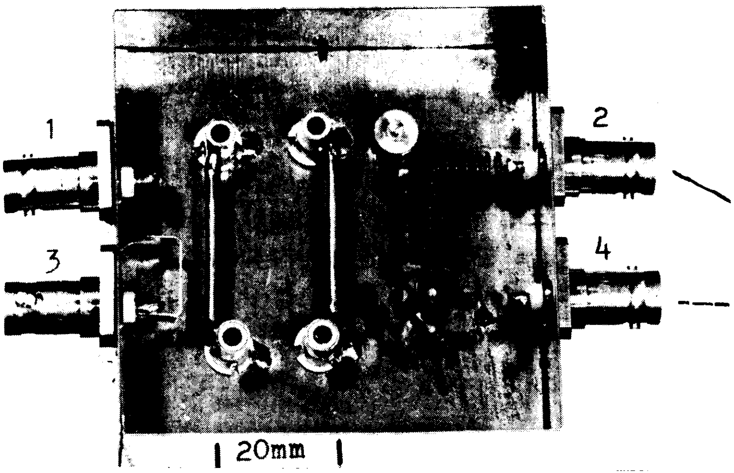


Figure 61

Overall view of mixer in its enclosure 72 x 72 mm (external dimensions)

f_Z = 144 MHz

1152 MHz, 3-5 mW

1: Test signal input
3: Antenna resp.preamp.

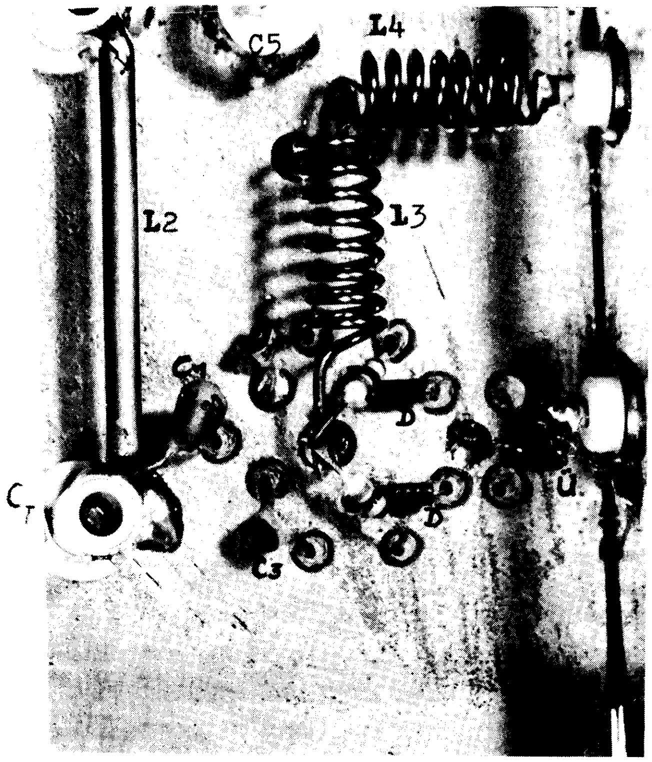


Figure 62

Enlarged section of balanced mixer showing IF and second RF stages

Schottky diodes:
Type HP 2900

C2/3 = 5.6 pF
C1 = 3.9 pF
C5 = 0.5-6 pF
CT = 0.5-6 pF

L3/4 = 6-8 turns (4mm diameter, 0.5mm Enamel Cu)

R1/2 and L3 are soldered to a ceramic mounting stud.

L2 = 30mm of 2.5mm diam. Cu.

Figure 63

Rear view of mixer assembly with the tuning capacitors 0.5-6 pF.

Mixing losses amount to approximately 8 dB.

This balanced mixer with Schottky diodes may be incorporated into a modular station concept (This allows easy replacement of all modules for whatever reason). On the other hand it could be integrated into a low-power transverter. The following photos show both the 70cm and 23cm mixers as part of low-power transverters up to approximately 0.5 watts. To have more than two transistors in one enclosure could result in wild oscillations and should be avoided. The intermediate frequencies of both mixers were placed inside the 10m band. L3/4 consist therefore of 29 turns (4mm internal diameter). C5 should cover about 10 to 50pF. The arrangement for the 70cm band is described in full detail in D.4.1 whereas the 23cm version was test-built on part of a printed circuit board according to "UKW-Berichte". In this case, however, the open construction as per D.4.3 is recommended.

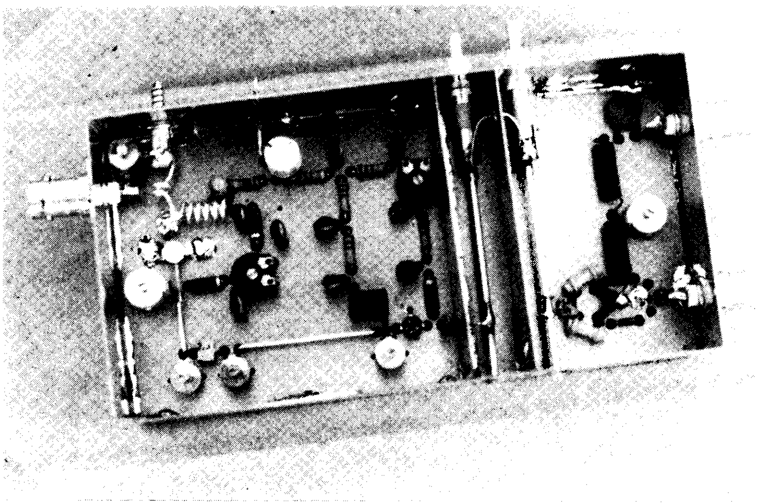
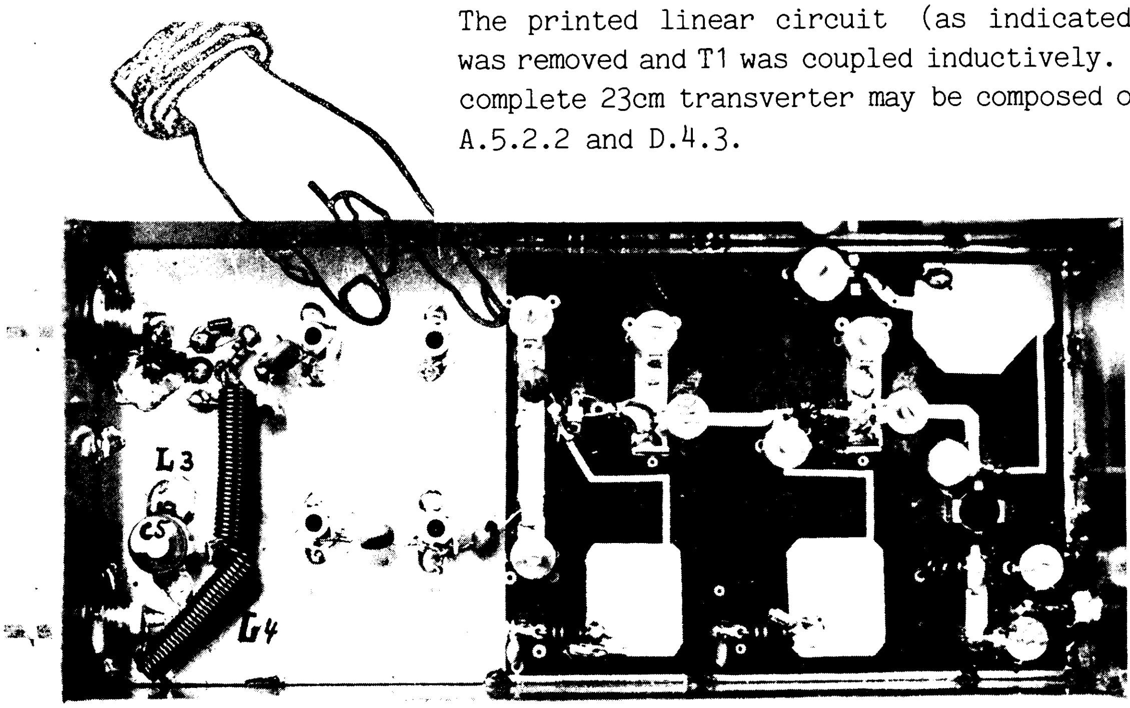


Figure 64
shows the described 70cm mixer as part of the transverter concept according to D.4.1.

Figure 65
Here a test model for the 23cm band is shown. The printed linear circuit (as indicated) was removed and T1 was coupled inductively. A complete 23cm transverter may be composed of A.5.2.2 and D.4.3.



A.5.3 Double-balanced mixer - general remarks

Double-balanced mixers have developed into standard components of communication systems, microwave modules and spectrum analyzers. When employed correctly they allow the amateur to achieve minimized distortion with a high degree of isolation from unwanted signals. The performance of a receiving system can severely deteriorate due to incorrect application.
A deeper understanding of the factors that give rise to interference and distortion as well as the knowledge of suitable routines for measuring the performance will aid the correct use of double-balanced mixers.
The following section will discuss briefly the theory of operation.

We shall start with the balanced mixer.

The balanced mixer shows good isolation between the RF-input and the oscilator input; but the isolation between RF- and IF-ports is poor. Let us first consider the circuit diagram:

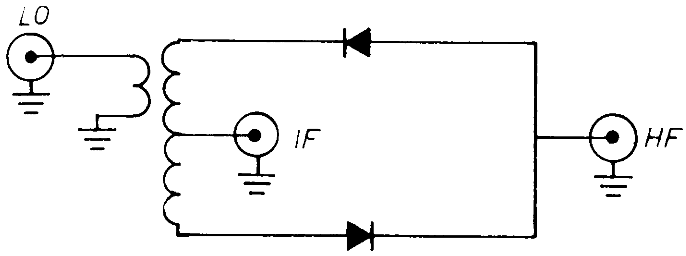


Figure 66
This presents a balanced mixer with good isolation between LO and RF ports but poor isolation between RF and IF.

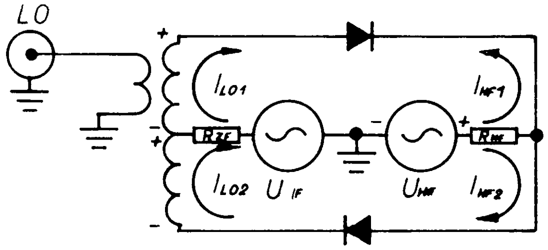


Figure 67
Balanced mixer. Voltages and currents according to the description.

Consider the polarity of the input transformer as shown. The directions of the Lo-currents would then be as shown for ILo1 and ILo2. The total current through the RF- and IF-sources is the sum of currents Lo and from the RF port.

Operation with applied signal:
The polarity (as indicated at the secondary windings of the Lo-transformer) will cause currents ILo1 and ILo2 to flow in counter-clockwise direction through the internal resistances of the IF- and RF-ports. Close examination of the current through the RF- resistance reveals complete cancellation of ILo assuming identical phase and amplitude of the ILo components 1 and 2. The same applies when considering the ILo current through the IF-port internal resistance. Again we observe complete cancellation. No Lo-power can therefore build up at either the IF- or the RF-ports. The balanced mixer will thus completely isolate Lo from both IF and RF ports.

What happens when a RF-signal is applied? The assumed polarity as indicated for the RF-port will give rise to the two currents IHF1 and IHF2. Both will add upon passing the IF-port and no cancellation will occur. A balanced mixer will thus provide no isolation between the RF- and IF-ports.

What is the situation between the RF- and IF-ports? As IHF1 will pass through the upper part of the secondary winding in opposite direction to current IHF2 through the lower part the magnetic fields will cancel and no voltage will be induced across the primary and the Lo-port. We therefore have separation between the RF- and Lo- ports.

The above mentioned considerations are based on the assumption that all current components have identical phase and amplitude. In all practical cases these values are somewhat influenced by

- variations in the transformer symmetry and
- unequal diode impedances

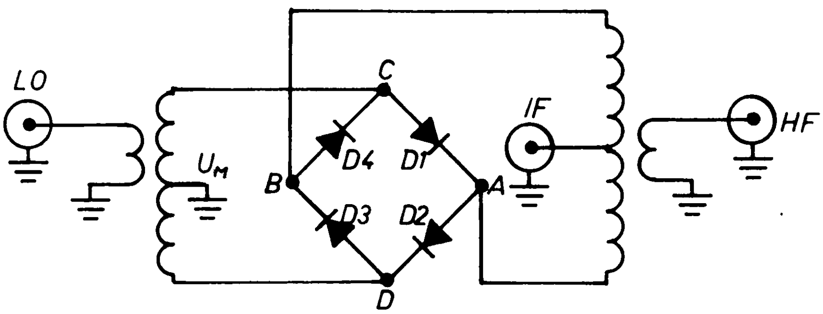
Beyond 1 GHz the capacitances of the wiring, the transformer windings and the physical layout will have an unfavourable influence on the balance. The symmetry will deteriorate with increasing frequency. For that reason manufacturers usually specify lower isolation values in their data sheet for higher frequencies.

Double-balanced mixers

A typical circuit diagram of a double-balanced mixer is shown in the following figure:

Figure 68

Typical circuit diagram of a double-balanced mixer.



Double-balanced mixers can provide improved values of isolation.

How is the balance achieved?

If both diodes D1 and D2 as well as the two secondary windings of the transformer are symmetrical then the voltage at point A will equal the voltage at the transformer center tap - namely zero. The same applies for diodes D3 / D4 and the point B. For that reason no voltage may pass that way towards the RF- and IF-ports.

How is the isolation between the Lo-, RF- and IF-ports achieved?
RF-input considerations: Let us assume that D4 equals D1 and D2 equals D3. In that case the voltages at points C and D are equal and no RF-signal will appear at the Lo port. Taking symmetry into consideration the same argument will hold for the IF-port; voltage U_{IF} is zero and no RF-signal may build up. The simplified diagram will help to clarify the argument:

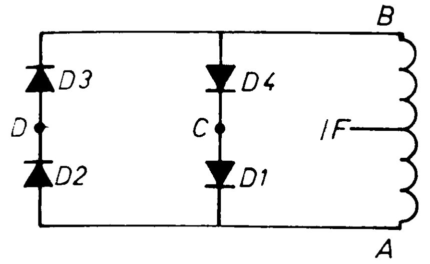


Figure 69
Simplified presentation of the double balanced mixer.

Again symmetrical transformer windings and identical diode parameters are assumed.

Optimized use of the mixer.
Imbalance of mixers and the resultant loss in isolation is caused by variations in diode depletion-layer capacitance and differences in the transformer windings etc.

Rule-of-thumb:
Increasing the operating frequency will cause the value of isolation to drop by about 5dB per octave.
Example: 45 dB of isolation at 150 MHz will leed to
40 dB of isolation at 300 MHz

Polarity of the RF-signal:
The properties of the double balanced mixer are independent of signal polarity.
Double-balanced mixers perform equally well as phase detectors, balanced modulators, amplitude modulators, pulse modulators and current controlled attenuators. A full description of all these applications is beyond the scope of this book. Merely the operation as attenuator will be described briefly.

The mixer as current-controlled attenuator:
Under normal operating conditions the double-balanced mixer (DBM) will exhibit excellent isolation between the RF- and Lo-ports due to its symmetrical arrangement. The theory of operation has been fully covered. If the balance is disturbed by applying a dc- voltage to the IF-port (which results in lowering the isolation) some fraction of the RF will pass from one port to the other. The degree of attenuation may be controlled by varying the dc-voltage applied to the IF-port. A current of 10 to 20mA may reduce the attenuation (insertion loss) between the RF- and Lo-ports to 3dB. Fast switching and control of attenuation is possible since this property is maintained from dc up to very high frequencies.

Double balanced mixers will suffer from saturation effects beyond a certain signal level (different value for each type produced by a manufacturer). This will lead to signal distortion. One possible method of measuring this effect is the application of two RF-signals to the mixer input port. We hereby come across a new definition:

Intercept point.

Two-tone third-order intermodulation distortion is a measure of the third-order products that are caused by a second signal applied simultaneously to the mixer input and appears at the mixer output. A common method to describe the ability of a mixer to suppress unwanted signals is to define the intercept point.
This is a theoretical point, describing the RF-signal level that results in equal power of signal and third-order products at mixer output.

It is convenient to describe the intermodulation products with respect to the input by stating the difference between the values in dB (relative). Example: A mixer may be specified with 60dB for two - 20dBm input signals. This means that this mixer will suppress the third order products by 60dB when two -20dBm signals (see G.3.1) are applied to the mixer input. Reducing the input signals by 10dB will reduce the third order products by three times that amount, i.e. 30dB. **The difference is 20 dB.** Two input signals of -30dBm would therefore yield 80dB of third-order suppression. Reducing both input signals to -40dBm (which is a further 10dB reduction) will result in 100dB suppression of third-order intermodulation products.

At what level would the signal power equal the third-order intermodulation products?

We started with two signals of -20dBm and third-order products down by 60dB (i.e. -80dBm). Raising the input signals by 30dB to +10dBm will triple the intermodulation products corresponding to 90dB. Add to this the basic level of -80dBm and the result will be 10dBm (10mW). Under these conditions desired signal and undesired products are equal.

All this may sound rather theoretical so a short hint towards the practical implication may help:
In exposed locations, in the vicinity of strong stations or as a result of excessive preamplification (see section D) signal levels may grow rather rapidly. Very quickly a situation can arise, where two signals of 0dBm or above exist at the mixer input. If intermodulation products are only 20db down then false signals with an S-meter reading of only 20dB below the desired signal will appear. Interfering signals at far-off frequencies may be suppressed by filter arrangements according to section C. The only protection against strong interfering signals inside the band, however, is a receiver concept with a sufficiently high intercept point.

Use the following rule-of-the-thumb to estimate the level of intermodulation products:

1. Obtain the 1dB compression point according to A.4.3.
2. Estimate the intercept point; near the lower end of the frequency band as specified by the maker it will be located about 15dB and towards the upper specified frequency limit roughly 10dB above the 1dB compression point.
3. Multiply the difference (in dB) between the intercept point and the RF-input level by the order of the harmonic in question.
4. Subtract the result from the intercept point to obtain the intermodulation level.

Example: Given: 1 dB-CP = + 1 dBm RF-input level
RF-level = - 10 dBm (lower end of frequency band)
Find the level of the third order intermodulation products!

Solution: 1. The 1 dB compression point is specified at +1 dBm
2. Intercept point at 1 dBm + 15 dB = 16 dBm
3. +16 dBm - (-10 dBm) = +26 dB
26 dB x third order = 78 dB
4. 16 dBm - 78 dB = -62 dBm
The third order intermodulation products have a level of -62dBm in this particular case.

For further illustration measurements in the form of photographs by "Mini-Circuits Laboratory" (from 1976) are presented.

Typical intermodulation products of double-balanced mixers under two-tone test conditions with varying input levels

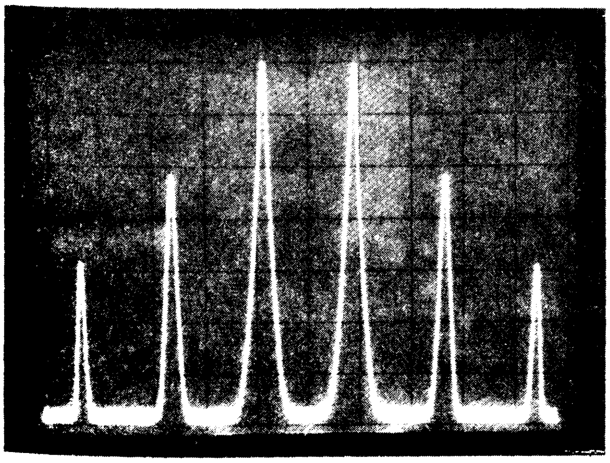
RF 1	abt.	180 MHz
RF 2	abt.	181 MHz
f_o	abt.	200 MHz
f_z	abt	20 MHz
and	abt.	19 MHz

Scale of all photographs:

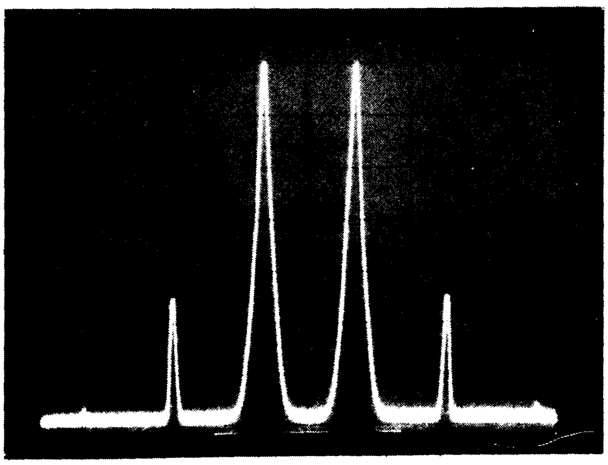
h = 0.5 MHz per line
v = 10 dB per line

Figure 70 Input signal level (of each RF-signal)

Ø dBm = 1 mW



-10 dBm = 0,1 mW

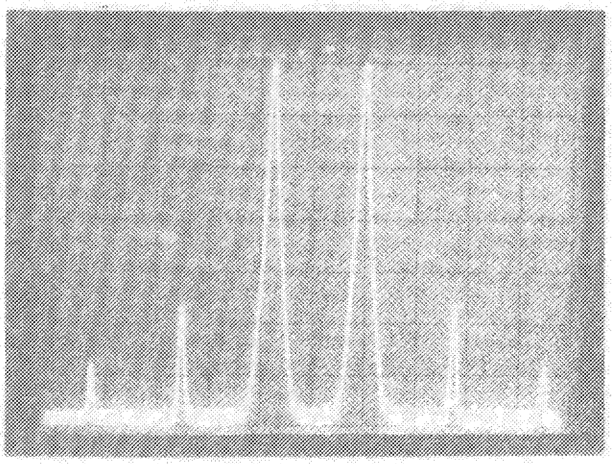


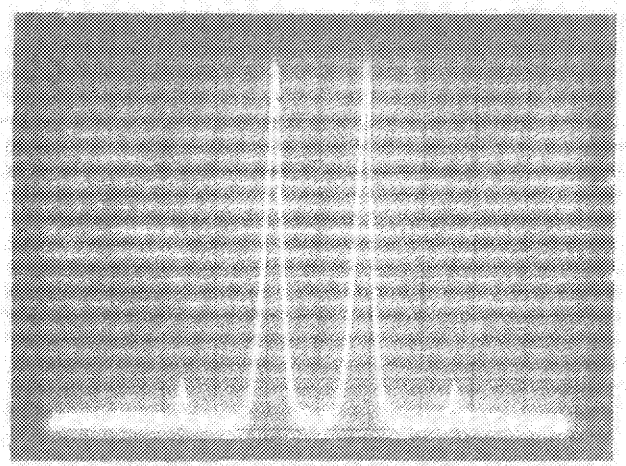
Modell: SRA-1 Pf_o = + 7 dBm = 5,01 mW

Figure 71

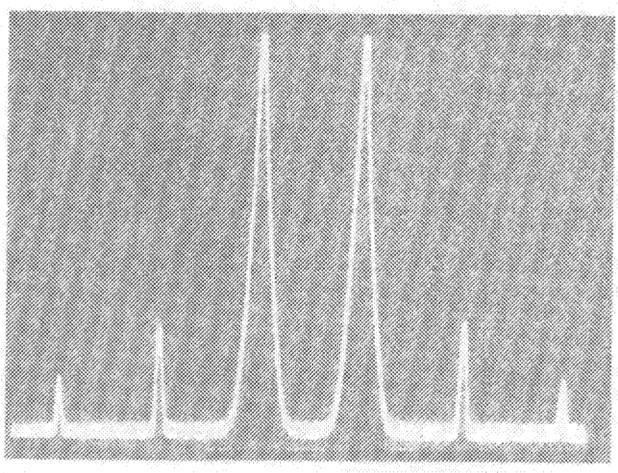
RF-level : Ø dBm = 1 mW

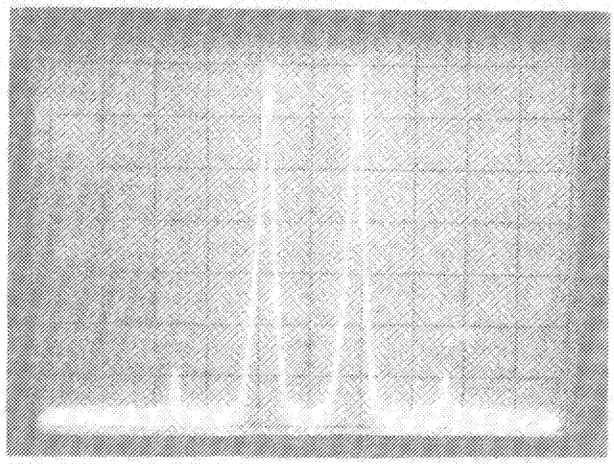
RF-level : -10 dBm = 0,1 mW



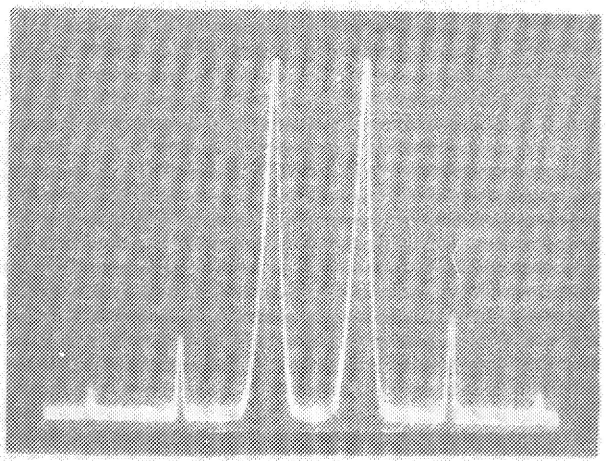


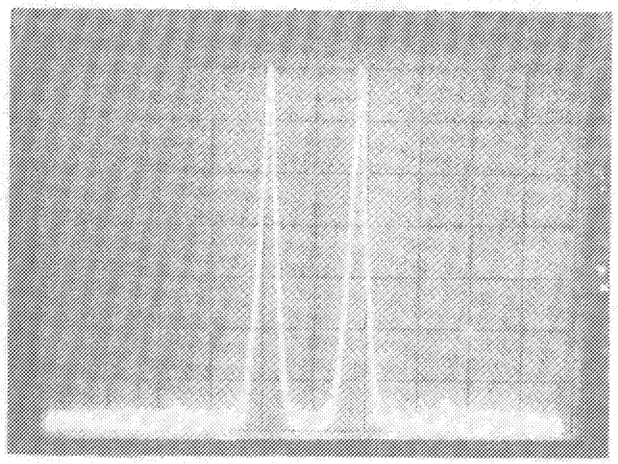
Modell: SRA-1H Pfo = 17 dBm = 50,1 mW





Modell: SRA-2H Pfo = 17 dBm = 50,1 mW



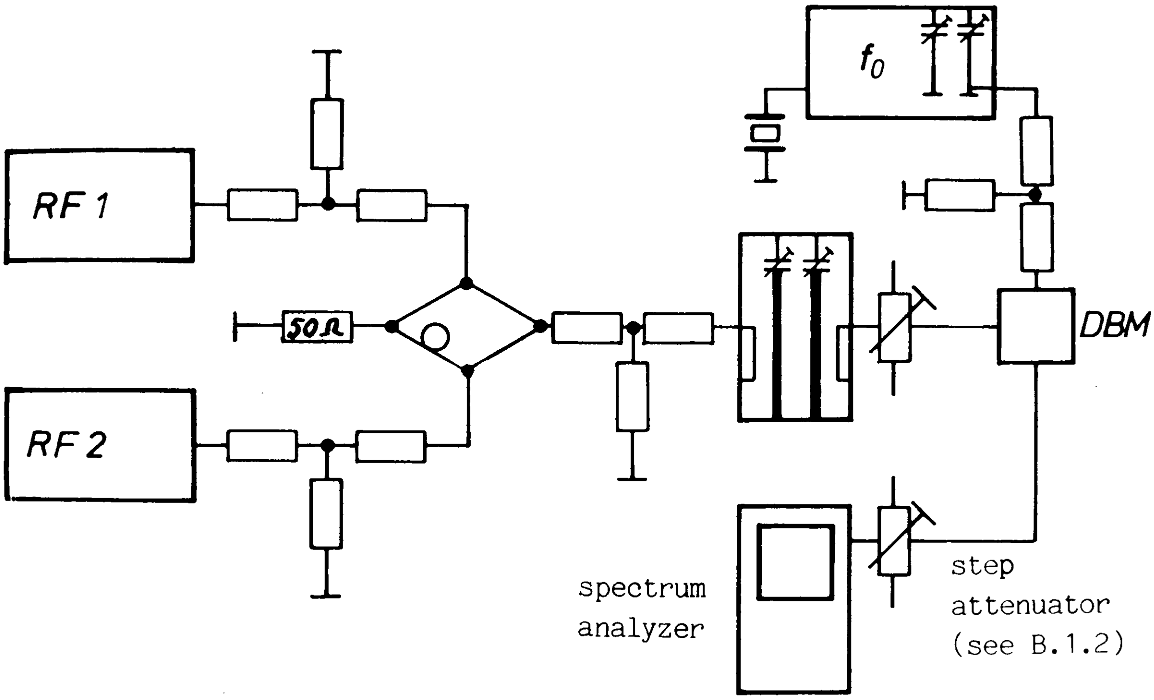


Modell: RAY-2 Pfo = 21,5 dBm = 141 mW

Test circuit for measurement of two-tone intermodulation distortion

After summing the two signal powers a low pass or band pass is required to prevent harmonics or spurious signals produced by the two generators from reaching the DBM. The power summing circuit will also help to decouple the two generators. Poor isolation between the two generators could cause harmonics and high order distortion that might be interpreted as distortion products generated by the mixer. Furthermore all modules must be separated by attenuators to ensure perfect 50 Ohm termination. Failing to do so could result in readings that differ by up to 20dB from the correct values.

Figure 72 Test circuit



A fixed-frequency signal genarator is described in B.8; tight coupling will allow the drawing of about 1mW of power.

Power splitters (and -adders) are described in B.1.3 and D.5.9.
Section C. covers the necessary filters.

For the f_0 generator see section D.6.
If no spectrum analyzer is available (hi) then a panoramic receiver according to B.5 (in conjunction with the switching attenuator in B.1.2) will permit the estimation of the result.

Reference:
MICRO WAVES PRODUCT DATA DIRECTORY

A.5.3.1 Double-balanced mixer for the 23 cm band

Having presented balanced mixers for 23cm in A.5.2 and the theoretical background of double-balanced mixers in A.5.3 we are now going to describe a module as built by DK1FI.

Again this double-balanced mixer may be used in both transmitting and receiving equipment. Its compact modular design ensures great versatility. No circuit details are given, as the mixer is described elsewhere as part of the 23cm transverter concept (section D.4.3). Selectivity at both input and output is provided by half- wavelength tuned circuits. Depending on the choice of intermediate frequency (28 or 144MHz), L1 is tuned to the f_o of either 1152 or 1268MHz. L2 is tuned to the signal frequency (f_n) of 1296MHz. Injection of f_o as well as extraction of f_n is performed inductively by coupling loops which are positioned roughly in the middle of the half-wave lines. There are individual screened cavities for either frequency. The IF-signal is supplied via a BNC-connector at the rear.

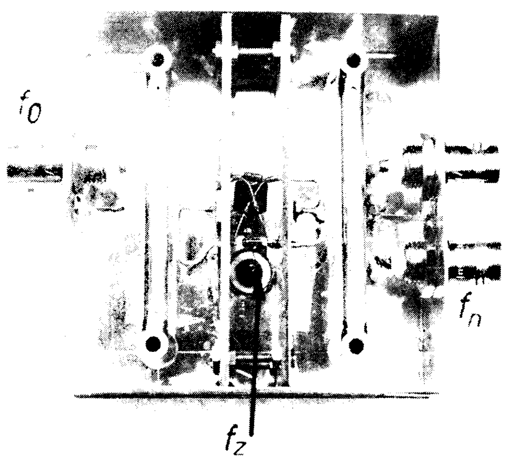


Figure 73

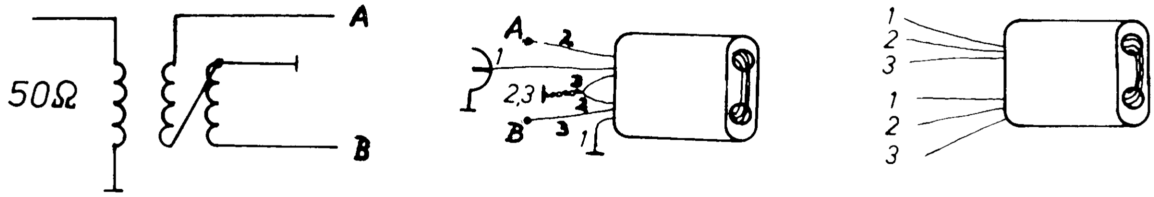
Total view of the mixer in its case with the dimensions:
72 x 72 x 30 mm

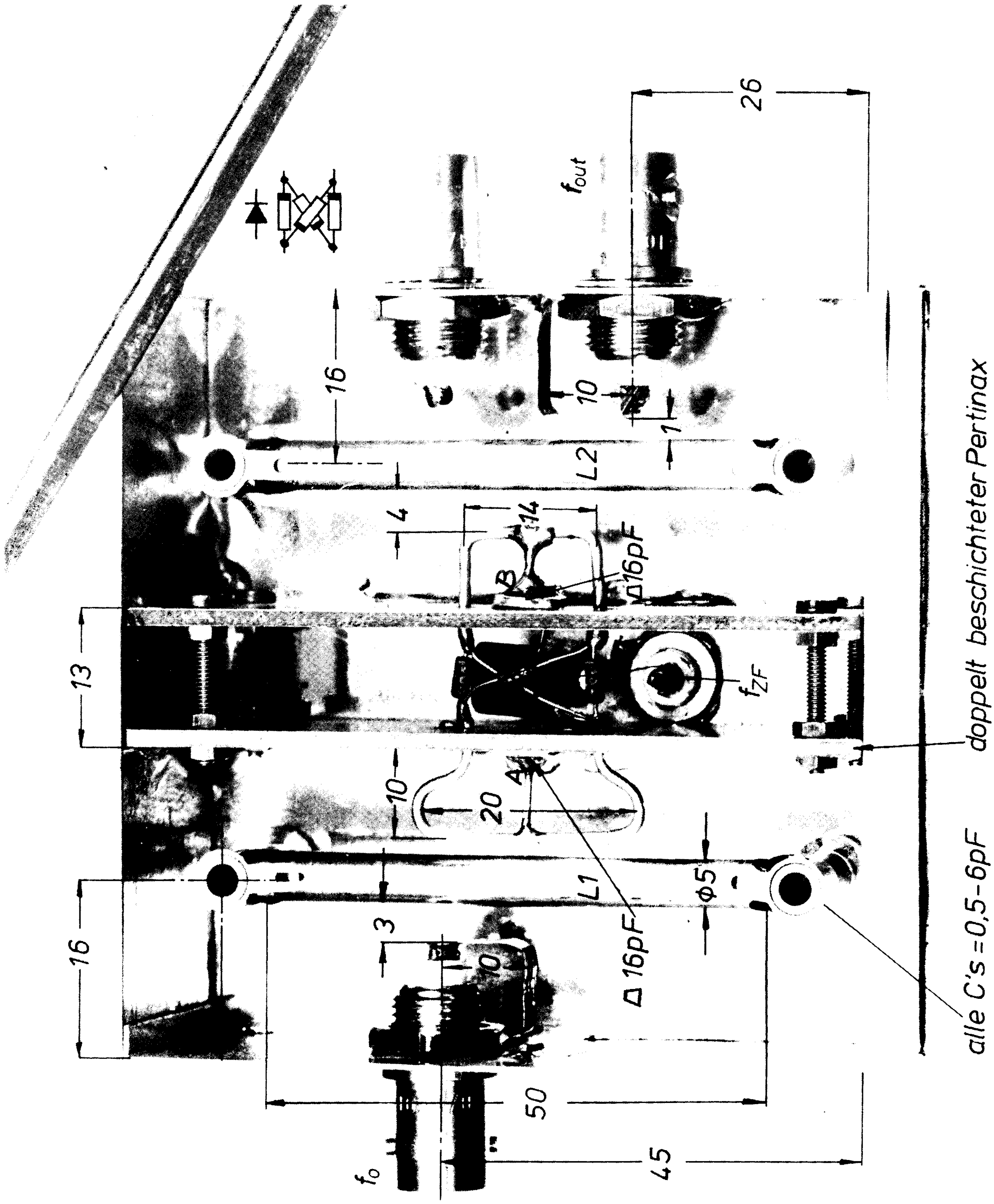
The second connector at the upper right corner allows the module to be used for both receiving and transmitting purposes. See D.4.3.

Figure 74

Rear view showing IF- connector and tuning capacitors 0.5-6 pF
The BNC-connectors are mounted half-way up the side walls.
Any material suitable up to 144MHz may be used for the two-hole tranformer core.
Examples are:
Neosid F10B 14.8.8 mm or equiv., such as balun in antenna input of old television set.

Figure 75
Enlarged view of the 23 cm band mixer in its case measuring 72x72x30 mm giving all relevant dimensions. The detail drawing of the transformer shows only one turn per winding. Three pieces of wire (0.3 mm Cu-enamel) are wound onto the core in parallel and its ends interconnected as shown.





A.5.4 Balanced mixer in stripline technique (stripline-mixer)

General information
Reference (1) describes an easy to build and reliably operating mixer in stripline technique.
In case of further interest the reader is referred to (2) which gives - amongst other things - a full treatment of the theoretical background.
The coupling network of a mixer consists of a) line circuits, b) resistors, c) directional couplers or d) toroidal core transformers.
The phase relations are obtained through ring- or bridge circuits. Special-purpose mixers with large dynamic range may contain eight or more diodes, whereas four diodes will suffice in standard versions. The choice of diodes is ruled by noise factor and efficiency. Types like the 1N21 used to be popular in amateur equipment; nowadays reasonably cheap Schottky diodes are being favoured. They warrant low mixing losses and tight tolerances when of the same type.
In balanced mixers the diode impedances change with the diode current which in turn is a function of applied oscillator level. The complex impedance of the IF-output port should be transformed to 50 ohms. To further reduce the intermodulation products this impedance should be the same for both the intermediate frequency and all unwanted products (sum frequencies). This matching will suppress all those signals that would otherwise be reflected from the mis-matched IF-port and generate new mixing frequencies. The necessary transformation can be achieved by means of L-, T- or Pie-networks.
Hybrid selectivity There are different coupling elements for feeding the RF- and oscillator signals to the mixer. Hybrid couplers split the applied signal into two signals with zero, 90 or 180 degree phase difference (see sections B.1.3 and D.5.9 on power splitting).
Although there exists a multitude of different circuits, only two arrangements lend themselves to amateur construction due to their simplicity:

- The 90 degree Hybrid (quadrature coupler) and
- the 180 degree ring hybrid (rat-race).

These types of coupling are employed predominantly in balanced mixers, power splitters and summing networks.

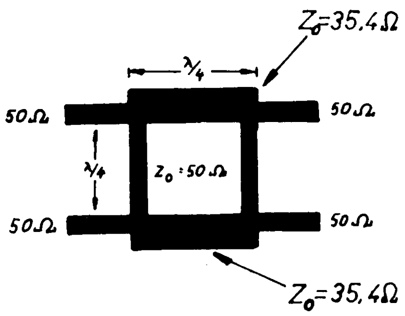


Figure 76
Quadrature coupler and its impedances at the various ports.

For practical examples see sections D.2.4 and D.2.5

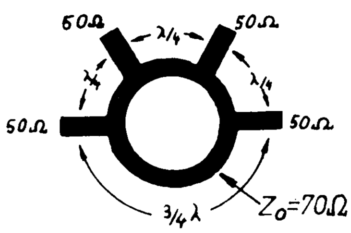


Figure 77
Six-section quarter-wave-length ring hybrid 180° coupler (rat-race).

Practical example in section D.4.2

References:
(1) Rat-race balanced mixer for 1296 MHz (WA6UAM) "hr" magazine July 77
(2) Unger/Harth "Hochfrequenz-Halbleitertechnik" Hirzel-Verlag Stuttgart

A.6 Active UHF-mixers.

The availability of components may lead to concepts that are not up to the state-of-art. This section is going to present several mixers that can be built with components from the scrap-box.

A.6.1 70 cm mixer employing germanium transistor AF279S

Not all that long ago mixers in TV-sets used to be based on germanium transistors AF239, AF279, AF280 or AF379. So these types may be found in many a scrap-box. They are particularly useful for the newcomer since they allow the construction of a mixer that is simple to build, that requires only few components and still fulfils the purpose of opening the door to the 70cm band.
When preceded by a preamplifier a very sensitive 70cm converter will be the result.
The mixing transistor is arranged in grounded base circuit. A transistor of the AF279S type (or equivalent) is suggested. The noise level of this mixer is of little importance since the overall noise figure is defined by the preamplifier stage. The indicated method of injecting the local oscillator signal (P_O) has proved stable and reliable. The 50 Ohm resistor terminates the oscillator supply cable. The oscillator signal is fed via the variable capacitor (6pF) to the emitter together with the RF- signal. The IF signal (28-30MHz) is singled out from all the mixing products by the bandfilter. No further details are given at this stage since this mixer is part of D.2.1.

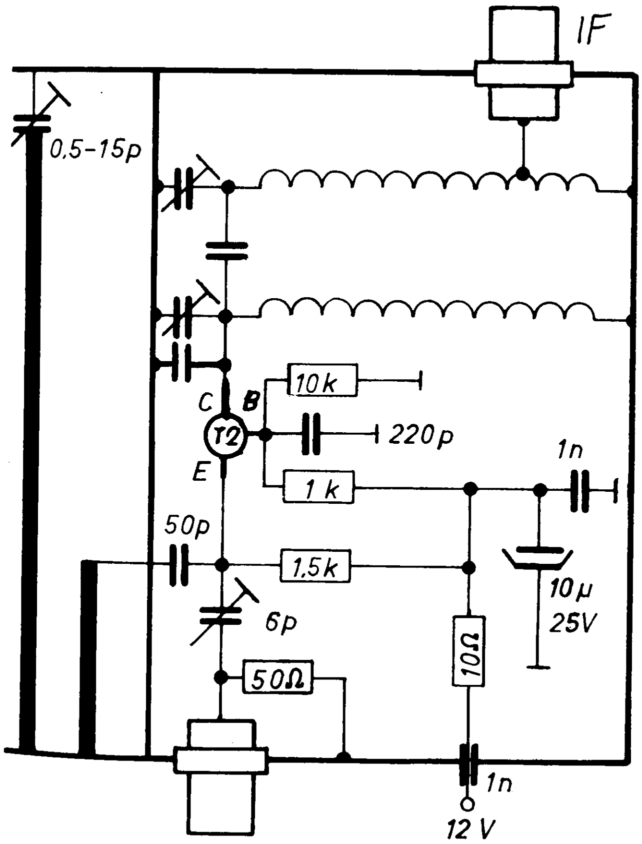


Figure 78
Circuit diagram of active mixer employing a germanium transistor of the AF279 AF280 AF379S family.

For further details see D.2.1

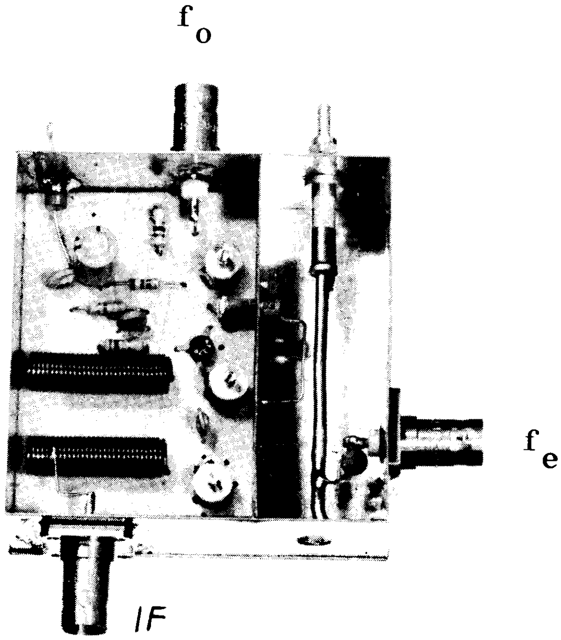


Figure 79
Construction of mixer in a case measuring 72x72x30 mm (see G.4).
In contrast to D.2.1 the mixer is based on a single side copper clad circuit board.

A.6.2 70cm mixer employing FET and dual-gate FET BF905

Before discussing the circuits the reader is referred to the mixer in D.2.2 utilizing the BFR34A. It may be incorporated into other concepts without a preamplifier (as shown there).
The following mixers will provide little selectivity. This could be improved by the addition of filters according to section C for the RF signal and by bandpass filters according to D.2.1 - D.2.3 in the IF channel.

The first circuit is a somewhat aged concept but is still well suited for experimentation with a FET capable of operating at UHF. With GaAs-FETs becoming available for amateur equipment this circuit arrangement might regain some of its popularity.

The second circuit originates from the winner of the 1979 BBT-contest (Bavarian field day). On request by the four responsible UHF-referees the station was checked for compliance with the contest rules and contained this mixer. OM Gerhard Schmitt (DJ5AP) offered the circuit diagram for publication in this Compendium (tnx dr Gerhard).

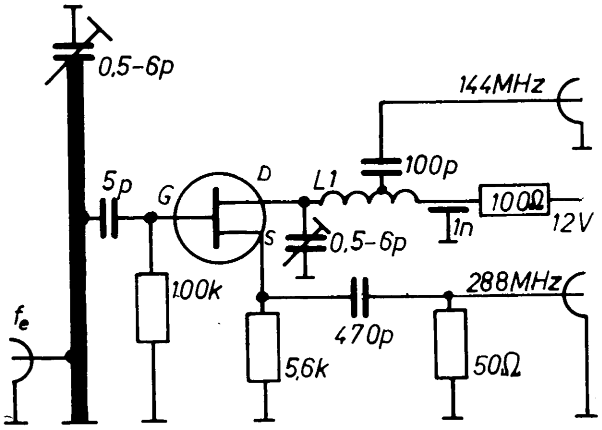


Figure 80
70 cm mixer with fieldeffect transistor capable of operating at UHF. (IF = 144 MHz).

L1 = 4 turns CuAg 1mm diameter

(Wound over a 5mm mandrel. Open construction, taped at 1 turn)

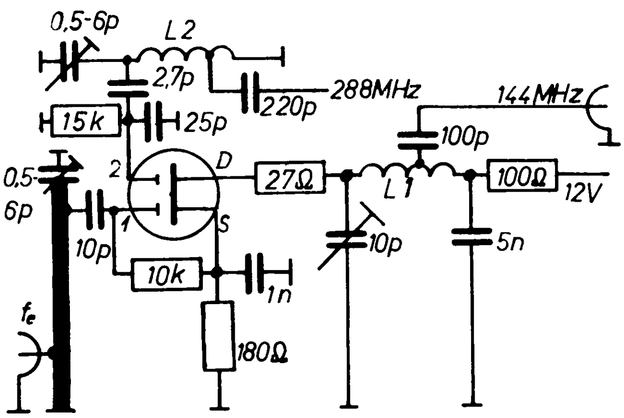


Figure 81
70cm mixer with BF905 according to DJ5AP (IF = 144 MHz)

L1 tuned to 144 MHz, 5 turns taped at 1 turn

L2 aprox.3 turns on 3mm mandrel taped at 1 turn

B Test-equipment and its construction

B.1 Pi- and T-section attenuators
(N. Schramm, DC9NI)

A. Introduction.

Attenuators are used to reduce the power level of high frequency signals and to decouple - repectively match - input / output impedances of interconnected stages.

a. **Defined attenuation**

Attenuators with precisely defined attenuation are required for the calibration of S-meters and for extending the measuring ranges of milliwattmeters and test receivers.

b. **Matching**

Attenuators with low values of attenuation improve both the matching of series connected transmitter stages and their electrical stability. The insertion of an attenuator between the preamplifier and the mixer of a receiver will improve its large-signal handling capability.

c. **Power reduction**

In case of excessive output power of a transceiver driving a power amplifier this power can be reduced by means of an attenuator. The resistors have to be rated for the power that must be absorbed.

The following fundamental aspects for the construction of attenuators should be born in mind:

1. Coaxial arrangement of components (for high precision attenuators).

2. Utilization of low inductance resistors suited for RF- application (do not use wire-wound or other types of resistors with the resistance material arranged in helical shape. Use composition [carbon] resistors instead).

3. Rating of resistors.

4. Attenuators should be identical when looking into either port.

5. Values of attenuation of more than 20 dB should be attained by series connecting several attenuators of lower attenuation value.

Figure 82 T-section attenuator circuit.

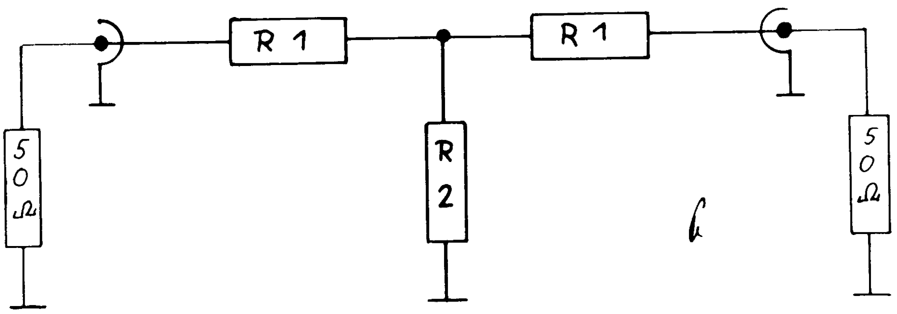
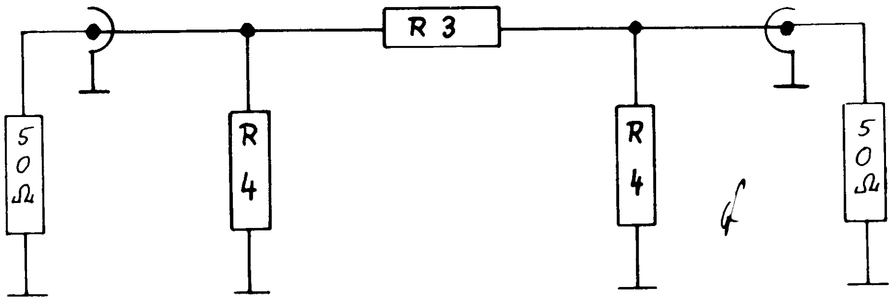


Figure 83 Pi-section attenuator circuit.



The choice of T- of Pi-section depends entirely on the resistors available to the amateur.

The following page lists the attenuation and the corresponding values of the resistances.

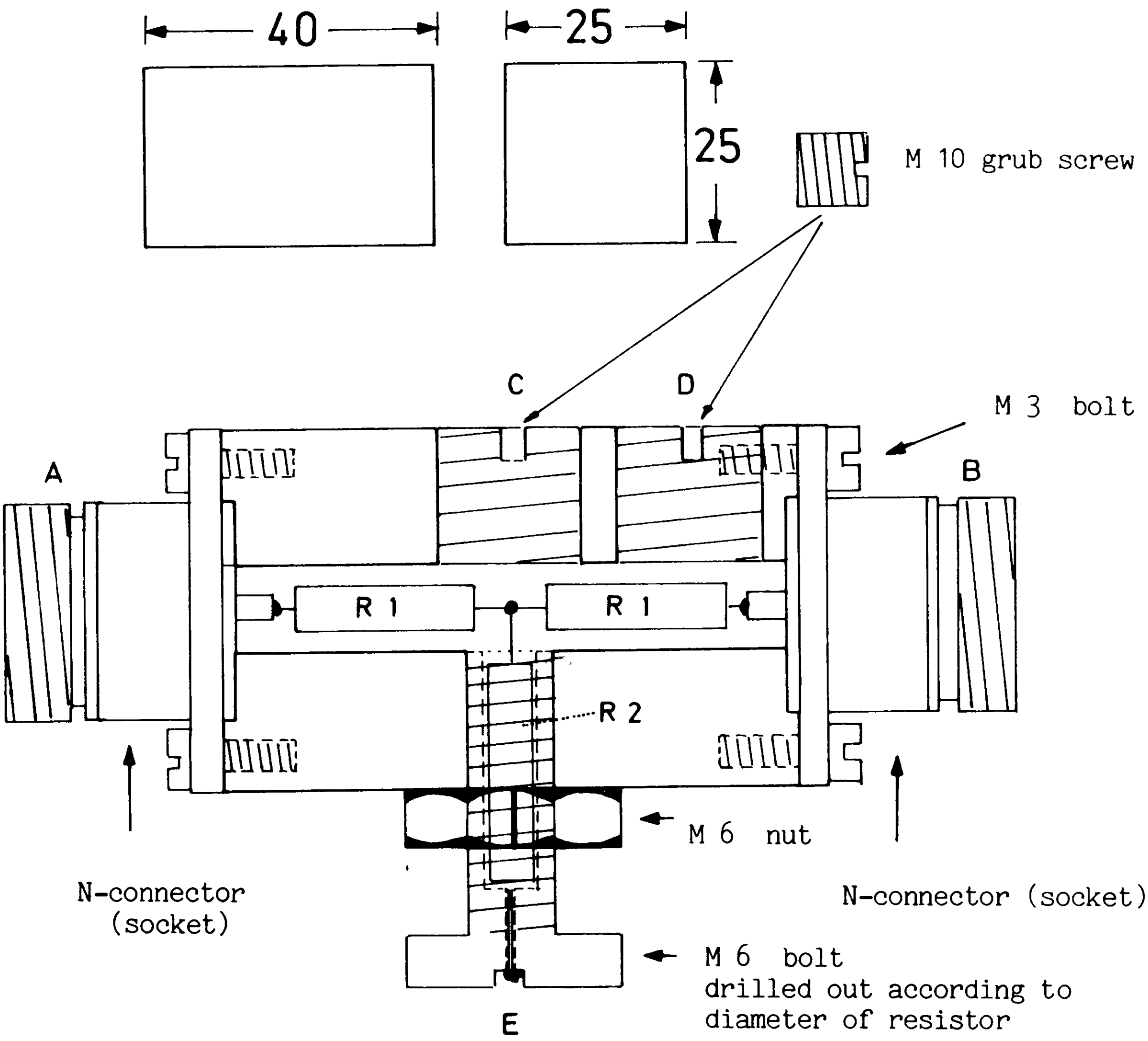
The following table lists the required resistance values (for 50 ohm impedance). The right-hand side of the table shows how approximate values may be obtained by combining standard resistors of the E 12 series

required attenuat.	calculated values [ohm] R 1	R 2	R 3	R 4	combination values [ohm] R 1	R 2	R 3	R 4
1 dB	2.9	433.3	5.8	869.5	12+12+ 12+12P	390+39S	12+12P	820+56S
2 dB	5.7	215.2	11.6	436.2	5.6	180+33S	22+22P	220+220S
3 dB	8.5	141.9	17.6	292.4	15+18P	120+22S	33+33P	270+22S
4 dB	11.3	104.8	23.8	221.0	22+22P	82+22S	47+47P	220
5 dB	14.0	82.2	30.4	178.5	39+22P	82	22+8.2S	180
6 dB	16.6	66.9	37.4	150.5	33+33P	33+33S	33+4.7S	150
7 dB	19.1	55.8	44.8	130.7	39+39P	47+8.2S	39+5.6S	120+10S
8 dB	21.5	47.3	52.8	116.1	22	47	47+5.6S	100+15S
9 dB	23.8	40.6	61.6	105.0	12+12S	22+18S	56+5.6S	100+4.7S
10 dB	26.0	35.1	71.2	96.2	47+56P	68+68P	56+15S	82+25S
11 dB	28.0	30.6	81.7	89.2	56+56P	15+15S	82	82+6.8S
12 dB	29.9	26.8	93.2	83.5	15+15S	27	82+12S	68+15S
13 dB	31.7	23.6	106.1	78.8	22+10S	12+12S	100+5.6S	68+10S
14 dB	33.4	20.8	120.3	74.9	33	10+10S	120	68+6.8S
15 dB	34.9	18.4	136.1	71.6	18+27S	18	270+270P	56+15S
16 dB	36.3	16.3	153.8	68.8	27+10S	33+33P	150	68
17 dB	37.6	14.4	173.5	66.4	27+10S	39+22P	155+22S	56+10S
18 dB	38.8	12.8	195.4	64.4	39	10+2.7S	180+15S	56+8.2S
19 dB	39.9	11.4	220.0	62.6	27+12S	22+18S	220	47+15S
20 dB	40.9	10.1	247.5	61.1	22+18S	10	220+27S	39+22S

P = shunt connection } { of resistances to obtain approximate value
S = series connection } { by combining standard resistors

B) Construction of a T-section precision attenuator

Figure 84



The design shown is suitable for frequencies up to the GHz range. When using common 0.5 watt composition resistors (of the Vitrohm type) the continuous rating will be around 0.7 watts or 3 watts for short SSB and CW transmissions. The rating may be doubled by filling the gaps between resistors and the mounting block with heat conducting paste.

List of materials:

1 off 25 x 25 x 40 mm brass.
2 off N- or BNC-connectors.
2 off grub screws M 10 x 10 brass-or bolts, heads sawn off and slotted
1 off bolt M 6 x 20 brass.
1 off nut M 6 brass.
3 off resistors, low inductance, according to table on previous page.

Order of construction:

a. Drill axial bore hole of 6 mm diam.

b. Drill transversal bore hole of 5 mm diam.

c. Open up one half of (b.) to 8 mm diam.

d. Cut M6-thread into 5 mm bore hole and M10-thread into 8 mm bore hole.

e. Drill bore holes for mounting studs into both faces according to chosen connector type. Cut M 3-thread into all holes and insert M 3 studs.

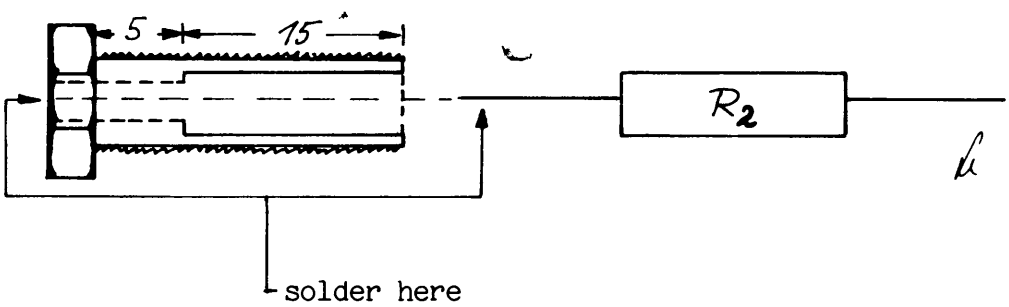
f. Drill 8 mm bore hole for plug D according to chosen connector. The hole must be positioned above the internal soldering lug of the coaxial connector. Cut M10-thread (see diagram).

g. Solder both resistors R1 to connector A. Install connector A by means of M3 nuts.

h. Install connector B. Solder R1 to connector B through plug hole opening D.

i. Drill out a M6 x 20 bolt (E): 3.5 mm diameter to a depth of 15 mm and to 2 mm in its upper part as indicated in the following diagram.

Figure 85



j. Insert R2 into this bolt and solder one end.

k. Screw bolt M6 x 20 into the block as far as necessary, then fasten by tightening the lock nut. Solder free end of R2 to the junction of the two R1 resistors. Attention: Cut R2 connecting lead very short!

l. Install plugs C and D.

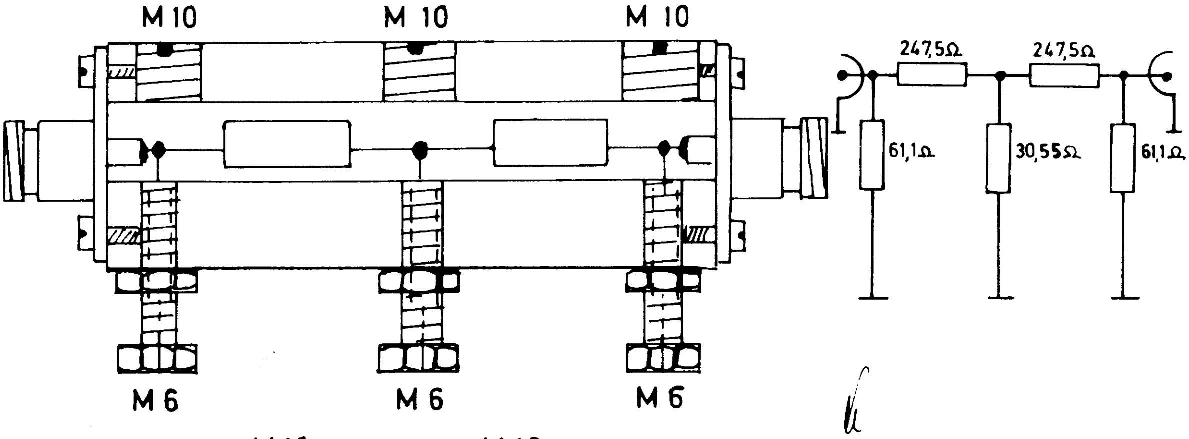
After assembly and prior to first use the attenuator should be calibrated by means of suitable test instruments. The attenuation should be measured (preferably for each frequency range of interest) and the attenuator labeled accordingly.

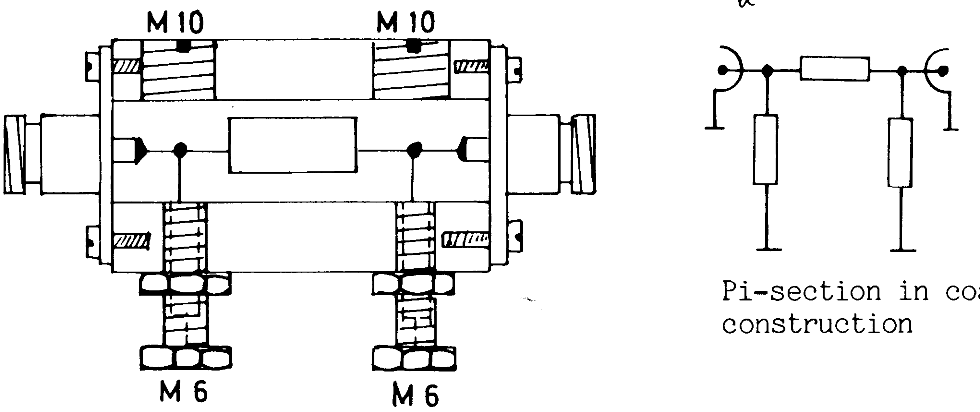
Suitable test equipment:
RF-millivoltmeter, spectrum analyzer or comparison against an attenuator of known attenuation.

Further examples of precision type attenuators:

Figure 86

40-dB Attenuator





Pi-section in coaxial construction

C) Attenuators employing BNC-connectors in tin plate casing

Attenuators of non-coaxial construction may be used up to 200 MHz without significant changes in attenuation. The latter is caused by inductances and capacitances of the resistor and its mounting components which will vary the effective value of resistance as function of the frequency (see section A.1). If an attenuator is required for the lower frequency ranges only, the following type of construction will suffice:

List of materials:
1 off case, tin plate (or brass), U-shaped, 18x18x40mm
2 off BNC sockets with flanges
3 off resistors according to table.

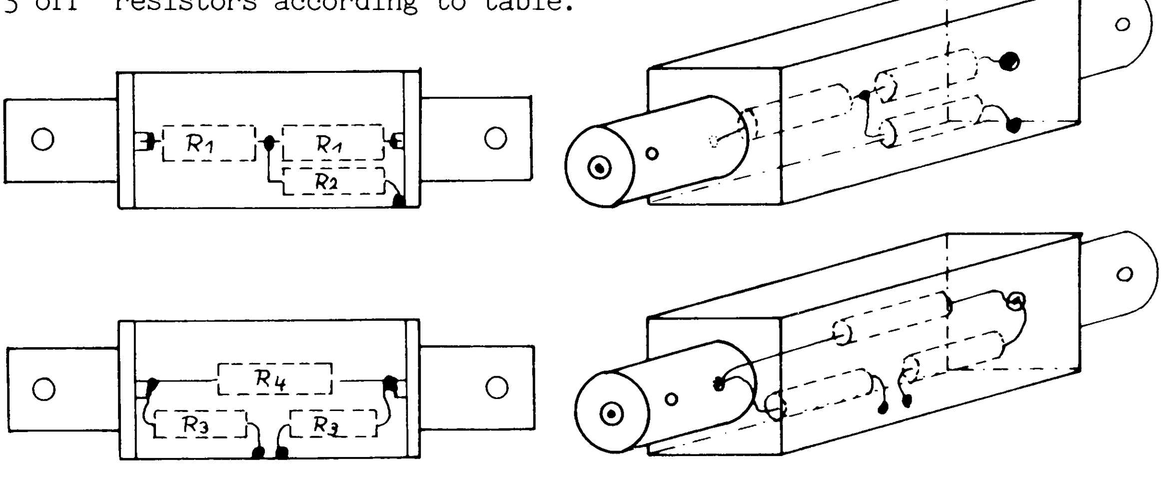


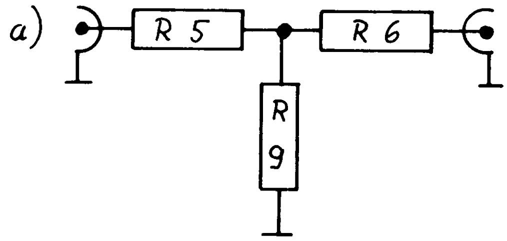
Figure 87

D) Attenuator of increased rating in tin plate casing (G.4)

A printed circuit board suitable for the standard casing is available which allows the construction of different types of attenuators depending on the required attenuation and the available resistors.
The following circuits are possible:

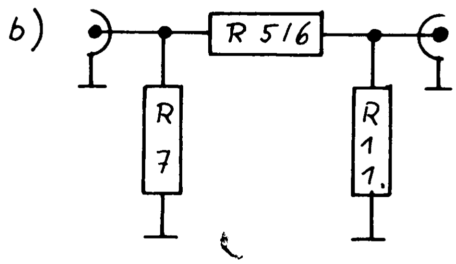
T-section:

Figure 88



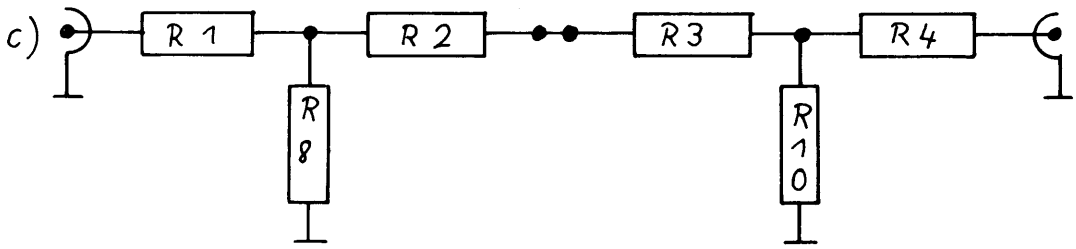
Pi-section:

Figure 89



Series connection of two T-sections:

Figure 90



Series connection of T- and Pi-section:

Figure 91

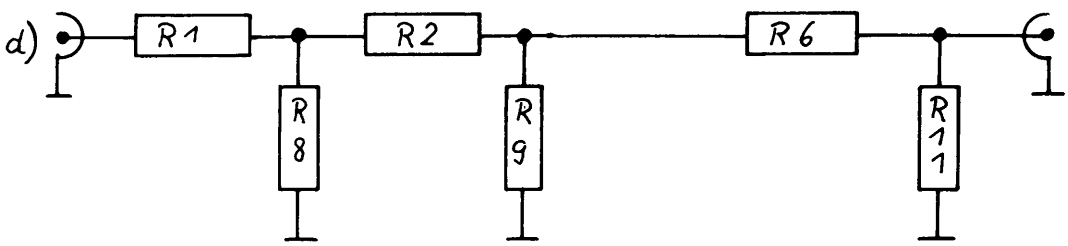


Figure 92 Universal printed circuit for attenuators

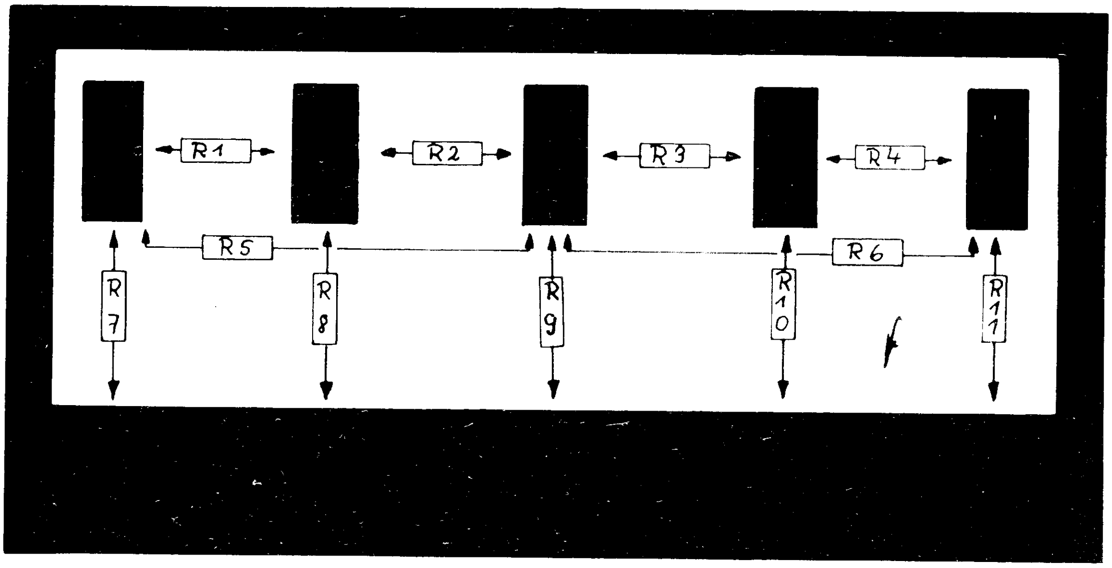
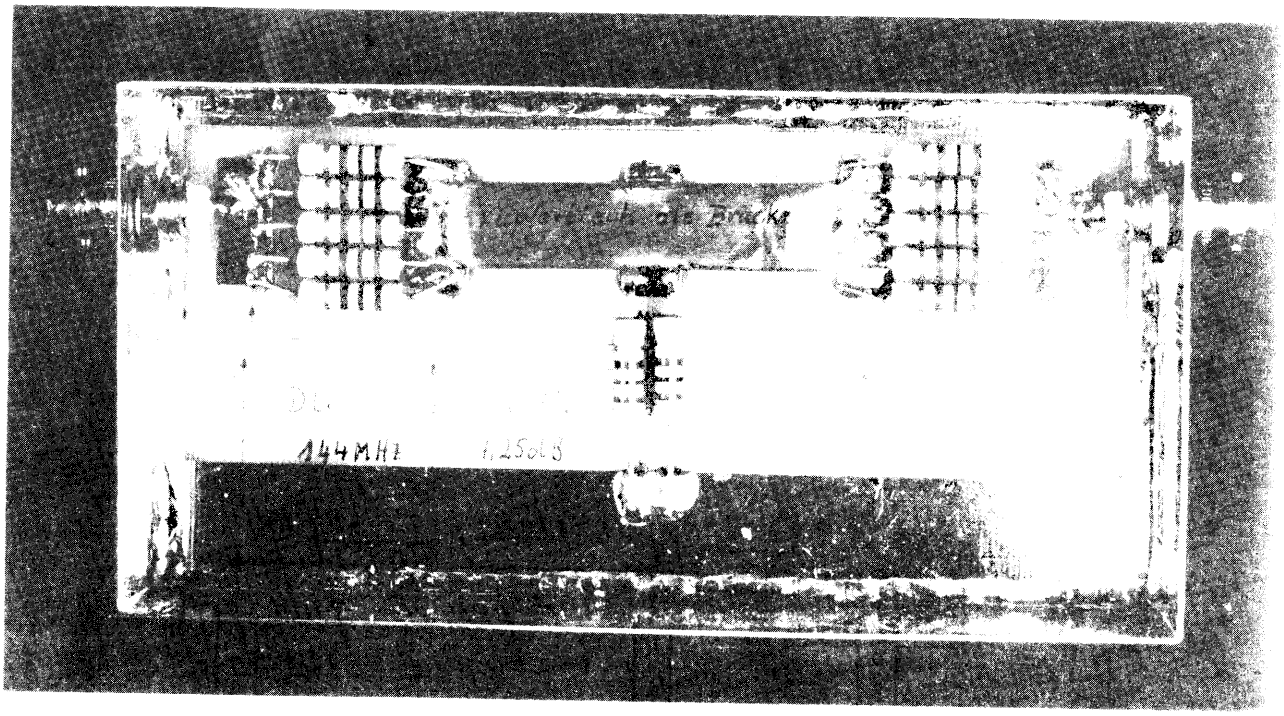


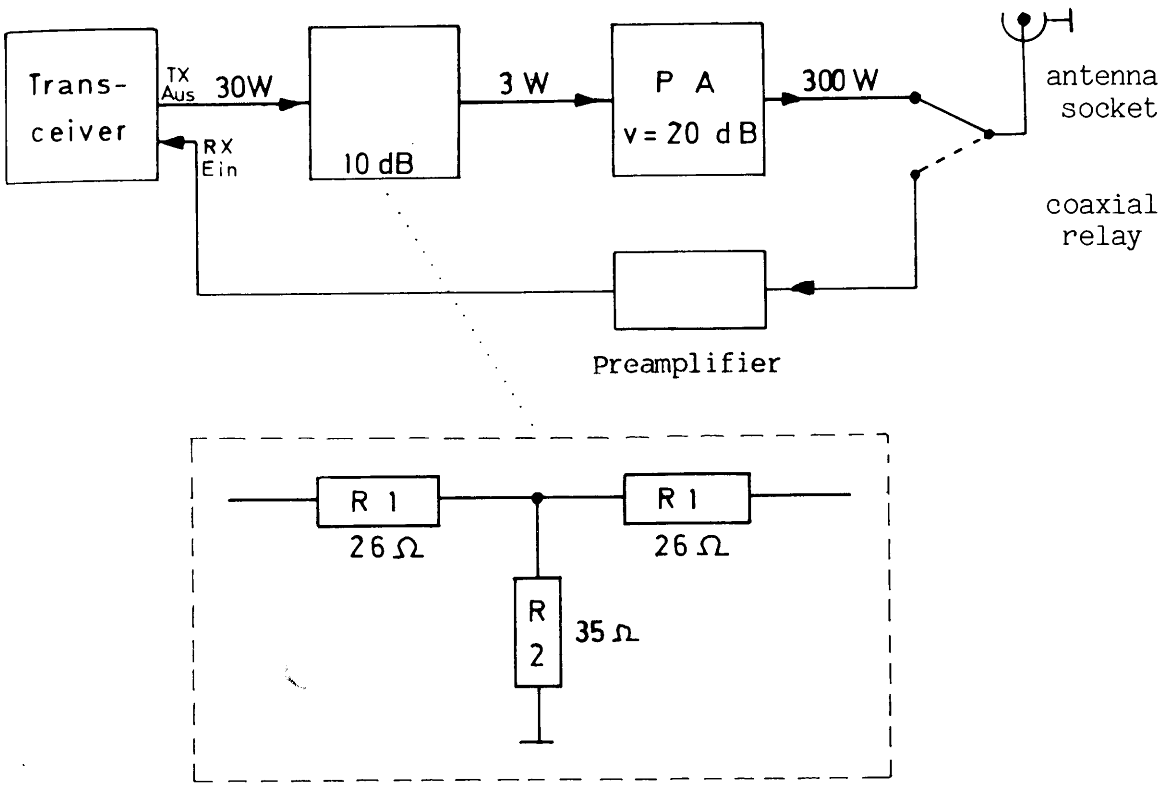
Figure 93 Practical example of attenuator in tin plate casing
dimensions: 146 x 72 x 30 mm.



E) Example for the use of a medium power attenuator

The block diagram shows the application of an attenuator. It is inserted between the driver and the power amplifier to reduce the power level.

Figure 94



Example:

Assume the outputpower of the transceiver to be 30 Watt. The driving power required by the PA is 3 Watt for full output. 30 Watt input power and 3 Watt output power require a 10 dB attenuator. For a T-section the table indicates R1 = 26 Ohm and R2 = 35 Ohm. From the calculations on the following pages the rating of the resistors may be obtained.

Result: for R1, P_{R1} = 15.6 Watt
for R2, P_{R2} = 17.05 Watt

Since the scrap box is not likely to contain resistors of 26 Ohm / 15.6 W and 35 Ohm/17.5 Watt they have to be combined from resistors with higher values and lower rating.

For instance 26 Ohm / 15.6 Watt = 10 resistors 270 Ohm / 2 Watt
35 Ohm / 17.05 Watt = 10 resistors 360 Ohm / 2 Watt

The slight deviation of the resultant resistance values and the higher rating is acceptable in view of the availability of the standard resistors.

F) Rating of attenuators

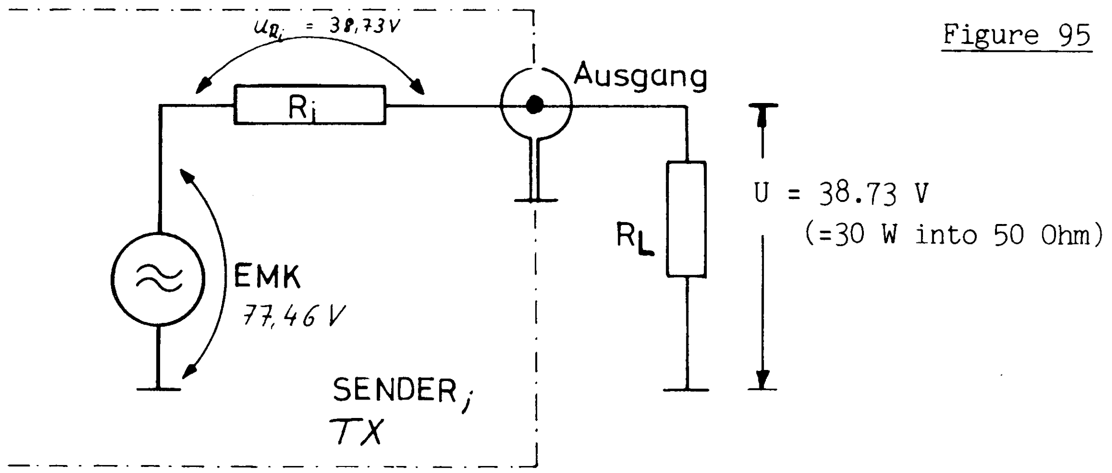
The internal resistance of a generator (transmitter) is responsible for variations in output voltage and current under changing load conditions. A thorough understanding is thus important before attempting any calculations.

Assumptions:

The generator (transmitter) has an internal resistance R_i of 50 Ohm and is capable of delivering 30 Watt into a load of 50 Ohm. This implies that the transmitter output voltage will be:

$$\boxed{U^2 = P \times R} \qquad U = \sqrt{30 \times 50} = 38.73 \text{ Volt}$$

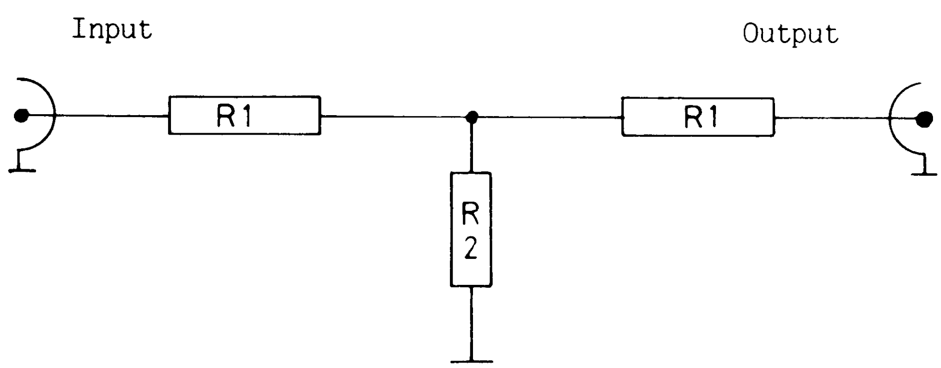
Figure 95



This voltage drop is caused by a current passing through the internal resistance R_i, too. Since $R_i = R_l$ the electro motive force inside the generator is twice the output voltage (power matching). A T-section is now connected to the output of this generator.

G) The T-section:

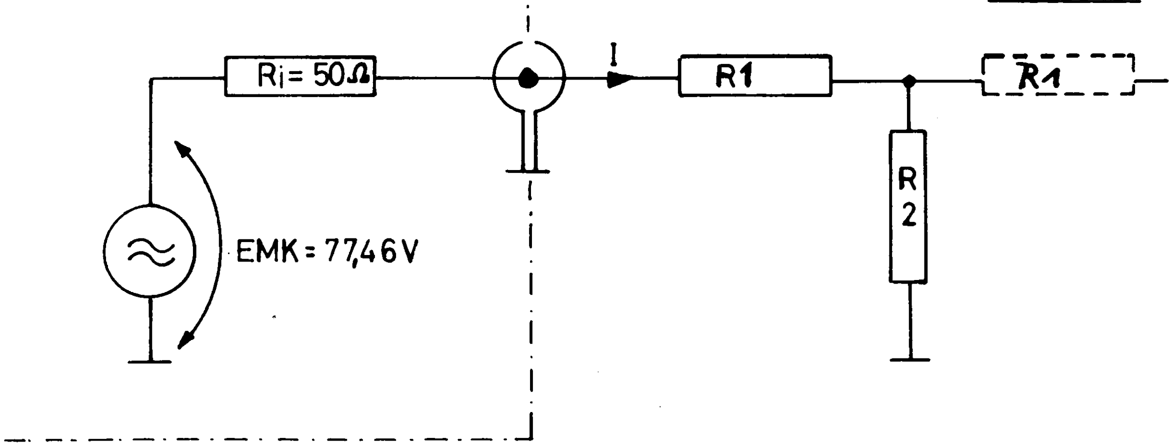
Figure 96



We consider two states for calculating the component rating:

1.) <u>The attenuator output is open-circuited</u> (R_1 towards the output is of no consequence).

Figure 97



Example: 10 dB attenuation, R_1 = 26 Ohm, R_2 = 35.1 Ohm

Calculation:

$$I = \frac{EMK}{R_i + R_1 + R_1} = \frac{77.46\ [V]}{50 + 26 + 35.1\ [Ohm]} = \underline{0.697\ [A]}$$

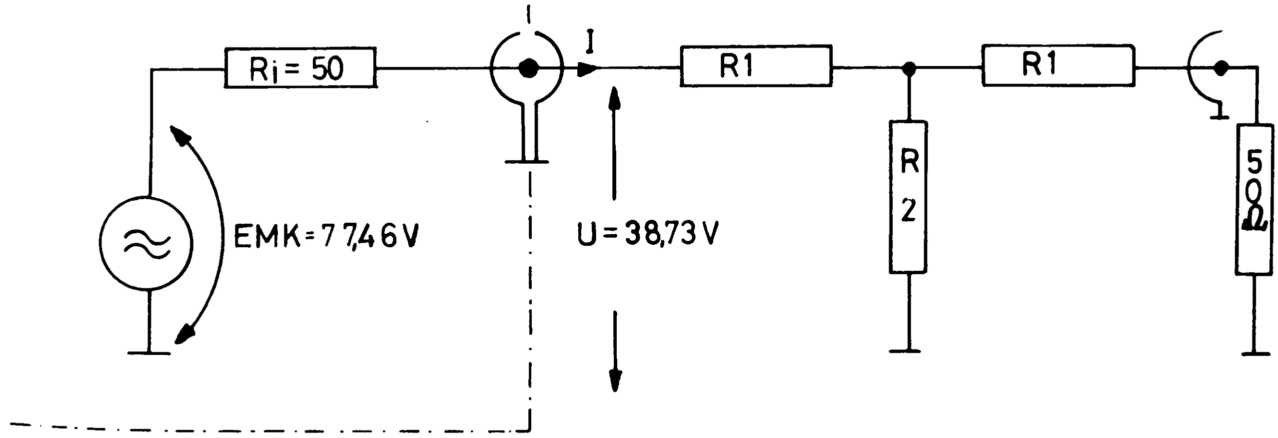
From this follows:

$$P_{R1} = I^2 \cdot R_1 = 0{,}697^2\ A \cdot 26\ Ohm = \underline{\underline{12{,}638\ W}}$$

$$P_{R2} = I^2 \cdot R_2 = 0{,}697^2\ A \cdot 35{,}1\ Ohm = \underline{\underline{17{,}05\ W}}$$

2.) <u>The output is terminated:</u>

Figure 98



With the attenuator output properly terminated the generator is (power-) matched to the attenuator/load combination and the generator output voltage will be one half of EMK i.e. 38.73 Volts in this example.

Calculation: $I = \frac{U}{R} = \frac{38.73\ [V]}{50\ [Ohm]} = \underline{\underline{0.775\ [A]}}$

$$P_{R1} = I^2 \times R_1 = 0.775^2\ [A] \times 26\ [Ohm] = \underline{\underline{15.62\ [W]}}$$

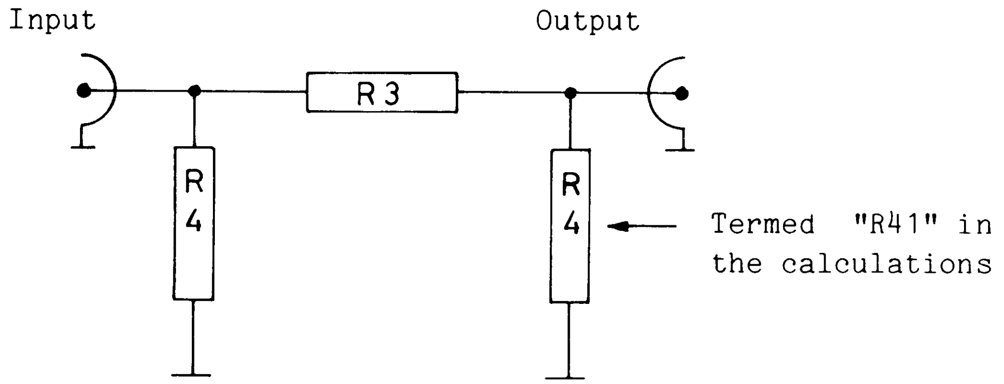
$$U_{R1} = I \times R_1 = 0.775\ [A] \times 26\ [Ohm] = \underline{\underline{20.15\ [V]}}$$

$$U_{R2} = U - U_{R1} = 38.73\ [V] - 20.15\ [V] = \underline{\underline{18.58\ [V]}}$$

$$P_{R2} = \frac{U_{R2}^2}{R2} = \frac{18.58^2\ [V]}{35.1\ [Ohm]} = \underline{\underline{9.84\ [W]}}$$

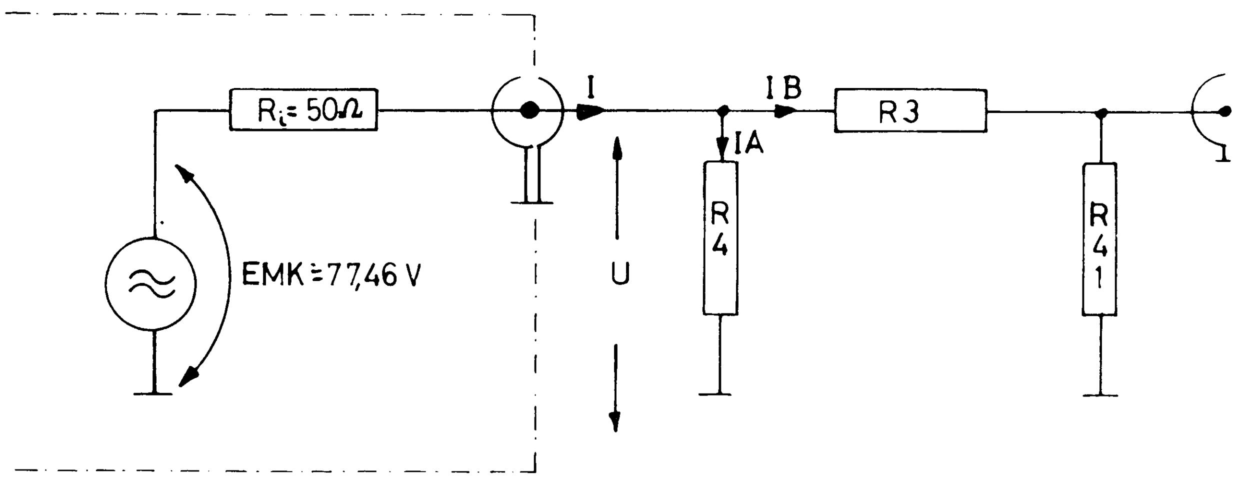
H) The Pi-section:

Figure 99



1.) The output is open-circuited:

Figure 100



Example:

10 dB attenuator,
R_3 = 71,2 Ohm
R_4 = 96.2 Ohm (referred to as "R41" in the calculations).

Calculation:

$$R = \frac{R4 \times (R3 + R41)}{R4 + R3 + R41} = \frac{96.2 \times (71.2 + 96.2)\ [\text{Ohm}]}{96.2 + 71.2 + 96.2\ [\text{Ohm}]} = \underline{\underline{61.09\ [\text{Ohm}]}}$$

$$I = \frac{EMK}{Ri + R} = \frac{77.46\ [V]}{61.09\ [\text{Ohm}]} = \underline{\underline{0.697\ [A]}}$$

$$U = R \times I = 61.09\ [\text{Ohm}] \times 0.697\ [A] = \underline{\underline{42.58\ [V]}}$$

$$P_{R4} = \frac{U^2}{R4} = \frac{42.58^2\ [V]}{96.2\ [\text{Ohm}]} = \underline{\underline{18.85\ [W]}}$$

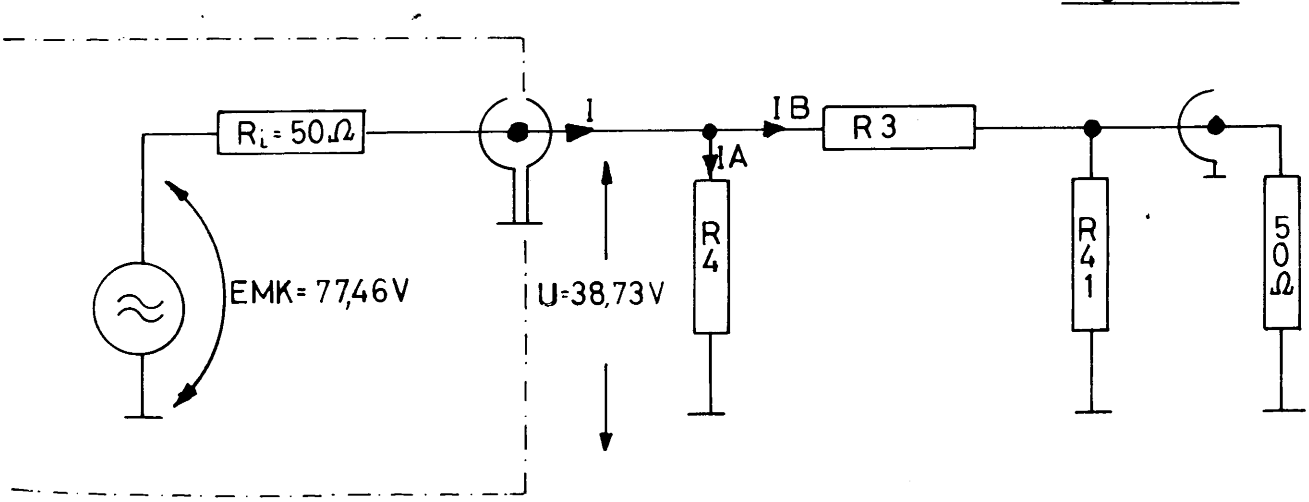
$$I_A = \frac{U}{R4} = \frac{42.58\ [V]}{96.2\ [\text{Ohm}]} = \underline{\underline{0.443\ [A]}}$$

$$I_B = I - I_A = 0.697\ [A] - 0.443\ [A] = \underline{\underline{0.254\ [A]}}$$

$$P_{R3} = I_B^2 \times R_3 = 0.254^2 [A] \times 71.2\ [\text{Ohm}] = \underline{\underline{4.59\ [W]}}$$

2.) The output is terminated in 50 Ohm:

Figure 101



Power matching; at 30 Watts the output voltage will be 38.73 Volts.

Example:

10 dB attenuator

R_3 = 71.2 Ohm

R_4 = 96.2 Ohm

Calculation:

$$P_{R4} = \frac{U^2}{R4} = \frac{38.73^2\ [V]}{96.2\ [Ohm]} = \underline{\underline{15.59\ [W]}}$$

$$I_A = \frac{U}{R4} = \frac{38.73\ [V]}{96.2\ [Ohm]} = \underline{\underline{0.403\ [A]}}$$

$$I = \frac{U}{R_{termination}} = \frac{38.73\ [V]}{50\ [Ohm]} = \underline{\underline{0.775\ [A]}}$$

$$I_B = I - I_A = 0.775\ [A] - 0.403\ [A] = \underline{\underline{0.372\ [A]}}$$

$$P_{R3} = I_B^2 \times R = 0.372^2\ [A] \times 71.2\ [Ohm] = \underline{\underline{9.85\ [W]}}$$

The results indicate quite clearly that the power dissipation of the resistors is going to be different for the open circuit and the correctly terminated case. For practical application the rating of each component should be chosen according to the highest value as obtained from the calculations.

For our examples this means:

<u>T-section:</u>

$P_{transmitter}$ = 30 [Watt]

Attenuation = 10 [dB]

R_1 = $\underline{\underline{26\ [Ohm]}}$, P_{R1} = $\underline{\underline{15.6\ [Watt]}}$

R_2 = $\underline{\underline{35.1\ [Ohm]}}$, P_{R2} = $\underline{\underline{17.05\ [Watt]}}$

<u>Pi-section:</u>

$P_{transmitter}$ = 30 [Watt]

Attenuation = 10 [dB]

R_3 = $\underline{\underline{71,2\ [Ohm]}}$, P_{R3} = $\underline{\underline{9.85\ [Watt]}}$

R_4 = $\underline{\underline{96.2\ [Ohm]}}$, P_{R4} = $\underline{\underline{18.85\ [Watt]}}$

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B.1.1 Terminating resistors and their application.

The dummy-load power meter in B.2 is an example of a typical terminating resistor; but there is a need for low power devices as well, in particular as they may be installed with even shorter leads which makes them suitable for the higher frequencies.

Ranges of application are:

- "Calibration standard" for calibrating a VSWR-meter (B.3.2) or for adjusting the directional coupler (B.4).
- Termination of a line with its characteristic impedance in the case of a power divider or power adder in D.5.9 or B.1.4 etc.

Due to component characteristics each terminating resistor will exhibit specific frequency-dependent properties which have been discussed in great detail in section A.1. This device will thus possess a particular VSWR related to a specific frequency. Terminating resistors are therefore well suited for calibration purposes concerning directional couplers and VSWR-meters. Home- constructed terminating resistors must, however, initially be checked for standing wave ratio, as the installed resistor is likely to have been taken from the scrap box. Poor quality material can not be expected to give satisfactory results at high frequencies. The occasionally long connecting leads and high capacitance mounting caps will result in a significantly frequency dependent impedance that has nothing to do with the imprinted value. The construction based on off-the-shelf low inductance cracked-carbon resistors yielded such excellent results up to the 2 GHZ region that we feel compelled to describe it - even more so, as it is cheap to duplicate.
Construction: The resistor is installed inside a BNC-connector type Radiall-141074 which was previously fitted to a coaxial cable RG58/U. The resistor is soldered to the centre pin with the shortest possible spacing. The other end is treated likewise and fully soldered.

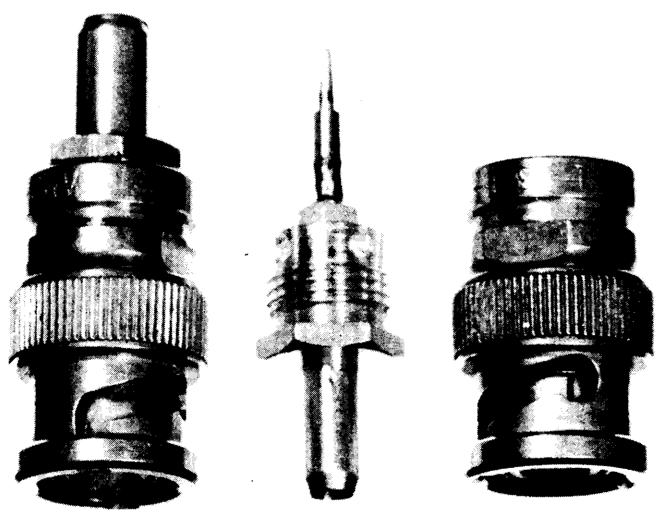


Figure 102

shows - from right to left - the plug body, the plug insert and the overall view of the connector as used for the construction.

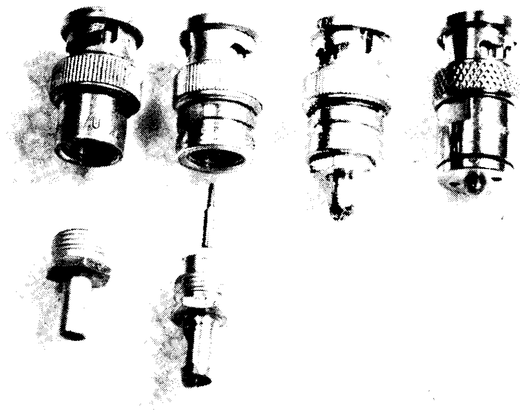


Figure 103

From left to right:

dismantled plug without centre pin,

Resistor mounted inside the plug insert,

original connector and

terminating resistor by RADIALL,
impedance is 50 Ohm.

The three black lines on the following figure represent the curves as obtained on the oscilloscope when measuring calibrated terminating resistors having voltage standing wave ratios of 1, 1.5 and 2 respectively. Afterwards the home-made device was connected to the measuring set-up. Its performance is represented by the white line as painted by the electron beam.

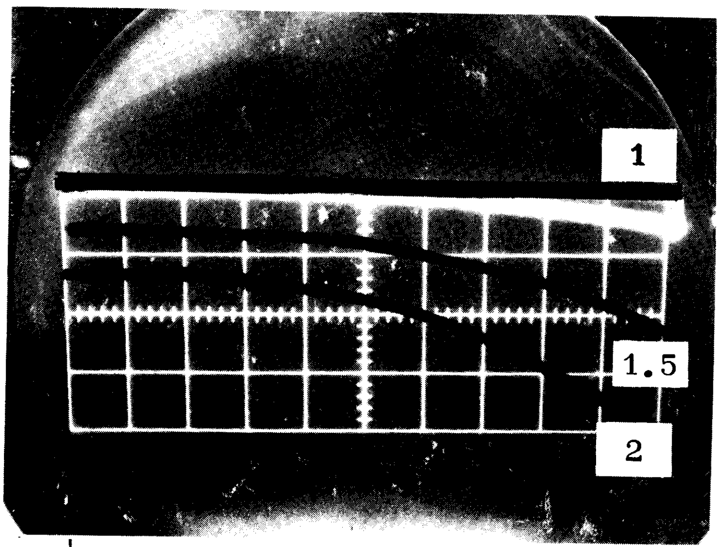


Figure 104

Standing wave ratio - respectively attenuation of the reflected power- of the home-made terminating resistor.

VSWR = 1

VSWR = 1.1 of home-made termination

VSWR = 1.5

VSWR = 2

h: 100 MHz per line.

B.1.2 A step attenuator, dc to 500 MHz

The following design is suitable for all frequencies from short wave up to the UHF-region and provides sufficient accuracy for amateur application. Resistors of the low inductance (carbon) film type should be chosen. Shunt connected resistors should be used whenever a specific value is not available. The individual push button switches should provide reliable contact and should have short connecting leads. The resistors are activated when the push buttons are depressed. When they are released the RF will be routed through a short piece of coaxial cable to the connecting (coaxial) cable and thus to the next switch. The case is a standard commercial tin plate unit measuring 146 x 28 x 28 mm, but any version with screened compartments could be used. The screening braids of the interconnecting coaxial cables are soldered to the screens.

Resistor values for 50 Ohm impedance.

R1 = 20 Ohm, R2 = 300 Ohm, R4 = 160 Ohm, R3 = 40 Ohm,
R6 = 100 Ohm, R5 = 75 Ohm, R8 = 70 Ohm, R7 = 250 Ohm.

These are not the exact values; but they are readily available.

Attenuation:

Figure 105

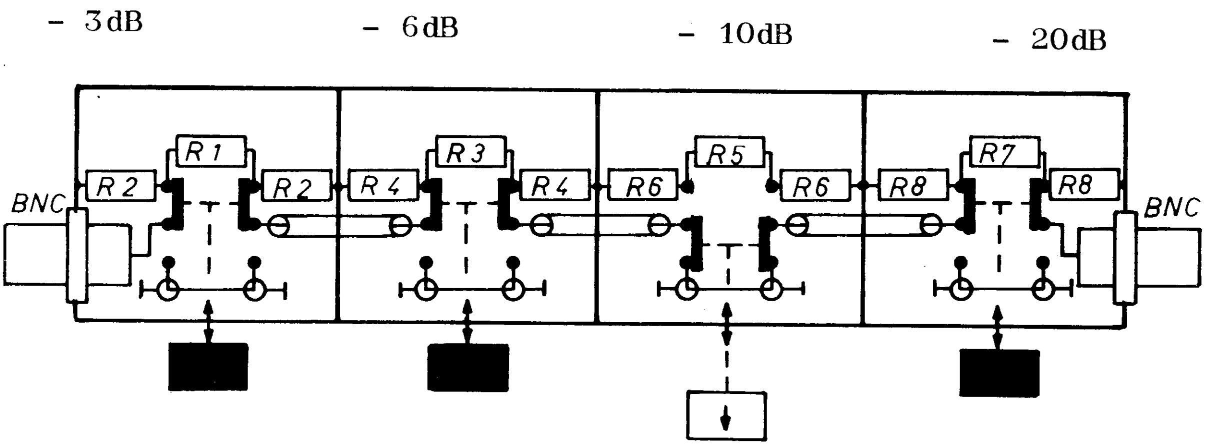
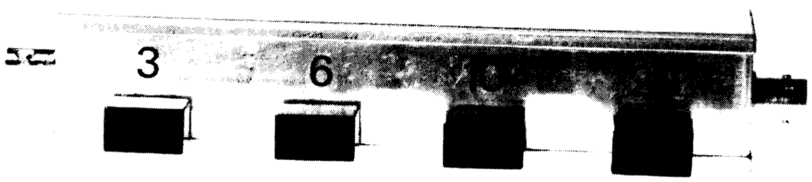


Figure 106



B.1.3 Power splitters/combiners and their application.

Until recently power splitters have been a rare sight in the amateurs shack although they can serve many purposes and solve many problems - such as:

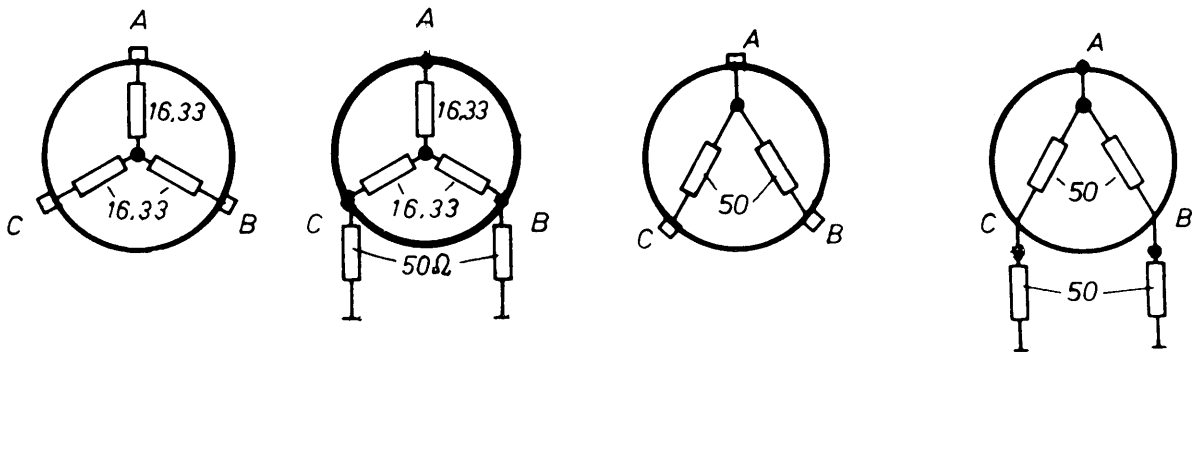
1) Measurement problems: Splitting the drive power of an amplifier to simultaneously measure the power by means of a power meter.
2) Splitting the driver output power of a short wave transceiver to share it between the 70 cm and 23 cm transverters for simultaneous operation on both bands (D.7, D.1.1, D.4).
3) Splitting some power (say: 10 Watt into 2 x 5 Watt) to simultaneously drive two - or more - power amplifiers. After amplification a considerable power level is available which could subsequently be recombined (D.5.9) and fed to a single antenna - or each amplifier could feed a separate antenna.
4) Distribution of power to feed several antennas etc. etc..

Generally speaking, we have to differentiate between two types of power splitters which will serve equally well as splitters or power adders: There are frequency-dependent and frequency-independent devices.

B.1.3.1 Frequency-independent power splitters/combiners

Frequency-independent power splitters may be composed of resistors. These resistors must be non-reactive - that is to say: Their resistance must remain constant for all frequencies. Resistors do - however - contain mounting caps and connecting wires which behave like inductors and capacitors (see section A.1.5) and could result in an effective resistance quite different from the value printed onto the component. If carbon-film or composition-type resistors with low holder capacitances and severely shortened connecting wires are being used the frequency- dependent component can be reduced. This type of power splitter may then be used from d.c. up to Giga Hertz frequencies - in other words, their properties are relatively frequency-independent.
The circuit diagram is as shown. Its main disadvantage is the fact that only one quarter of the power entering at port A will be delivered to each of the two output ports B and C. This is obvious since the power (for instance at B) is taken off a point correponding to one half of half the input power as indicated by the right-hand diagram. Whereas this arrangement will only split the power from A to B and C, all ports may be interchanged in the case of the arrangement shown to the left.
Some applications are described in sections D.7 and D.1.1.

Figure 107 Power splitters composed of carbon film or composition-type resistors. Aim for the shortest possible connecting leads!



B.1.3.2 Frequency-dependent power splitters/combiners

Two types of frequency dependent power splitters for sharing the RF-power between two mixing diodes have already been mentioned in section A.5.4:

1) The quadrature-hybrid (90° phase shift) and
2) the ring-hybrid (180° phase shift).

The practical aspect is covered by the converter and mixer descriptions of sections A.5.2, A.5.4, D.2.4 and D.2.5.
The main difference between the devices described here and those that are mentioned in B.1.3.1 is the fact that no phase shift between the various ports will occur when using non-reactive resistors; however, there is always a power loss of at least 6 dB (two port power splitter) and even more in the case of multiple port arrangements.
Frequency-dependent devices can only be used over a relatively narrow frequency band. They have the advantage of evenly sharing the power between the ports or - conversely - combining the input powers. This opens a wide field of possible applications to the amateur. Furthermore, good isolation can be achieved between the various ports. One particular arrangement which is easy to construct is the 180°-four-port hybrid network. It is known under the following names:
Magic Tee, Hybrid junction, 3 dB-coupler, Hybrid coil, Iso-T, Null-T, Power divider and power combiner - to mention just a few.

The 180° four-port hybrid network

This easy to construct low-loss four-port network is well suited for combining power amplifiers or antennas.

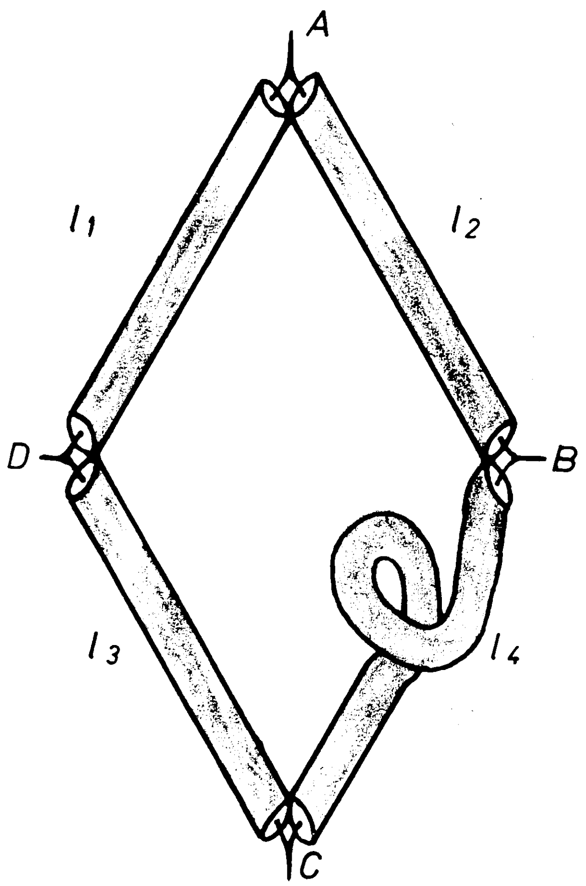


Figure 108 shows the four-port network composed of sections of coaxial cable.

All input and output ports are terminated according to supply cable impedance (i.e. 50 Ohm)

Depending on the selected input port, the phase conditions will vary between the output ports. For considering the phase relations, the ouput ports are taken as reference and the 90° shift due to the $\lambda/4$ sections is neglected

Input	Output ports			
	B	D	C	A
A	0	0	zp	-
B	-	zp	- 180	0
C	-180	0	-	0
D	zp	-	0	0

Phase shift in degrees.
zp indicates zero power due to power cancellation (path length difference is $\lambda/2$)

The port opposite the input port must be terminated in 50 Ohm.

Dimensions of cable sections of the 180° four-port hybrid network:

L1 and L3 = $\lambda/4 \times V$ (V = velocity factor of cable material)

L4 = $3\ \lambda/4 \times V$ (Z_K = characteristic impedance of cable)

Z_K = $Z \times \sqrt{2}$ (Z = characteristic impedance of cables feeding the network, i.e 50 Ohm).

If 50 Ohm coaxial cable is used throughout the station, then Z_K will come to $50 \times \sqrt{2} = 70.7$ Ohm.

Slight deviations are acceptable and 75 Ohm cable according to RG-59/U specification or television-type coaxial cable may be utilized.

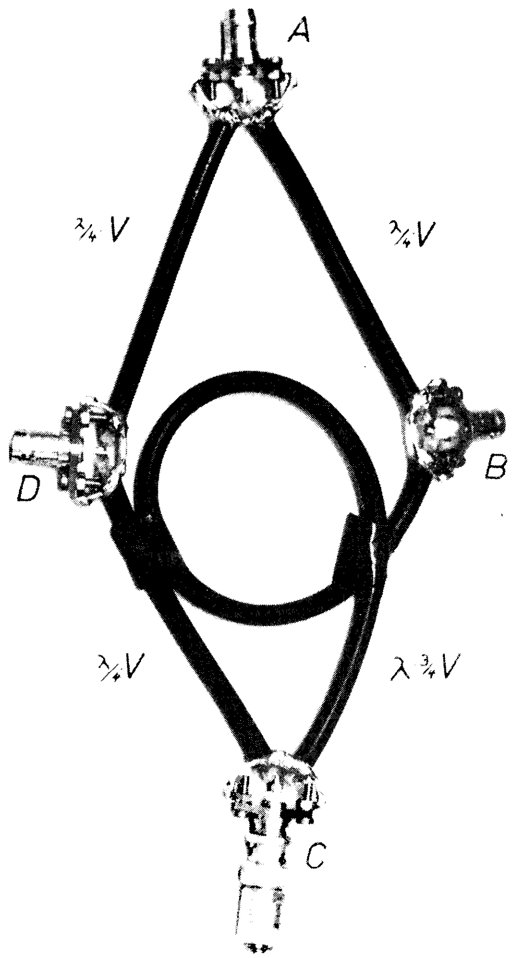


Figure 109

Four-port hybrid network made of sections of 75 Ohm cable, type RG-59/U.

The coupler was arranged to give zero degree phase difference between output ports.
(A = input, B/D = output).

Theoretically there will be no power at port C. It is terminated in a 50 Ohm resistor.

The $3\ \lambda/4$ section is rolled up to form a loop as shown in the photograph.

When intending to use this network for combining the output of power amplifiers delivering more than 40 Watt then a cable type with suitable power rating must be chosen.

The cable dimensions are measured from flange to flange of the BNC-connectors. If certain types of cable possess different velocity factors then the correct length must be found by trial and error. The sections should be shortened in 5 mm increments or be replaced by correspondingly longer pieces. The whole assembly may be installed in a tin plate chassis 146 x 72 x 30 mm as shown on the next page.

Cable dimensions corresponding to the 70 cm band: $\lambda/4 \times V = 100$ mm

$3\ \lambda/4 \times V = 300$ mm

The circuit developed by DC9NL employs a different method of power splitting. In OM Dieter's arrangement all cable sections are of equal length ($\lambda/4 \times V$). The TV-quality coaxial cable has a characteristic impedance of 75 Ohm. This set-up only allows power splitting from A to B/C as 0° coupler.

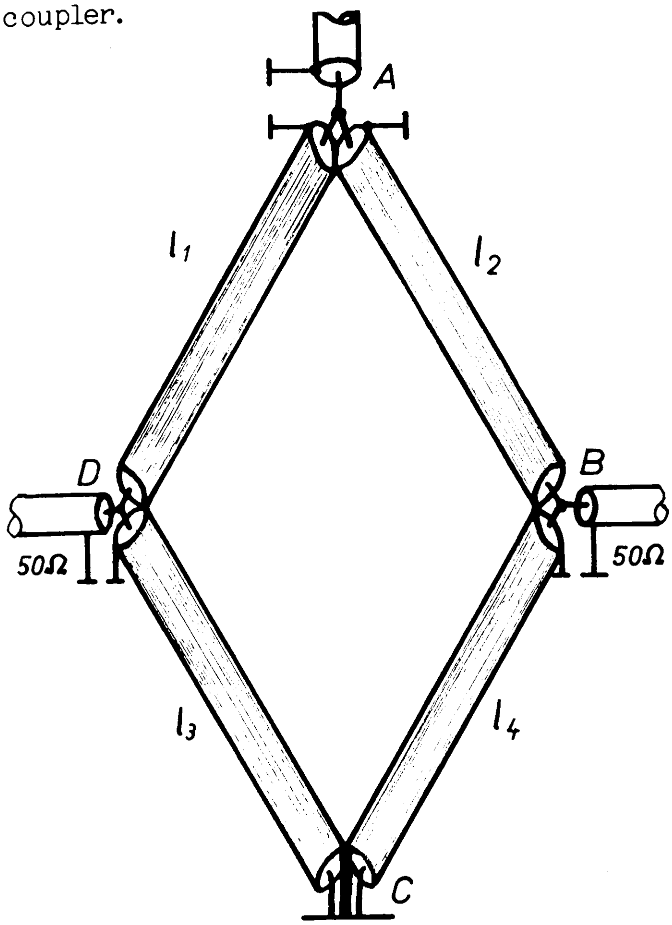


Figure 110

Power splitter / combiner according to DC9NL.

Power entering the circuit at port A will be shared amongst output ports B and D.

The phase relations at ports B and D are identical (0°).

In this case output C is grounded.

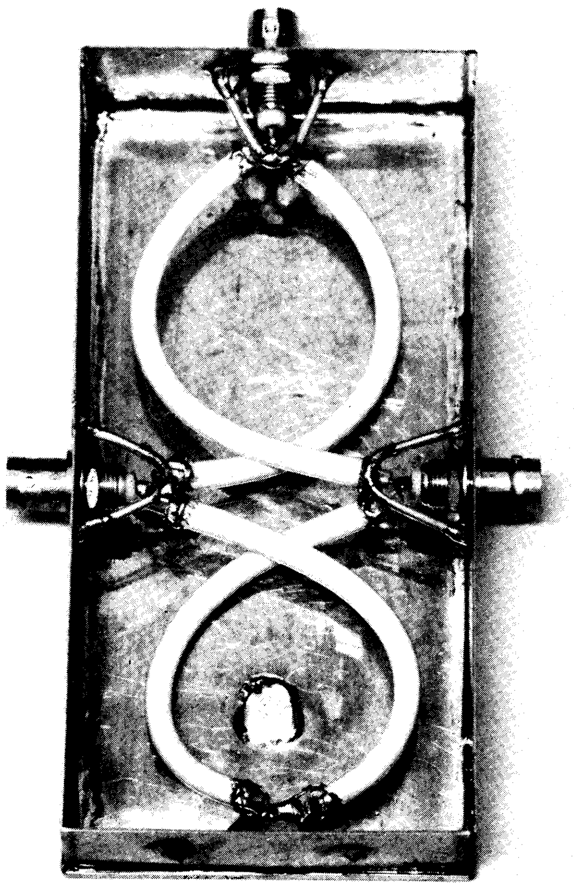


Figure 111

Installation of the circuit in a chassis 146 x 72 x 30 mm.

The outer conductor is soldered to a wire bridge which is then soldered to the chassis.

Because of the cable dimensions the sections were arranged in loops. No detrimental effect was observed.

All previously mentioned frequency-dependent power splitters/combiners are based on sections of coaxial cable. This technique is adequate for frequencies from short wave up to the 70 cm band. Beyond this the dimensions are getting rather critical and every millimeter will severely shift the resonant frequency. In this case the printed circuit technique will give better results. Figure 112 shows such a version for the 1152 MHz oscillator frequency. The velocity factor is defined by the dielectric properties of the base material (Epoxy resin, see A.2.4). This arrangement is also used in the transmitting converter D.4.2.

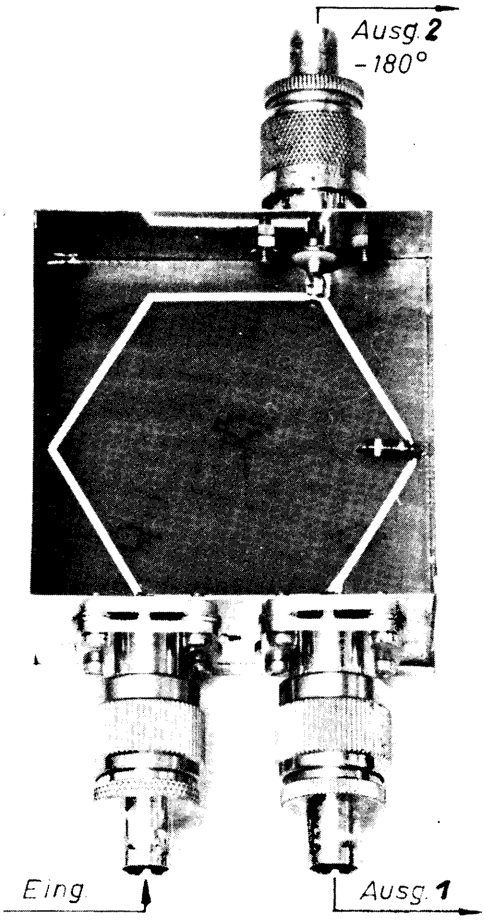


Figure 112

Power splitter/combiner in stripline technique (6 x $\lambda/4$ circuit). The base material is epoxy resin.

Taking output port 1 as reference, output port 2 will lag by 180°.

Each $\lambda/4$ section is 33 mm long. The corresponding value for 1296 MHz is 29 mm.

The width of the conductor was chosen to be 1.4 mm.
The thickness of the epoxy board is 1.6 mm.

For measuring purposes the visible adapters (N to BNC) were installed.

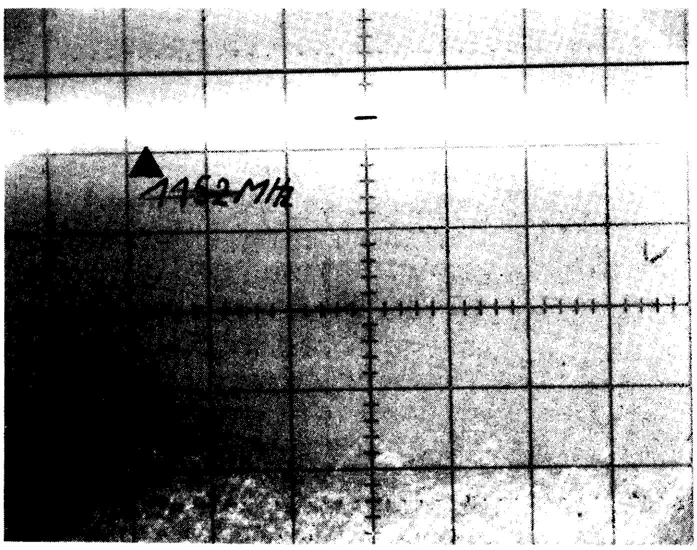


Figure 113

demonstrates the power as delivered to both output ports for the region from 1-1.65 GHz.

P_{in} = black line

P_{out} = white lines

Vert.scale = 5 dB/line

At 1152 MHz both outputs are 3 dB below the input. At all other frequencies the power distribution is uneven.

Again $Z_K = Z \times \sqrt{2}$
$= 70.7$ Ohm

As mentioned before, there should be no power at the port opposite the input. However, due to imbalance and stray coupling a certain amount of power (about 20 to 30 dB below the input power) will appear at this point. Assuming an input power of 10 Watt and 20 dB of attenuation this results in 100 mW being present at the output. After terminating output ports B and D with fixed resistors and calibrating the whole arrangement, this four-port network - in conjunction with a mW-meter according to B.2.1 - could be used as wattmeter according to B.2.1.

Several short-wave transceivers deliver an output power of 10 Watt; by means of a four-port network (A = input, B and D terminated in 50 Ohm at 5 Watt) about 100 mW are available at port C. This could be used for control purposes or to simultaneously drive two transverters (70 and 35cm bands). The examples given in D.7 and D.1.1 are supplemented by the following suggestions:

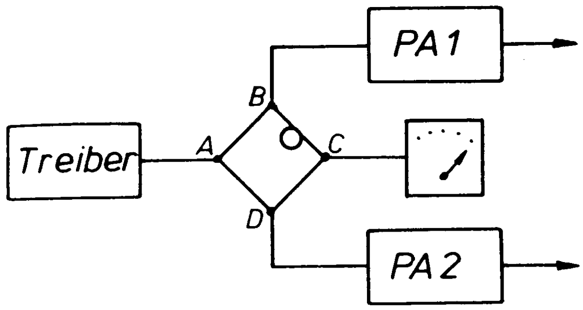


Figure 114

Employing the unused power at point C for control purposes.

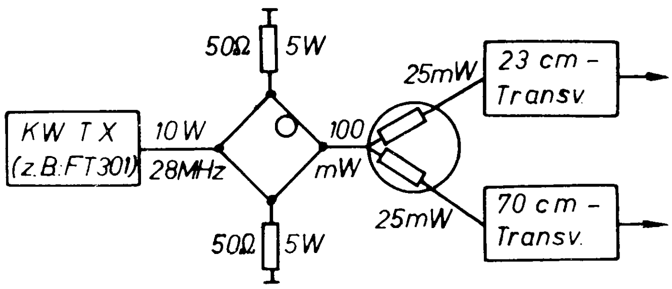


Figure 115

Simultaneous supply of drive power to two transverters.

$\lambda/4 \times V = 1.76$ m
(for a velocity factor of V = 0.66)

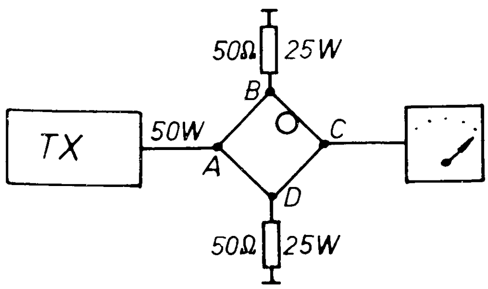


Figure 116

Making use of the four-port hybrid network to extend the range of a mW-meter up to Watts.

Literature: (1) Anzac, RF and Microwave Components.
(2) Möhring,Empfangstechnik im UHF-Bereich,LOEWE OPTA Bd.III

B.1.4 UHF absorption frequency meter up to 2.5 GHz

A similar version of this frequency meter has been described previously in "UKW-Berichte". The following contribution should simplify the construction. The use of a Schottky diode has resulted in improved sensitivity and a much sharper resonance indication. As an example, an input power of 0.1 mW at 1296 MHz results in a voltage of 50 mV being delivered to a 50 000 Ohm/Volt instrument.

Theory of operation:

The measuring device is a cavity into which an inductance is inserted. Thus a trough line resonator is formed. Any signal which is passed through the cavity will set up a stray field which is coupled via the tuning rod (L) to the Schottky diode. The voltage across the diode will be maximum when the tuning rod has tuned the trough line to the signal. After calibration a scale outside the cavity will allow to determine the frequency of the signal. This device will operate up to 2.5 GHz. As indicated by the diagram, a second voltage maximum will occur if the tuning rod is pushed in too far (example figure C, i.e. at one third the correct frequency) without that particular frequency being fed to the device. Thus faulty readings may result. To avoid this, the tuning rod should be withdrawn fully before commencing any measurements. The first indicated frequency upon inserting the tuning rod will be the correct reading of either the signal or one of its actually existing components.

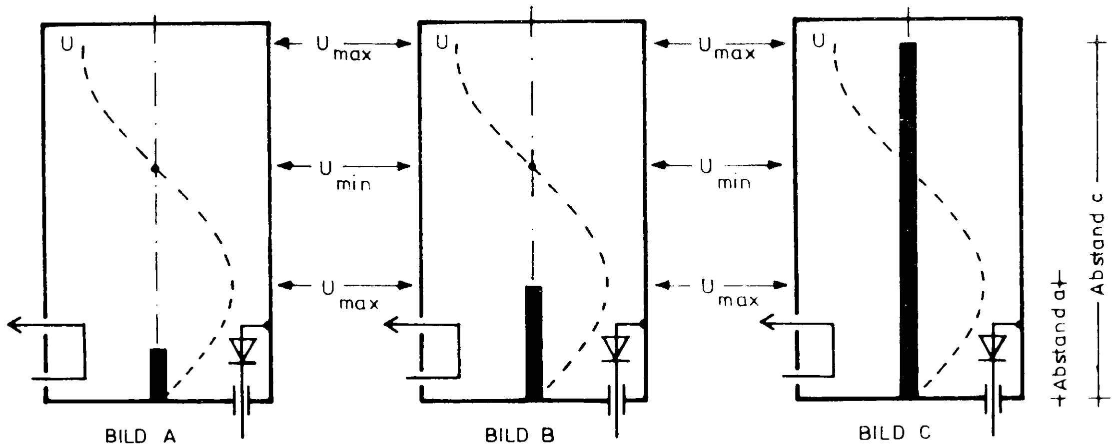


Figure 117

A) Low voltage transfer (out of resonance)
B) Maximum voltage transfer (resonance)
C) Erroneous frequency reading (f_R), f = indicated frequency x 3

Example:

With a signal frequency (f) of 1500 MHz a further resonance will be indicated at 500 MHz.

Figure 118
Interior of absorption frequency meter. A small strip of brass sheet with one end suitably shaped is soldered to the cavity end plate. It provides good contact and proper support of the tuning rod. The coupling loop is formed from a strip line (brass or copper, 2 mm wide) and is installed 3 mm off the cavity wall. No cover is required since the cavity is higher than it is wide. The resonant cavity is thus easily accessible.

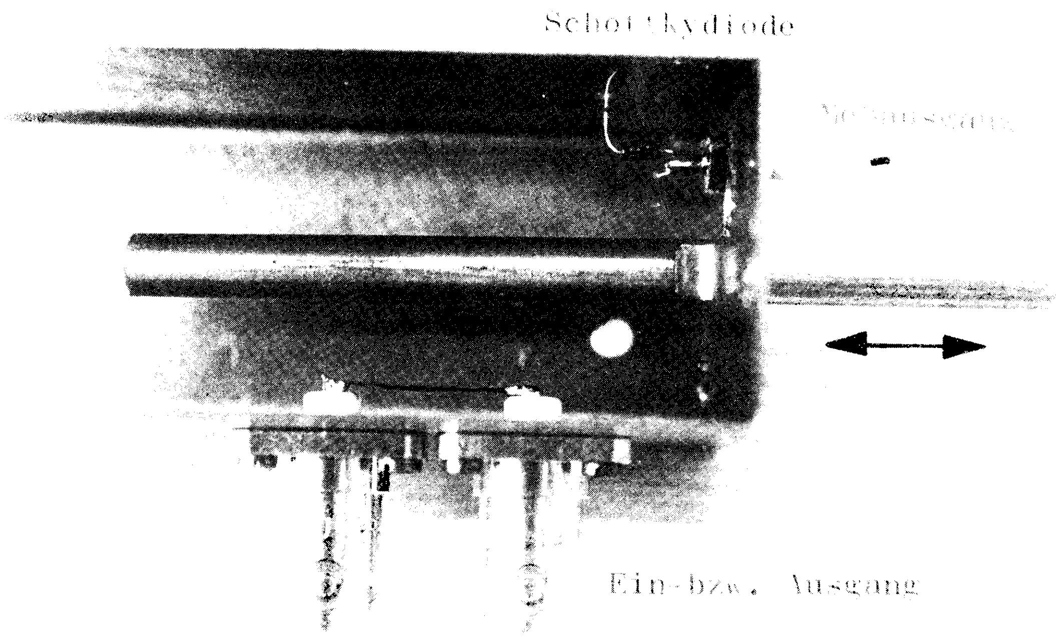
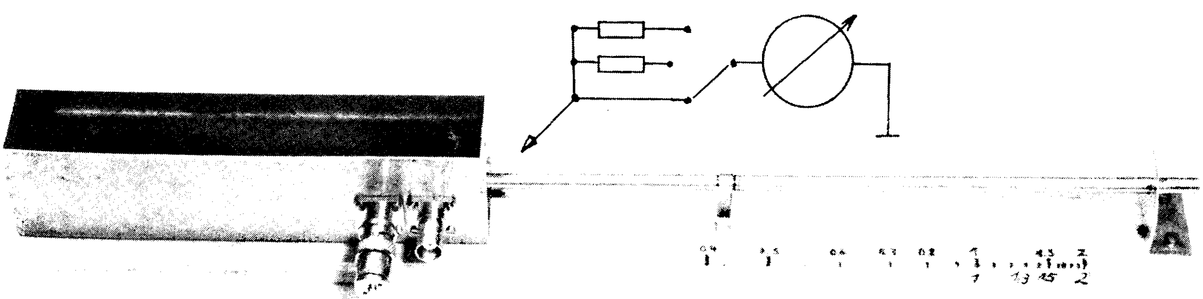


Figure 119
Overall view of frequency meter. The case is constructed by bending a piece of 1 mm brass sheet into a U-shaped trough. Small plates are soldered to the ends (as shown in the photograph). The case dimensions are: length = 165 mm, width = 30 mm, hight = 45 mm. The pointer is crimped around the tuning rod and soldered. The rear support ensures correct positioning of the rod and acts as a guide. The wooden base plate measures 440 x 50 x 15 mm. The output is terminated in a 50 Ohm resistor (see B.1.1).



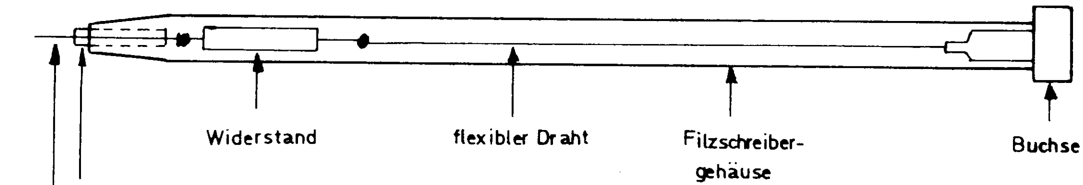
B.1.5 Test probes and alignment tools
(N. Schramm, DC9NI)

1. Test probes with (a) and without (b) decoupling resistor.

   List of materials:

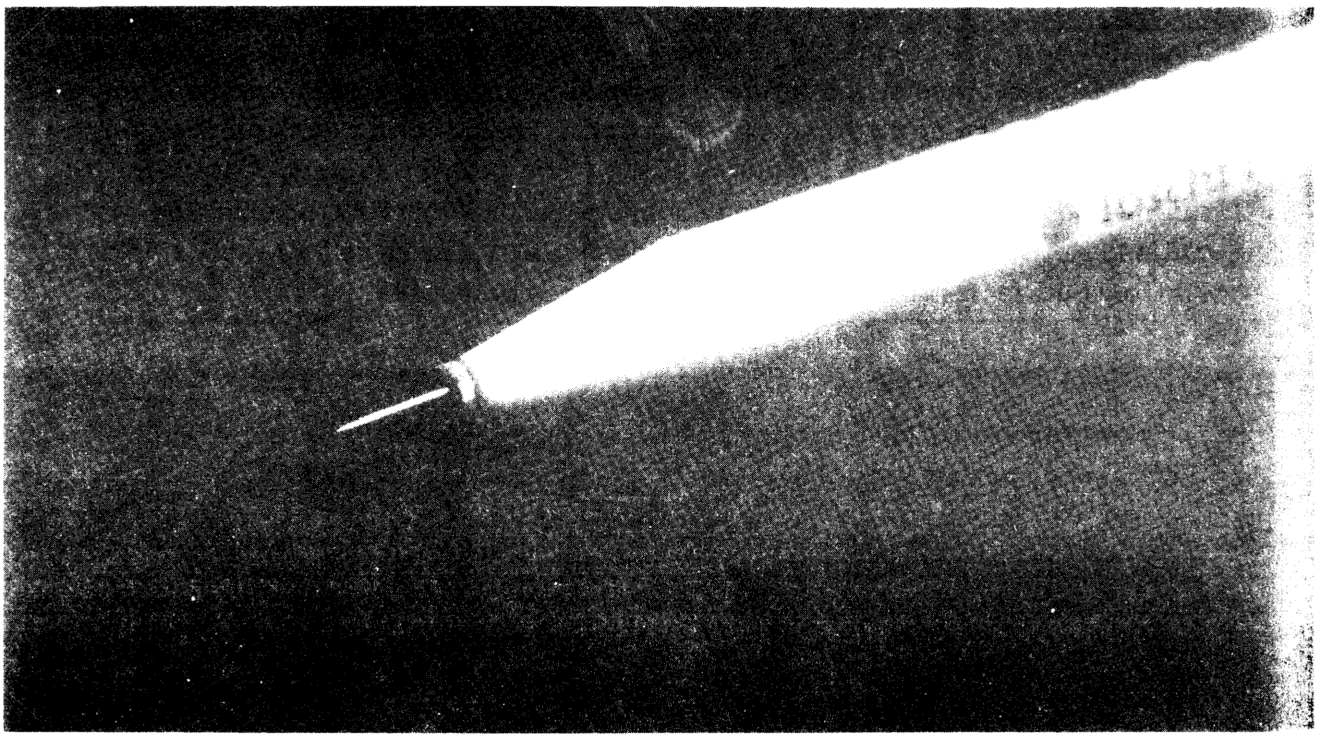
   1 old felt-tip pen
   1 pin
   1 insulated socket, 4 mm
   20 cm of flexible wire
   1 carbon film resistor 100KOhm (for "a" only)

Figure 120



Stecknadel = pin
Filzspitze = felt-tip
Widerstand = resistor
Flexibler Draht = flexible wire
Filzschreibergehäuse = felt-tip pen body
Buchse = socket

Figure 121



Advantage:

Test probes with decoupling resistors are essential for voltage measurements at RF-test points. Capacitive loading of the circuit-under- test is avoided.

Disadvantage:

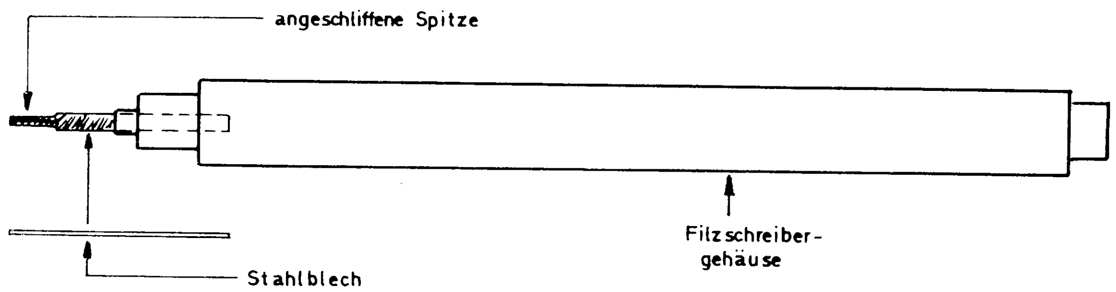
Inaccurate readings will result when used in conjunction with voltmeters of low internal resistance. Test probes with 100 kOhm decoupling resistors should therefore be used only if the voltmeter has an internal resistance of 10 MOhm.

2.) Alignment tools:

a) For inductors

Figure 122

angeschliffene Spitze = tip, ground to shape
Stahlblech = steel sheet
Filzschreibergehause = felt-tip pen body



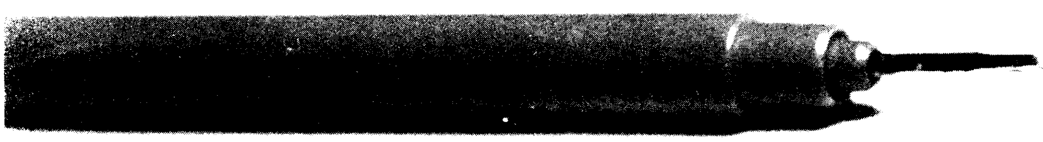


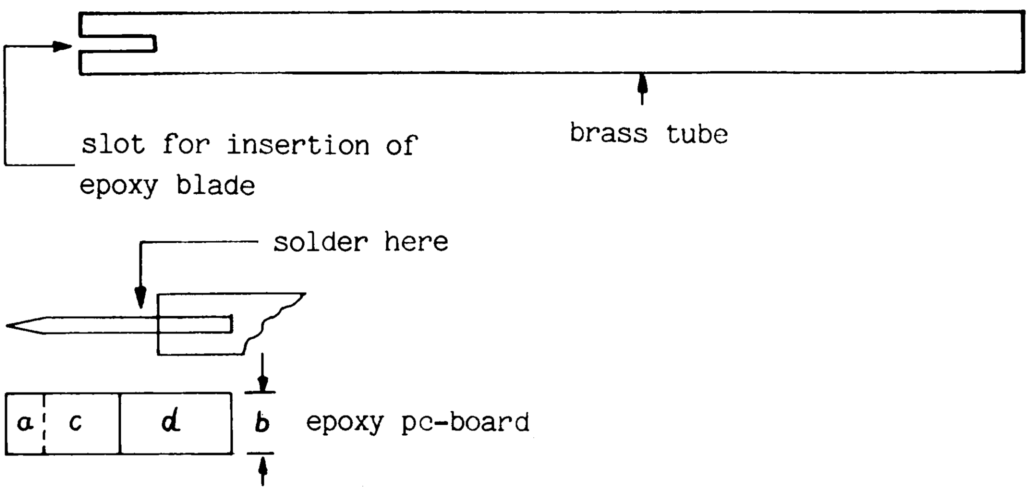
Figure 123

<u>List of materials:</u>
1 old felt-tip pen
1 piece of steel sheet (old saw blade!)

Remove felt tip and insert metal tip. Grind to fit the various core sizes.

b) <u>For capacitors:</u>

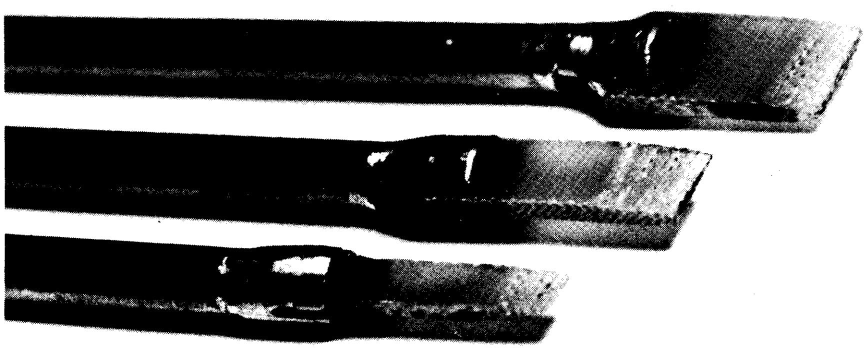
<u>Figure 124</u>



a = wedge-shaped tip
b = width (according to variable capacitor)
c = remove copper layer
d = retain copper layer

<u>Figure 125</u>

Various sizes of home made alignment tools



<u>List of materials:</u>
1 piece of brass tubing 6 mm diam., 15 cm long, one end slotted.
1 piece of copper clad printed circuit board, base material is epoxy.

3.) RF probes

The construction of multiplier stages in oscillators and transmitters calls for stage-by-stage adjustment. Depending on the nature of the coupling point one of the rf probes described below is required.

Figure 126
Various types of home made rf probes.

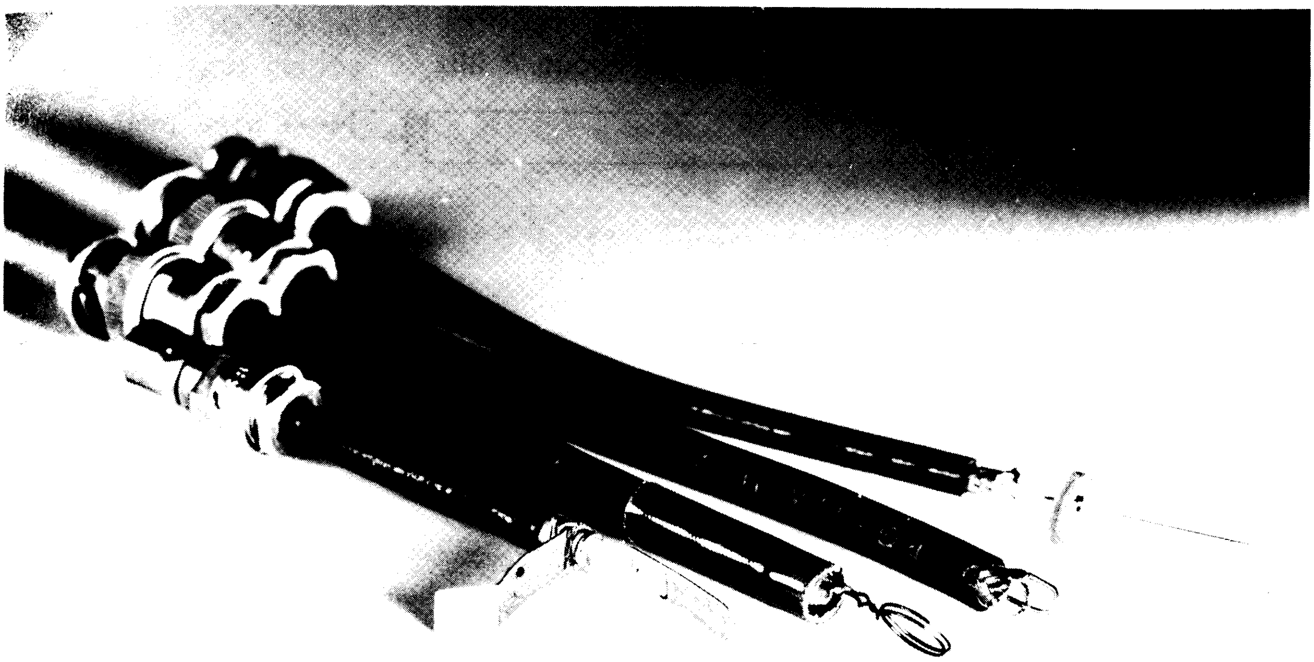
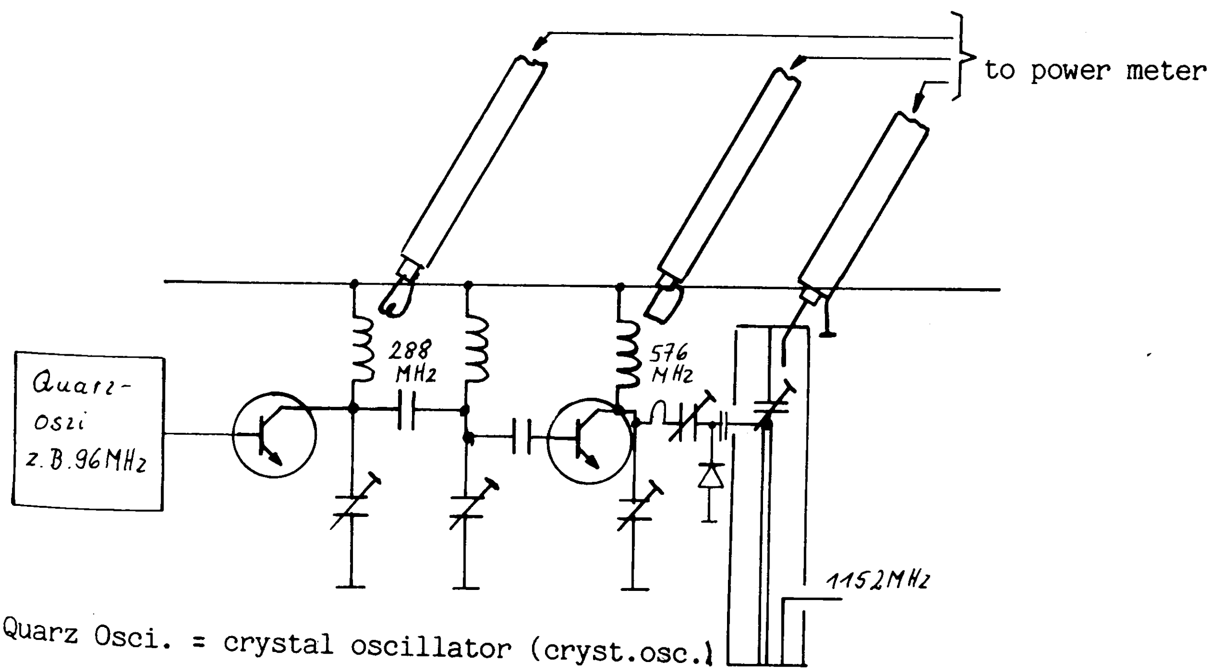
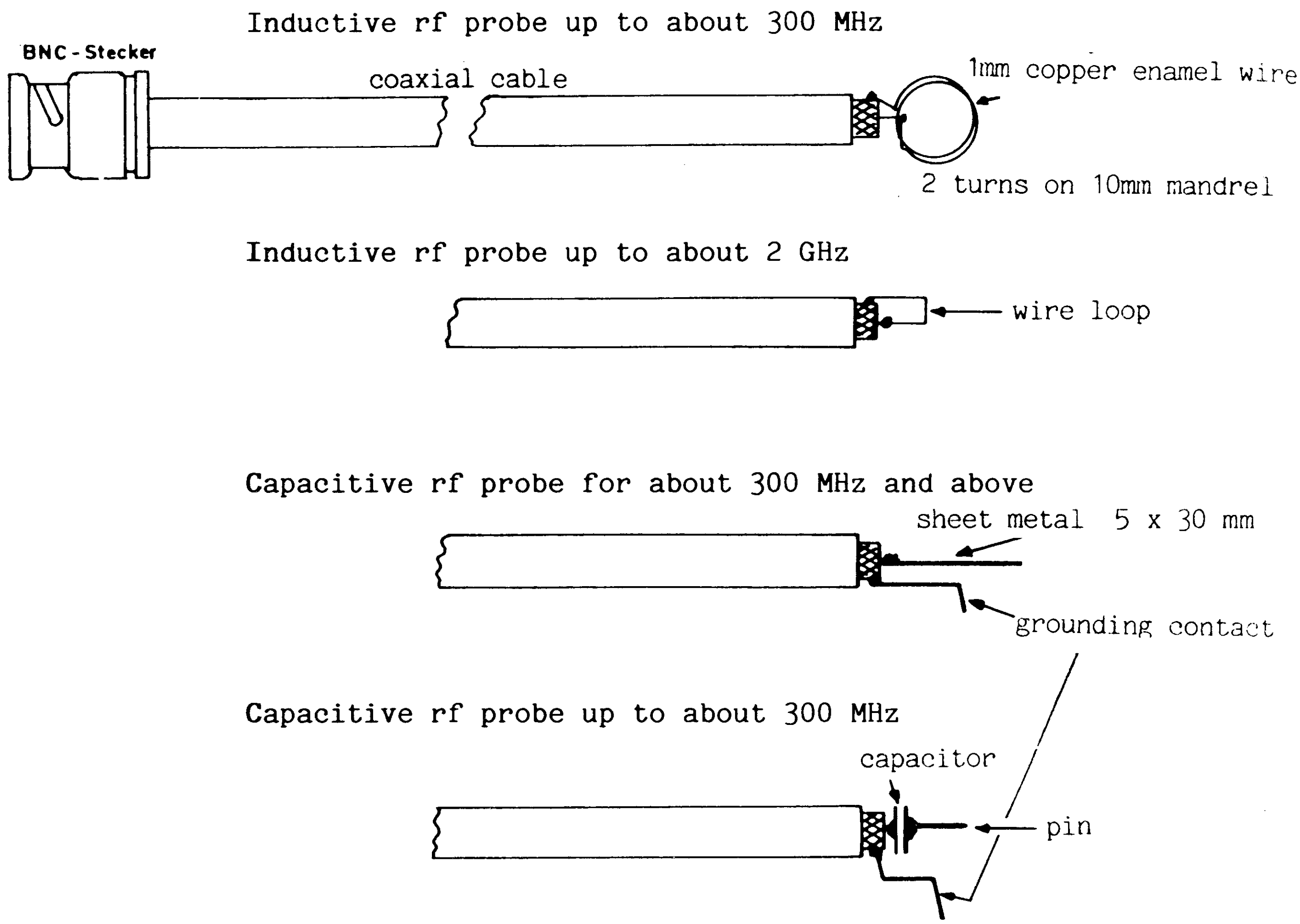


Figure 127 The correct choice of rf probe for the various coupling points in a transmitter chain is demonstrated.



Quarz Osci. = crystal oscillator (cryst.osc.)

Figure 128 RF probes for various frequency ranges.



4.) HF-test probe

When adjusting hf-circuits an hf-test probe is required. It should have high impedance to avoid loading the circuit-under-test. The construction of a suitable hf-test probe is subsequently described.

List of materials:

1 old felt-tip pen
1 piece of brass tubing, 6 mm diameter, length according to felt-tip pen
1 meter of screened cable
2 capacitors of 4.7 nF, 63 Volt
2 diodes OA 85 or equivalent type
1 piece of springy sheet metal (such as switching arm of old relays!) to provide ground contact.
1 bolt M 2 x 25 suitably shaped or 2mm nail, 25 mm long.
1 piece of insolating material, see drawing.

Figure 129 A practical HF-test probe.

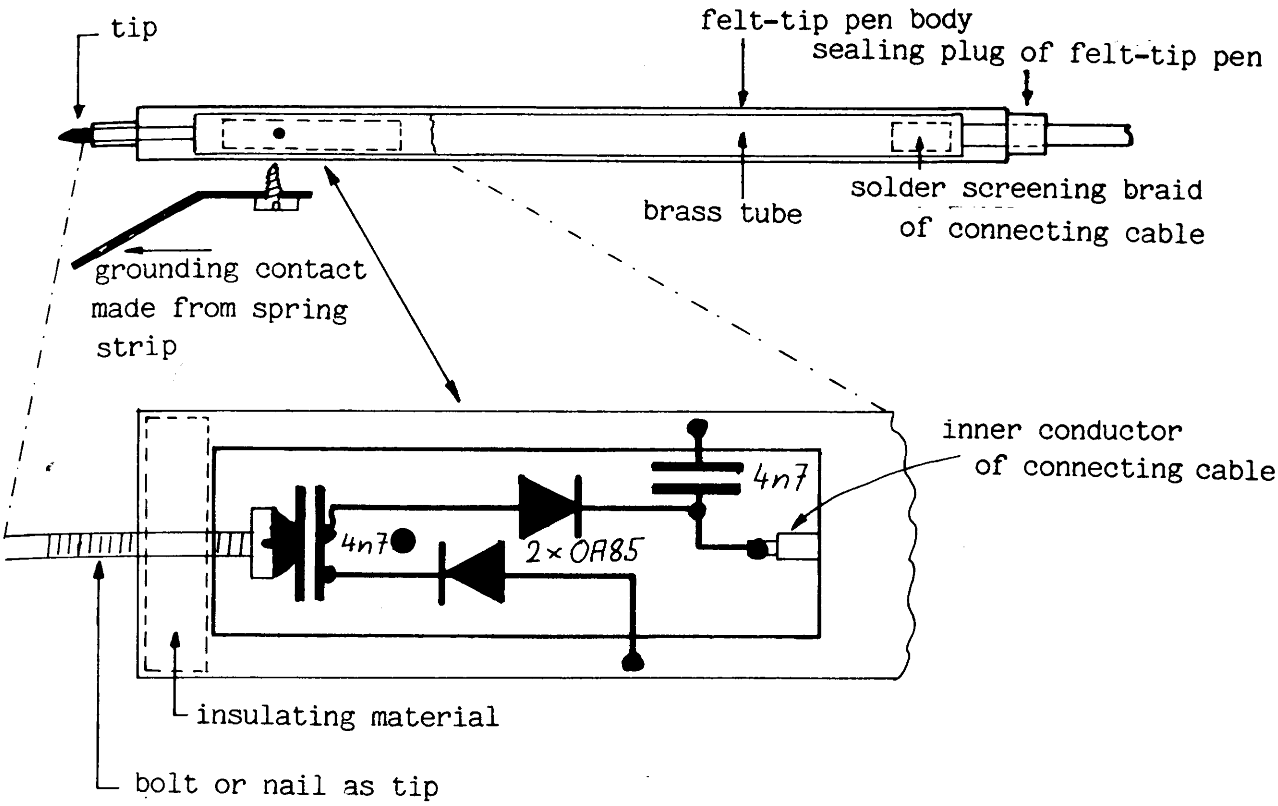
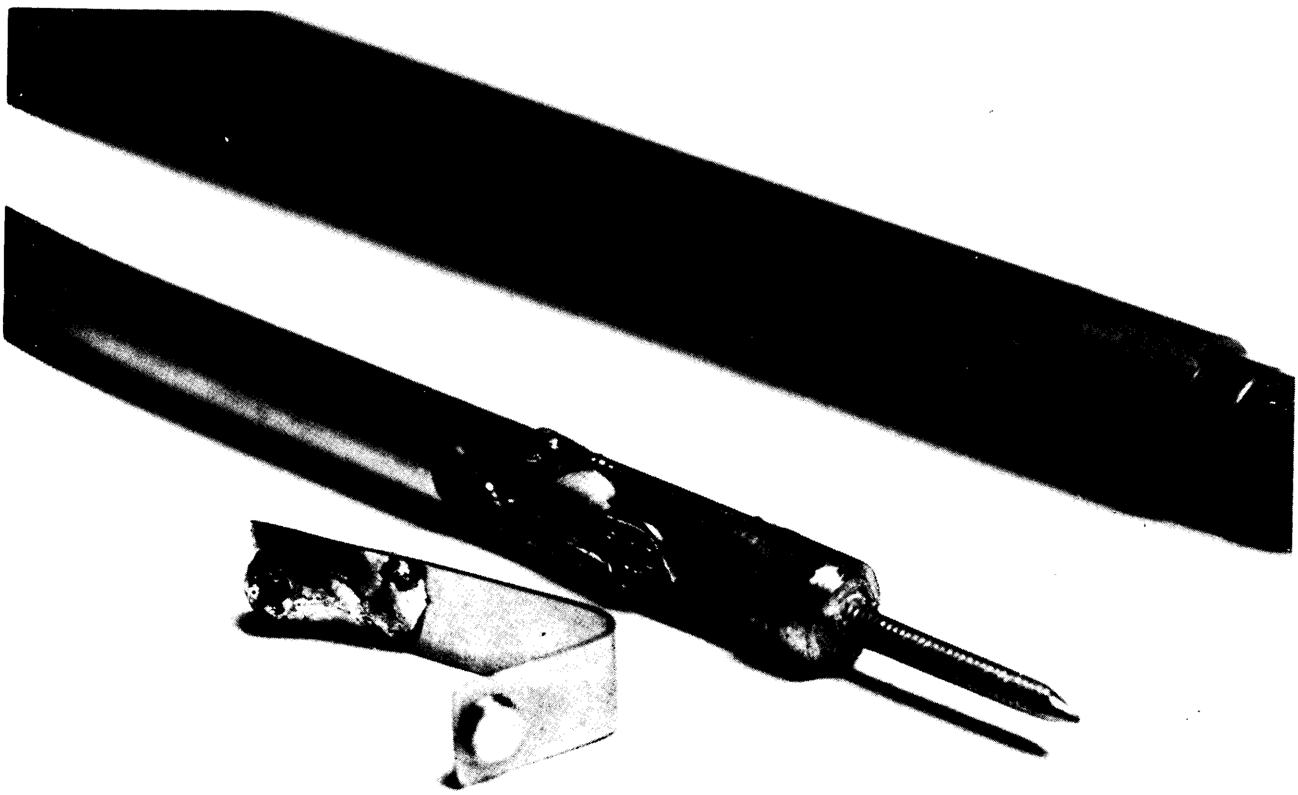


Figure 130 Photograph of the described hf-test probe.



B.1.6 A transistor dip oscillator for 270 to 500 MHz

(J. Koferl, DC9RK)

The dip oscillator described in this section covers the frequency range from 270 to 500 MHz and employs two oscillator heads. A circuit with unknown resonant frequency will absorb energy from the dip oscillator when both are tuned to the same frequency. It is thus possible to estimate the resonant frequency of a tuned circuit. This procedure is well known from the grid-dip-meter.
This transistorized dip oscillator was designed for the 70 cm band. The frequency range may be extended to 190-600 MHz by additional oscillator heads. In that case the coupling capacitors, the oscillator coil and the varicap diodes will have to be replaced. However, OM Josef did not supply any details for these modifications.
The two supply voltages required by the stages are stabilized and temperature compensated according to reference (1). The 12 Volt ac is supplied by a small encapsulated transformer. The required 25 Volts dc is obtained by voltage doubling, filtering and Zener diode stabilisation. Temperature compensation is achieved by two series connected diodes 1N914.

Theory of operation:
When the oscillator operates, a small part of the rf-voltage is taken off the emitter. It is then rectified and passed on to the dc amplifier BC237A. The 100 KOhm potentiometer is utilized to set the emitter current to 0.8-1 mA. The overall sensitivity is improved by biasing the diodes. To this effect the 1 KOhm potentiometer is adjusted such that the 1 mA meter will just indicate zero current. During this setting-up procedure the oscillator must not oscillate. This is achieved by short- circuiting inductor L.
Apart from the coil winding data the construction of both oscillator heads is identical. The cases are made up from copper clad pc-board. The component layout is indicated by the diagrams that are drawn to scale. The oscillator heads are tuned by means of the varicap diodes BB105. The resonant frequency is determined by inductance L, the voltage dependent capacitance of the tuning diodes and by the two coupling capacitors (CK1 and CK2).
If a tuned circuit with identical resonant frequency is brought into the vicinity of the oscillator coil, a certain amount of power will be drawn from the oscillator. Consequently, the rf voltage at the emitter will be reduced which in turn reduces the control voltage to the dc amplifier BC237A and a negative current dip is indicated. The approximate frequency can be read off the dial of tuning potentiometer P1 (10K) - assuming, that it had been calibrated beforehand. One end of L1 is soldered to ground (side wall) and the other end is soldered to the "hot" side (A) of an insulated support (formed by a piece of pc-board).

The capacitor between base and emitter is of the disc- or trapezoid-type and has no soldering lugs. The screening of the transistor is soldered to ground as well. The choke in the tuning voltage circuit consists of two turns wound on a single-hole 5 mm ferrite bead. A similar ferrite bead (FP) is used between the base of BFY90 and the 10K/5.6K voltage divider. In this case, however, the connecting wire of one resistor is merely threaded through the hole. The rf section of the head is separated from the supply socket by means of a partition.
Power is supplied to the oscillator head through a standard 5-pole screened diode cable.

Figure 131

Circuit diagram of oscillator head. All feed-through and blocking capacitors are 1 nF.

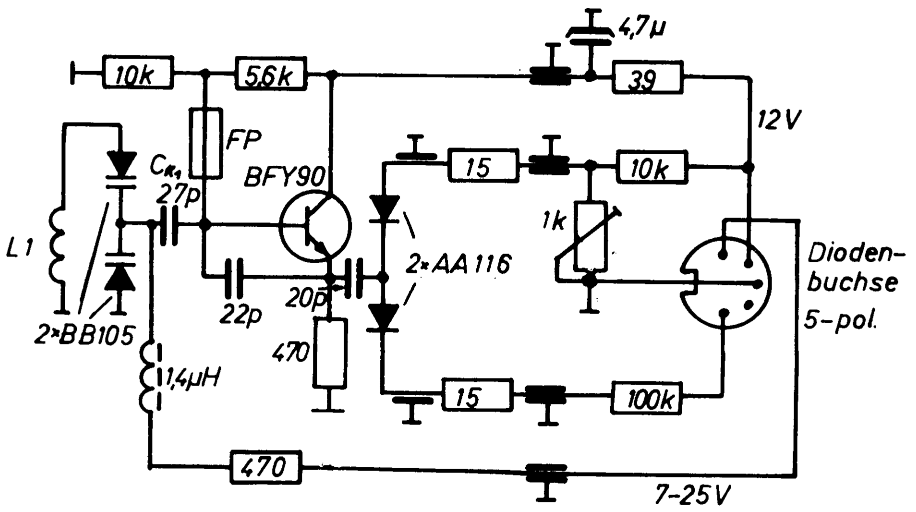


Figure 132

Circuit diagram of power supply and indicator unit. The resonant frequency of the unknown circuit can be read off the previously calibrated dial of potentiometer P1.

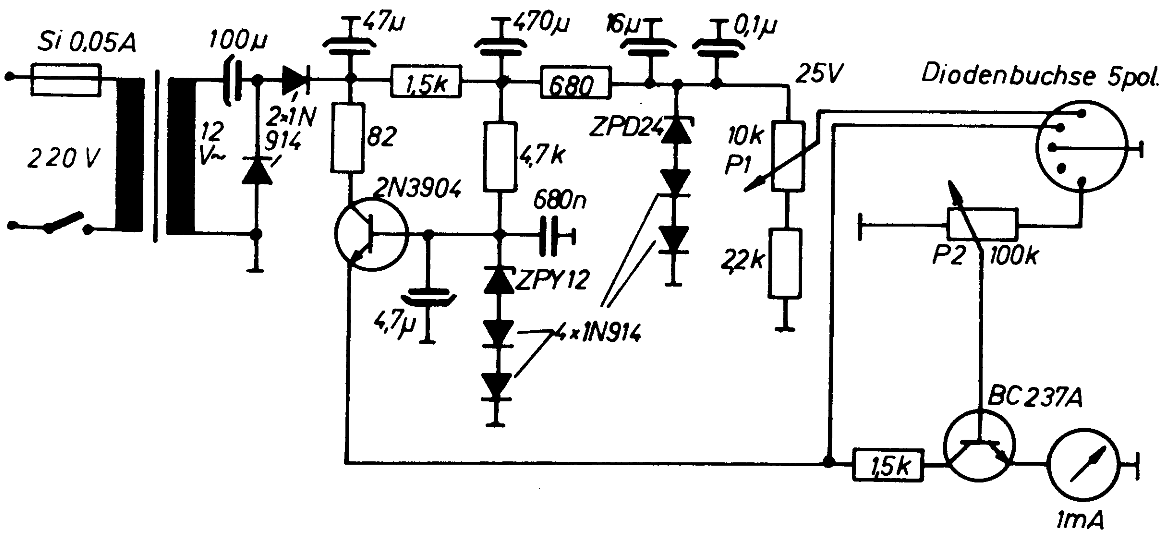


Figure 133
Plane view and side view of oscillator head with component symbols.
(Drawn in actual size)

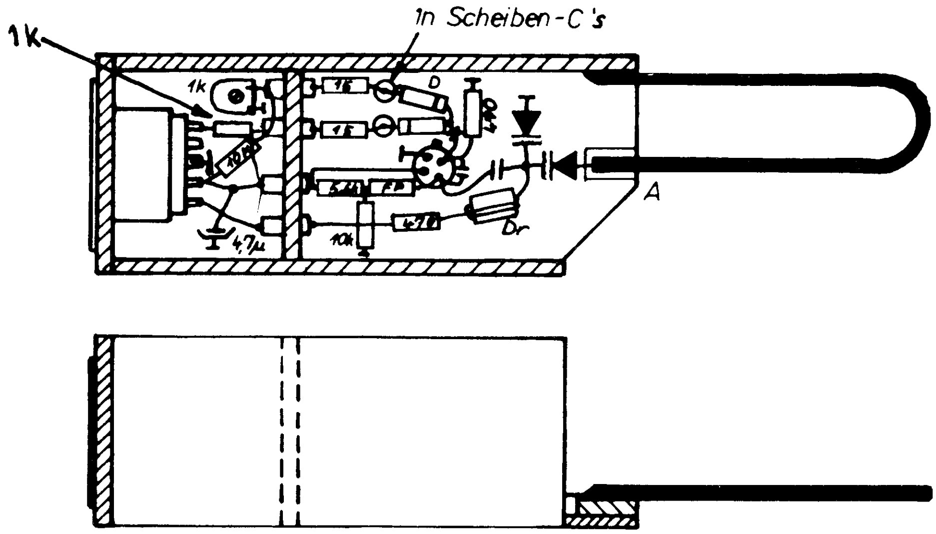


Figure 134
Front view and location of
feed-through capacitors in
partitioning wall.

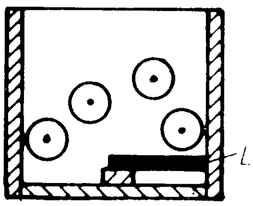


Figure 135
Dimensions of the inductors for both frequencies
(CuAg, 1.5 mm diam.)

Bibliography: (1) VHF/UHF-Technik DUBUS

B.2 A dummy load power meter for VHF and UHF

When a high frequency cable is terminated in its characteristic impedance, the sum of all power components (i.e. including the unwanted spurious products) will cause a voltage drop across this termination which - after rectification - may be used to feed a voltmeter. The reading is a function of the total power applied.

The arrangement described hereafter consists of ten shunt-connected low inductance film resistors (each having 550 Ohm, 2 Watt) as termination. The addition of two resistors (330 Ohm, 1 Watt) will reduce this value according to the following equation:

$$R_g = (55 \times 660) / (55 + 660) = \underline{50.76 \text{ Ohm}}$$

An old milk tin (55 mm diam.) houses the assembly. The BNC-connector is mounted in the centre and the feed-through capacitor is mounted 5 mm off the rim of the lid that had previously been cut off. On the reverse side of the lid the resistors are soldered to the spigot of the connector making the connecting wires as short as possible. The resistors are arranged in a star-like manner and their free ends soldered to the rim of the lid. The influences of capacitances and inductances of the metering circuit on the termination is reduced by connecting the diode to the center tap of the voltage divider. The output is loaded by a 5 KOhm resistor. This reduces the effects which differing internal resistances of voltmeters could have on the readings even when the rf-power remains constant.

D is a Schottky diode, but any other diode suitable for VHF/UHF may be employed.

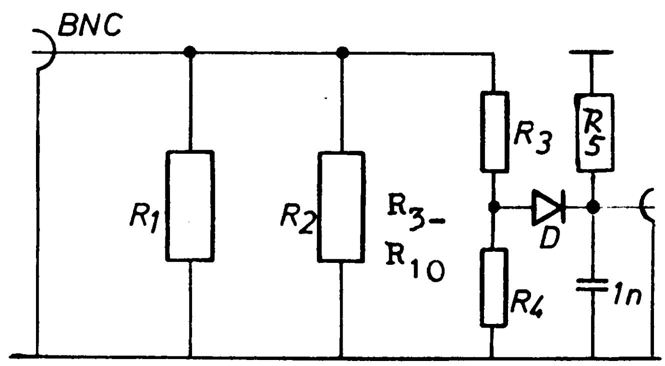


Figure 136
Circuit diagram of dummy load power meter.
R_1 to R_{10} are 550 Ohm at 2 Watt each
R_3 and R_4 are 330 Ohm at 1 Watt each
R_5 is 5 KOhm

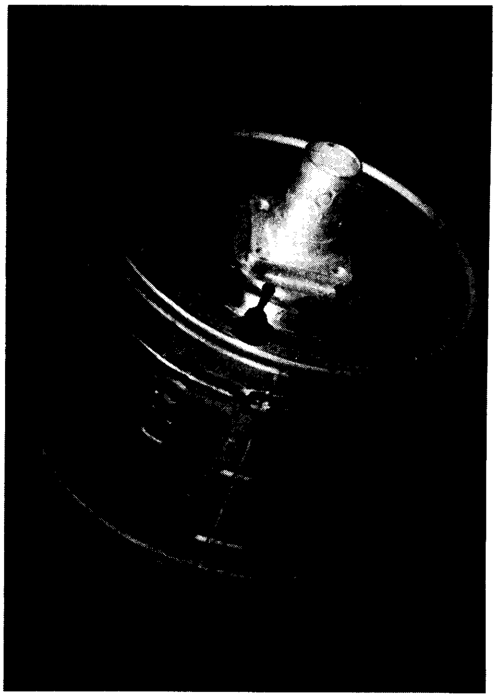


Figure 137
Photograph of finished measuring head. This arrangement could be installed in combination with a voltmeter in a single casing. Recalibration of the dial allows direct reading of the power.

The empty milk tin is kindly donated by one's favorite YL or XYL...(hi).

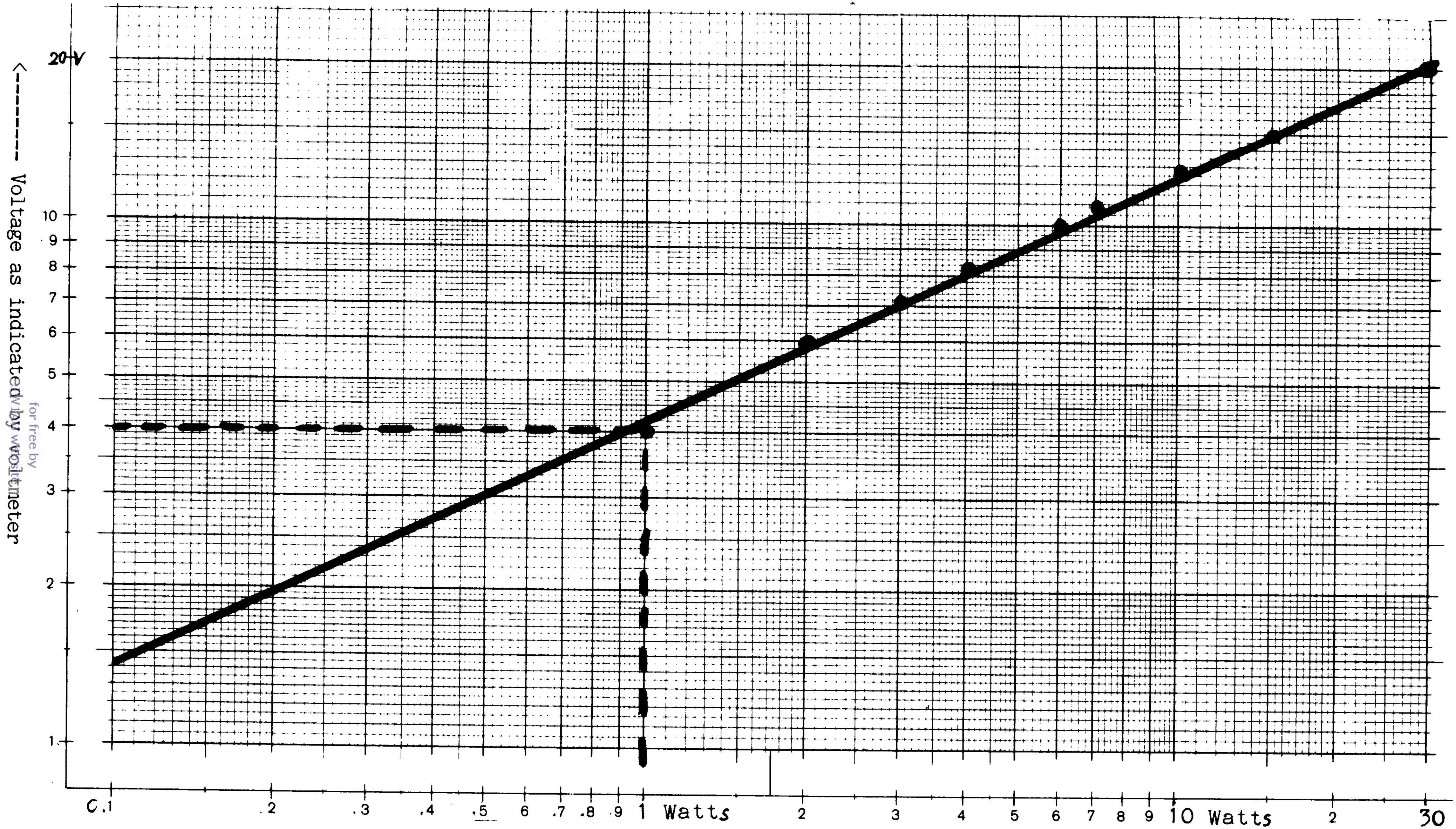
When using the specified resistors a rating of 12 Watt continuous power and 25 to 30 Watt intermittent (up to 10 seconds) will result. For higher rating ten resistors of 550 Ohm, 5 Watt or 20 resistors of 1100 Ohm in a correspondingly larger tin could be used. This could, however, result in a poorer VSWR due to different component properties and longer connecting leads when installing the resistors in the tin (see section A.1). The arrangement as described above exhibits a VSWR of 1.1 throughout the 2 m band and 1.15 on 70 cm.

If the same voltmeter is used consistently, then the 5 KOhm resistor could be disregarded.

Calibrating the power meter:

The local amateur radio club will be only too happy to purchase a power meter just for you and to help you out... It is then connected to a 2 m transmitter having an adjustable output power. The power is set to give a 1 Watt reading on the borrowed instrument. The transmitter is now connected to the dummy load power meter. The voltage is read off the voltmeter dial and the reading is entered onto a piece of graph paper. Example: 4 Volts corresponding to 1 Watt input power. This procedure is repeated for 2, 3, 4 Watt etc. Finally all points are linked by a line to give the intermediate values. Check the result by choosing any power level and verify by means of the borrowed instrument. If all readings are confirmed the calibration process is finished for the 2 m band and may now be repeated for the 70 cm band.

Figure 138 Example for plotting the voltage readings against transmitter output power. Since the component properties may vary each time the dummy load power meter is duplicated, the voltage-versus-power line may be deviate as well. When designing the dummy load for higher power ratings, the voltage divider should be changed to - perhaps - 500 and 160 Ohm.



B.2.1 A mW power meter up to UHF

"If you want to be sure of a measurement, only measure once!" This statement highlights the problem of taking measurements. There is, however, no denying the fact that a certain minimum of measurements cannot be avoided. Otherwise no amateur is likely to successfully turn theory into home-made working equipment.

Signal levels in the various stages of mixers, converters, amplifiers and oscillators are increasingly specified and cannot be adjusted without the use of a power meter for the milli-watt range. Not everybody has access to a commercial grade instrument; the following description proves, that there are simple means for performing measurements that will give satisfactory results in the 28 to 1296 MHz range.

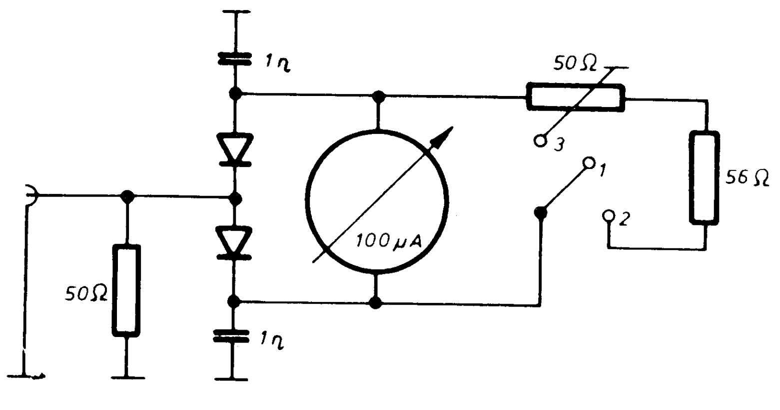
The basic idea is to rectify and measure the rf-voltage across a resistor which is caused by the total rf power supplied to this resistor.

Circuit diagram 1

The rf voltage that has to be measured exists across a 50 Ohm resistor. It is rectified and doubled by means of two diodes. The 1 nF capacitors remove the remaining rf component. The resultant dc voltage is sufficient to operate a voltmeter having the specified internal resistance R_i. Three ranges are available and are selected by shunting the meter with load resistors. The calibration curves are shown in figure 142.

Figure 139

Circuit diagram 1



Diodes: AA 116

Instrument: Moving coil, 100 uA full scale, R_i = 1.5 KOhm

Figure 140
Detail of the rf head. The power is taken directly off the BNC connector and is rectified. This keeps the inductance of the connecting leads to a minimum. R is a 50 OHm carbon film resistor. The two disc capacitors (1 nF) double as solder lugs to support the remaining diode leads.

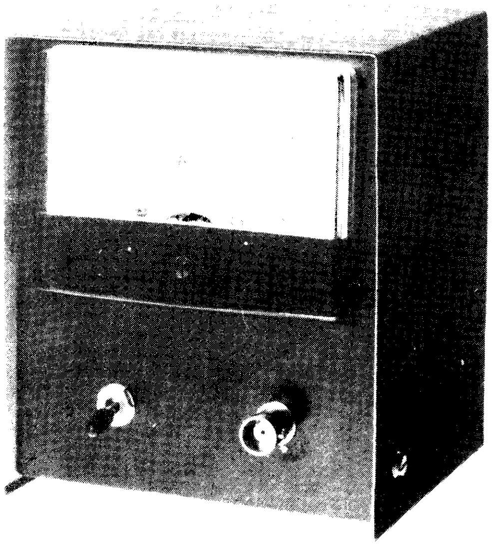


Figure 141
The completed power meter. Front panel and base plate consist of copper clad printed circuit board and are painted.

Dimensions:

Hight = 120 mm
Width = 100 mm
Depth = 100 mm

Instrument:

Face: 86 x 64 mm
Flange: 45 mm diameter

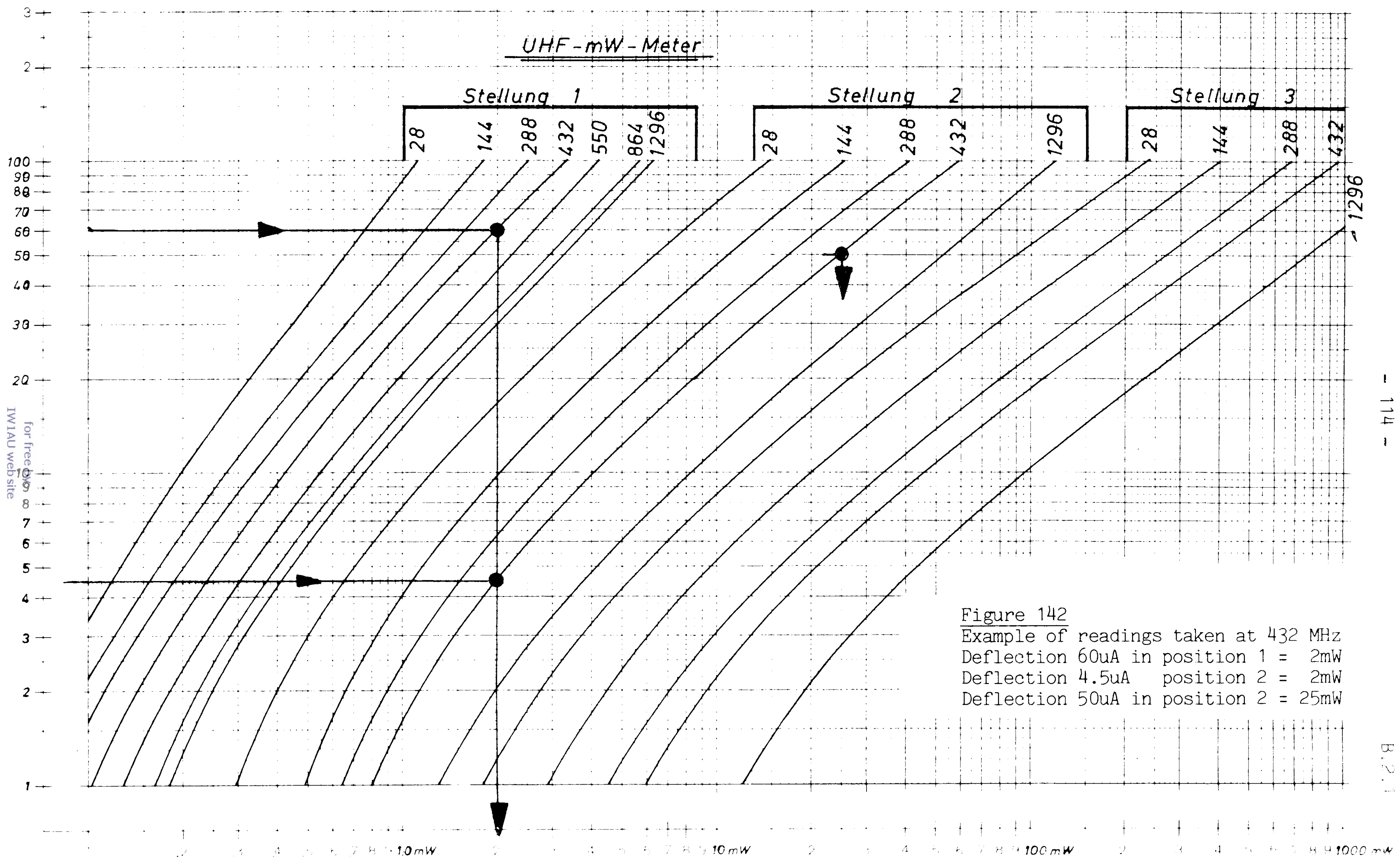


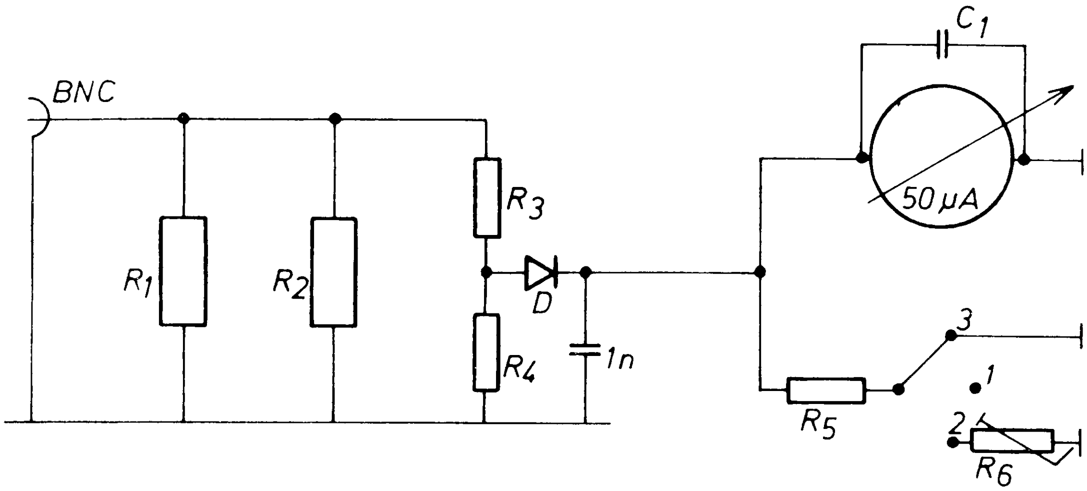
Figure 142
Example of readings taken at 432 MHz
Deflection 60uA in position 1 = 2mW
Deflection 4.5uA position 2 = 2mW
Deflection 50uA in position 2 = 25mW

Circuit diagram 2

In this arrangement the diode is connected to a 1:1 voltage divider as in B.2. Again the connecting leads should be kept as short as possible. The trade-off is a comparatively low frequency dependence of the power indication.

The 50uA instrument (Ri = 1500) can be shunted to extend the range. There is, however, no point in having more than three ranges.

Figure 143



R_1, R_2 = 110 Ohm, 1/4 Watt
R_3, R_4 = 330 Ohm, 1/8 Watt
C_1 = trapezoid or feed-through capacitor
D = Schottky or other UHF diode
R_5 = 10 Ohm
R_6 = 100-15- Ohm (see text)

Both circuits can be constructed separately from the instrument unit. The rf head should be equipped with a BNC plug to which a light case of printed circuit board is soldered. The output lines are internally filtered and brought out to connectors. The whole rf head may thus be plugged into the output terminals of objects that require measuring. This way losses and matching problems of connecting cables are avoided.

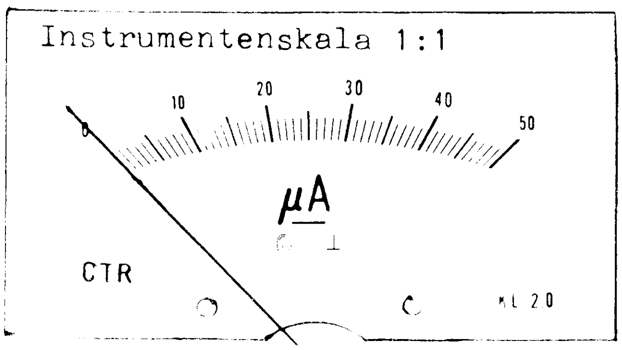
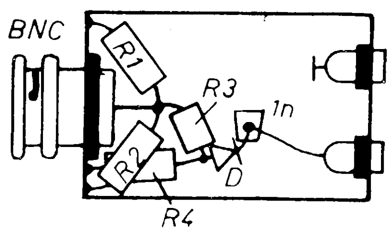


Figure 144
Diagram of a rf head



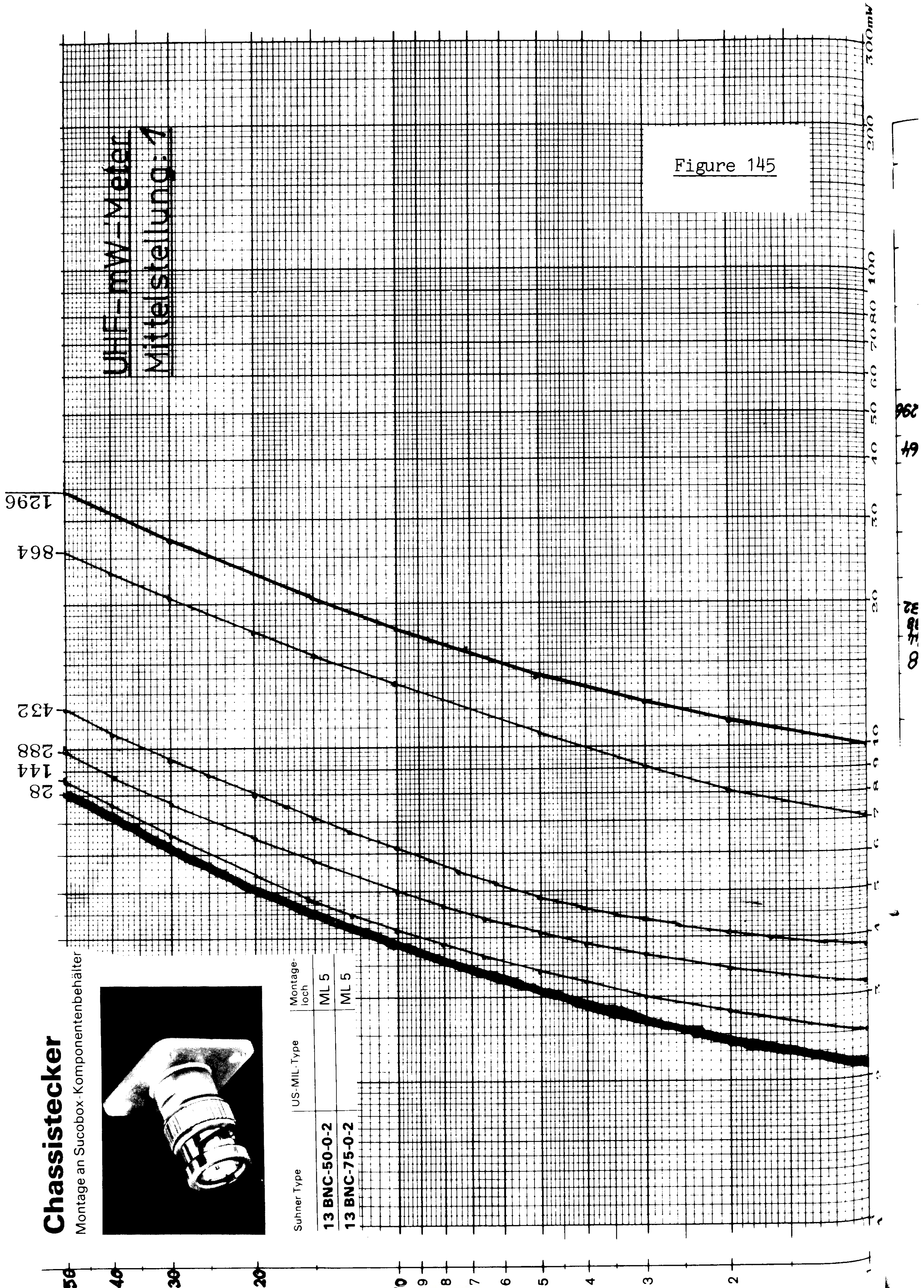
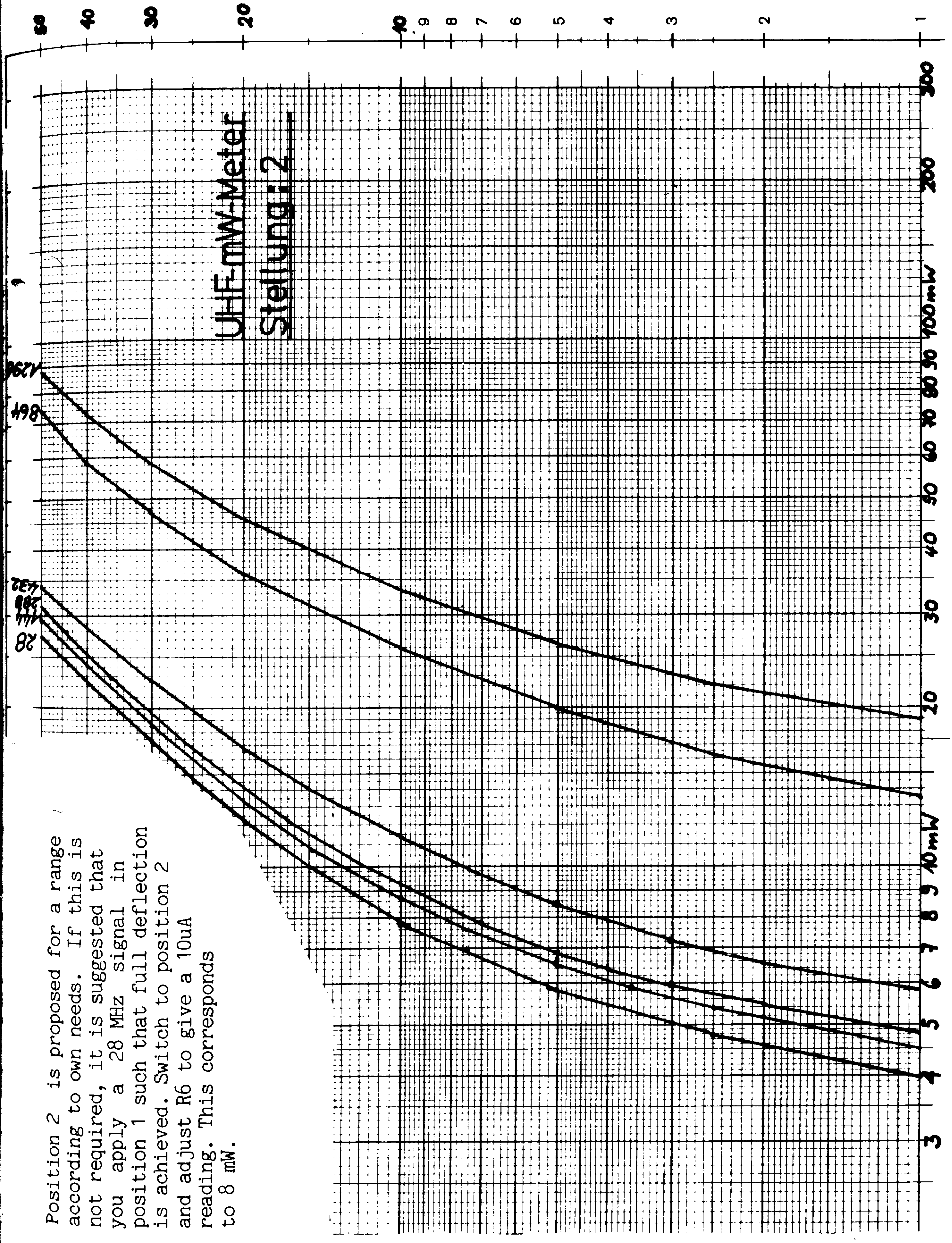


Figure 145
UHF-mW-Meter
Mittelstellung: 1
1296
864
432
288
144
28
300mW
200
100
Chassistecker
Montage an Sucobox-Komponentenbehälter
Suhner Type
US-MIL-Type
Montage-loch
13 BNC-50-0-2
13 BNC-75-0-2
ML 5
ML 5

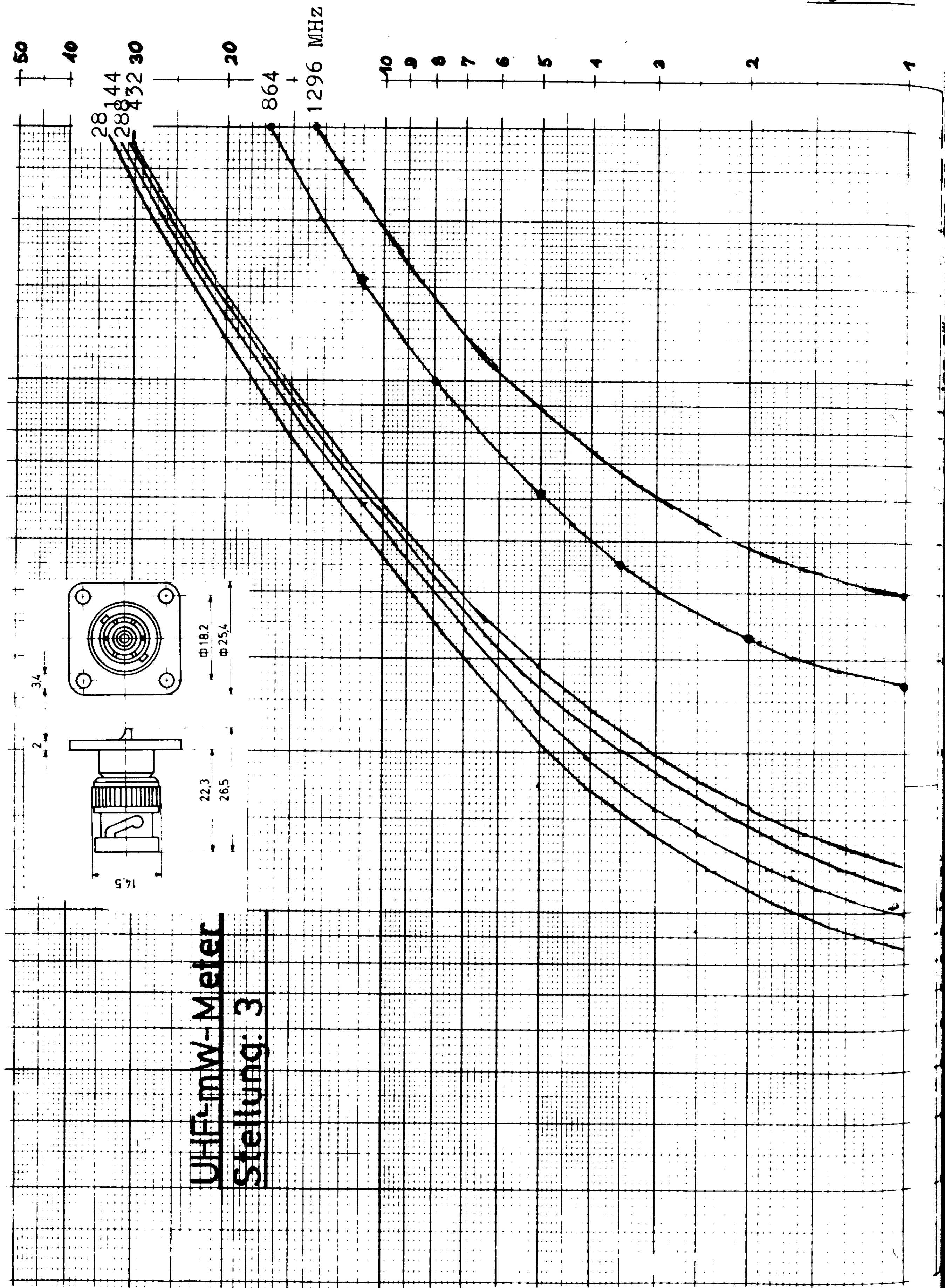
Figure 146



Position 2 is proposed for a range according to own needs. If this is not required, it is suggested that you apply a 28 MHz signal in position 1 such that full deflection is achieved. Switch to position 2 and adjust R6 to give a 10uA reading. This corresponds to 8 mW.

<------ uA reading

Figure 147



<------ uA reading

B.3 Standing wave and power measurements

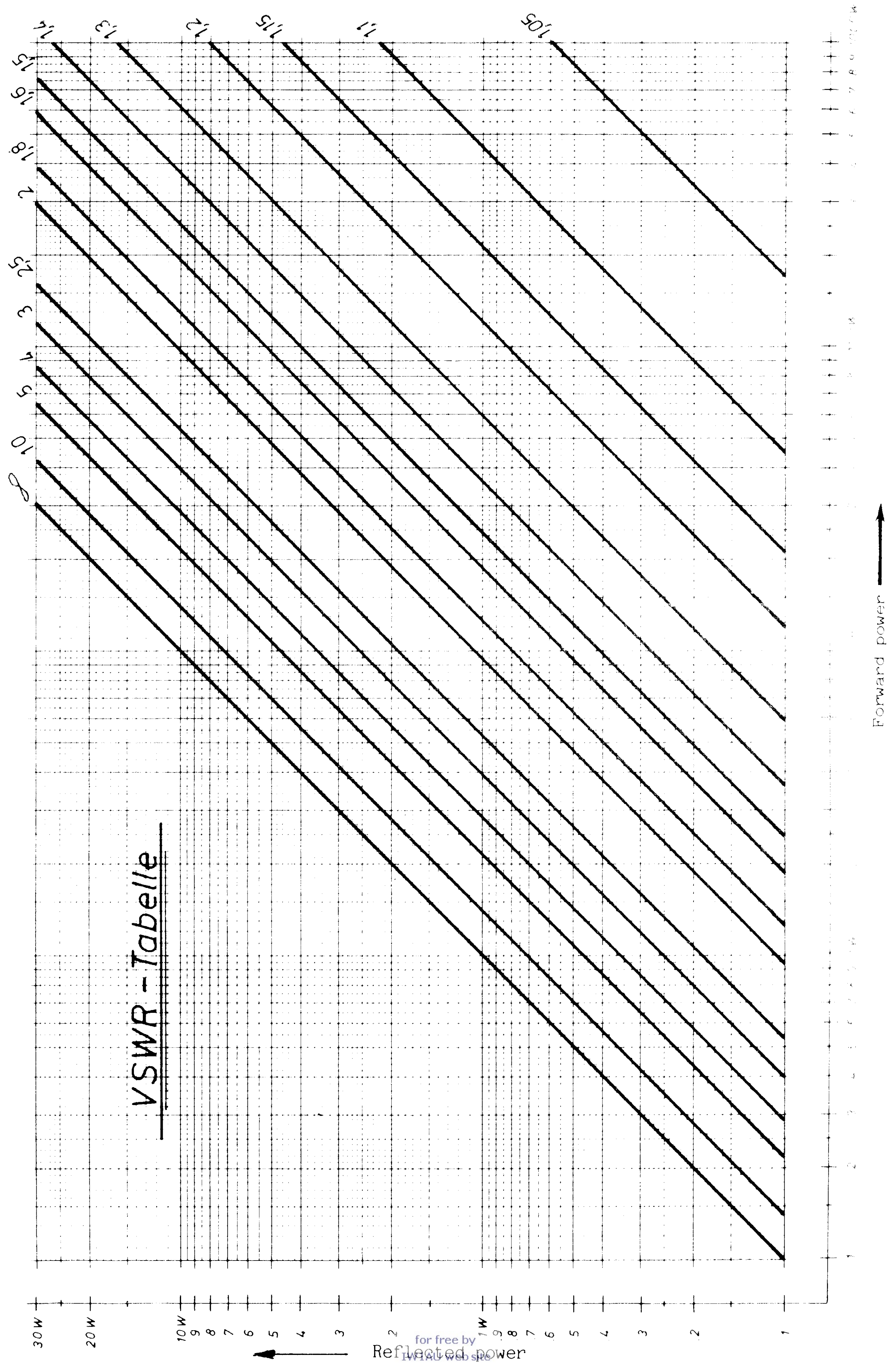
At UHF two techniques are commercially applied to measure RF power. We have to differentiate between feed through and dummy load power meters. In dummy load power meters the RF load is part of the measuring equipment. Over the design frequency range they will constitute a low reactance termination of the usual 50, 60 or 75 Ohm. With this type of matching only a minute part of the RF power will be reflected. The main part of the RF power will heat up the load resistor (real power). This heating effect is used to imbalance a very sensitive bridge circuit. When feeding dc power to the reference resistor of the bridge until the bridge is once again balanced the two powers are equal. Since the dc power can be measured without great difficulty it is thus possible to establish the RF power. Simpler but frequency dependent yet quite satisfactory for amateur application is the method of rectifying and measuring the RF voltage across the load resistor. The power meters described in sections B.2 and B.2.1 are based on this principle; short interconnecting leads and lowest possible inductive and capacitive components of the load resistor are essential.

Feed through power meters are inserted into the connection between the RF source and the actual load (antenna, terminating resistor etc.). They permit the simultaneous measurement of forward and reflected power on some sort of directional coupler principle (B.4). The SWR meter as an essential component of any amateur radio station is a typical example of this type of instrument.

$$\text{Voltage standing wave ratio} = \frac{1 + \sqrt{\text{refl.power / forward power}}}{1 - \sqrt{\text{refl.power / forward power}}}$$

The table on the following page indicates the VSWR (see G.1) for known values of forward and reflected power. It also indicates the reflected power for known values of forward power and VSWR. The voltage standing wave ratio should be measured at the load to avoid errors (B.3.3)

A.0.14 R & S elektronische messgeraete und messysteme 1978 (Rhode und Schwartz).


VSWR - Tabelle
1,4
1,3
1,2
1,15
1,1
1,05
1,5
1,6
1,8
2
2,5
3
4
5
10
∞
Forward power
30 W
20 W
10 W
9
8
7
6
5
4
3
2
1 W
9
8
7
6
5
4
3
2
1
Reflected power

B.3.2 VSWR meter for the 2m, 70cm and 23cm bands

Several different VSWR meters at widely differing prices are commercially available - but hardly any of them will work satisfactorily on 70cm - not to mention 23cm. The majority of manufacturers utilize the directional coupler principle (B.4) and employ two instruments to indicate both forward and reflected power simultaneously. The vast majority of amateurs possesses such a VSWR meter. It was therefore attempted to modify these meters in such a manner that they may be used on UHF without impairing their possibilities in the 2m band. The modification results in a slight reduction of sensitivity which might even be advantageous at power levels beyond 100 Watt.

Building instructions:

Remove the complete coaxial cable assembly including the connectors. Instruments and potentiometers remain unaltered. Install two N-connectors. Mount a piece of coaxial cable according to the desired impedance (50, 60 or 75 ohm) between the two connectors after previously removing the outer insulation. In our case a piece of RG 214/U (50 ohm, see section G.1) was chosen. Thread a piece of insulated twin lead under the outer conductor braiding of the coaxial cable so that entrance and exit are 25mm apart. A knitting needle will come in handy for this operation! The wiring is indicated in figure 148. The outer conductor is brought as close as possible to the N-connectors and is soldered to each of them on two opposite sides. The measuring line could also be installed inside the transmitter or near the antenna connectors.

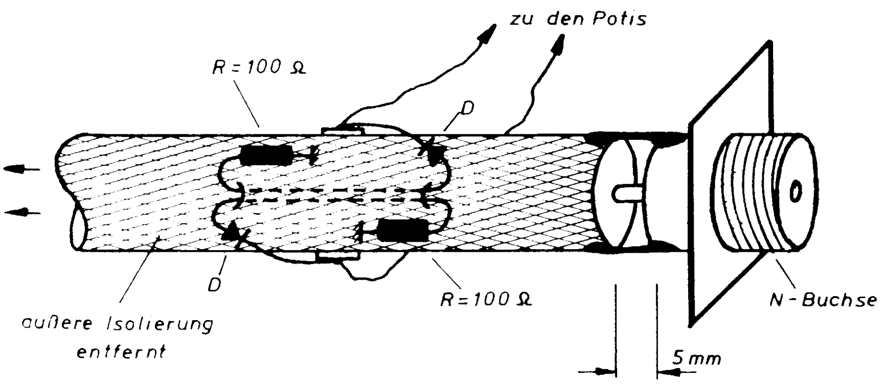


Figure 148
This diagram shows both the component layout and the wiring. The insulation of the coax cable is removed.
R = 100 ohm at 1/4 watt
D = UHF diode, (Schottky, HP5082, HP2805 or equivalent)

The "cold" ends of the resistors are soldered straight to the outer conductor braiding. Following the diodes, trapezoidal or disc capacitors (approximately 1nF) are used for RF blocking and are also soldered as close as possible to the braiding.

B.3.3 Errors in VSWR measurements

Now that we possess a suitable VSWR meter it should be checked prior to carrying out a measurement. For this purpose the output is connected to a UHF terminating resistor of the correct characteristic impedance (Z). The amateur should standardize on one characteristic impedance (50, 60 or 75 Ohm) for all RF-interconnections throughout his station - including the test instruments. Upon the application of RF power the indicated VSWR should be less than 1.1 - even after swapping input and output.

Inserting the VSWR meter

It is convenient to insert the VSWR meter right after the transmitter. However, since the antenna feed line is always subject to losses, the true VSWR will always be worse than the indicated value. The correct sequence would be transmitter - feed line - VSWR meter - antenna. The following example illustrates the generation of measuring errors.

Frequency:	432 MHz
Transmitter output power:	100 Watt
VSWR meter reading:	1.5
Reflected power according to table B.3.3:	4.0 Watt
Attenuation of antenna feed line:	3.0 dB

An attenuation of 3.0 dB implies that only 50 Watt are available at the antenna input. Since the reflected power was measured to be 4 Watts at the location of the VSWR meter and since the one way attenuation was 3 dB a total of 8 Watts must have been reflected by the antenna input. The correct values for entering the table are thus 50 Watts forward power and 8 Watt reflected power. This corresponds to a VSWR of 2.33 at the antenna input.

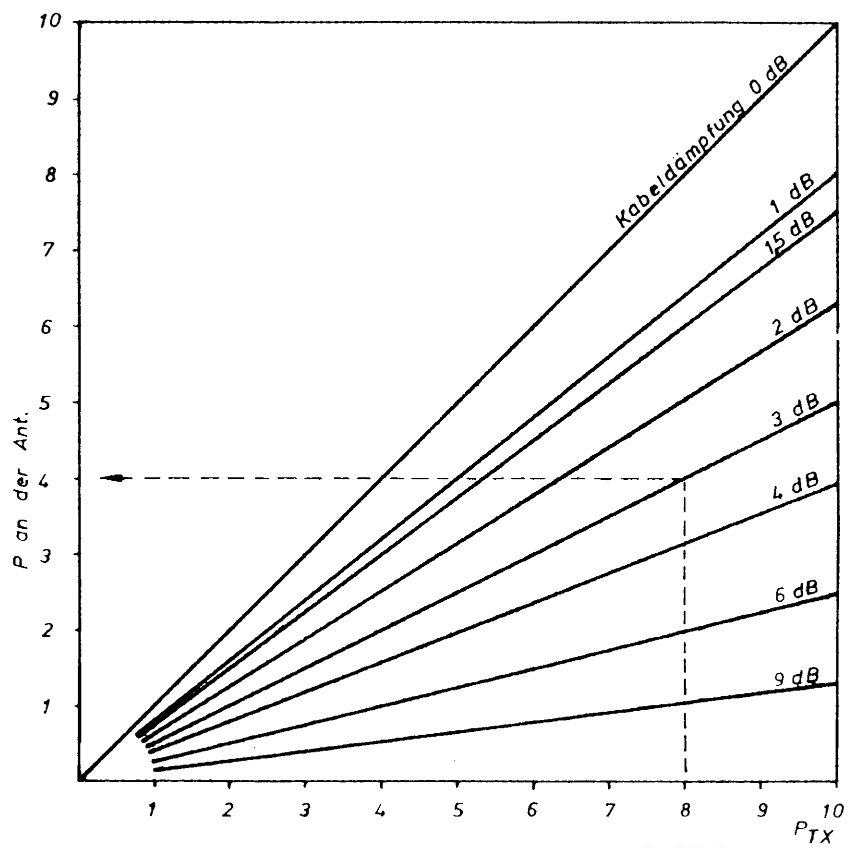


Figure 149
Table for the estimation of power losses due to feeder attenuation.

Worked example
TX-power:
80 resp. 8 Watt

Attenuation: 3 dB

Power at antenna:
40 resp. 4 Watt

B.4 Directional couplers and their application

Directional couplers are finding increasing use in UHF amateur radio. Since the theory of operation has been covered in various publications, only a brief discussion of the essentials is required.

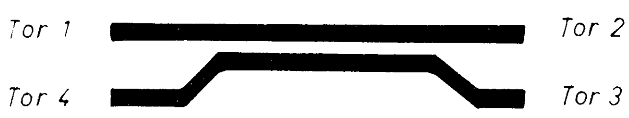
As the name "directional coupler" implies, a defined fraction of the power flowing in a particular direction is split off. To be more specific, directional couplers are used to distribute energy which is travelling to and fro on a RF line either evenly (3 dB coupler) between two ports or unevenly so that only a small fraction is diverted towards a measuring device.

There are many applications for directional couplers. Some typical amateur radio examples are listed below:

1) Power splitting to drive two power amplifiers.
2) Coupling line in VSWR meters.
3) Re-routing of a small fraction of power in a transmitter chain for indication, tuning or control purposes.
4) Insertion of frequency counters and panorama receivers.
5) Detector head for automatic protection of valves and transistors in final amplifiers in case of poor VSWR.
6) Insertion of signal generators and coupling of additional antennas (stacking).

B.4.1 Basic diagram of the directional coupler

Figure 150

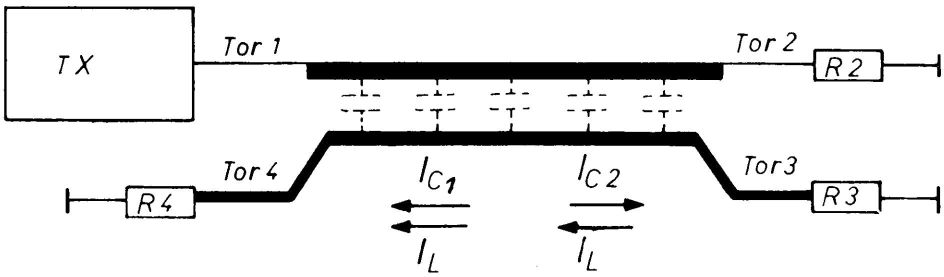


The directional coupler consists of a primary line (port 1 - port 2) and an auxiliary arm (port 4 - port 3). Depending on the arrangement (spacing between the lines) and the length of the auxiliary arm a certain fraction of the power entering the primary line will be diverted. The length of the coupling line (as part of the auxiliary arm) will be a quarter wave length [illegible] or less.

B.4.2 The directional coupler - principle of operation

According to reference (1) primary line and auxiliary arm are coupled both inductively (length of the lines) and capacitively (spacing of the lines).

Figure 151 Equivalent circuit of directional coupler



The distributed capacitance between the two lines gives rise to two currents (IC1 and IC2) flowing in opposite directions along the auxiliary arm whereas IL (caused by the inductive coupling) will flow in one direction only. It is thus obvious that the two currents IC2 and IL will cancel one another whilst IC1 and IL will add up. At port 4 a power level corresponding to the incident (i.e.forward) power reduced by the coupling attenuation will be available.

B.4.2.1 Coupling attenuation

The coupling attenuation (also referred to as nominal coupling or coupling factor) depends on length and spacing of the two interacting lines and is the ratio of input power into port 1 (incident or forward power) and the power available at output port 4 - expressed in dB. Correct termination of all ports is assumed. Shortening the length of the auxiliary line or increasing the spacing between the two lines will result in an increase of coupling attenuation. Coupling attenuation and insertion loss are therefore closely related.

$$\text{Coupling attenuation (dB)} = -\ 10 \log \frac{\text{power at auxiliary port 4}}{\text{incident power into port 1}}$$

B.4.2.2 Insertion loss

The insertion loss depends on the power available at port 4. The following theoretical relation applies:

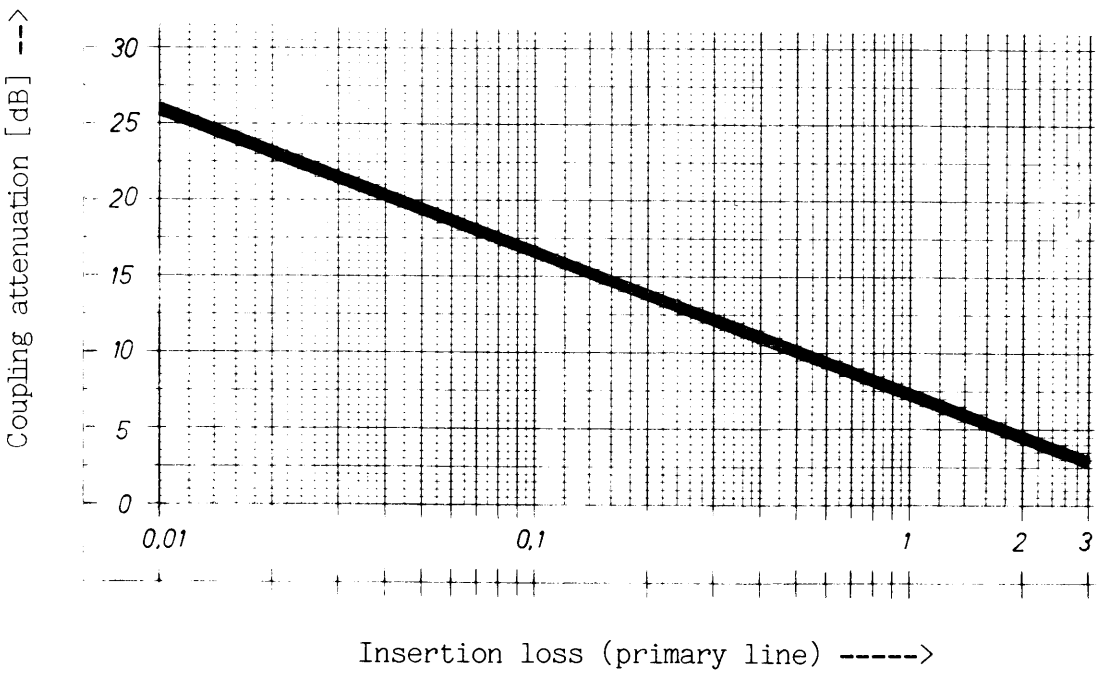
$$\boxed{P_{\text{port 1}} = P_{\text{port 2}} + P_{\text{port 4}}}$$

A comparison of power levels in connection with usual values of coupling attenuation should help the understanding. An input power of 10 Watt into port 1 is assumed.

Theoretical insertion loss	Coupling attenuation	Power level at port 4
3 dB	3 dB	1/2 P1 = 5 Watt
1.2 dB	6 dB	1/4 P1 = 2.5 Watt
0.46 dB	10 dB	1/10 P1 = 1 Watt
0.04 dB	20 dB	1/100 P1 = 0.1 Watt

All other values may be read off the following diagram.

Figure 152 Coupling attenuation as function of insertion loss.



B.4.2.3 Directivity

No power should be available at port 3 (figure 151) if the opposing current components have equal amplitude. Due to asymmetries in the construction - or perhaps unequal amplitudes - actual directional couplers will invariably deliver a certain amount of power to the isolated port. Assuming the transmitter to be connected to port 1 directivity may therefore be defined as:

Directivity [dB] = 10 log (P4/P3)

B.4.2.4 Isolation

Isolation describes the attenuation between input port and isolated port according to:

Isolation [dB] = - 10 log (P4/P1) or:

Isolation [dB] = Coupling attenuation [dB] + Directivity [dB]

B.4.3 Directional coupler construction (single auxiliary arm)

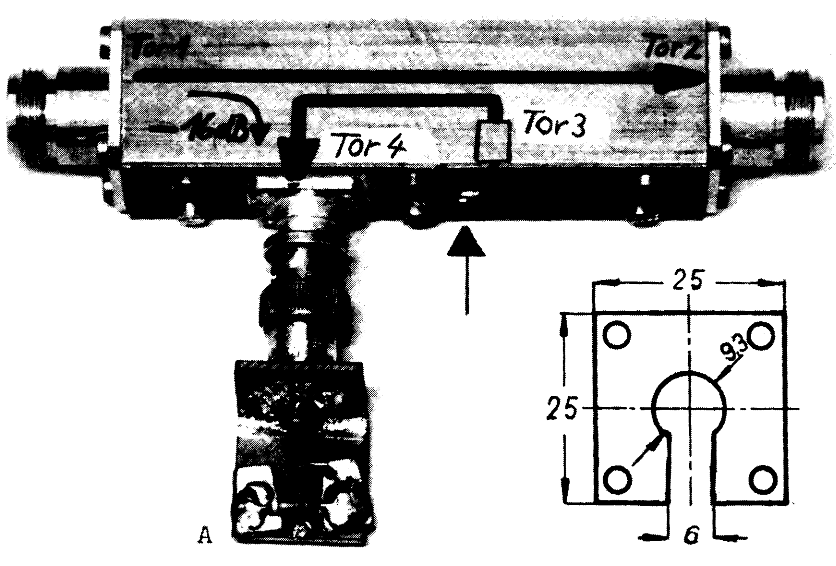
Section B.3.2 contains the description of a VSWR meter with a coupling line representing the simplest possible two - way directional coupler arrangement of acceptable performance. This device is useful in all cases where a relative indication will suffice. The rectified signal - rather than the RF - is available when terminating port 4 in a UHF diode.

This type of construction is prone to line flexing leading to unstable values of coupling attenuation. Precise measurements therefore demand a more elaborate construction.

The home-made directional coupler described below consists of a brass bar 25 x 25 x 90 mm. It was drilled out and slotted to produce the cross section indicated in figure 153. The square faces are drilled out and tapped to permit the installation of the two N-connectors. The inner conductor (Cu, 4mm diam.) is positioned in the center of the 9.3 mm bore and soldered to the connector spigots. The slot is closed up by a cover which is fixed by means of six M3 bolts. The coupling line (LK) consists of a piece of wire (CuAg, 2.5 mm diam.) running from the BNC-connector to the 50 ohm terminating resistor. LK has a length of 20 mm and is positioned parallel to the primary line.

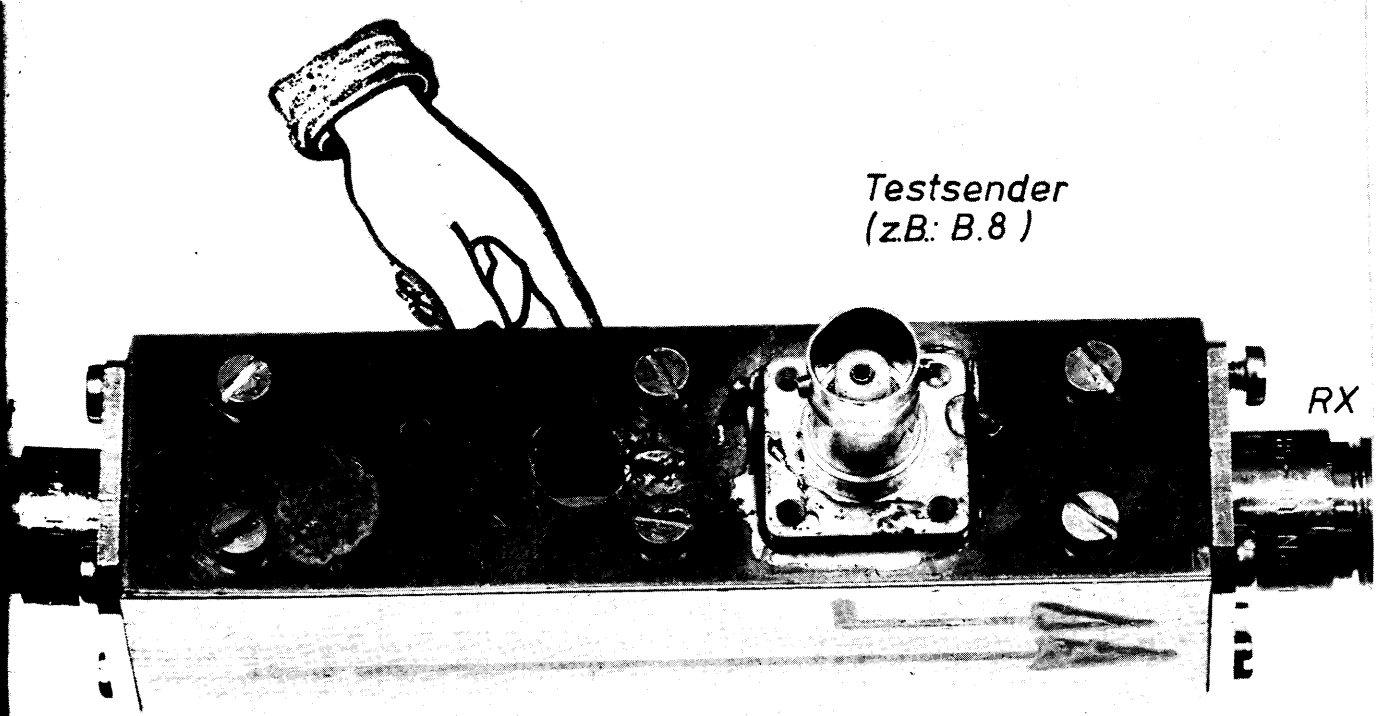
Figure 153

Home-made directional coupler for the frequency range 144 - 1296 MHz. The coupling attenuation is 22.6 dB at 432 MHz and 16 dB at 1296 MHz.



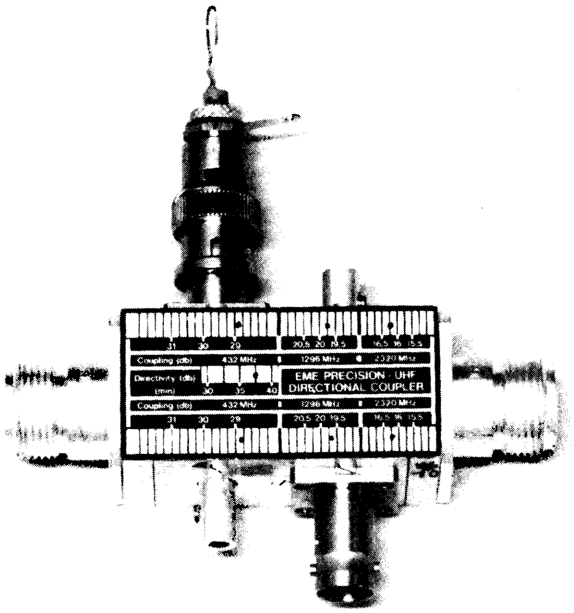
Output port 4 is shown with a detector head already connected. It contains a 50 Ohm termination followed by a Schottky diode, a blocking capacitor 1 nF, a 100 Ohm resistor and an additional 1 nF capacitor to ground. The rectified RF signal (for indication purposes) is available at capacitor "A". Through an opening in the cover (arrow) the coupling can be tightened by forcing LK closer to the primary line or loosened by pulling the free end of the resistor - thus increasing the spacing between coupling line and primary line. In that case the free resistor lead must remain in contact with the edge of the opening. Having adjusted the coupling attenuation by means of a mW power meter at port 4 (and table G.3.1) to the desired value and at the frequency band of interest the resistor lead is bent and soldered (as pointed out in figure 154).

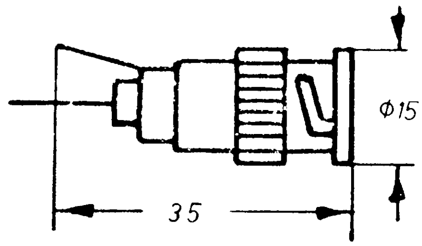
Figure 154
Home-made directional coupler shown in an arrangement for measuring receiver sensitivity. The antenna (Ant) remains connected to the receiver (RX). A signal generator (for instance B.8) is connected to the BNC-connector. The hand-sign points towards the solder joint between the resistor lead and the cover.

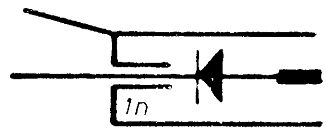


decreased coupling and increased coupling

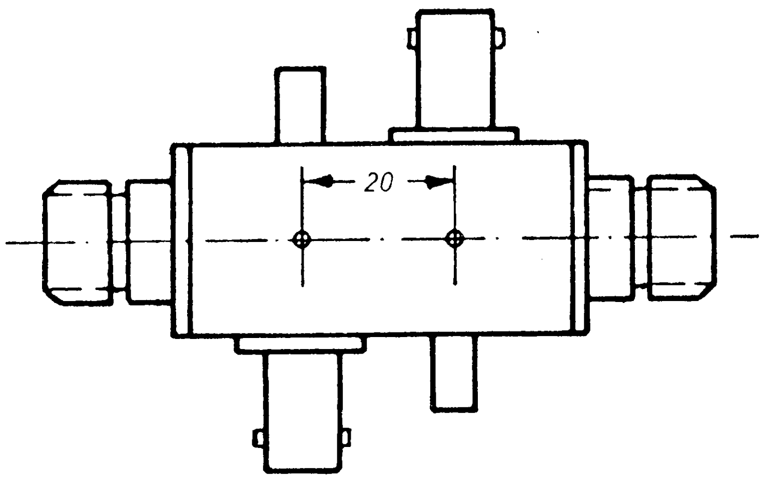
Figure 155 shows a twin auxiliary arm directional coupler as offered by EME (K.Mueller, DC3CT). Frequency range: 432-2320 MHz. The values of coupling attenuation are approximately 30, 20 and 16 db corresponding to the 70 cm, 23 cm and 13 cm bands. The specified directivity is 30 dB. When connecting a BNC-plug into which a diode has been installed to any of the output ports a dc voltage - suitable for indicating or measuring purposes - is available. Otherwise the RF-signal is available at the ports. The photograph shows only one output port being equipped with a detector by EME.

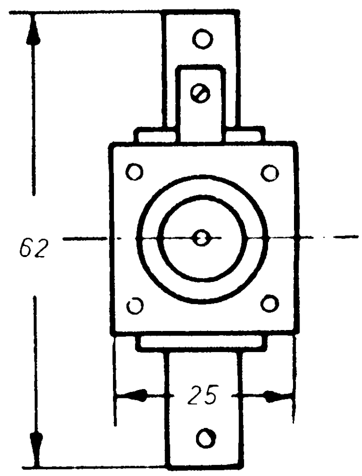


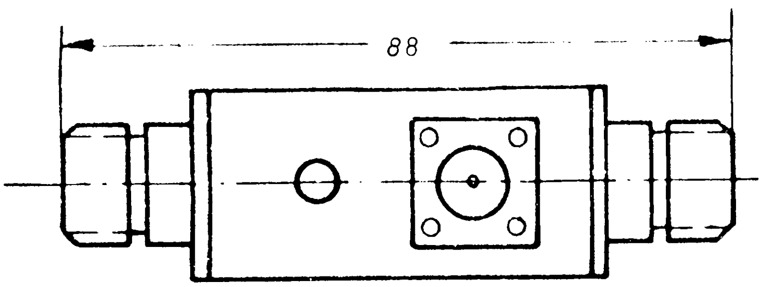


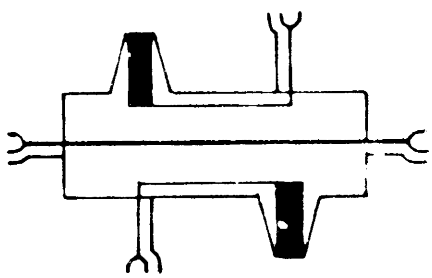


All dimensions in mm.
Drawn to scale 1 in 1









B.5 A low cost panoramic receiver

It is the dream of many an amateur to possess a panoramic receiver. The cost of such a device is , however, usually beyond the hobby budget. As outlined in the following description it requires only comparatively little effort to make this dream come true. In this particular case all amateur bands from 10 metres up to 23 centimetres can be switch-selected and displayed. Upon completion of this project the owner will find instant answers to all of the following questions:

1. What goes on inside the selected frequency band?
2. Rough indication of another station's operating frequency
3. Spurious radiation of a station
4. Type of modulation / operating mode
5. Percentage of modulation
6. Symmetry of frequency modulation
7. Relative signal strength of various stations

Any amateur can call himself a lucky man if he is in possession of an oscilloscope and an old valve-equipped or transistorized receiver for the 10m band. In this case he only needs three things for complete happiness:

one resistor,
one varicap diode and
one capacitor.

B.5.1 The panoramic receiver - theory of operation.

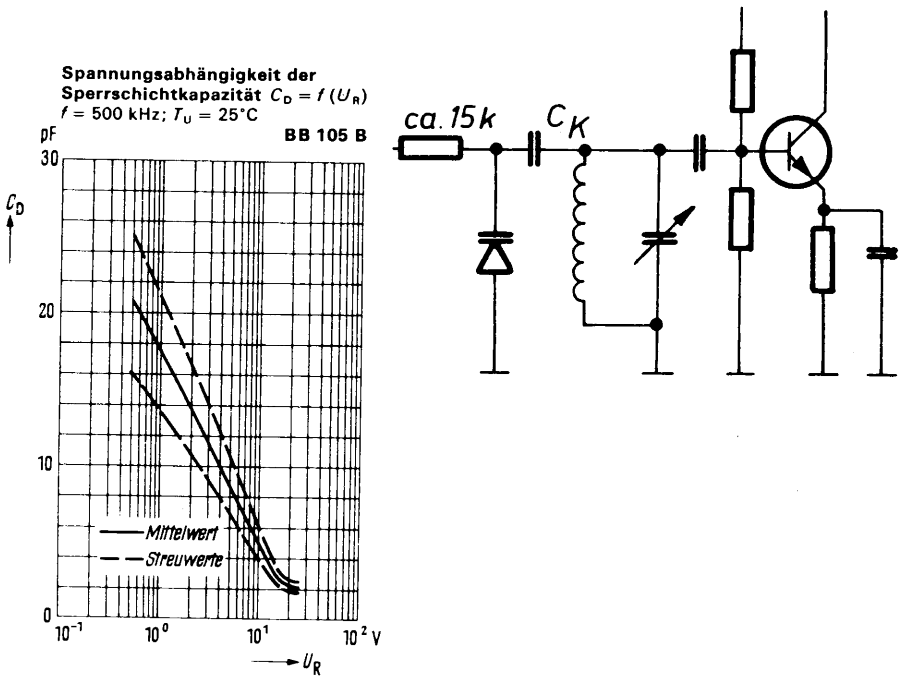
A panoramic receiver consists of two main modules - the receiver and the display unit.
The received frequency (fe) is defined by the local oscillator frequency (fo) and the intermediate frequency (fz). For tuning across the whole frequency band the local oscillator frequency must be swept over a corresponding range of frequencies. Recent receiver designs achieve this tuning by means of feeding a tuning voltage to a varactor diode which is connected across the tuned circuit of the oscillator. The variable tuning voltage will vary the capacitance of this diode and consequently control the local oscillator frequency.The variable tuning voltage could either be adjusted manually or be made to sweep over a certain range.

Such a voltage is produced by a sweep generator. Its output is a voltage that starts from some initial value, increases linearly with time to a maximum value, after which it returns again to its initial value. The time required for the return to the initial value is called the return time, the restoration time, or the flyback time. Since this return time is usually very much shorter than the rise time the output will be a sawtooth voltage.

The relation between applied voltage and capacitance variation of the planar silicon varactor diode is demonstrated by the following diagram. The right - hand side of figure 156 indicates the method of coupling a varactor diode to a tuned circuit.

Figure 156

Junction capacitance (CD) as function of applied voltage (UR).
(Solid line represents typical value, broken line indicates tolerances).
Frequency = 500 KHz, junction temperature = 25 ° centigrade.
The circuit diagram shows the coupling of a varactor diode to a tuned circuit.

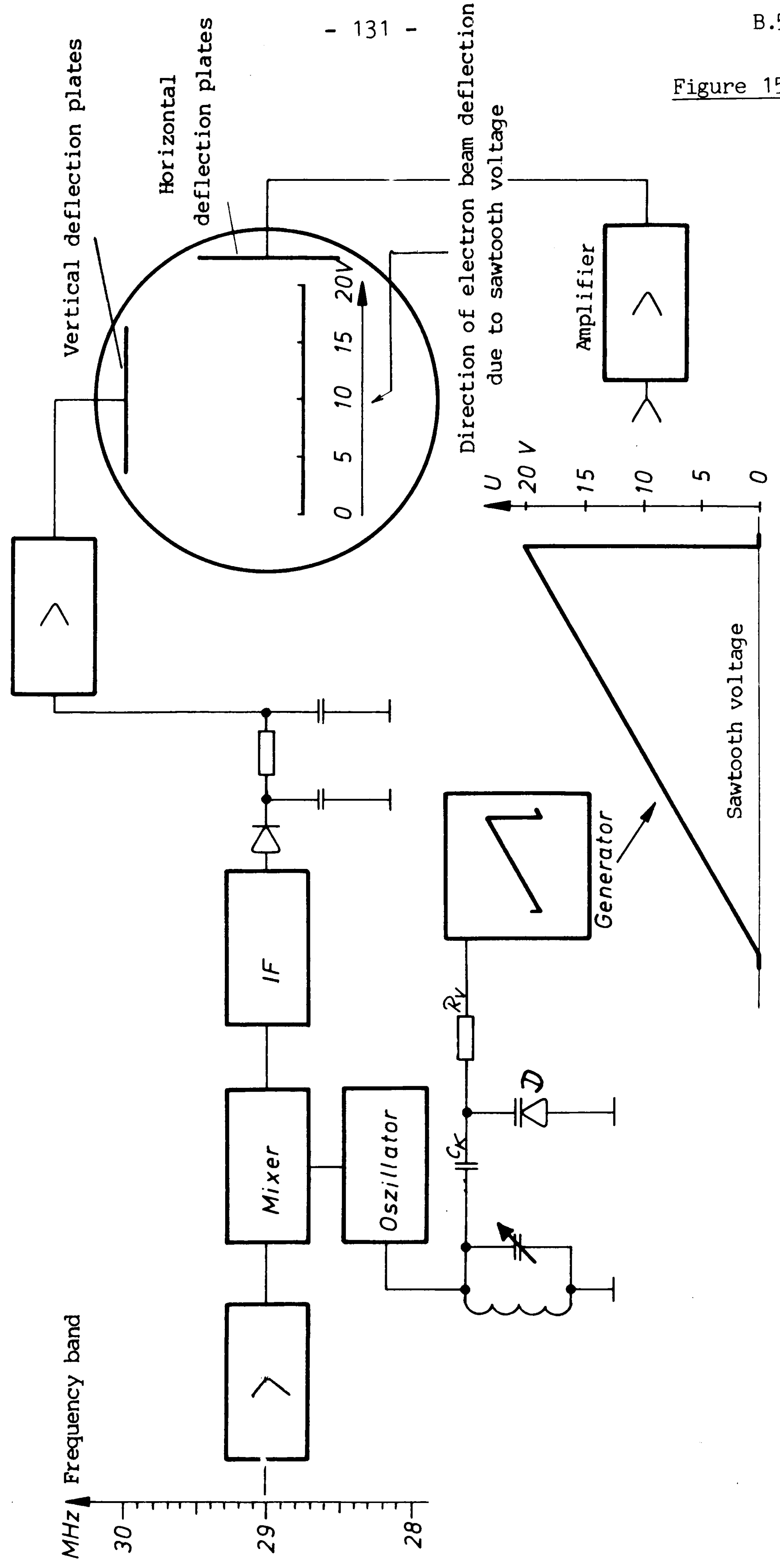


B.5.2 Circuit description

An oscilloscope (in this case a Hameg HM 107) serves as a display unit. The receiver is a 28-30 MHz unit (MB 108 or equivalent) with a varactor diode shunt connected to the variable capacitor of the local oscillator according to figure 156. The sweep voltage is generated inside the oscilloscope where it is used for the horizontal deflection of the electron beam. This sawtooth generator is the central part of our panoramic receiver. Let us first consider the block diagram of such a receiver.

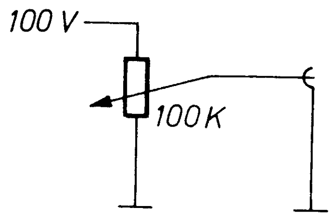
Figure 157

Functional diagram of panoramic receiver



B.5.2.1 Extracting the sawtooth voltage

The voltage which is used for the horizontal deflection is rather high (several hundred volts) and would lead to the instant destruction of the varactor diode. Transistorized horizontal deflection amplifiers will usually supply a suitable voltage. In case of valve amplifiers a suitable voltage should be available at the control grid. In any case, this voltage should not exceed 25 volts. A voltage of approximately 100 volts could be reduced by a potentiometer (100Kohm, 0.25 watt) which is installed inside the oscilloscope. The attenuated voltage is brought out to the front of the scope. Because of instrument inertia the slowest possible sweep frequency should be selected when trying to measure the voltage.



B.5.2.2 Modifying the receiver

As indicated in figure 156 the diode is connected to the hot end of the local oscillator tuned circuit via a 1 nF coupling capacitor. This will lower the oscillator frequency. This may be compensated by readjusting the inductance. A screened cable (audio frequency, both ends grounded) is used to feed the sweep voltage (25 volts max.) via the 15 kohm resistor to the tuning diode. This resistor prevents oscillator signal leakage. The demodulated IF signal (receiver in AM-mode) is routed through a screened AF cable and fed to the vertical input of the oscilloscope. Its internal Y-amplifier is quite sufficient to amplify the demodulated signal and to deflect the electron beam along the Y-axis.

B.5.3 Signals deflecting the electron beam

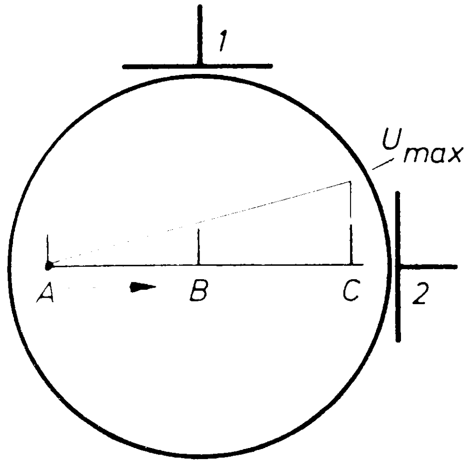


Figure 158
With no external voltage supplied the electron beam will illuminate a spot in position "A".
A sawtooth voltage applied at deflection plate 2 will cause the beam to move along the X-axis, reaching position "B" with approximately half, and position "C" with the full voltage intended for this purpose.

Referring to the sawtooth voltage applied to the tuning diode this implies:

a) Low voltage = high capacitance (CD) = lower band limit
b) High voltage = low capacitance = upper band limit

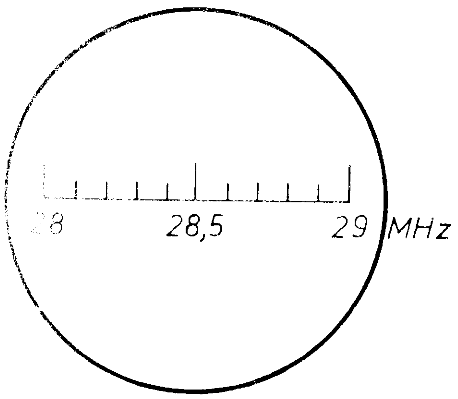


Figure 159
This variable shunt connected capacitance will therefore permit the tuning of the desired range of frequencies. Adjusting the sweep range to cover 1 MHz and setting the lower band limit to 28 MHz by adjusting the oscillator inductance, one will have a panoramic receiver covering the 28 to 29 MHz band. A signal inside this range will be demodulated and supplied to plate no.1.This will cause the beam to be deflected in Y-direction. The amount of deflection is a measure of the signal amplitude and the position will indicate the signal frequency.

B.5.4 Frequency range, signal width, deflection time

Upon removal of the variable capacitor only the capacitance variation (ΔCD) of the diode will define the range of frequencies (Δf) that can be received. Δf is influenced by the following factors:

1. ΔCD according to sawtooth voltage provided.

2. ΔCD of the diode. Different types of diodes exhibit different capacitance properties when subjected to tuning voltages. Diodes employed at low frequencies result in wider frequency ranges than those that are employed at UHF. The type BB 105 diode used in one of the trials permitted the display of a 200KHz section inside the 10m band. Preceded by a 2m converter, and upon adjusting the variable capacitor of the local oscillator, practically the whole range of repeater frequencies could be displayed - to give an example.

3. Δf_0 of the receiver. A 28MHz band oscillator shunted by ΔCD can only be swept across a comparatively narrow range of frequencies. The 135MHz oscillator (of a 2m receiver) shunted by ΔCD will, however, tune across several MHz.

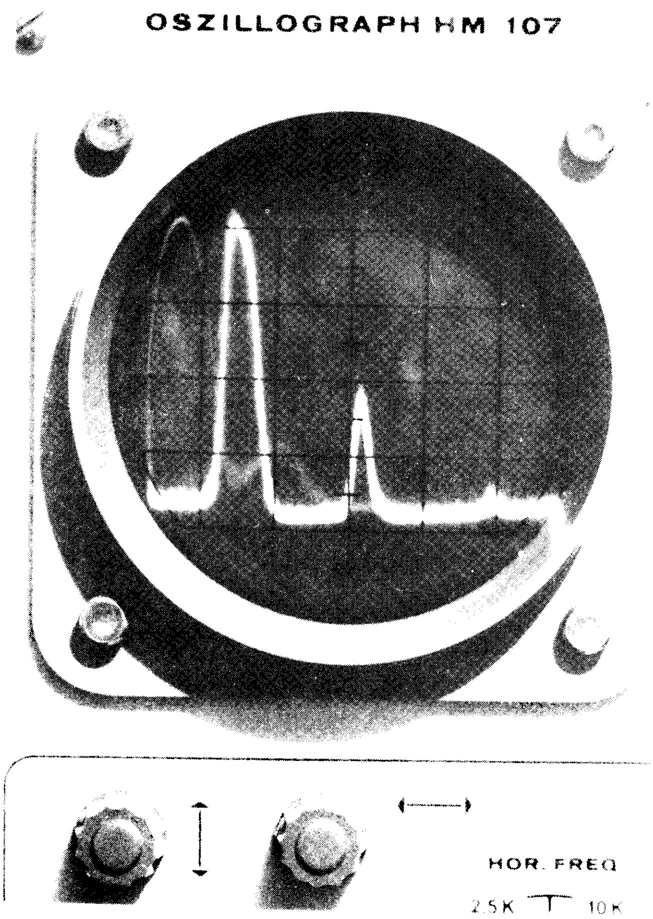


Figure 160

Presentation of repeater output signals on the display (HM 107) of the described panoramic RX. THe vertical deflections represent the frequencies 145.8, 145.75 and 145.7MHz (from left to right).

Signal width

The width of the displayed signal depends essentially on the bandwidth of the intermediate frequency filters. The use of a crystal filter will result in steeper edges of the displayed signals.

Deflection time

The deflection time is adjusted by means of the "Sweep frequency", "Sweep rate" or "Horizontal frequency" depending on the oscilloscope used. The sweep frequency should be around 50 to 100 per second. From 40 sweeps per second onwards the movement of the beam is no longer noticable and a steady line apears on the scope. The desired frequency band as defined by CD, CK and fo is scanned at the same rate. A faster sweep rate is neither necessary nor is it recommended since ringing of steep skirted filters could occur.

The tunable IF strip 28 to 30 MHz (MB108 as produced by SEMCO several years ago) is well suited for this purpose - as is any other 10 m RX of reasonable quality. Combining it with any converter designed for an intermediate frequency of 28 to 30 MHz will permit the supervision of any desired amateur band or range of frequencies.

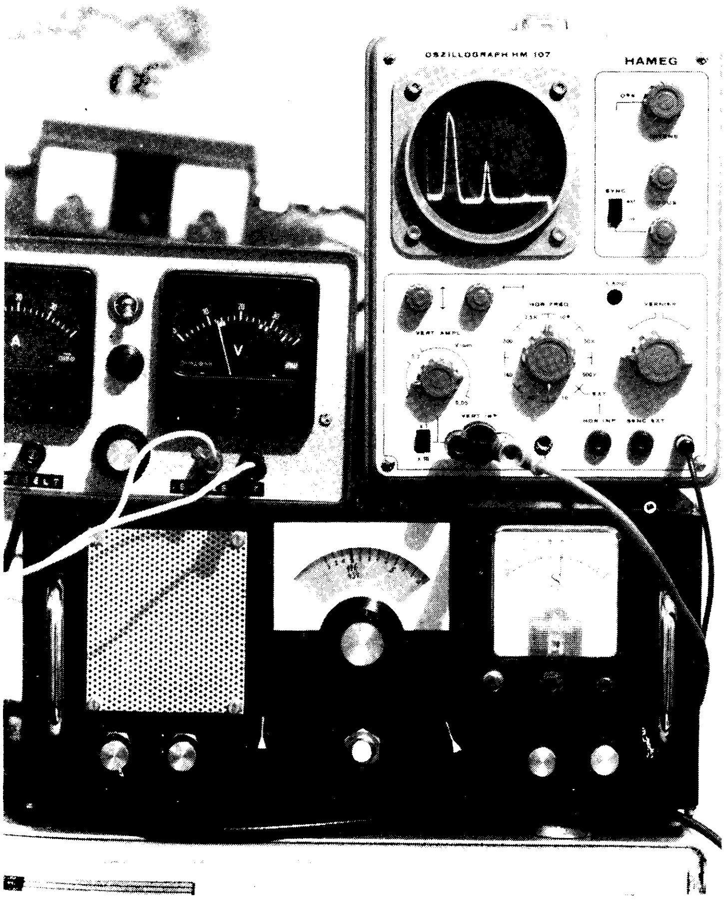


Figure 161

The photograph shows the arrangement at DJ 9 HO. The receiver is a modified portable station with converters for the 2m, 70cm and 23cm bands installed.

Possible applications are suggested in section B.6. In this particular case the sweep range is only 60 KHz. It may be extended by adding further varactor diodes in parallel. Suitable diodes are all devices designed for UHF or VHF tuning purposes. (BB 105, BB 109g or equivalent). Increasing the sweep range will narrow the vertical deflection and make the signals look like spikes.

B.6 Suggestions for using the panoramic receiver

Apart from the suggestions made in section B.5 such as instant information on band activity, transmit frequencies and mode of operation of stations working inside the selected frequency range this concept could be used as tuning aid, modulation monitor or for band surveillance. When connecting a converter to the 10 meter panoramic receiver it will immediately become apparent to which part of the band the converter tuning was optimized.

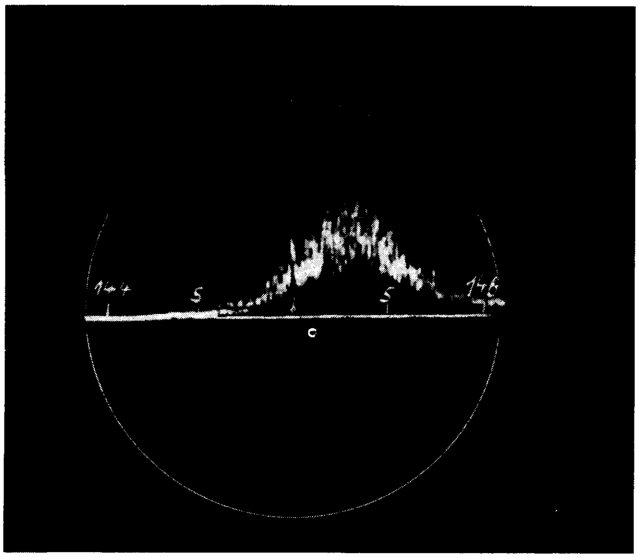


Figure 162

It can be seen that the converter has maximum gain around 145.4 MHz. Broad band tuning is reached when the noise level as indicated by the vertical deflection is constant over the whole width of the display.

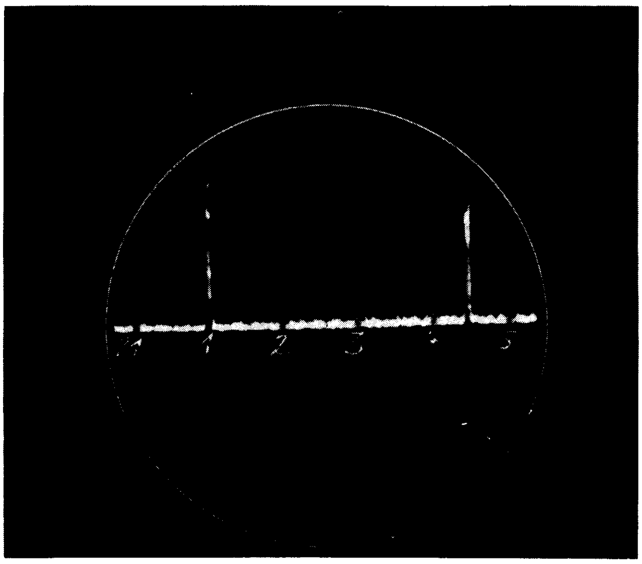


Figure 163

A test set up was used to display the activities on the 21 MHz band. Two signals of differing field strength can be seen. A SSB or CW signal will only appear when speech is transmitted or when the transmitter is keyed. Rotating the antenna beam direction with a panoramic receiver connected to it is a new and exciting experience!

The application as modulation monitor scope is demonstrated taking a FM signal as an example. The reader is reminded of the fact that in a frequency modulated signal all information is conveyed in the variation of the carrier frequency. Interfering signals such as ignition sparks etc. that might exist on top of the carrier are removed by the limiter being part of the receiver. This assumes that the rf signal delivered to the receiver is sufficiently strong.

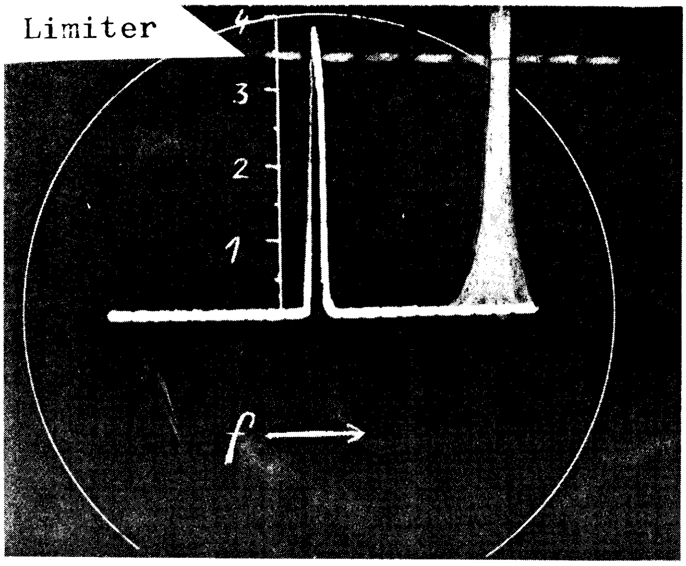


Figure 164

An unmodulated carrier is shown. Assuming the limiter to have a threshold corresponding to 3.5 on the scope additional interfering signals are cut off. The representative IF bandwidth of a NBFM receiver is indicated to the right (ΔZF).

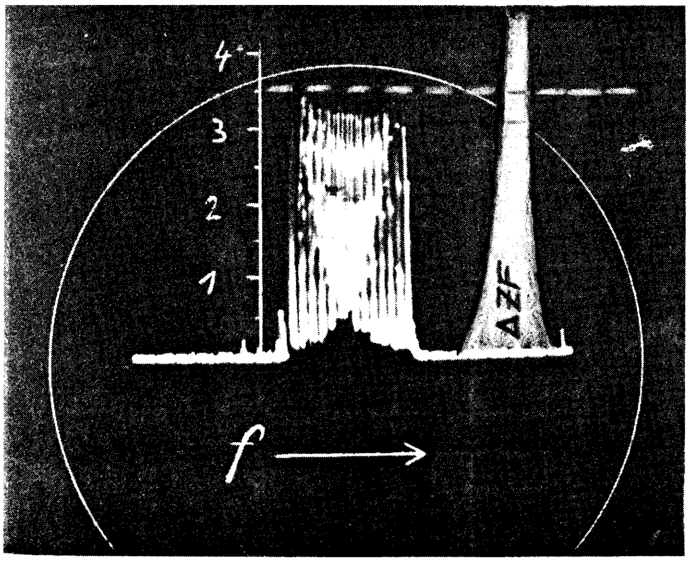


Figure 165

The above mentioned carrier is now modulated by a single tone, the deviation being 50 KHz. The amplitude reduction indicates that the transmitter power is now spread out over the entire spectrum. The narrow band receiver will interpret this as a reduction in signal strength. The effect of the limiter (threshold at 3.5) will be lost.

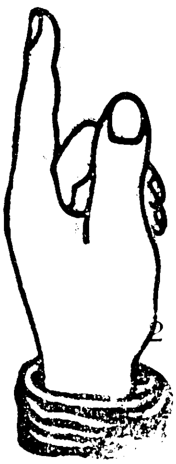
In contrast to AM and SSB a higher voice level will not lead to a higher field strength at the receiver. As experienced many times in repeater operation a lower voice level could occasionally lead to improved filling of the receiver pass band.

B.7 Discarded TV-set as panoramic display

The realisation of the panoramic receiver concept according to section B.5 requires - apart from other things - an oscilloscope as display unit. Not all amateurs are in possession of such an instrument. To give all amateurs the chance of assembling a panoramic receiver a different concept had to be - and was - found. This involves the use of a discarded TV set which, after slight modification, will fulfil the functions of the display unit and the sawtooth generator.

W A R N I N G ! H I G H V O L T A G E ! D A N G E R !

The modification of the TV-set is a simple task; it does - however - involve certain risks and must be performed by a TV-specialist of the local club.



Reason:

1. Early TV-sets are not equipped with a mains transformer. Depending on the sense of inserting the mains plug into the wall socket the TV-chassis could be connected to the life phase. The remedy is an isolating transformer. Otherwise voltage differences up to the mains voltage could be present when interconnecting the various components of the panoramic receiver. This could result in damage to components, short circuits or in the operator serving as the earthing strap!

2. Apart from the mains voltage that could exist at the chassis there are other very high voltages (beyond 10 KVolt) which could lead to lethal electric shocks when carrying out the modification in an unprofessional manner.

B.7.1 Modifying the TV set

The sweep frequency of the electron beam inside the cathode ray tube is horizontally 15626 Hz and vertically 50 Hz - resulting in 625 lines completely covering the face of the tube. This is undesirable for the display unit of the panoramic receivier. As mentioned in section B.5 a sawtooth voltage having about 50 Hz is required for the horizontal deflection. At present this type of deflection is provided for the vertical sense. This is a simple problem. To start with, the two wires leading to the horizontal deflection coils are unsoldered and connected to a spare deflection coil acting as a dummy load. This could be taken from another old TV set and is finally mounted inside the set selected for the modification.

Alternatively a wire wound resistor (10 Ohm, 20 Watt) could be used. In the case of transistorized TV sets the resistor should suffice. Having completed this task the horizontal deflection will be missing - leaving just a vertically deflected narrow electron beam.

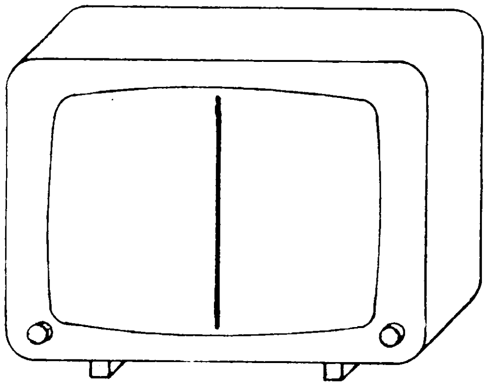


Figure 166
Upon disconnection of the horizontal deflection coil only the vertical deflection remains operative.

Next, the deflection coil assembly is unclamped and rotated through 90° until the electron beam paints a horizontal line onto the screen. The sense of rotation defines whether the lower limit of the displayed frequency band is to the right or to the left. Obviously this is a personal choice and can be altered at any time by rotating the coil through 180°.
Further tasks are:

- to extract the sawtooth voltage from the TV set and
- to install a dc amplifier.

B.7.1.1 Extracting the sawtooth voltage

The 50 Hz sawtooth voltage that was used for the previously vertical - and now horizontal - deflection is tapped off the oscillator amplifier stage via a 100 KOhm resistor and fed to point "e" of the network according to figure 167. Output "d" will have to be connected to the Varicap diode of the receiver.

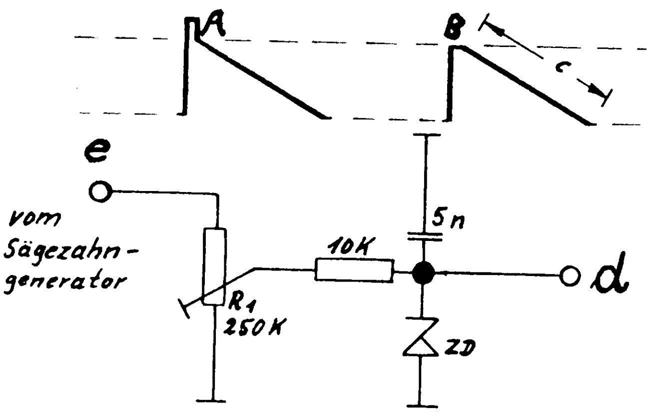
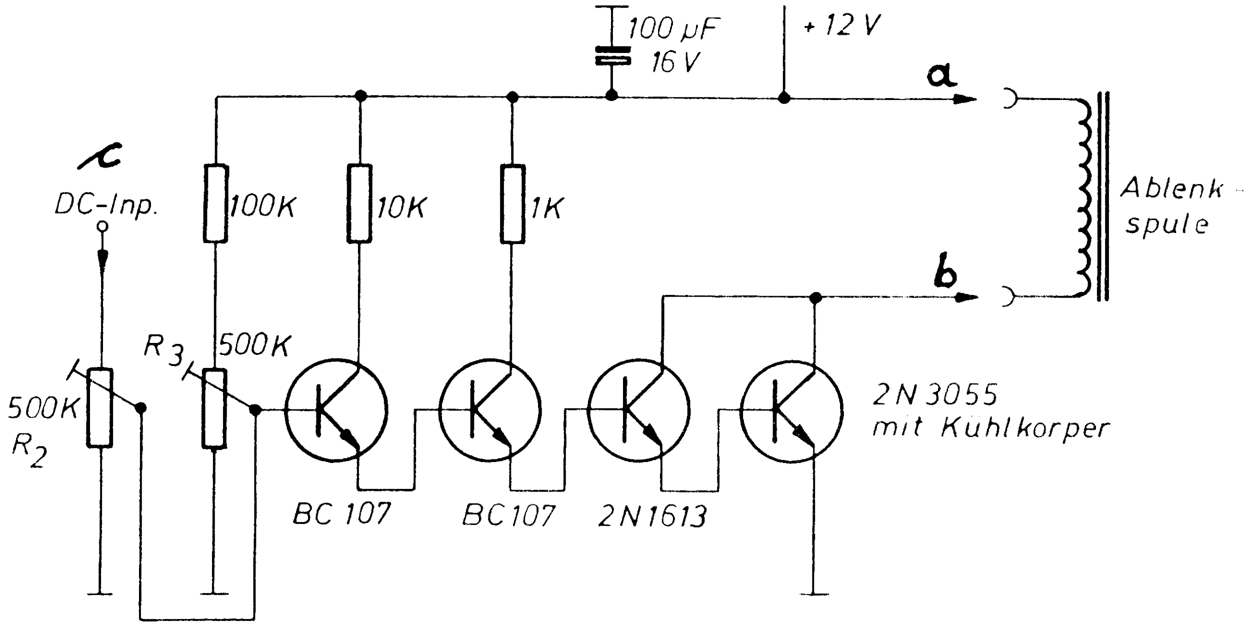


Figure 167
R_1 reduces the voltage to the desired value as defined by U_{ZD}. The Zener voltage should have a value of between 10 and 25 Volts - depending on whether a 2 m or a 10 m receiver is used.
A = sawtooth wave form as supplied by TV set.
B = Wave form with correct setting of R_1.
C = Part of wave form used for controlling ΔC.

B.7.1.2 Installation of the dc-amplifier

In a TV set the previously horizontal deflection of the electron beam requires a power level of typically 10 to 12 watts. Obviously the power delivered by the demodulator following the IF amplifier will be considerably less. Thus a dc-amplifier is required to raise any small signal to a power level sufficient for the vertical deflection of the electron beam. This is achieved by four transistors arranged according to the following circuit diagram:

Figure 168 Deflection amplifier DC-V1



It is recommended to instal the amplifier inside the TV set. Two screened cables will connect the receiving section with the display module: The first cable (l_1) connects point "d" of the network module through a 10 Kohm resistor with the varicap diode that had been installed in the local oscillator. The second cable (l_2) feeds the demodulated IF signal to the dc-amplifier. It is advisable to use plugs and sockets with additional ground connections to be able to modify the concept if so desired. The linearity of the display depends on the wave form of the sawtooth voltage and on the diode characteristics. If all rf-carrying components (converter, receiver) have constant gain throughout the displayed frequency range then the vertical deflection of the display can be calibrated in 10 dB steps. The step attenuator (B.1.2) comes in handy for this task. The amplifier board (DC-V1) is again laid out to fit a case according to section D.6. The subsequent diagrams (figures 169 and 170) depict pc-board and component lay-out.

The concept based on the TV set can be employed to automatically indicate exceptional propagation conditions. By means of steel strip several photo diodes are mounted to the TV set such that they may be adjusted both horizontally and vertically. When placing the diodes in a predetermined height (= field strength) above the weak background noise and in horizontal positions corresponding to selected beacon frequencies then the electron beam will reach the diodes as soon as the conditions improve.

An amplifier following the photo diodes and driving a tape recorder or a buzzer will alert the amateur or perhaps make him leap out of bed.
Should the deflection occur downwards terminals "a" and "b" should be changed around.
That's all! Up to this moment several versions of this arrangement have been successfully built and comparable systems are commercially available. So, roll up your sleeves and get cracking!

Figure 169 Print side of pc-board (by DL5RAD)

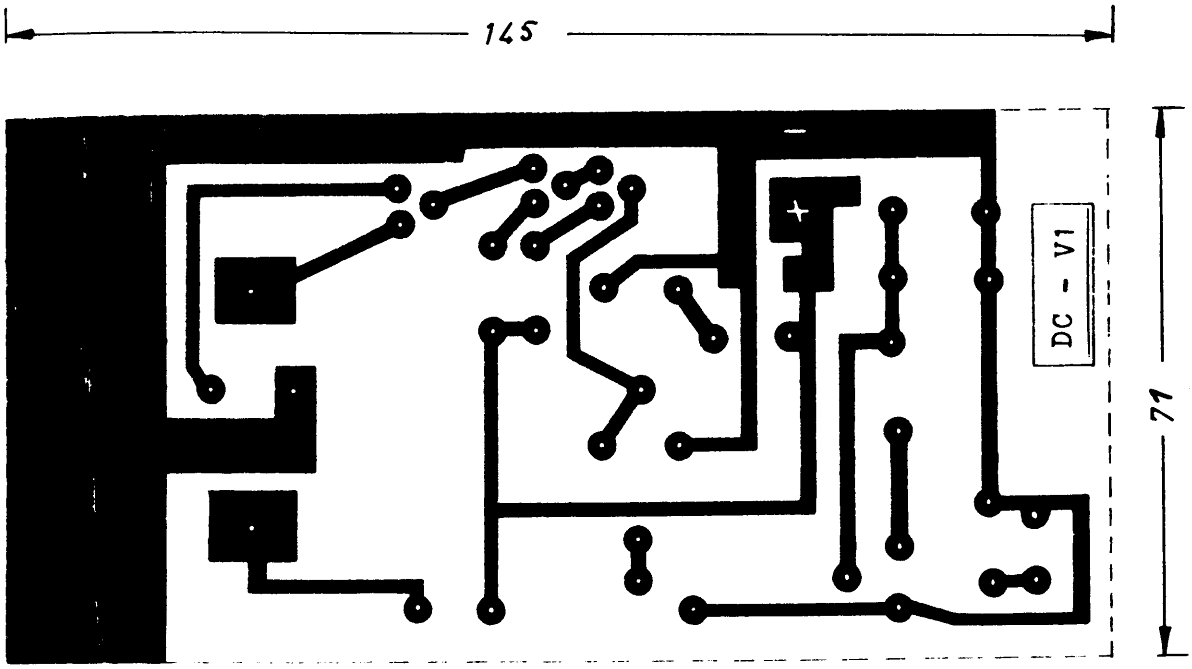
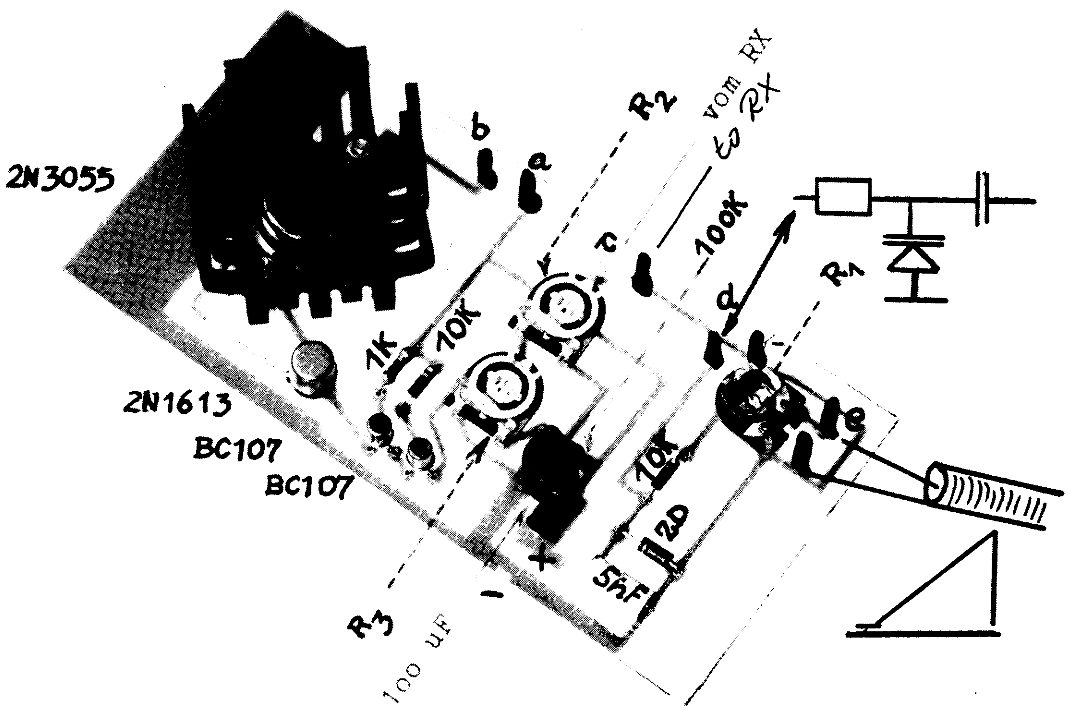


Figure 170 Completed amplifier DC-V1



R2 = sensitivity control
R3 = limits of vertical deflection of display

B.8 Simple fixed frequency generator for RX alignment up to 5.6 GHz
(M. Kuhne, DB6NT)

For the alignment of receiving systems beyond 144 MHz the amateur has to rely on beacons or signal generators. If neither of these signal sources was available it used to be difficult to optimize a receiving system - in particular on the VHF, UHF and SHF bands.
The crystal oscillator (28.8 MHz) as built by OM Michael and described below represents a fixed frequency signal generator. Since it produces harmonics that may be controlled by means of a built-in attenuator it is well suited for alignment and optimization of converters and receiving systems on the amateur bands between 2m and 6cm.
For continuous monitoring of the receiver sensitivity it could remain connected to the receiver input via a directional coupler inserted into the antenna cable and be switched on as required.
Since there is a fair chance of tuning to the image frequency when trying to use a noise generator for alignment purposes this module with its stable set of frequencies will be a great help.
Upon adjustment of the coupling loop such that a 2m receiver (144.0MHz = 28.8MHz x 5) MHz indicates a signal level P_e = 45dB above noise then the following power levels are to be expected:

2 m	45 dB
70cm	47 dB
23cm	30 dB
13cm	20 dB
9-6cm	12 dB (depending on receiver sensitivity)

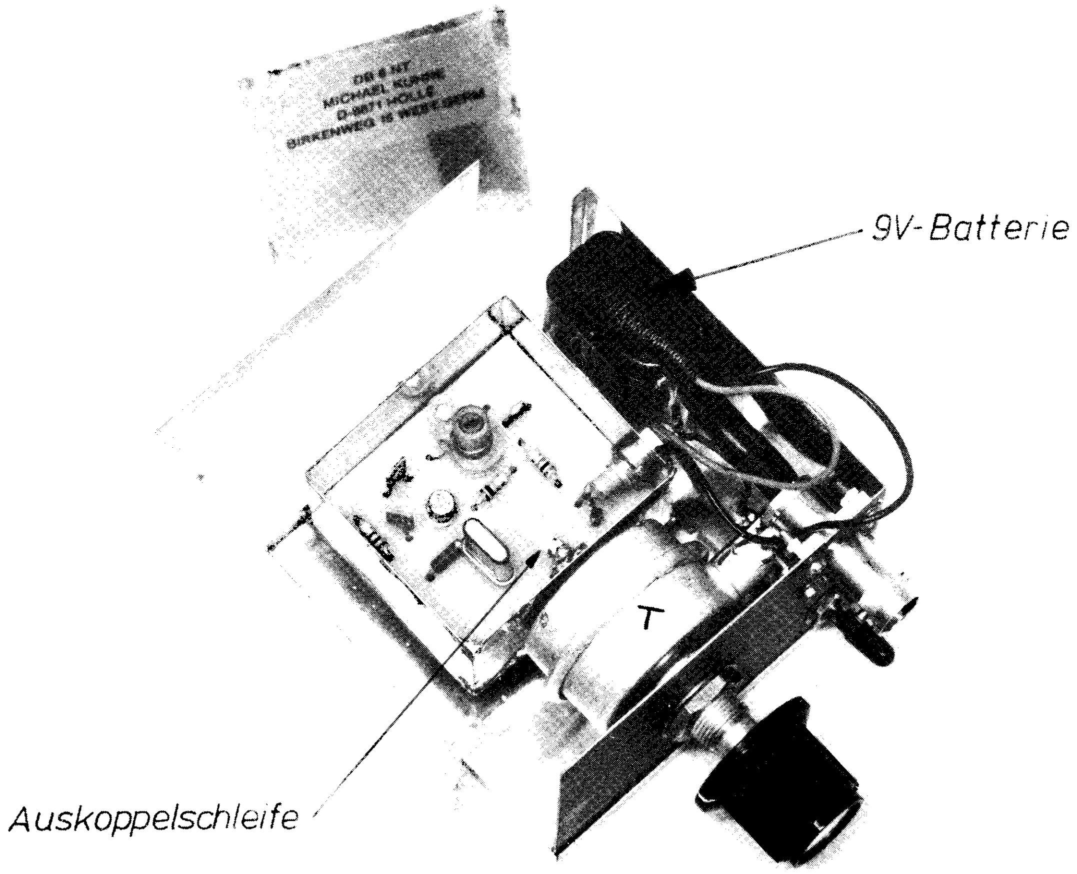
By means of the built-in adjustable attenuator (Preh, suitable for frequencies up to 1 GHz, supplier: NEUTRON) signals may be attenuated to inaudible level. If the oscillator was calibrated against a precise frequency counter then accurate frequency markers up to 5670 MHz (28.8 x 200) are available.

Construction:
To avoid stray radiation the oscillator (with transistor T1 = 2N981 by TEXAS Instruments) is surrounded by two cases. The inner case houses the oscillator board. The outer case encloses the inner case, the 9 Volt battery and the attenuator. The oscillator circuit - having proved its qualities in D.6.2 - works reliably and relatively frequency stable with supply voltages of between 7 and 15 Volts without additional stabilisation.

Crystal:
OM Michael employes a series resonant crystal, 28.8MHz, HC-18/U
Supplied by:

SSB-Elektronik
Karl-Arnold-Str. 23
D-5860 Iserlohn

Figure 171 Construction of fixed frequency generator
The photograph shows the oscillator compartment, the coupling arrangement, the power supply (9 Volt battery), the output connector, the on/off switch and the outer housing (72 x 72 x 50 mm).
The attenuator input is tightly fitted to an opening in the oscillator compartment and soldered from the inside. Upon verification of the correct oscillator operation the compartment is closed and soldered along all edges. Afterwards the sides of the outer housing are soldered and the two covers are installed. Prior to this the length of the coupling vane had to be adjusted to the intended power level. In this particular case it measures about 10 mm in length.



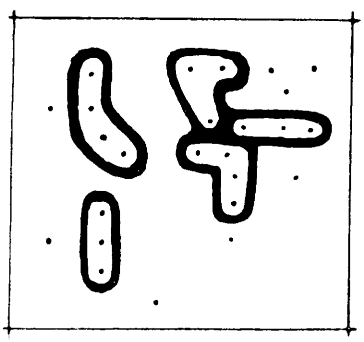


Figure 172
Printed ciricuit of the simple fixed frequency generator

As shown by the photograph copper cladding is required on one side only.
Again it is stressed that this circuit works well with the 2N918 and the reader is advised to stick to this type.

Figure 173 Circuit diagram of fixed frequency signal generator
The box marked "T" represents the adjustable attenuator by PREH. The input terminal of this attenuator protrudes into the oscillator compartment. Here a 50 Ohm resistor serves as matched termination. The coupling vane soldered to the input defines the energy that is drawn from the oscillator. Increasing its length will increase this energy. Thus the energy may be set to satisfy one's particular requirements.
L_1 = 10 turns on 6mm diam. coil former with core (tuned to f_Q).

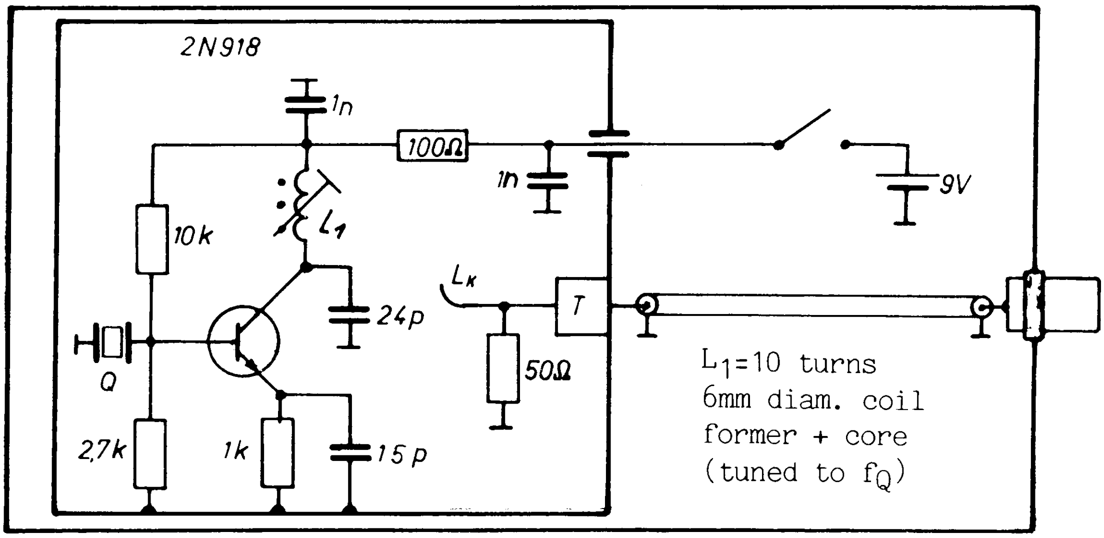
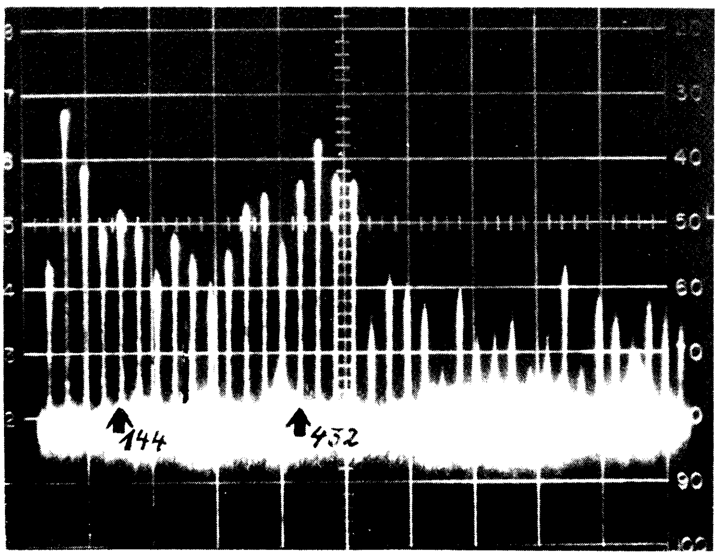


Figure 174 Frequency spectrum as generated by the fixed frequency generator between 1 and 1000 MHz. Clearly, there is a significant reduction in power level of the higher harmonics. The frequencies 144 and 432 MHz have been marked. All others may be computed (f_Q x 2,..3,..4 etc.).
The coupling vane should be lengthened if the module is predominantly used for adjusting equipment for the 23cm band or even higher frequencies.



Range: 1-1000 MHz

Scale:
horizontal = 100 MHz/div.
vertical = 10 dB /div.

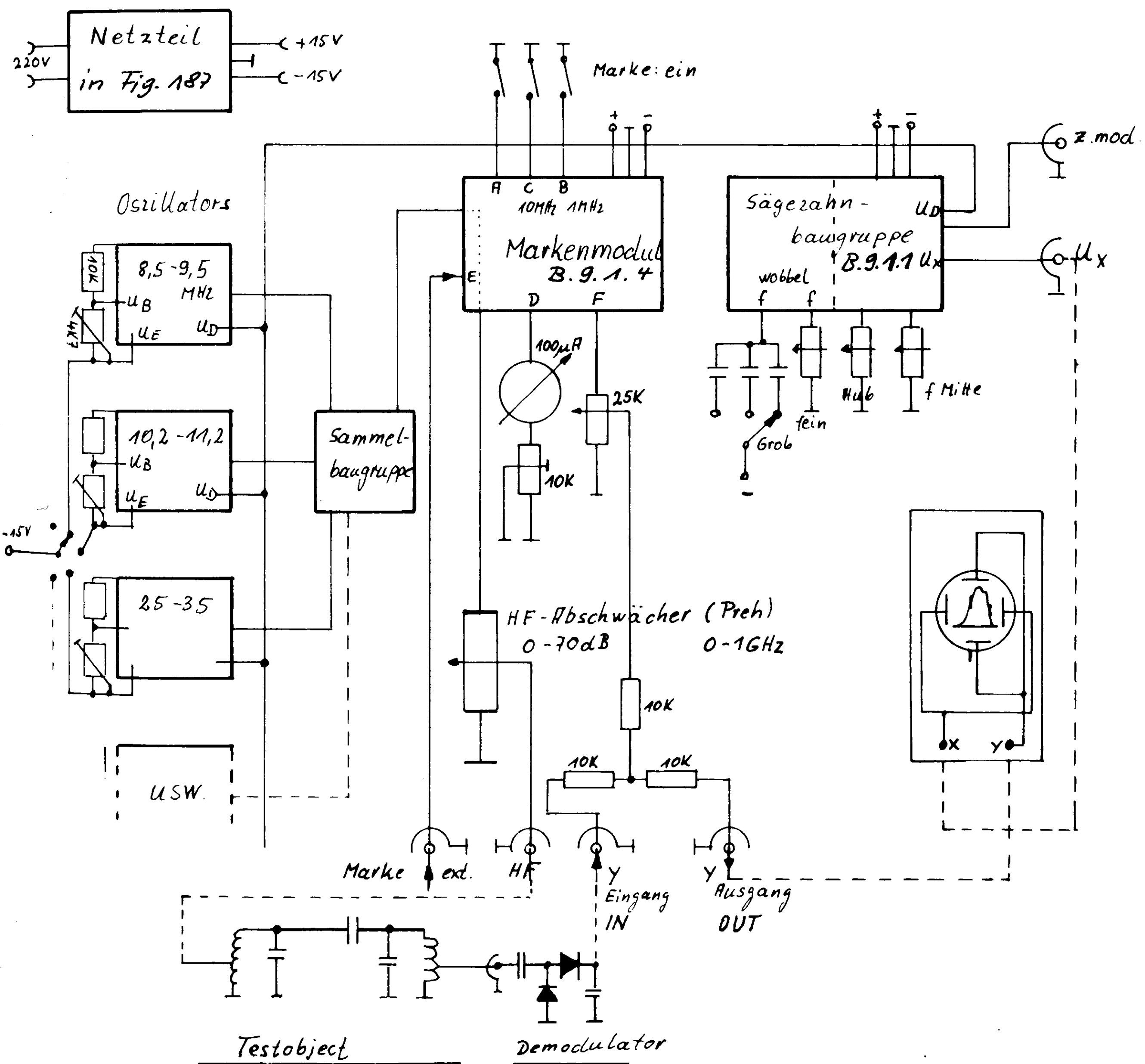
B.9 Sweep oscillator for 8.5 MHz to 1.3 GHz
(N. Schramm, DC9NI)

The following describes a sweep oscillator system that is composed of standard modules and may easily be duplicated.
The following modules are required and described in detail:

1. Sweep voltage generator (B.9.1.1)
2. Sweep oscillators (B.9.1.2)
3. Range selector (B.9.1.3)
4. Marker generator/mixer (B.9.1.4)
5. Power supply (Fig.187)

The concept makes use of an oscilloscope as display module in conjunction with the demodulator shown below.

Figure 175 Block diagram of sweep oscillator concept.



B.9.1 Description of modules

B.9.1.1 The sweep voltage generator

A sawtooth voltage is required to periodically sweep the oscillator frequency while simultaneously deflecting the electron beam of the oscilloscope. The frequency of this sawtooth voltage may be varied by means of an external capacitor (C) and a potentiometer (P1). The sawtooth voltage is generated by the timer NE555 (IC1). At pin 3 of IC1 a signal for blanking the electron beam of the oscilloscope during flyback is available. IC2 serves as buffer. From the output of IC2 the signal is taken to IC3 and IC8 which permit control of the origin - respectively the horizontal position. IC4 and IC9 raise the signals to the required amplitudes. By means of potentiometer P2 (preferably - but not necessarily - of the helical type with several revolutions between stops) the mean voltage for the tuning diode may be adjusted. IC7 is a buffer. The sawtooth voltage as supplied by IC5 and the dc voltage from IC7 are fed to the "sweep width" potentiometer P3 and routed via buffer stage IC6 to the tuning diode. IC8 and IC9 provide horizontal positioning and amplification for the X-deflection of the oscilloscope.

Alignment of the sweep voltage generator module
Required instruments:
Oscilloscope with dc input.

1. Connect output of IC4 with Y-channel input of oscilloscope. Set range to 5 Volts/cm. Adjust horizontal position of electron beam (screen centre, input short-circuited). Adjust P4 so that the sawtooth voltage is symmetrical with respect to the zero-line. Adjust P5 to achieve 28 Volts peak-to-peak.

2. Disconnect Y-input from IC4 and connect with output (U_X) of IC9. Symmetry of the displayed sawtooth voltage with respect to zero-line is adjusted through P6.

3. Connect output of IC9 (U_X) to X-input of oscilloscope. Adjust P7 so that the sweep width is 90% of screen width.

Figure 176 Circuit diagram of sweep voltage generator

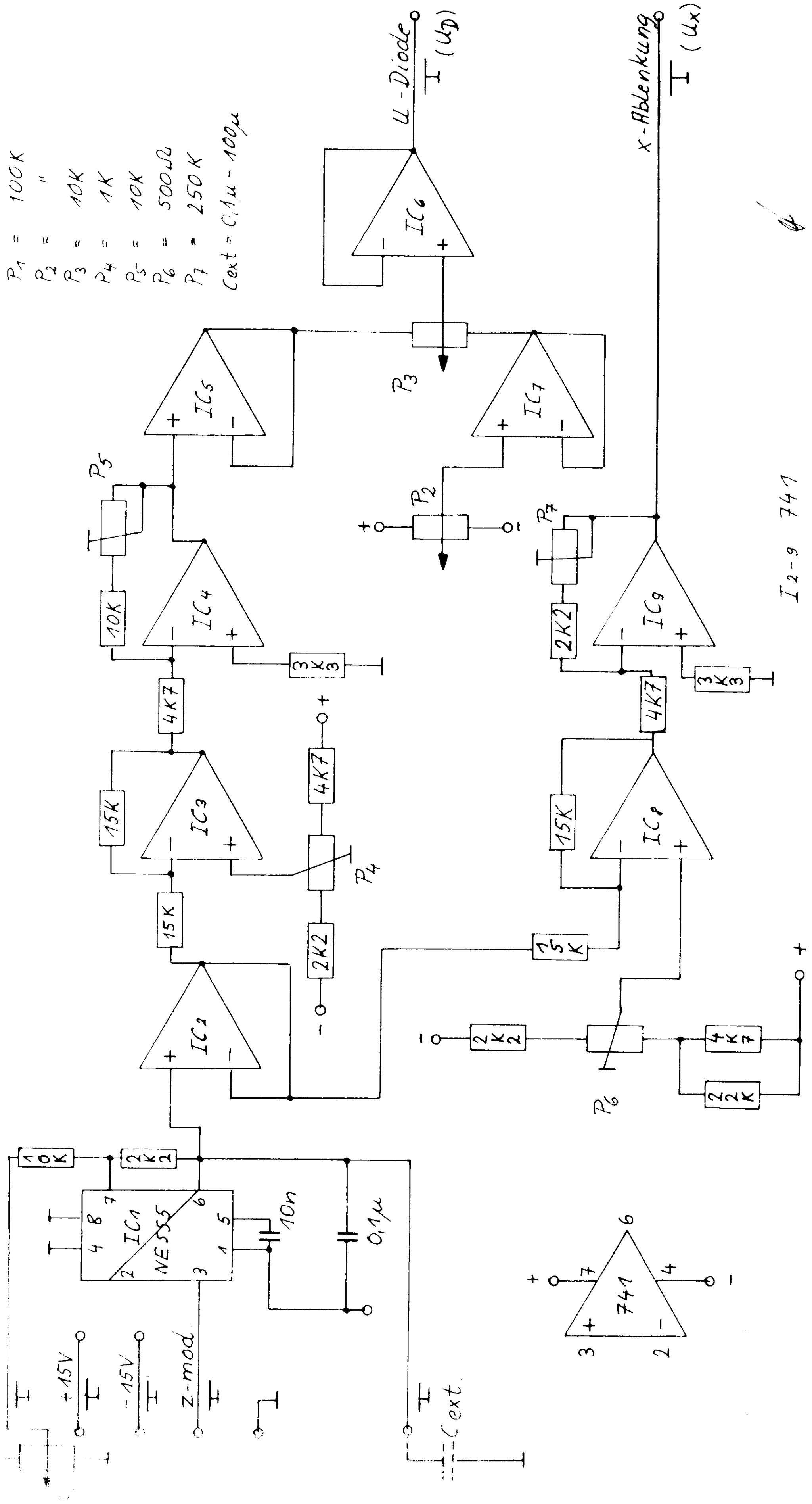
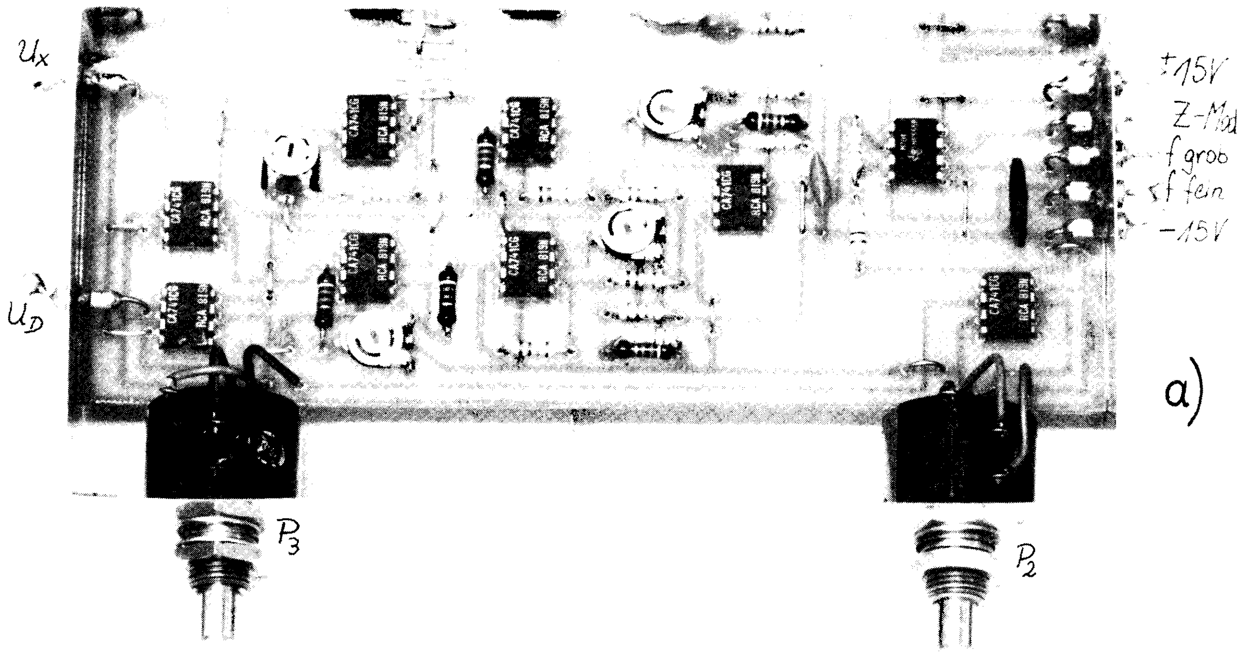
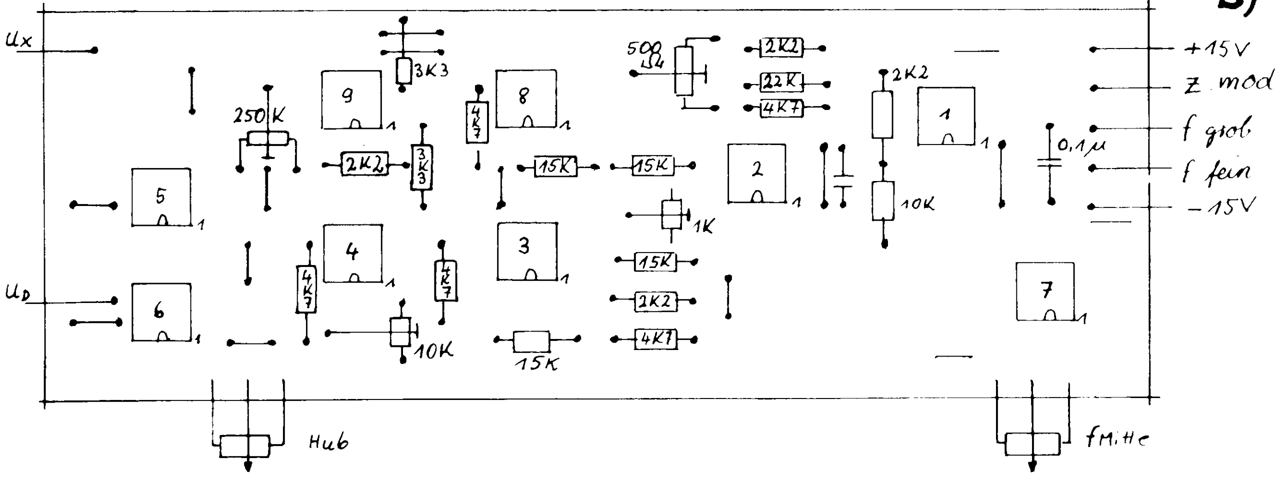


Figure 177 a) Component side of sweep voltage generator
b) Component location and numbering of integrated circuits
c) Printed circuit board underside (actual size).

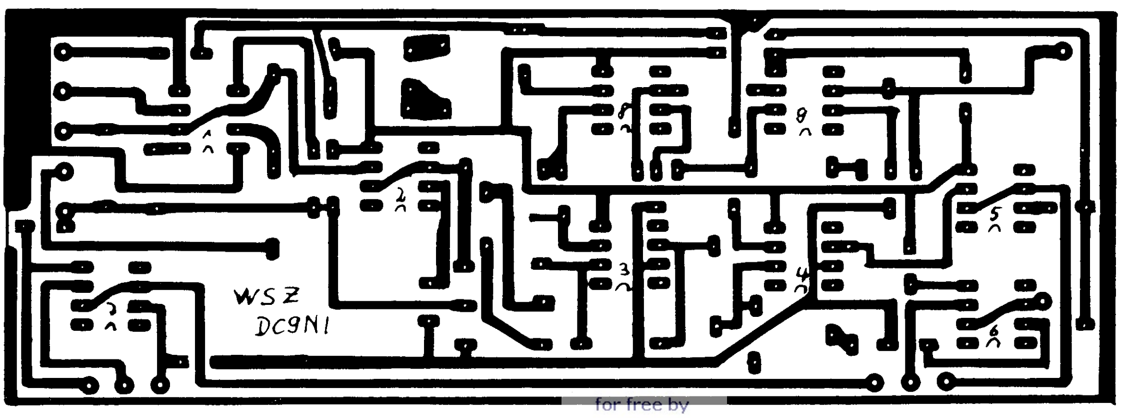


a)

b)



c)



B.9.1.2 The sweep oscillator

The oscillator circuit is a negative impedance converter in grounded collector configuration. The base voltage is generated externally and is supplied via a choke. Tuning is achieved through a varicap diode - the anode of which is tied to the emitter potential. The sweep voltage (U_D) is fed to the cathode via a choke. The unusual voltage supply:

Ue = -15 Volt $\qquad$ U_D = - 15 to + 15 Volt

was chosen so that a simple power supply +15/-15 V would suffice. Output impedance matching is achieved by means of a variable capacitor C_X. The oscillator output power is corrected through R1 and the T-section attenuator R2, R3 and R4. If the appropriate oscillator module is selected by means of the diodes of the range selector module then R5 has got to be installed. Otherwise oscillator switching must be done through a rather expensive coaxial switch.

Alignment of sweep oscillators

Required instruments:

Milliwatt power meter, frequency counter, completed sweep voltage generator

1. Connect milliwatt power meter to BNC-output connector of sweep oscillator module.

2. Set U_D with "centre frequency" potentiometer P2 of sweep voltage generator to zero volts.

3. Adjust Cx to give maximum output power reading.

4. Set external 4K7 potentiometer to give maximum output power reading.

5. Should the output power stay below the desired value (1-200 mW are possible) then R1 must be varied. Should this be insufficient the T-section attenuator R2, R3, R4 must be modified (i.e. reduce attenuation if output power is too low).

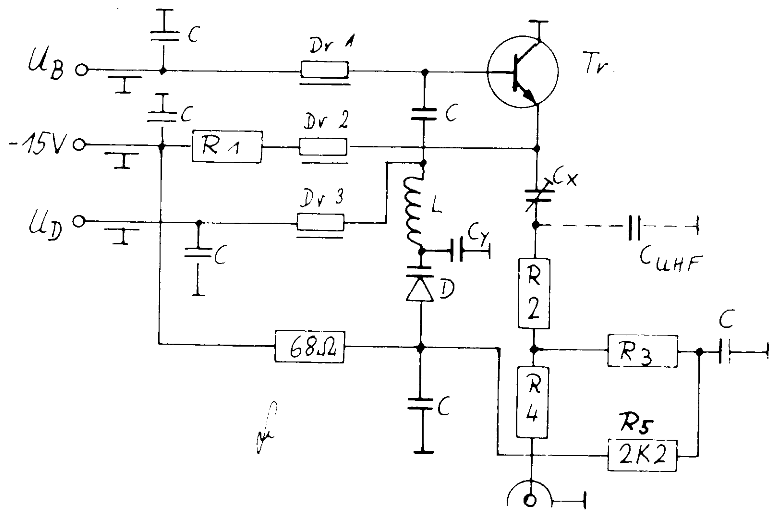
6. Connect frequency counter to output. Adjust "sweep width" potentiometer to minimum sweep width. P2 is set to yield U_D = -15 V. The lower frequency limit of any oscillator module is defined by L and C_Y and is set by proper adjustment of their values.

7. If necessary the steps 2-4 should be repeated.

8. Upon termination of the alignment procedure the inductor cores should be secured using glue.

Figure 178

Circuit diagram of one oscillator module. The basic diagram applies to all frequency ranges. The component values for the various frequencies are given in the table below.



C = 4.7 nF up to 30 MHz

C = 1 nF beyond 30 MHz

D = BB 105

Component specification for specified frequency ranges:

Frequency [MHz]	Transist. Type	R1 in [Ohm]	R2/R4 in [Ohm]	R3 in [Ohm]	Cx in [pF]	Choke	L turns/wire diam. [mm] /coil diam. [mm]
8.5-9.5	2N4014 *)	75	15	68	abt. 100	100 turns 0.15 mm on 220K	55 / 0.15 / 7 RF-braid with core
10.2-11.2	2N4014 *)	75	15	68	abt. 100	100 turns 0.15 on 220 K	50 / 0.15 / 7 RF-braid with core
25-35	2N4014 *)	120	33	27	5-15	150 uH	30 / 0.15 / 4 RF-braid
100-170	2N4014 *)	68	15	68	3-15	Valvo 6 hole	6 / 0.8 / 7 copper enamel
350-550	BFR90	68	15	68	3-15	Valvo 6 hole	bridge 1 mm to replace L
1150-1350	BFR90	75	10	10	3-15	Valvo 6 hole	see instructions on figure 181

C_{UHF} is 20 pF in range 350 - 550 MHz or 12 pF in range 1150 - 1350 MHz. C_y is 5 pF in range 25 - 35 MHz only. In all other cases these additional capacitances are missing.

*) or BFR90, BFR91, BFR34 and BFR35

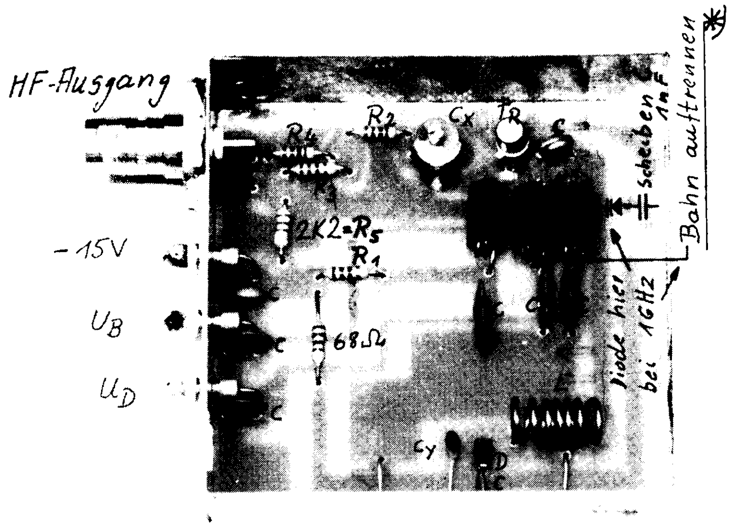


Figure 179
Component location of an oscillator module - in this case 100-170 MHz. In the 350-550 MHz module inductor L is replaced by a wire bridge. In that case the printed strip conductor provides the required inductance.

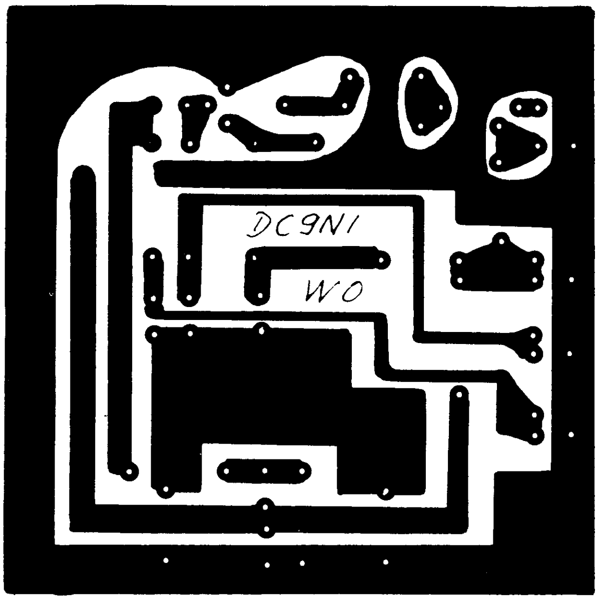


Figure 180
Underside of pc-board for frequencies ranging from 8.5 to 550 MHz. For higher frequencies the pc-board should be modified according to fig.181.

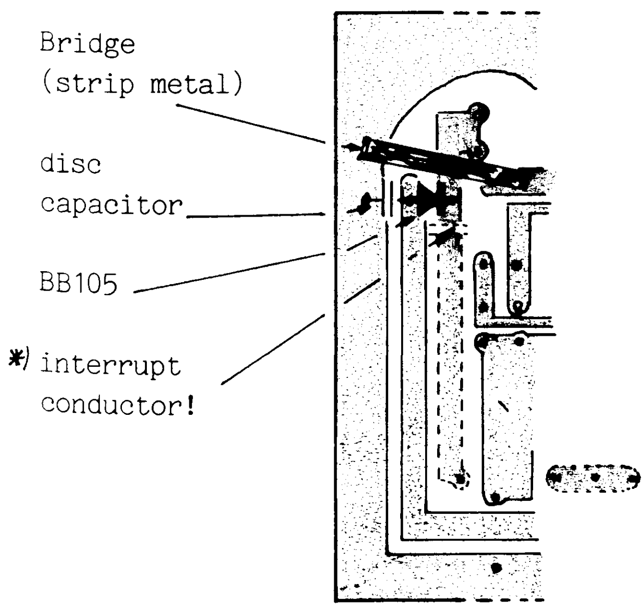


Figure 181
Underside of pc-board for frequency range 1150 to 1350 MHz. The conductor that was in the lower range used as inductor must be shortened to lower the inductance to the required value.

B.9.1.3 Range selector module

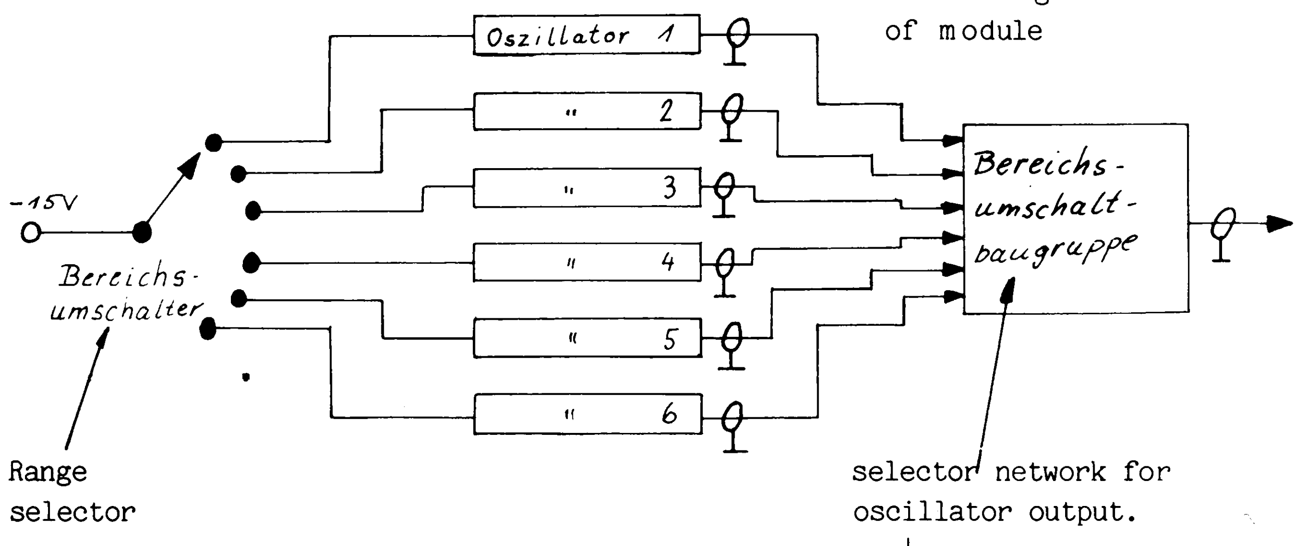
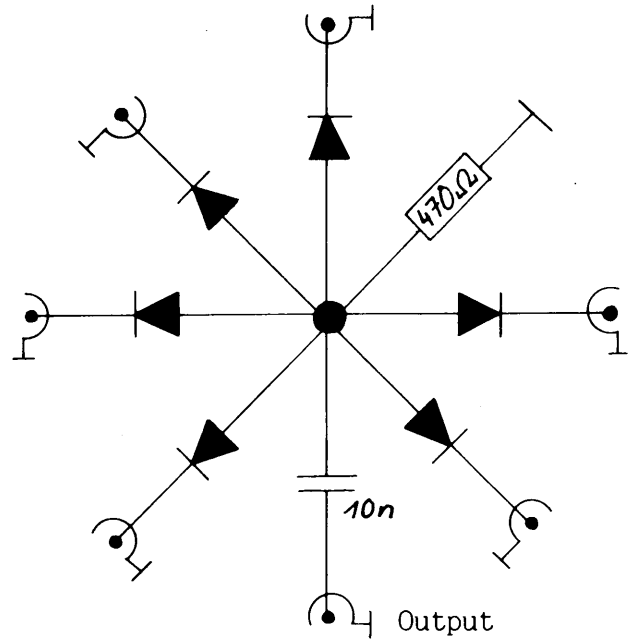


Figure 182
Block diagram of module

Figure 183
Internal arrangement of selector network for oscillator output selection.

Diodes: 1N4148 or BA182



If range selection should be performed without the use of expensive coaxial switches then the range selector module with its output selector network must be used. R5 must be fitted to the oscillator boards.
This module consists of a tin plate case measuring 72 x 72 x 30 mm. One BNC connector (output) is mounted in the centre and a number of BNC connectors (input, according to the number of oscillators) is installed concentrically around it.

Theory of operation:
A dc voltage passes via R5 of the turned-on oscillator to the corresponding network diode which will become conductive. Thus the rf signal is passed on to the output. The voltage drop across the 470 ohm resistor is sufficient to turn all other diodes off.

B.9.1.4 The marker-generator/mixer module

Theory of operation:

This module provides frequency markers in 10 MHz and 1 MHz steps which are derived from a 10 MHz crystal oscillator. The 10 MHz signal of the oscillator stage is passed through a buffer stage (BC237) and on to gate I of IC2 (7400). Upon closure of switch "A" the 10 MHz pulses are passed on to gate II which acts as buffer. The gate II output signal is then routed to gate III for the 10 MHz pulses and via IC3 (7490) to gate IV for the 1 MHz pulses. IC3 is arranged as divide-by-ten circuit.

The selected 1 MHz or 10 MHz spectrum passes through the 47 Ohm resistor and to the mixing diode D1 where it is mixed with the rf signal of the sweep oscillator.

The beat frequency of rf signal and marker spectrum is routed through a low pass filter and on to the operational amplifier IC4 for amplification. The amplified and amplitude-limited signal is available at its output.

Alignment and test

Required instruments:

Oscilloscope, sweep oscillator module (such as 25-35 MHz) and sweep voltage generator, frequency counter.

1. Connect frequency counter to collector of T2, adjust C_T of marker generator to give a 10 MHz reading on counter.

2. Connect oscilloscope (Y-input) to test point 1 (MP1). Operate switches "B" and "C". A 1 MHz resp. a 10 MHz signal should appear.

3. Connect sweep oscillator output to rf input (marked HF). Turn "sweep width" potentiometer of sweep voltage generator to maximum width. Connect X-output of sweep voltage generator to X-input (extern) of oscilloscope. Connect marker output "F" to Y-input of oscilloscope. Switch on 10 MHz markers and record amplitude (U_{SS}). Switch on 1 MHz markers and adjust amplitude to same value as 10 MHz markers by means of potentiometer "P".

This finishes the construction and the setting up procedure of the modules. They may now be interconnected according to the block diagram and installed in a cabinet.

Section B.9.2 contains suggestions for the application of the sweep generator.

Figure 184 Circuit diagram of marker generator/mixer module

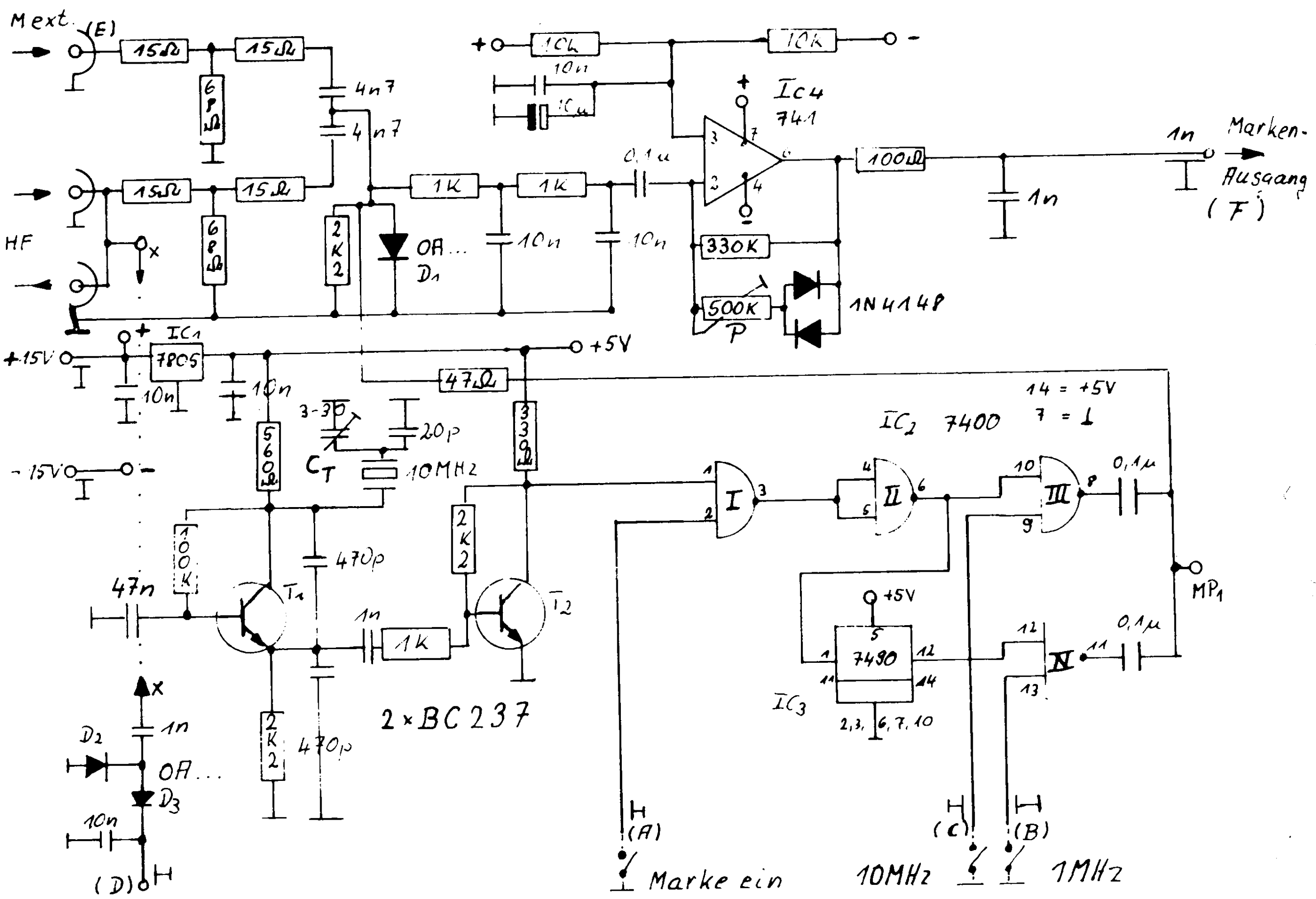


Figure 185 Underside of pc-board to fit a case measuring 146 x 72 x 30 mm (C.4).

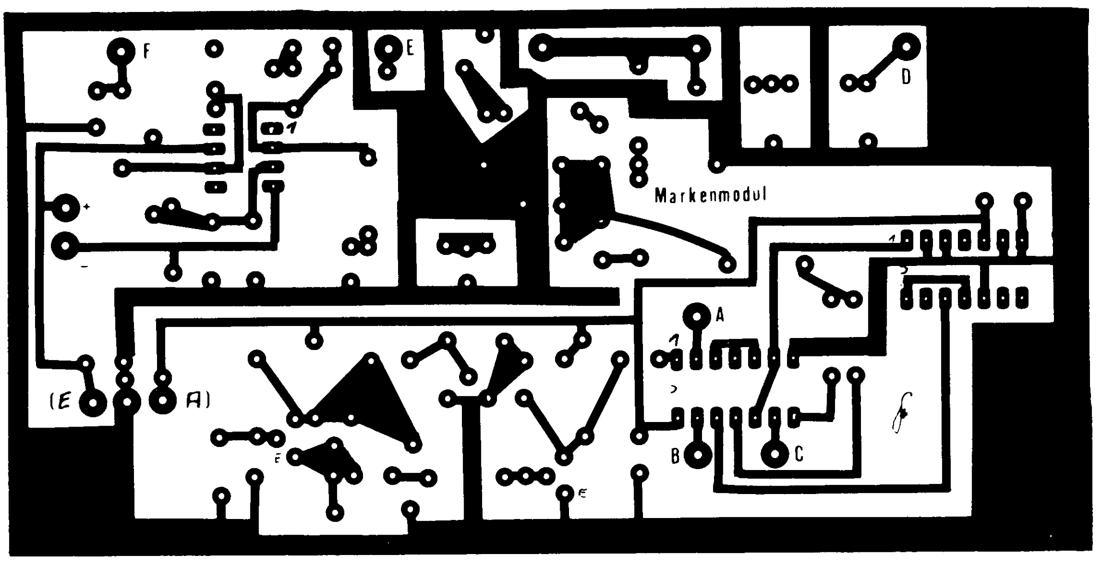


Figure 186 Component location diagram of marker generator/mixer

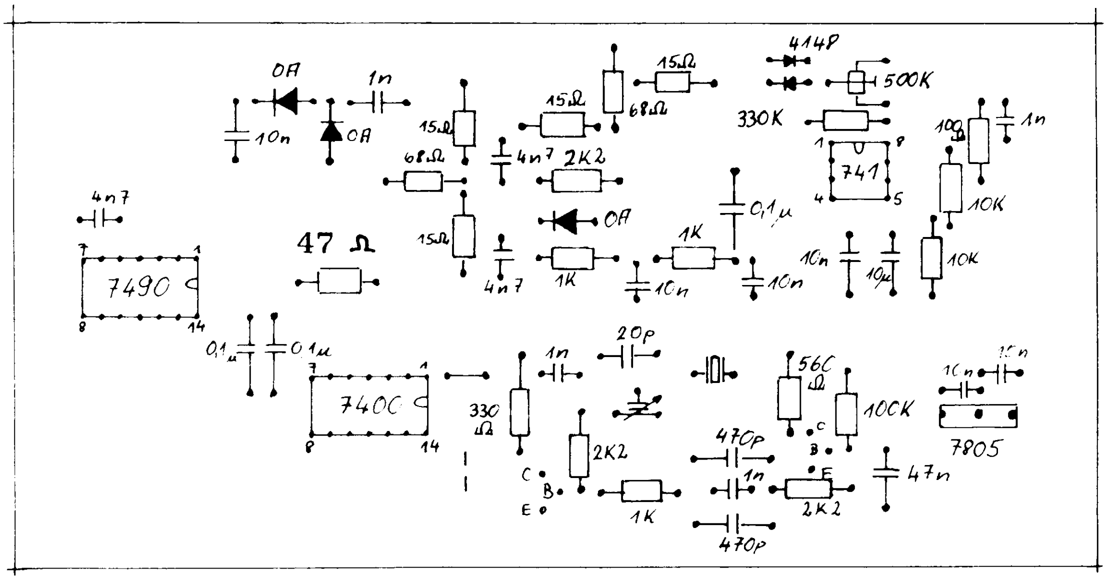
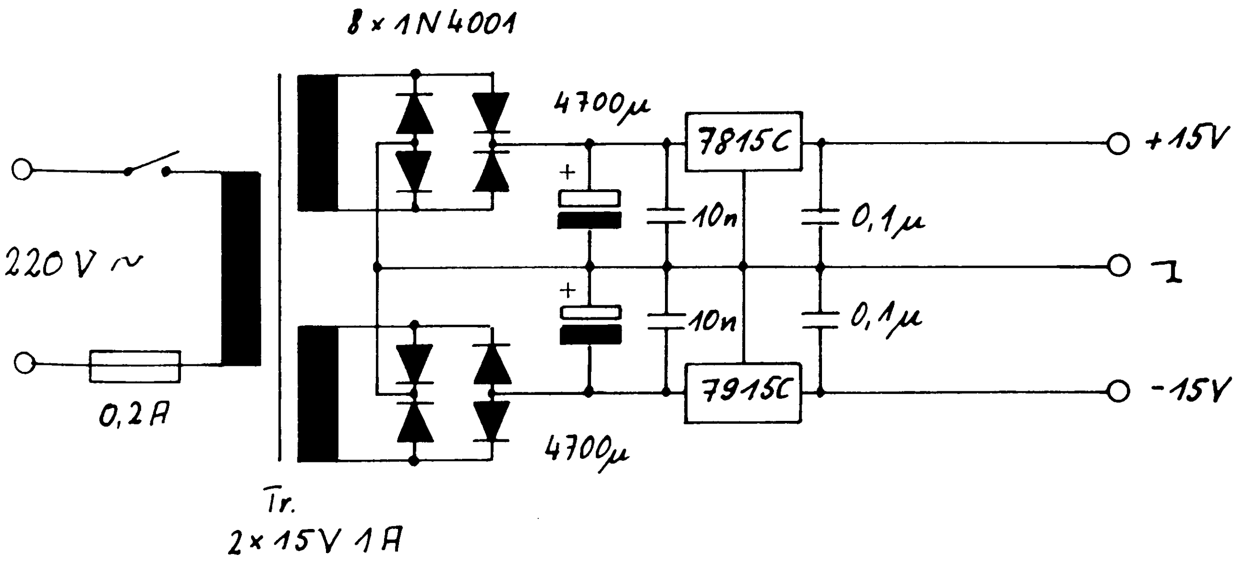


Figure 187 Suggested power supply for complete system



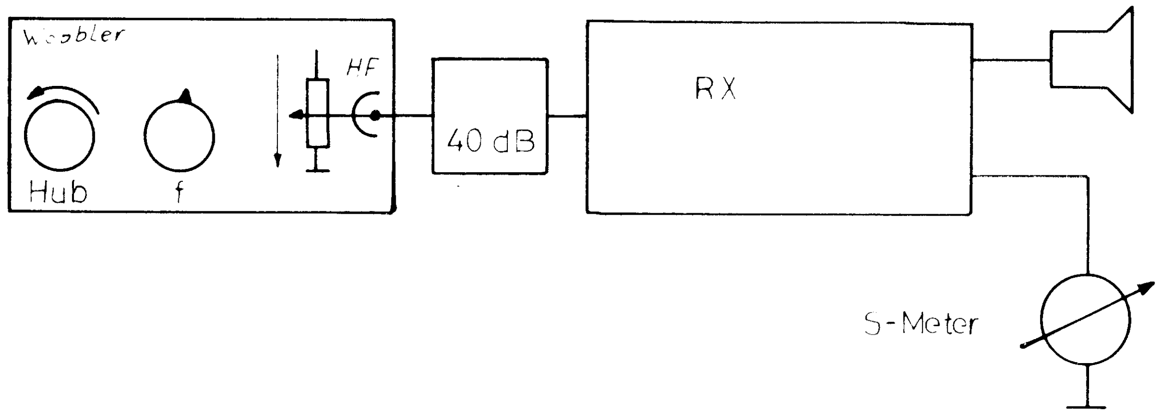
B.9.2 Using the sweep oscillator of section B.9 - examples

B.9.2.1 Receiver alignment

Introduction:

For the alignment of receiver sensitivity one requires a rf signal which is either unmodulated or modulated according to the mode of operation (AM, FM, SSB). This signal should have a sufficiently stable frequency and its amplitude should be adjustable.

Figure 188



Procedure:

Connect sweep oscillator output through a 40 dB attenuator to receiver input.

Reduce sweep width to zero.

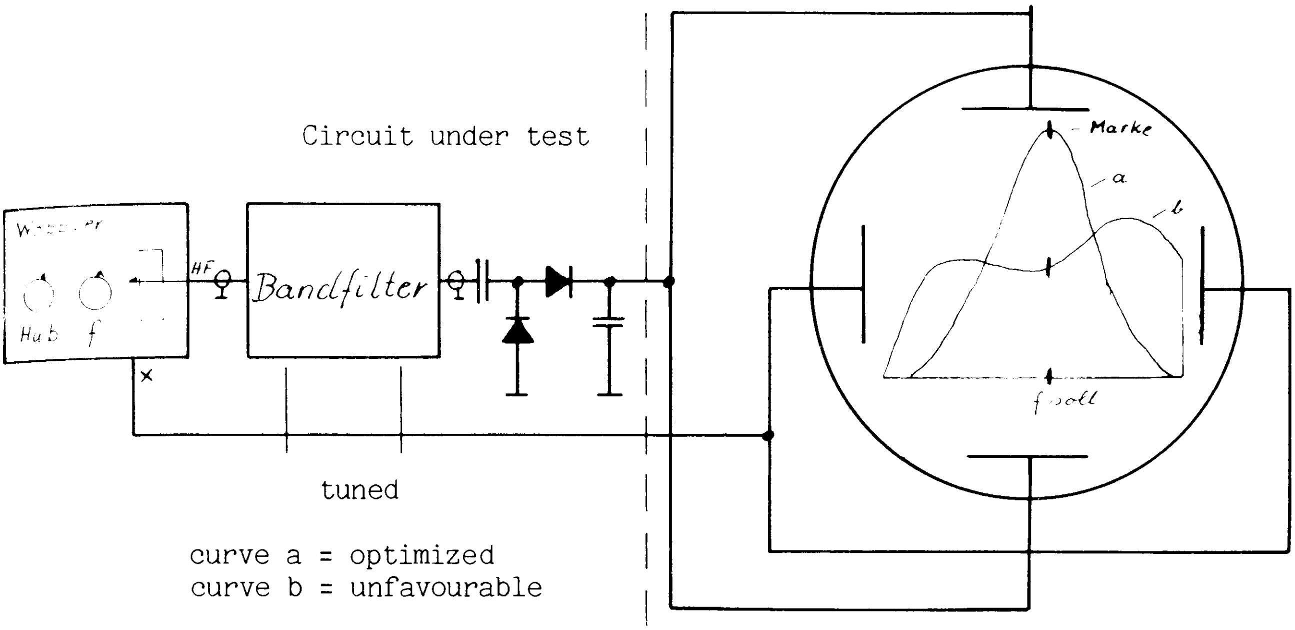
Set frequency of sweep oscillator to desired value.

Align the circuits under consideration (Use S-meter for control purposes).

Match sweep oscillator output to receiver S-meter deflection by suitably reducing the output power.

B.9.2.2 Measurement of frequency response

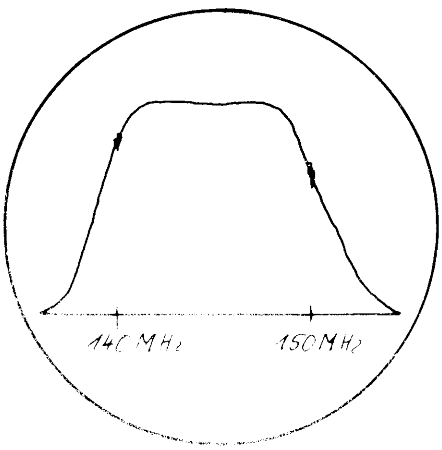
Figure 189 Method



Introduction:

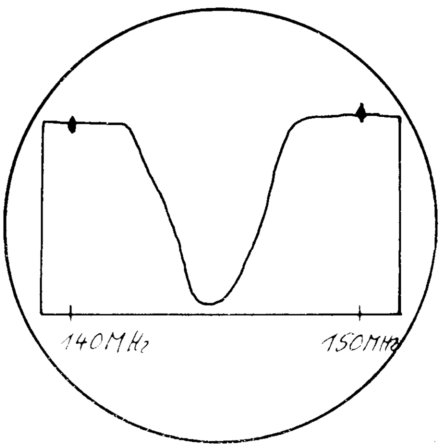
The rf signal generated by the sweep oscillator sweeps across the desired frequency range according to the sweep rate. The circuit under test is connected to the sweep oscillator and will exhibit frequency dependent attenuation (in case of passive networks as shown in B.9.2. a and b) or frequency dependent gain (for active circuits according to B.9.2.2. c, d and e). The output signal of the circuit under test is rectified by means of a rf demodulator and fed to the Y-input of the oscilloscope. Depending on the nature of the circuit different types of curves will be displayed on the scope and some examples are given below:

Figure 190



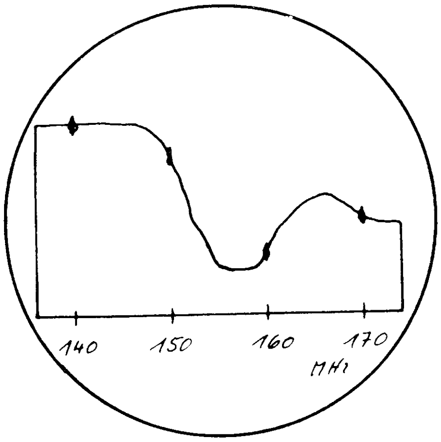
Band-pass

Figure 191



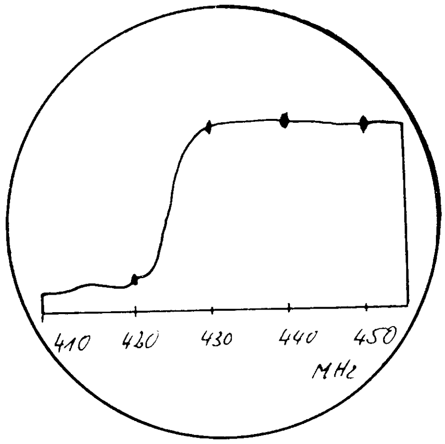
Band-stop

Figure 192



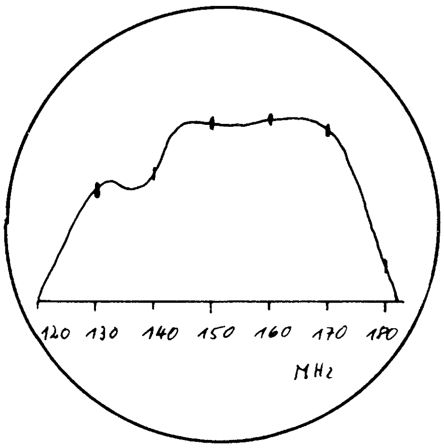
Low-pass

Figure 193



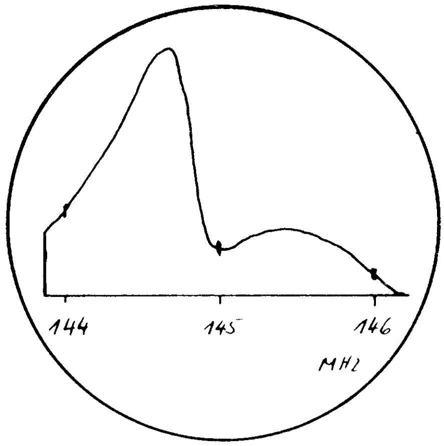
High-pass

Figure 194



Amplifier

Figure 195



2m PA

Warning!!

When performing measurements on amplifier stages coaxial cables with high-quality screening must be used. Due to poor screening radiation might pass through the screening mesh and falsify the displayed frequency response curve. It is thus recommended to use coaxial cable having two layers of screening mesh for measuring purposes (such as RG 55 B/U, RG 214, RG 223/U). This improves the screening by further 10 dB.
When sweep-testing transmitters they must work into a dummy load to avoid any radiation outside the authorized frequencies and consequent interference.

B.9.2.3 Measurement of matching properties

Assuming that no slotted trough line is available the most suitable method is to employ a fixed measuring probe (diode rf head) for measuring the standing wave ratio (ripple factor s) at the transmitter output. Since frequency bands rather than individual frequencies are allocated it is preferable to employ a sweep oscillator for the measurements. Corresponding to the set sweep width either a narrow or wide frequency band may be covered. When feeding the diode output of the rf-head to an oscilloscope the ripple factor across the swept range will be displayed. To display the voltage representing the standing waves a leader cable is used. Its length depends on the number of maxima and minima one intends to display across the desired range of frequencies. Increasing the length of this cable connecting the signal source and the test device will increase the number of displayed "waves". Thus the resolution per unit of frequency is improved. In other words: If one intends to measure the matching properties across a specified frequency band a minimum cable length is required. This length of leader cable (l_M) may be computed according to this equation:

$$l_M = \frac{300}{4 \times f} \times 1.2$$

Δf = desired frequency band
1.2 = safety factor

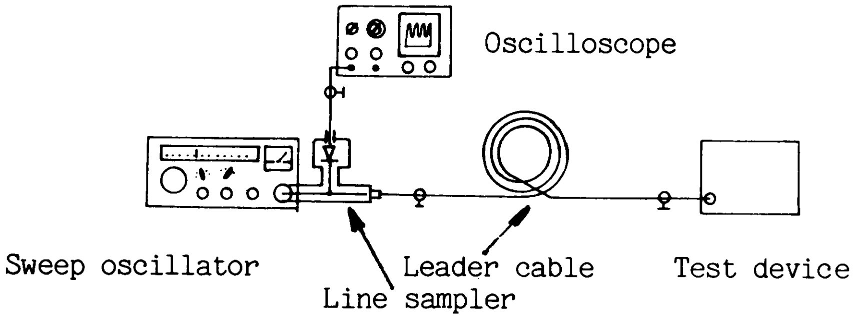
In this case one complete wave is stretched across the frequency band, i.e. one maximum and one minimum.
If "n" complete waves are supposed to be traced the length should be calculated according to:

$$l_M = 150 \times n \times v \times \left(\frac{1}{\Delta f}\right)$$

where n = number of waves (usually 5 to 25)
v = velocity factor of coaxial cable

It is particularly important to use coaxial cable with low losses at VHF/UHF and no discontinuities in its impedance.(RG—214/U-1 is barely adequate having impedance discontinuities $\Delta Z \leqslant 1\ \%$).
RG 58 and RG 213 are totally unsuited for this purpose.

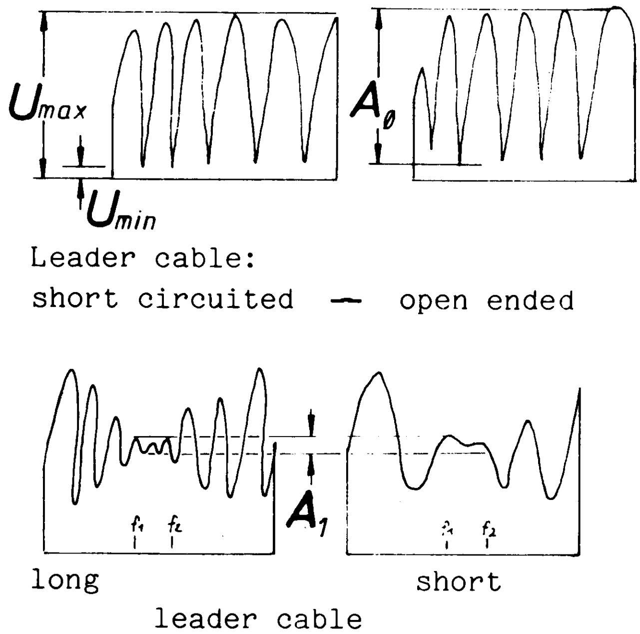
Figure 196
Method:



When carrying out measurements on the open-ended or short-circuited leader cable the oscilloscope will provide both ripple (decreasing with increasing frequency) and reference (zero line) on the oscillogram. Its distance from the corresponding minimum is a measure for the cable attenuation (d) which may be calculated according to:

$$d = 10 \log \frac{U_{max} + U_{min}}{U_{max} - U_{min}} \text{ [dB]}$$

Figure 197
Presentation of ripple on the oscilloscope



The test set up is checked by terminating the cable in a non-reactive resistance of a value corresponding to the characteristic impedance of the cable, the test instrument input and the test device (for instance 50 Ohm)
Any residual ripple caused by discontinuities and cable tolerances may be compensated for by means of a tuning device.
(see section E.3.1).

To obtain the ripple factor the amplitude A_0 of the incident wave must be measured (test device disconnected). Expressed as reflection coefficient this corresponds to 100%. Next, the test device is connected to the leader cable and amplitude A_1 is measured - respectively the percentage of the reflected wave is calculated (= reflection coefficient).

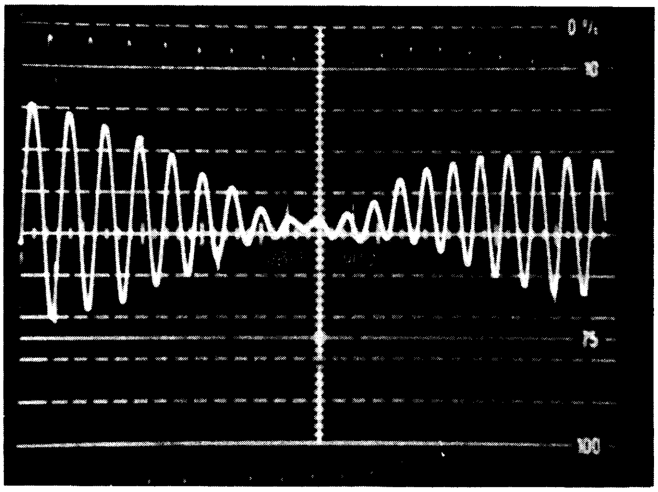
$$r = \frac{A_0}{A_1} = \frac{U_{incident}}{U_{reflected}} \times 100 \quad [\%]$$

The standing wave ratio may be calculated from the reflection coefficient according to:

$$s = \frac{r + 1}{r - 1} = \frac{1}{m}$$

For reference to power see section B.3.

Figure 198 indicates the matching characteristics (reflection coefficient) of a Hirschmann TV antenna modified for the amateur band (see section E.7)



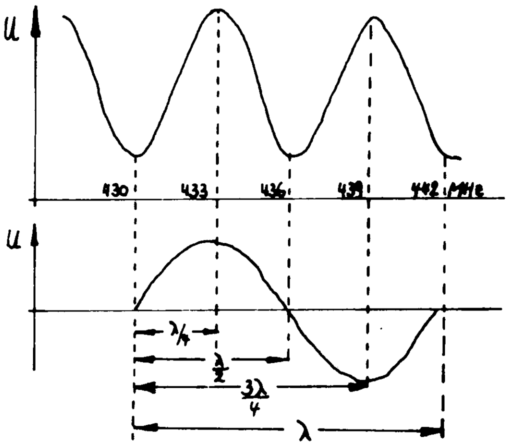
Locating impedance discontinuities (faults) in antenna feeders:
If a feeder is correctly terminated in a non-reactive resistance corresponding to its characteristic impedance the approximate location of an impedance discontinuity may be calculated from the frequency difference Δf between maximum and minimum values (switch on the frequency markers!). This Δf relates to an electric length of $\lambda/4$ between sweep oscillator output and position of impedance discontinuity.

$$\lambda = \frac{300}{f_{MHz}} = \text{electrical distance}$$

The physical length is obtained through multiplying the electrical length by the cable velocity factor.

Figure 190 shows the voltage as traced on the oscilloscope.

Standing wave on cable (R < Z) due to impedance discontinuity (R = impedance discontinuity, Z = characteristic impedance)



B.9.2.4 Panoramic receiver

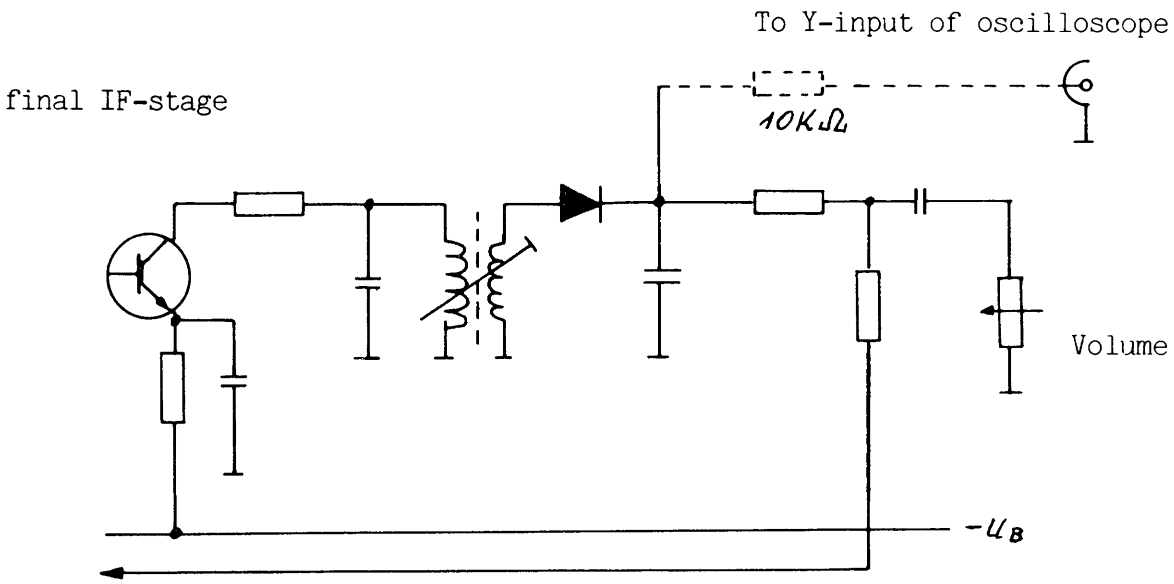
Purpose:

To visualize the activities across a significant frequency band and to detect spurious radiation a panoramic receiver is a rather convenient tool (see also section B.5).

As mentioned in section B.5 the electron beam is deflected in X-direction according to the sweep voltage fed to the X-channel. The synchronized sweep frequency as generated by a sweep oscillator is fed to a mixer (ring modulator) and its output entered into an IF-amplifier. A broadband preamplifier preceeding the mixer compensates for the mixing losses. The signal required for driving the Y-channel is taken off the receiver AM-demodulator.

Figure 200

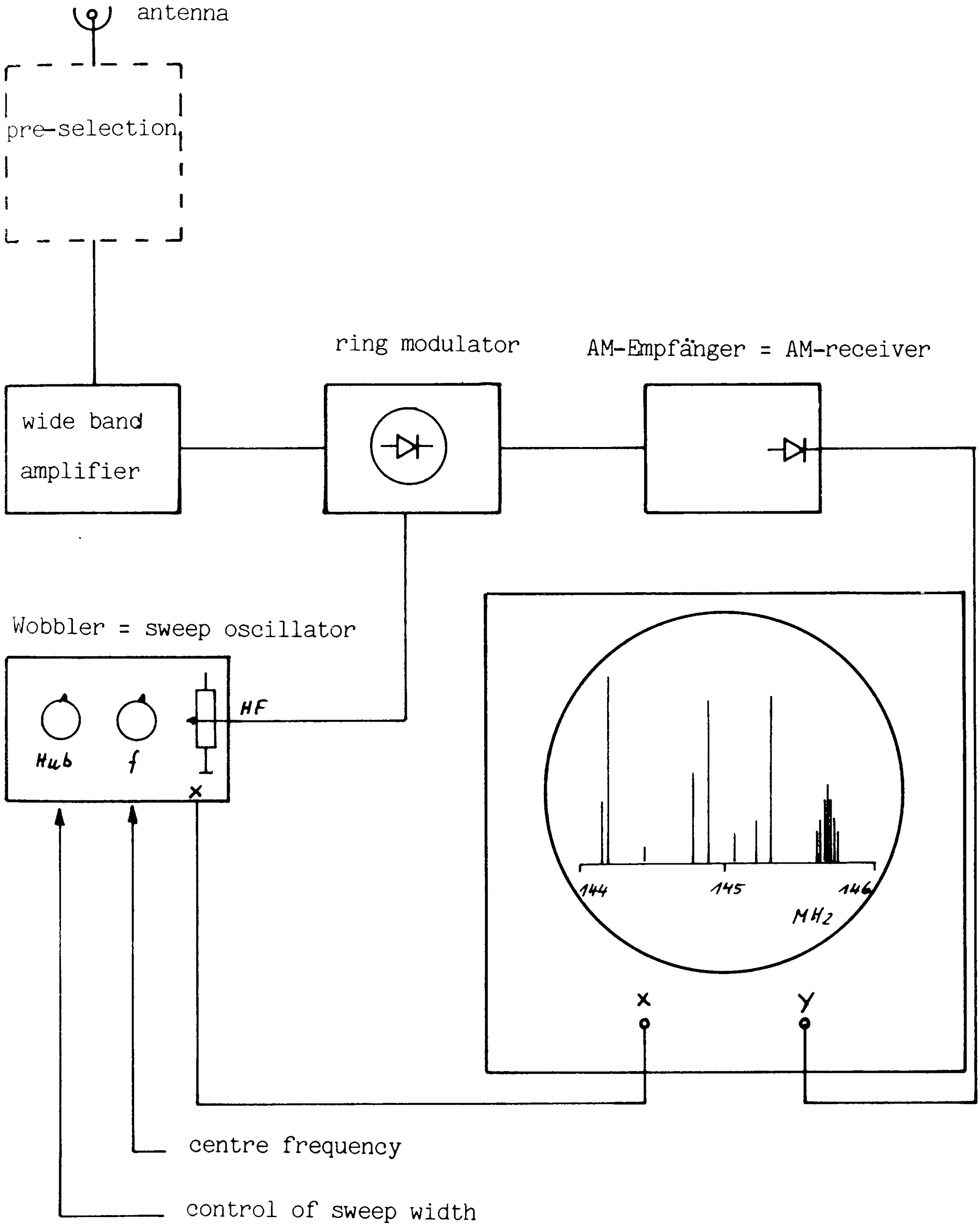
shows the required modifications to an-AM receiver to permit its use as IF-strip of a panoramic adaptor.



List of materials:

1. Sweep oscillator (sweep voltage module and oscillator module)
2. Broadband amplifier (cheap TV-type antenna preamplifier)
3. Ring modulator or converter head according to D.2
4. AM-receiver as IF-amplifier (for instance cheap CB-set re-adjusted for 28 MHz.
5. Oscilloscope
6. Possibly band filters for the frequency bands concerned (preselection).

Figure 201
Block diagram of a wide band panoramic receiver based on a sweep oscillator.



C. Filters for signal selection and TVI-suppression

The RF-signal produced by a generator or a transmitter stage does not always possess the favourable properties as those shown in D.6.1. Even bandpass coupling will pass spurious signals and harmonics - even more so, if the components are closely spaced and if no screening is provided. Not only the neighbourhood will show little sympathy for having the "goggle box" upset by any of these by-products; the amateur, too, will notice the detrimental effect in both receiving and transmitting applications. Since power measurements (see B.2) are based on the sum of all power components (including by-products) leading to a voltage drop across a resistance including filters for selective measurements is recommended.

Figure 202 represents the frequency range from 1 to 2000 MHz. The vertical deflections indicate the spectrum of an experimental circuit designed as receiving mixer on 23 cm. In spite of bandpass filters and a metal case, undesirable by-products and harmonics of f_Q are present.

To deal with these problems several circuits are going to be discussed in detail and their construction will be described.

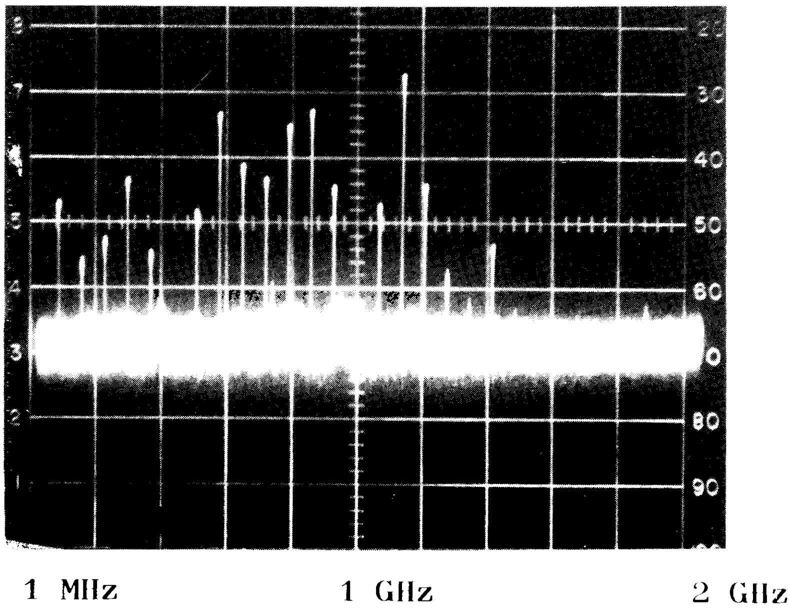


Figure 202
Frequency spectrum from 1 to 2000 MHz of experimental circuit based on f_Q = 73 MHz.

horizontal scale:
200 MHz / line

vertical scale:
10 dB / line

C.1 144 MHz four-section filter for harmonic suppression

Well designed 2m transceivers have values of harmonic suppression of 70 to 80 dB. Assuming an output power of 10 Watts on 2 meters the harmonics will be 80 dB down on 70 cm - i.e. at -60 dBm. A modern 70 cm receiver will read such a signal S9 + 30 dB! Simultaneous operation on both bands will be rather difficult - at least on 70 cm! If mounted directly at the transmitter output the filter described below will - in conjunction with decoupling the antennas - reduce the harmonics by a further 60 dB. Additional effort makes little sense since at this stage stray coupling through the screening mesh of the coaxial cable becomes a possibility. It is therefore advisable to separate the cables concerned.

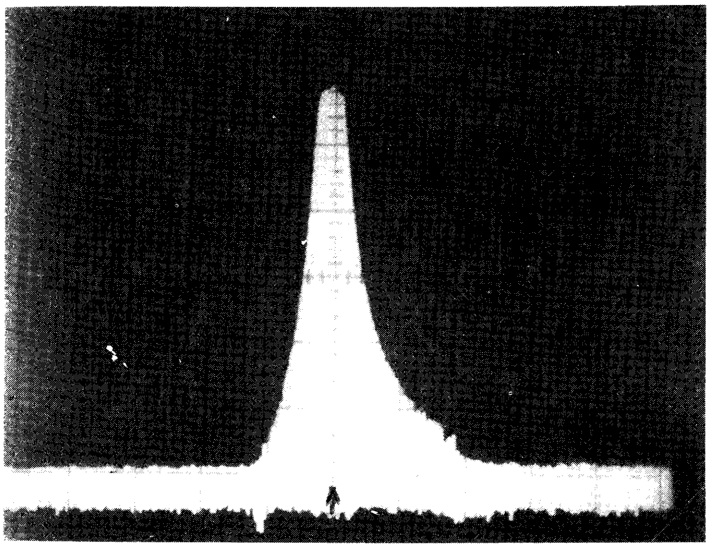


Figure 203
Frequency response of the filter over the range 95-195 MHz.

Insertion loss: 1.5 dB

Scale:
horizontal = 10 MHz/div
vertical = 10 dB/div

f_{res} = 145 MHz

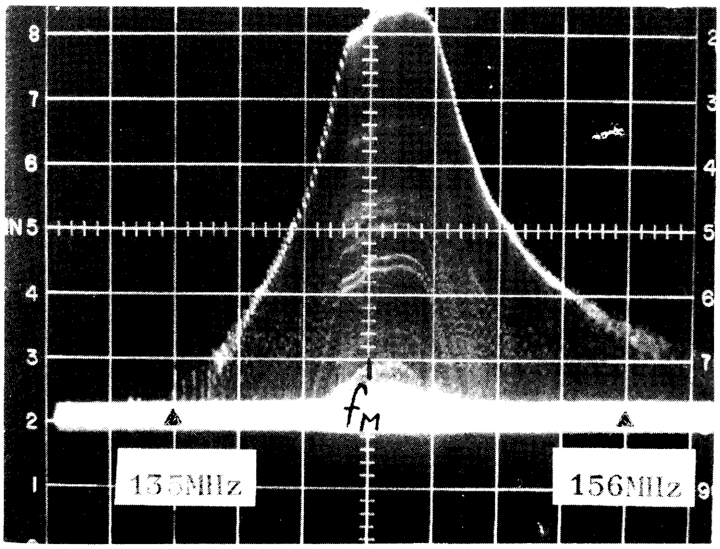


Figure 204
Frequency response of the filter over the range 129-159 MHz

Scale:
horizontal = 3 MHz/div
vertical = 10 dB/div

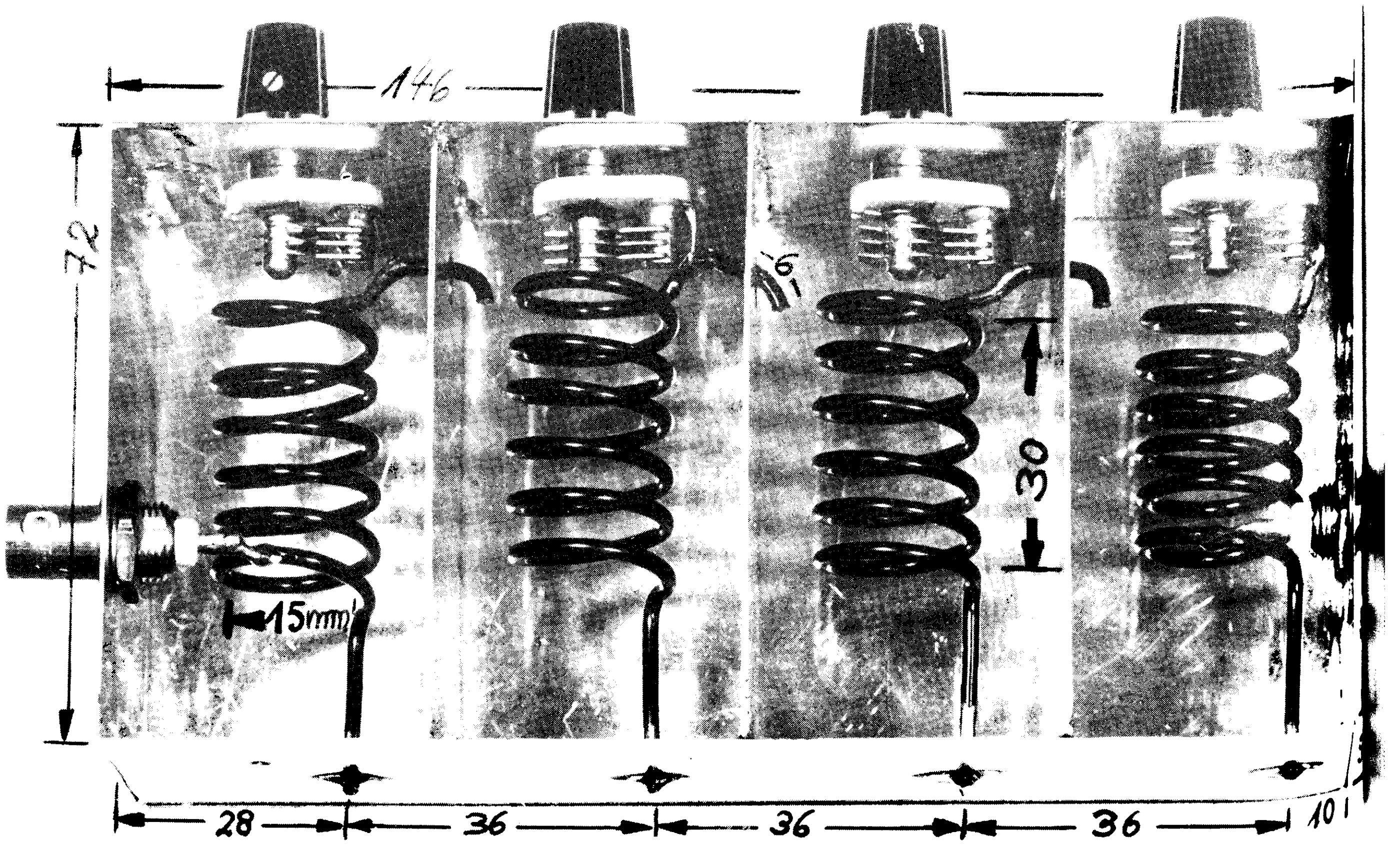
centre freq. = 144 MHz

attenuation / frequency
- 3 dB at +/- 1 MHz
- 10 dB at +/- 2.5 MHz
- 20 dB at +/- 3.5 MHz
- 40 dB at +/- 7.5 MHz
- 60 dB at +/-15 MHz

Filter input and output is through BNC connectors. Ceramic trimmers may be used as tuning capacitors in which case the power rating is reduced. Capacitors with air insulation are certainly preferable. The general arrangement of the filter is clearly shown in figure 205.

The filter consists of four helical resonators in individual compartments. Each helix is made up of six turns (copper wire, 2mm diam.) wound over a 15 mm diameter mandrel. The hot ends are soldered to the air trimmer capacitors so that the ends protrude approximately 6 to 7 mm into the neighbouring compartments. The length of these ends has a direct influence on the filter bandwidth.
With the shown capacitors that could be obtained cheaply 25 Watts can be handled without any difficulties. Different capacitors (for instance with more than 0.5 mm spacing between the plates) will enable the filter to handle up to 50 Watts. Higher power levels require a case made up from thicker material.
As may be seen in the photograph the cold ends of the helices are threaded through holes drilled into the base plate of the filter box. The BNC connectors are mounted 28 mm up from the base plate. The straight part of each inductor has a length of approximately 20 mm. A quarter turn further-on the inductor is soldered to the BNC connector. These connectors are not placed symmetrically but are mounted 12 mm off the open side. The partitions are arranged such that the whole box (length 146 mm) is divided up evenly according to the photograph. Each compartment will then measure about 36 mm.

Figure 205 Construction of the receive/transmit filter for the 2m band. The air trimmers have a capacitance of 2 to 15 pF. The photograph shows the filter in larger than actual size.
The internal dimensions of the box are 145 x 72 x 50 mm (see G.4).
Only one of the covers needs to be soldered.



C.3.1 Single-section 70 cm bandpass filter

A filter consisting of a single tuned circuit will in many cases provide sufficient attenuation to spurious products and harmonics of oscillators and injection frequency generators. The frequency response will not reach those values that are attainable with multi-section arrangements but are acceptable in many applications.

If the majority of the unwanted signals are below the frequency of the desired signal then a capacitively coupled filter will give the required result. Otherwise, see next description.

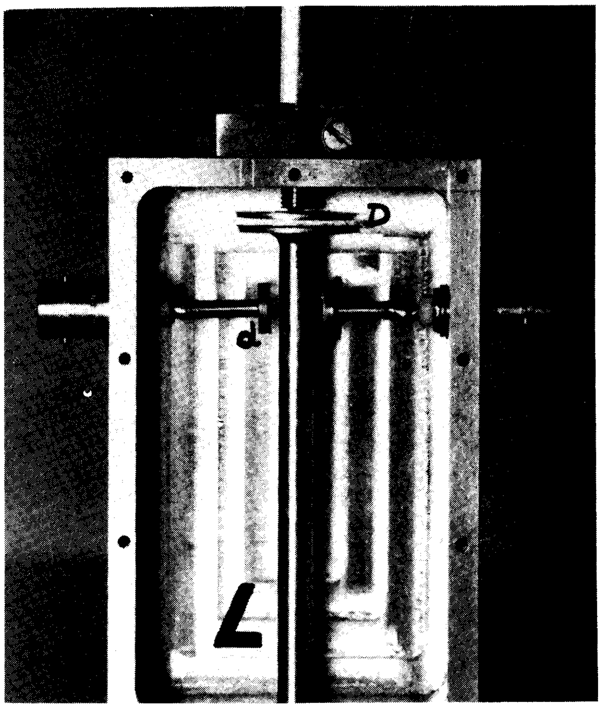


Figure 206
Capacitively coupled single tuned circuit filter

Internal dimensions:

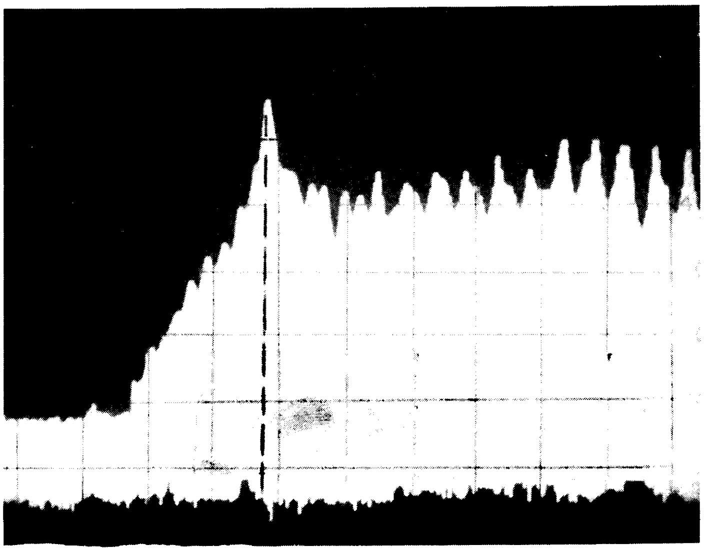
height	90 mm
width	55 mm
depth	25 mm

L = 6 mm diameter
D = 15 mm diameter
d = 10 mm diameter

F_{Res} = 404 - 432 MHz, tunable

Insertion loss = 1.5 dB

L is bolted to base plate from outside.

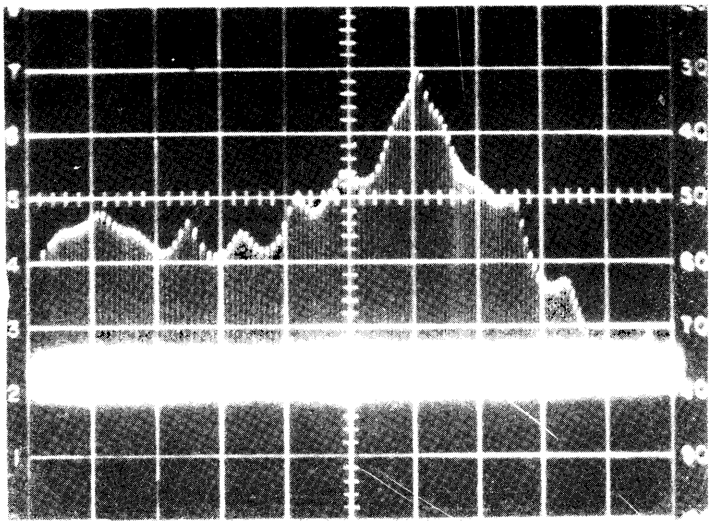


1 MHz f_{Res} 432 MHz 1000 MHz

Figure 207
Frequency response of the filter tuned to 432 MHz.

Scale:
Horizontal = 100 MHz/div.
Vertical = 10 dB/div.

Evidently, a simpler type of construction could be chosen for such a filter. Copper clad material or tin plate is quite acceptable for 70 cm waves without excessive losses. The following circuit will suppress predominantly signal components having higher than resonant frequencies. The case is a commercially available unit of dimensions 107 x 35 x 28 mm.
Inductor (L) and the capacitor are mounted coaxially.



250 MHz f_{Res} 550 MHz

Figure 208 shows the frequency response of the inductively coupled filter.

Scale:
horizontal = 30 MHz/div
vertical = 10 dB/div

f_{Res} = 432 MHz

Insertion loss = 1.5 dB

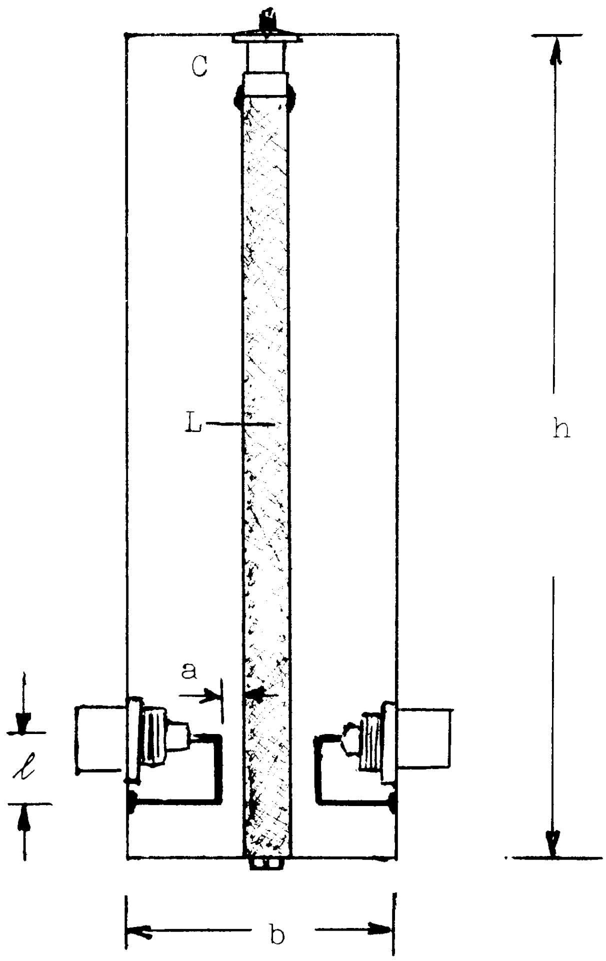


Figure 209

Dimensions:
- h = 107 mm (110 mm are possible)
- b = 35 mm
- t = 28 mm
- L = 6 mm diam.
- l = 10 mm
- a = 3 mm
- C = 0.5 - 6 pF

Length of C should be approximately 15 mm.

C.3.2 Twin-section 70cm bandpass filter

The two tuned circuits are capacitively coupled by means of two metal strips (3 mm wide) soldered to the inductors, overlapping for about 5 mm with 2 mm spacing, positioned inside an opening in the partitioning wall. Figure 210 represents the frequency response between 1 MHz and 1 GHz.

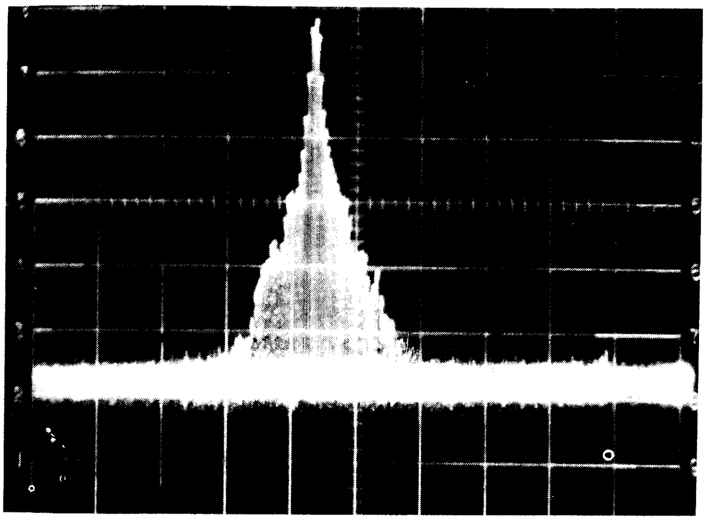


Figure 210
Frequency response

Scale:
horizontal =100 MHz/div
vertical = 10 dB/div

f_{Res} = 432 MHz

insertion loss = 1.2 dB

frequency / attenuation
± 40 MHz / - 28 dB
± 120 MHz / - 53 dB

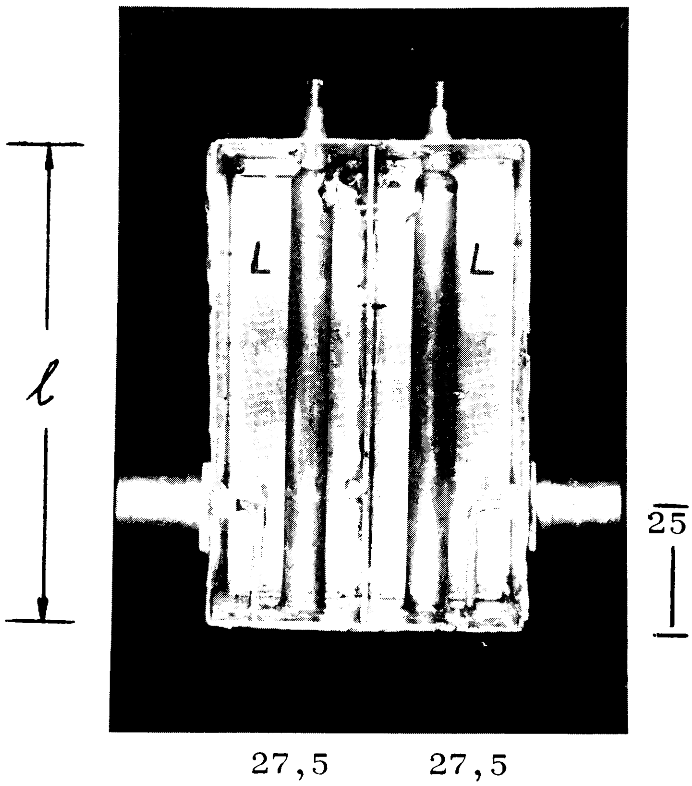


Figure 211

Dimensions:
depth : 25 mm
l = 90 mm
L = 6 mm diameter
C = 0.5 - 12 pF
(both capacitors)

The case was constructed from brass sheet (1mm) and soldered. The cover is screwed to nine M3 nuts soldered to the case (at corners and half-way along each side, not shown on the photograph).

All dimensions in mm

C.3.3 Three-section 70 cm filter

The construction of this filter proves that comparatively little effort can result in a very effective filter. The case may be constructed of pc board. In that case the copper cladding must be on the inside and the partitions must be made of double sided pc board or sheet metal. Tuning is achieved through variable capacitors. This has the advantage that the filter may be resonated at 404 MHz or 432 MHz simply by turning the tuning knobs to previously marked positions. The case (tin plate) is commercially available; both cover plates are removable (see G.4).

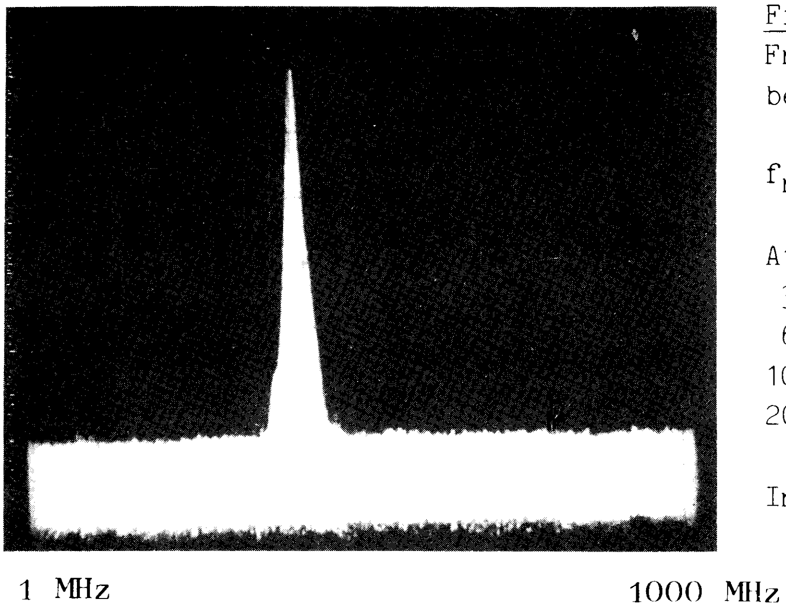


Figure 212
Frequency response between 1MHz and 100MHz

f_{res} : 432 or 404MHz

Attenuation at bandwidth:
3 dB at 2 MHz
6 dB at 7 MHz
10 dB at 11 MHz
20 dB at 20 MHz

Insertion loss: 1.3 dB

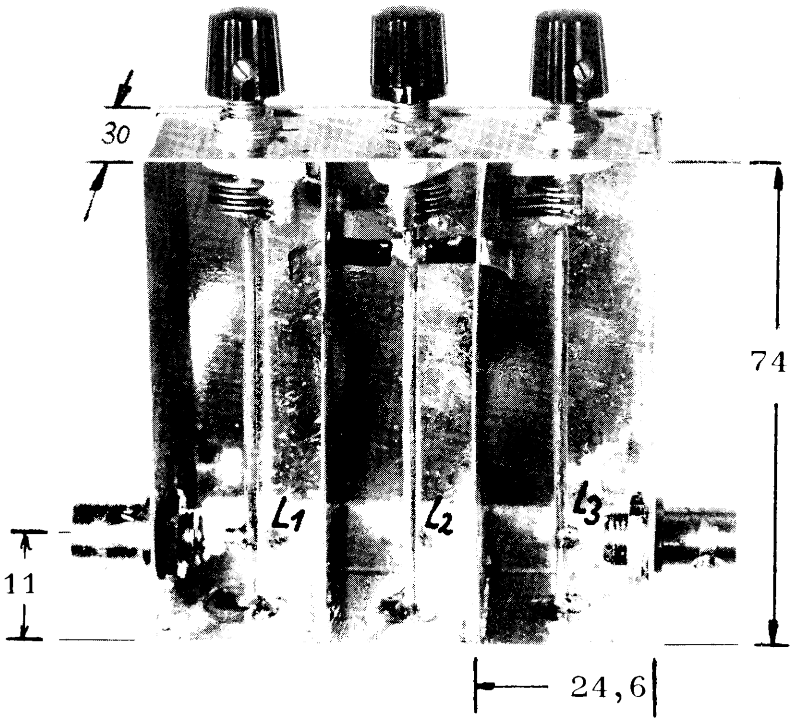


Figure 213
Photographic presentation of three section 70cm filter giving all essential dimensions.

l = 65mm CuAg 1.5mm diam.

C = 1 - 15 pF miniature variable capacitor or ceramic trimmer.

All dimensions in mm.

l_k = 45 mm solder-joined to L2 (see photo)

C.4 Filter arrangements for the 23cm band

For frequencies corresponding to the 23cm band or above filters and resonant circuits should no longer be constructed of mechanically unstable materials. Slight twisting of the case or perhaps even slight temperature changes might shift the resonant frequency in an uncontrollable manner outside the desired range.
The higher mechanical effort will result in improved stability and will pay off in the long run.

Brass was chosen for all the wide band filters described below; it is easier to obtain and less costly than copper and has a lower temperature coefficient. Since tin is a poor conductor it will provide significant transfer resistance to 23cm waves and screw mounting was chosen for assembling all filter components - appart from input and output coupling loops.

Originally galvanic coupling (A.3.1) was used throughout all filter arrangements. But it soon became apparent that the location of the tapping points on the inductors is rather critical. Some sort of sliding clamp on the brass rods would have been one possibility to obtain correct coupling whereas the type of coupling used in conjunction with interdigital filters that are open on both sides would have been another solution.
The following designs of totally enclosed compact filters are, however, based on inductive coupling which poses fewest problems when trying to duplicate them.

Pass bands: We have to differentiate between wide band and narrow band filter arrangements. Wide band filters will pass the complete 23cm band without requiring further tuning whereas narrow band filters will select specific sections of the band (for instance the SSB section only) or specific frequencies such as to provide selectivity for 23cm repeaters. Both concepts thus serve different purposes. The narrow band characteristic is usually tied to the disadvantageous property of increased insertion loss. A possible solution could be the application of a wide band filter with its pass band somewhat shifted to one side. In transverters it thus makes sense to shift the pass band slightly upwards to obtain improved attenuation of the oscillator signal at 1152 or 1268MHz and to further suppress the image frequency.

Finally it is up to the individual OM to decide what sort of effort he is prepared to make; this may be partly governed by the existing equipment of which his station is composed. Several solutions were tried out and evaluated and will be described below.

C.4.1 Twin section filter for the 23cm band

The filter case consists of four pieces of brass stock and two cover plates. It is assembled using counter sunk bolts (M3) at positions marked (<). The threads are cut into the parts labeled "B". The eight taped holes (M3) as seen in figure 214 (and on the reverse side) are used for fixing the cover plates (76x50x1mm). Brass bolts (M5, thread length 25mm) serve as tuning elements. The sides labeled"A" were taped to facilitate the mounting of the BNC connectors. Should this prove impossible then the connectors could simply be inserted and clamped by means of small bolts with counter sunk heads.

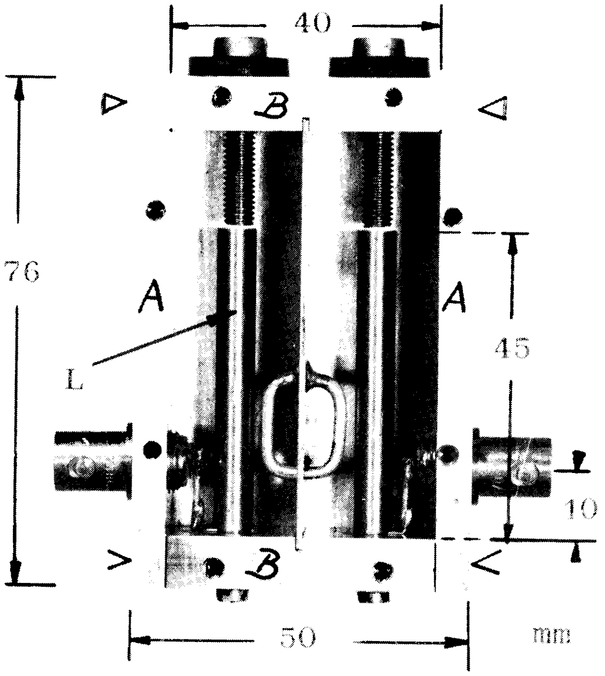


Figure 214

External dimensions without cover: 76 x 50 x 20 mm

Covers: 76 x 50 x 1 mm

Part A: 76 x 20 x 5 mm

Part B: 40 x 20 x 8 mm

The partition (65x20x1mm) is retained by slots that were cut into parts "B". Coupling is provided by a wire loop (2mm diam. CuAg) soldered at its upper end and passed through a 5mm diameter hole 10mm off the base plate.

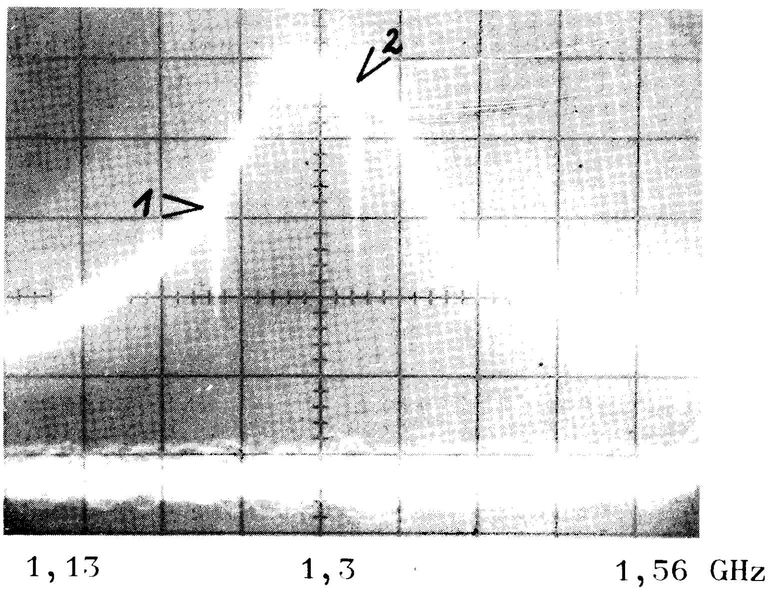


Figure 215 shows the frequency response of the filter.

Z = 50 Ohm

Insertion loss: 0.7 dB

Scale: h = 52 MHz/div.
v = 10 dB/div.

Frequency markers:
1 = 1268 MHz
2 = 1360 MHz

Attenuation at
1268 MHz = -14 dB
1360 MHz = -30 dB

Soldering tags are clamped between the base plate and the two lines (l=45mm, diam.= 6mm) when assembling the filter. Short pieces of wire (1mm diam. CuAg) connect the soldering tags to the BNC connectors and provide input and output coupling. Length of L_k is 11mm, spacing from respective line(L) is 1.5mm. Rubber grommets are placed between case and tuning bolts to provide a firm hold.

C.4.2 Three-section filter for the 23cm band

This filter has narrow-band properties and was intended to be used as bandpass (1296 MHz) or to filter out injection frequencies between 1152 and 1268MHz. Optimized coupling and the specific arrangement of the lines ensure minimum losses in conjunction with narrow bandwidth. The filter construction is governed by the internal dimensions which are 60x60x20mm. L_2, L_4 (6mm diam.) and the case consists of solid brass. L_2 and L_4 measure 54.4mm each and L_3 is 5omm long. Again all components are assembled by means of bolts inserted from the outside. Capacitive tuning is provided through M5 brass bolts and lock-nuts. Soldering tags are inserted at the cold ends of L_1 and L_3 and are clamped between the lines and the filter base using M4 bolts. Both cover plates (measuring 70x75mm) are screwed to the frame which is composed of four parts (comparable with C.4.1 and C.4.3) using eight screws per cover. All other details are explained by the photograph.

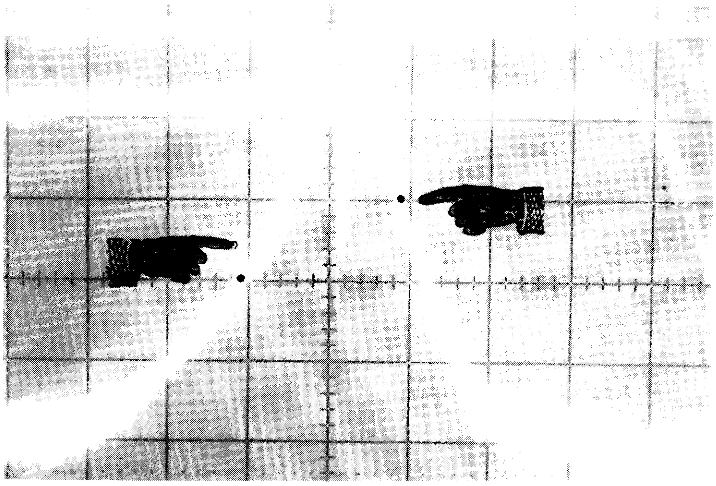


Figure 216
Frequency response and return loss (see E.4.6) of the three section filter.

Upper trace:
Zero (reference) line and return loss at f_{res} (1296MHz) = 26dB corresponding to:
VSWR = 1.1

Lower trace:
Frequency response
Scale: h = 1 - 1.5 GHz
v = 10 dB / div.

Markers: 1.2 and 1.3 MHz

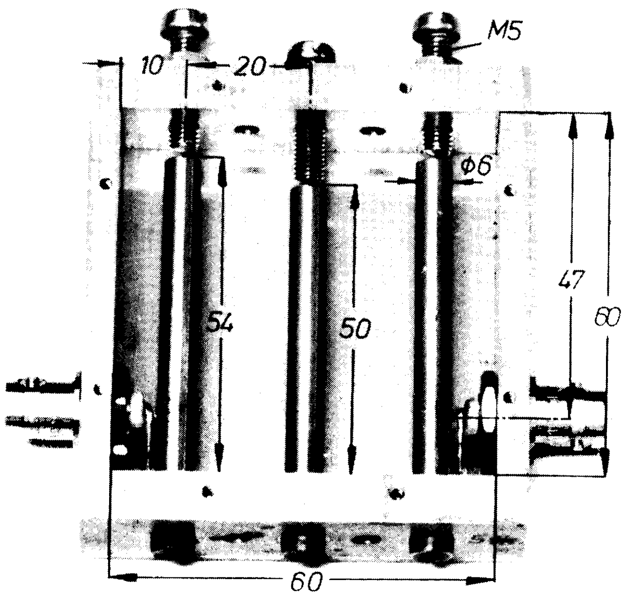


Figure 217
Construction of the three section filter (tuned to 1296 MHz) and dimensions.

L_2/L_4 = 54.5mm, 6mm diam.
L_1/L_5 = 13 mm
L_3 = 50 mm, 6mm diam.

C.4.3 Five-section filter for the 23cm band

This filter - composed of five tuned circuits - exhibits low insertion loss and low ripple throughout the 23cm band. It was designed as broad band filter for transmitters and for measuring purposes. Inspite of the large bandwitdh the nearest injection frequency of 1268MHz will be attenuated by more than 40dB once the filter has been carefully adjusted. Tuning is provided through M5 brass bolts (l = 25mm). They are spring loaded by means of helical bronze springs - thus ensuring stable conditions. All parts are screwed together as described in C.4.1. The soldering tags under L1 and L5 are connected to the BNC connectors using 1.5mm CuAg wire as shown in figure 219. Each of the two cover plates (146x74x1mm) is held by 16 bolts (M3). L1 - L5 consist of 6mm diam. brass rod - each measuring 50mm in length are, again, held by M3 bolts screwed in from outside.

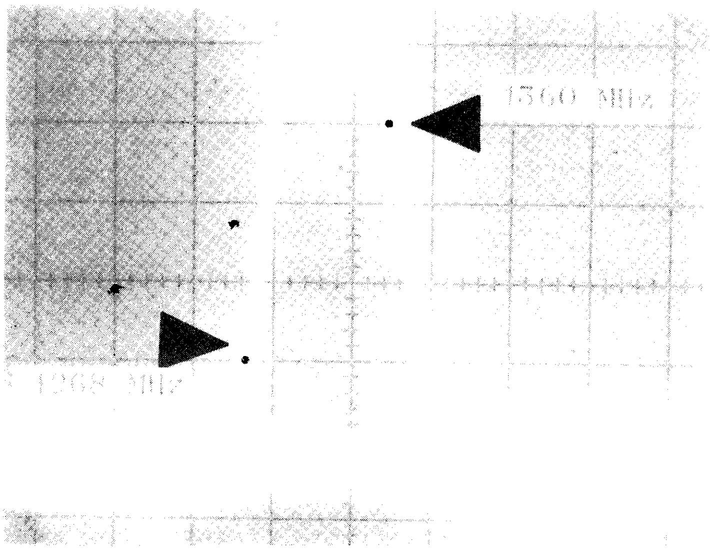


Figure 218

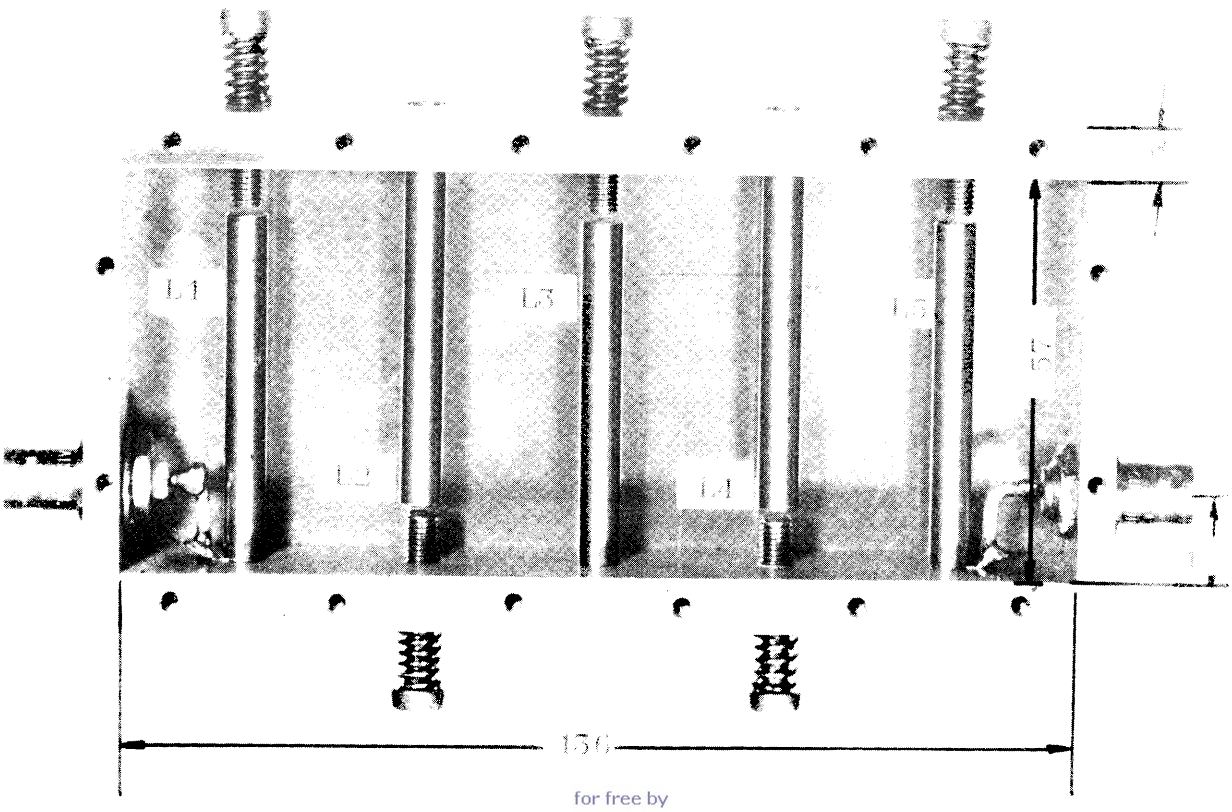
Frequency response of the five-section filter for the 23cm band.

Scale: h = ~ 50 MHz/cm
v = 10 dB/cm

Insertion loss: 0.9 dB

VSWR = 1.06

Figure 219



C.4.4 23cm interdigital bandpass filter

Reference(1) contains descriptions of interdigital filters for the 23cm-, 13cm- and 9cm bands. They had been constructed by OM Dieter Vollhardt (DL3NQ) and were analyzed by DJ5ZE, DL8JT and DL1BU.

To complement the description of filters in brass technique it was decided to construct a filter having five tuned circuits according to reference (1). The same sort of insertion loss (approximately 2.4dB as stated) was achieved. Inspired by the comparatively low insertion loss of the brass filters it was attempted to improve on the original design by DL3NQ. The materials used remained practically the same. Assuming that tighter coupling would reduce the attenuation the spacing between the lines was reduced. The resultant dimensions are indicated in figure 222. Once again M5 bolts of 25mm length provide for tuning. They protrude approximately 3mm into the cavity (f_{res} = 1296MHz). Prior to any tuning attempts they should be set to this position since it is impossible to find the resonance range quickly without the help of sensitive instruments.

The resultant insertion loss was found to be 2.4dB. A significant improvement of the properties seems to be impossible as long as the construction is based on aluminium. The filter design is already optimized.

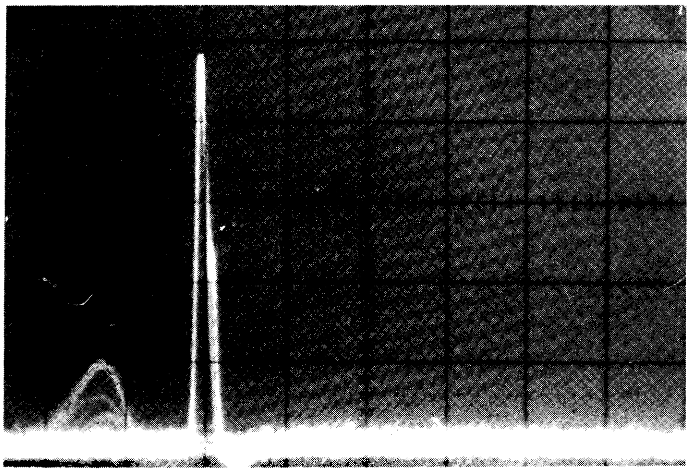
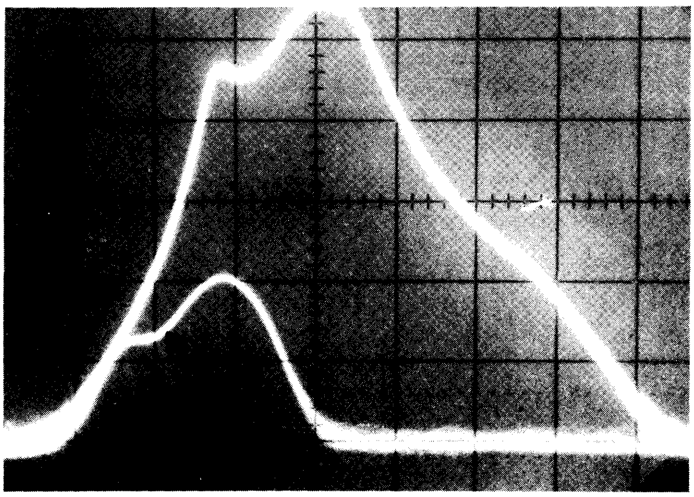


Figure 220

Frequency response of the five-section filter in aluminium-technique for 1 - 2 GHz.

Scale: v = 10 dB/div.
h = ~ 100 MHz/div.



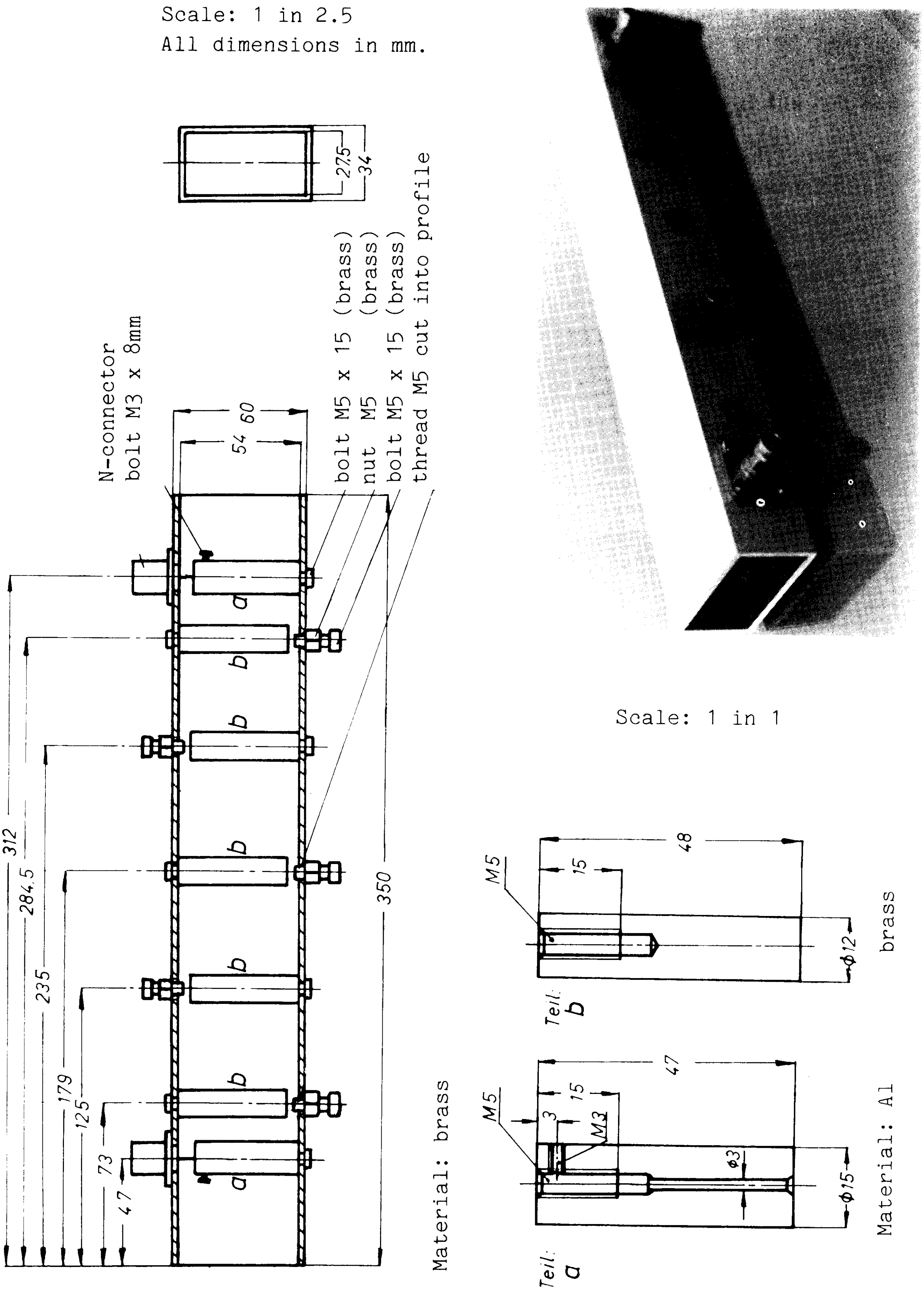
1276 MHz 1316 MHz

Figure 221 is a stretched display of the frequency response.

Scale: h = 5 MHz/div.
v = 10 dB/div.
f_{centre} = 1296 MHz

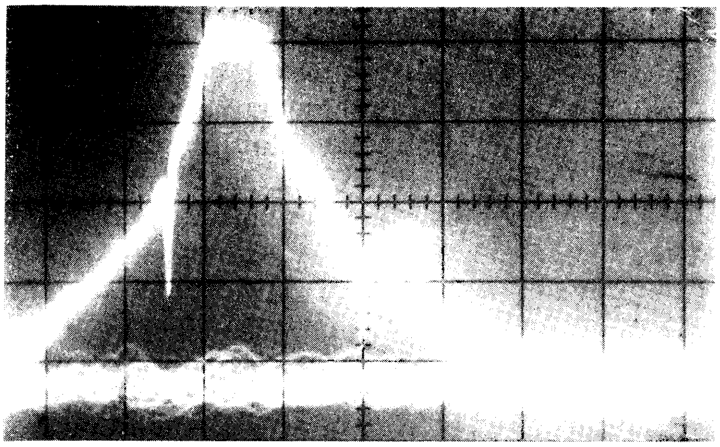
The lower trace (arrow) was erroneously displayed by the instrument and should be disregarded. The efficient application of the filter in case of interference is described in C.4.6.

Figure 222 Construction of five-section bandpass filter for 23cm waves. The described and analyzed filter employes N-connectors.



Reference: (1) UKW-Berichte (Baiersdorf) Vol.2/77 pages 97 ff.

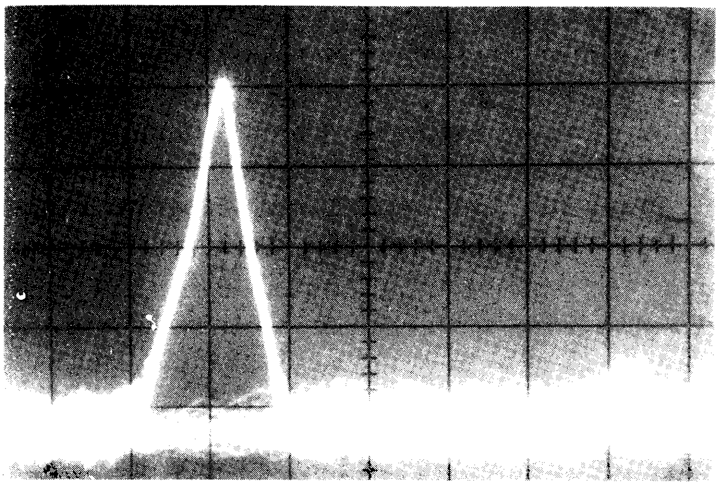
C.4.5 Summary of measured performance of all 23cm bandpass filters



Twin-section filter according to C.4.1

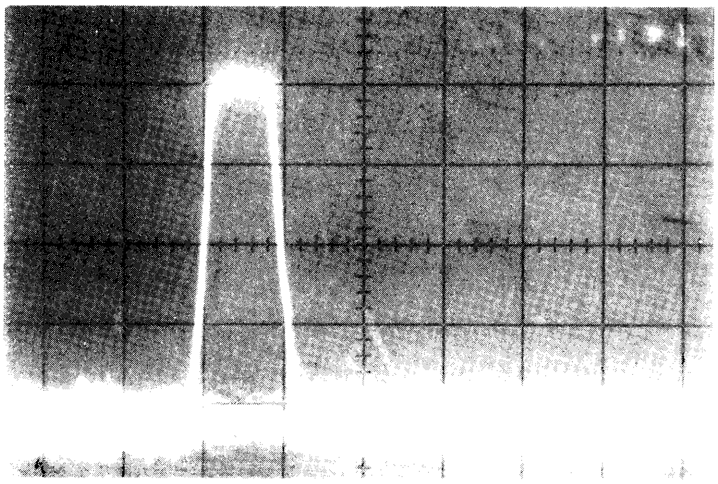
Insertion loss: 0.7 dB

Markers: 1250 MHz
1500 MHz



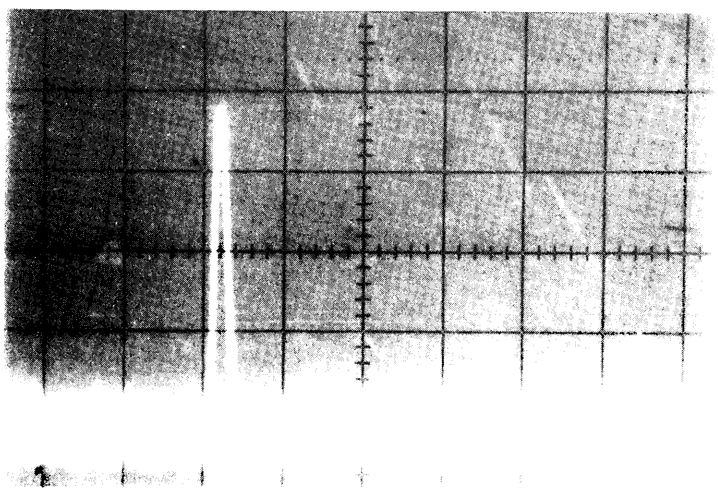
Three-section filter according to C.4.2

Insertion loss: 0.5 dB



Five-section filter according to C.4.3

Insertion loss: o.9 dB



Five-section filter in aluminium technique according to C.4.4

Insertion loss: 2.4 dB

Scale of all diagrams:
horizontal = 100 MHz/div.
vertical = 10 dB/div.

1 GHz 1,5 GHz 2 GHz

C.4.6 Reduction of RADAR interference

Many 23cm band communication links break down due to the interference caused by signal sources inside or just outside the amateur band at - sometimes - very high power levels. The amateur at least believes that they radiate inside the amateur band. It can easily be observed in AM that these interfering signals originate from radar stations and drive the S-meter needle against the stops whenever the radar antenna points to one's own location. These interferences may become continuous if one happens to live in the vicinity of such a radar. As is generally known radars have rather high output power levels - of the order of one Megawatt or more.

If one intends to fight these interfering signals one has to study some literature first to gain a better understanding of the nature of these signals.

One will read about radar spectrum and will come across pulse repetition frequency (PRF), pulse duration (PD), scan period and so on. A multitude of new terms and expressions will come down on the amateur. One will soon appreciate that the radar spectrum (total width of the spectrum) is governed by the steepness of the radar pulse. Furthermore one will realize that there are additional maxima (sidelobes) and minima at regular intervals on either side of the centre frequency (carrier). The distance of these minima from the carrier frequency depends on the pulse duration (PD) of the radar signal according to

$$f_N = \frac{1}{PD}$$

f_N = Frequency spacing between carrier frequency and first minimum in MHz

PD = in usec

To further clarify the problem a Hewlett Packard signal generator was set to a carrier frequency of 1300MHz. The output was amplitude modulated such as to yield 400 pulses per second with a pulse duration of 4usec. This signal was fed to a spectrum analyzer and resulted in the following spectrum being displayed:

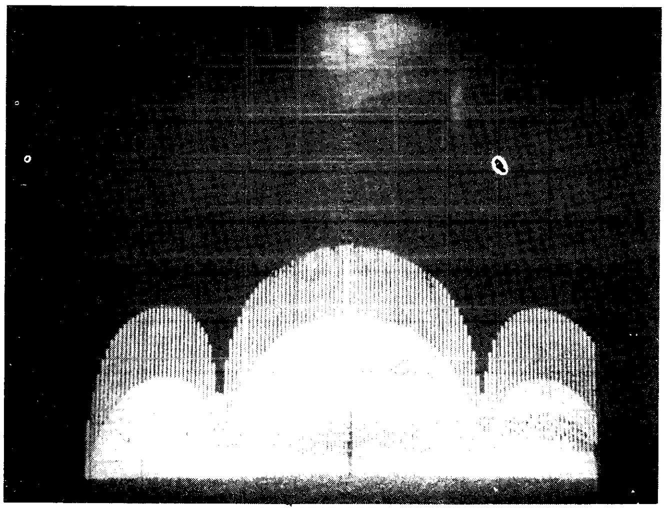


Figure 223 Presentation of a pulsed signal on a spectrum analyzer.
(PRF = 400Hz, PD = 4usec)

h: 100KHz/division
v: 10 dB/division

The spacing of the spectral lines corresponds to the PRF (not all lines are shown).

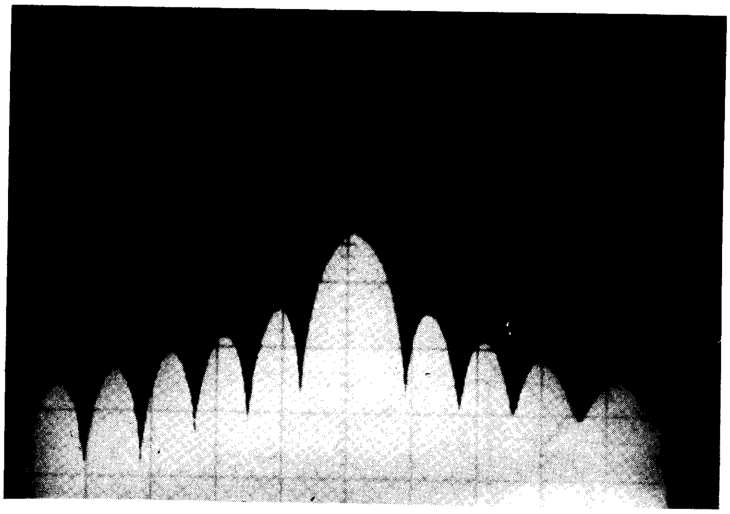
The first minima are located 0.25MHz on either side of the carrier frequency.

Setting the spectrum analyzer to display a wider band will show that the higher order sidelobes will decrease in amplitude. This reduction will be rather marked near the carrier frequency and will level out as one moves away from the carrier.

Let us first analyze a total range of 3MHz (i.e. plus and minus 1.5MHz either side of the carrier). Power ratios may be read off the vertical scale.

Figure 224 shows the spectrum of a pulsed signal generator, carrier frequency = 1300MHz. The position of the minima is defined by the pulse duration (PD) of 4 usec.
The width of the main lobe is 2/PD.

h: 300KHz/division
v: 10 dB/division



1300 MHz

The centre frequencies of the first sidelobes are separated by:

$$f_{A0/A1} = \frac{1}{PD} + \frac{1}{2\ PD}$$

The equation given for f_N applies for all other spacings. We shall not go into the calculation of the sidelobe amplitudes A1, A2,... with respect to the mainlobe amplitude Ao. Instead the calculated values are listed below:

Ao	A1	A2	A3	A4	A5	A6	A7	A8	A9
0dB	-13.3	-17.8	-20.8	-23.0	-24.8	-26.2	-27.4	-28.5	-29.5

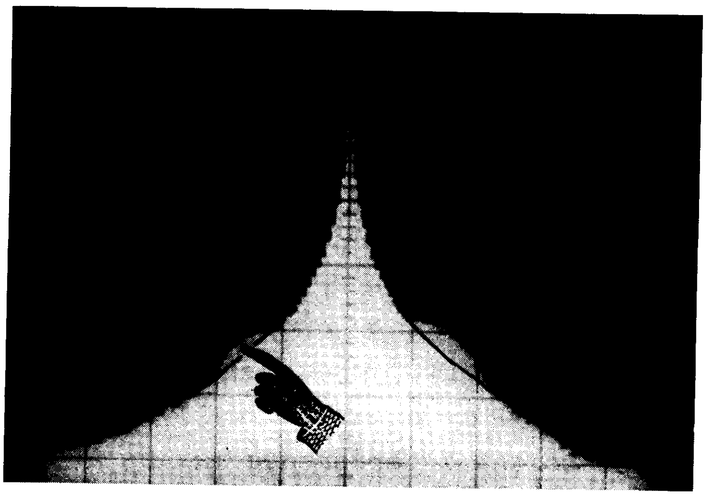
A10: -30.37dB A20: -36.18dB A30: -39.63dB A40: -42.9dB A50: -44.0dB

The levelling-off of the power reduction between higher order sidelobes is clearly noticeable.
Summary: In spite of the relatively wide spectrum of a radar signal the power will be concentrated in a comparatively narrow band. Employing a narrow band filter (C.4) may reduce the sensitivity by 2 to 3dB (insertion loss); but only 15MHz off the main lobe (carrier frequency) the radar signal will be attenuated by something like 40dB (i.e. 1 in 1000) which is an improvement of 8 units on the S-meter.

Figure 225 Here it is shown that in spite of the large bandwidth of a radar signal most of the energy is radiated within a narrow band.

h: 3 MHz/division
v: 10 dB /division

The uneven power distribution shown in this figure is caused by the signal generator.

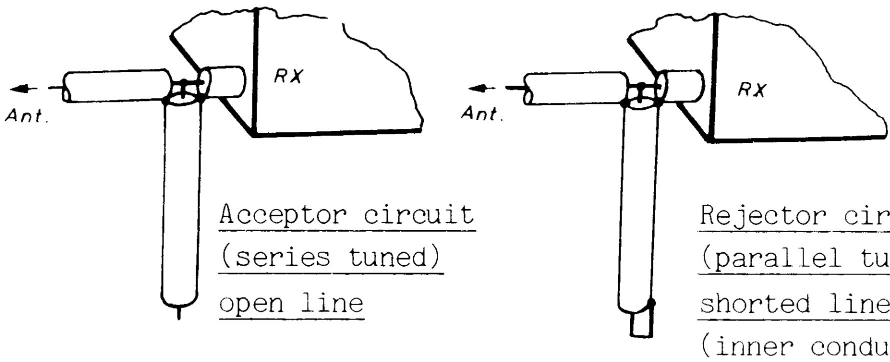


Literature: Zinke/Brunswig: Lehrbuch der Hochfrequenztechnik
Skolnik: Radar Handbook

C.5 Acceptor and rejector circuits composed of rf-cable sections

The most convenient method of keeping an interfering signal away from the input of a receiver or TV-set without too much mechanical effort is the application of quarter wave length stubs. A stub may be arranged as acceptor or as rejector circuit (trap) and should be connected as near as possible to the antenna input of the receiver suffering from interference.

Figure 226
Acceptor and rejector circuits from $\lambda/4$ coaxial cable sections



Acceptor circuit
(series tuned)
open line

Rejector circuit
(parallel tuned)
shorted line
(inner conductor soldered to outer conductor [shielding])

The length of the line may be calculated according to:

$$l = \frac{c \times v}{f \times 4}$$

c = velocity of light
v = velocity factor of cable (see G.1)
f = resonant frequency (i.e. interference)

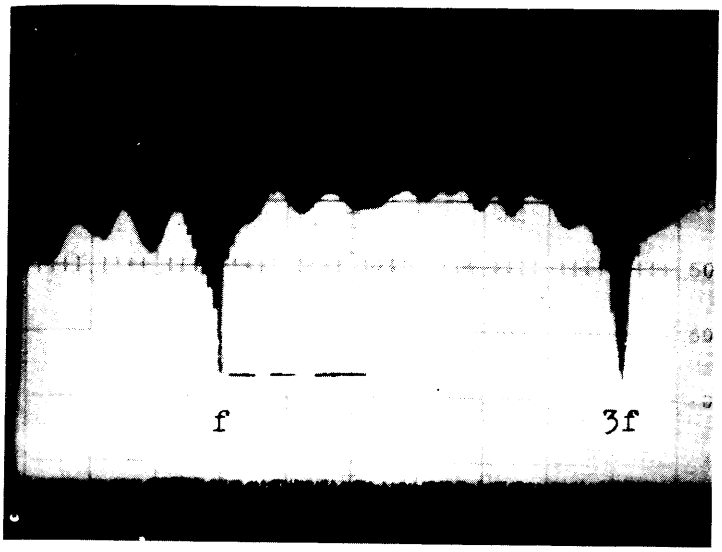
or simplified:

$$l = \frac{300 \times v}{f_{(MHz)} \times 4}$$

If the "open" stub is tuned to the 2m band and attached to the TV-set (resp. the antenna amplifier) of one's neighbour a reduction of the 2m signals of 25 to 30dB may be expected. At the same time all uneven harmonics of the design frequency (3f, 5f. etc.) will be attenuated as well. Since the UHF amateur is particularly concerned with f and 3f a photograph may clarify the situation.

In many cases the problem is not caused by the amateur radio apparatus but by the broad band antenna amplifiers feeding a number of TV-sets or by the broad band input stage of a single TV-receiver lacking any components for signal selection. This statement is frequently verified by the interference free operation of one's own television. The following experiment will illustrate how the own signal that gives rise to the interference may be attenuated.

A panoramic receiver (set for 1-500 MHz) was connected to a signal generator that could be tuned slowly across the 1-500MHZ range. An open stub (according to the diagram) consisting of 35cm ($\lambda/4 \times v$) of RG58/U was inserted into the connection between signal generator and receiver.

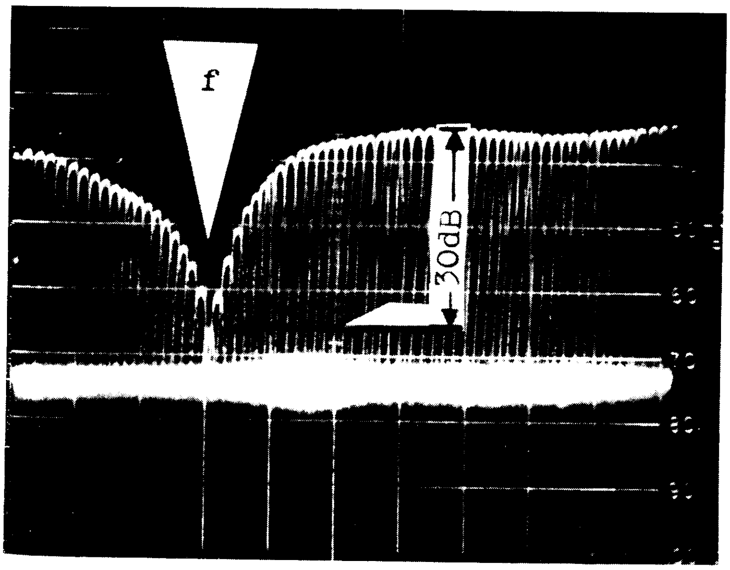


1 MHz 500 MHz

Figure 227 covers the frequency range 1-500MHz as displayed by the panoramic receiver and shows the attenuation effect of the open stub (acceptor circuit) at frequencies f = 144MHz and 3f = 432MHz.

v: 10 dB/division
h: appros. 50 MHz/div.

Ripple at the upper power limit was caused by the signal generator. The resultant attenuation will suffice in many instances where TVI is caused by amateurs or taxi car transmitters. An enlarged proportion of figure 227 will give a good idea of the skirts of the response curve.



124 144 164 184 204 MHz

Figure 228 indicates the width of the the stop band at frequency f.

h: 10 MHz/division
v: 10 dB/division

Obviously this type of stub arrangement will attenuate only those interferences that are due to insufficient overload capability of the receiver. Protection against spurious radiation can only be provided through a filter at the transmitter output.

If the quality of the previously good and clear TV picture is reduced upon correct insertion of the stub one should investigate whether 3f 5f or 7f fall into the television frequency band. It may become necessary to tune to a different channel or turn the antenna to a different TV- station. When arranged as rejector circuit a broad band around the design frequency will be enhanced. Attenuation will be observed at the even integer multiples of f (2f, 4f, ...etc.).

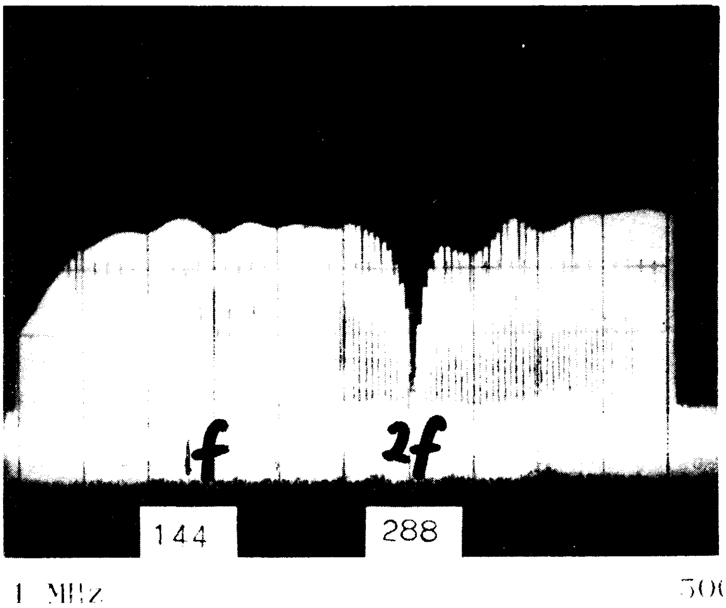


Figure 229 Frequency response of the stub as rejector ciricuit. i.e. shorted $\lambda/4$ stub.

h: approx. 50MHz/div.
v: 10dB/division

Frequency range:
1 - 500MHz

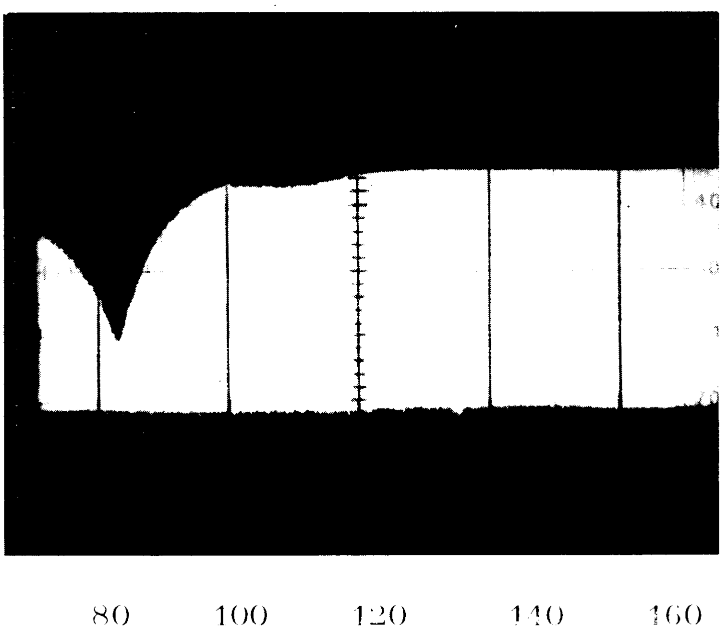


Figure 230
Further example:
A 2m band amateur receiver is overloading due to a local UHF broadcast station.
The OM connects a 60cm section of open circuited RG58/U cable to the receiver input and achieves the shown filtering effect.
Suspect spurious radiation if no effect is observed!

C.5.1 Stacking of acceptor circuits

Combining several acceptor circuits will lead towards increased attenuation and a rather narrow stop band for interference suppression. The following results were achieved with an arrangement according to figure 231. If it is intended to suppress a complete frequency band then several stubs of decreasing lengths to cover the range between lower and upper frequency limits should be combined. In that case l_2 should have a value between l_1 and l_3.

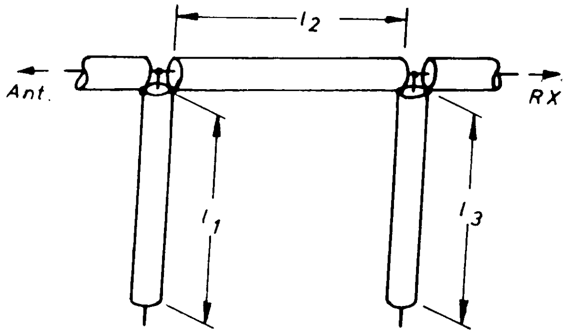


Figure 231
Stacking of two equal length stubs
$l_1 = l_2 = l_3$
Calculations as in C.5

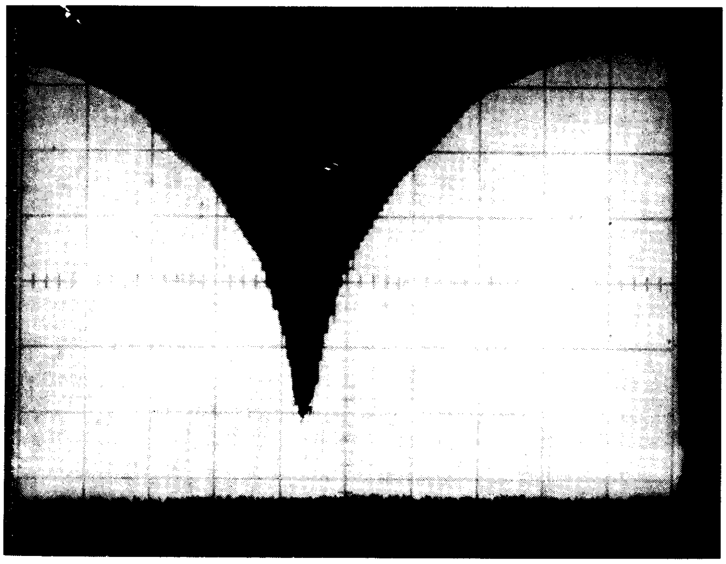


Figure 232
In this case a 48MHz oscillator signal had to be suppressed.
Attenuation at resonant frequency is 56dB.

h: 3MHz/division
v: 10dB/division
f_{middle} = 500MHz

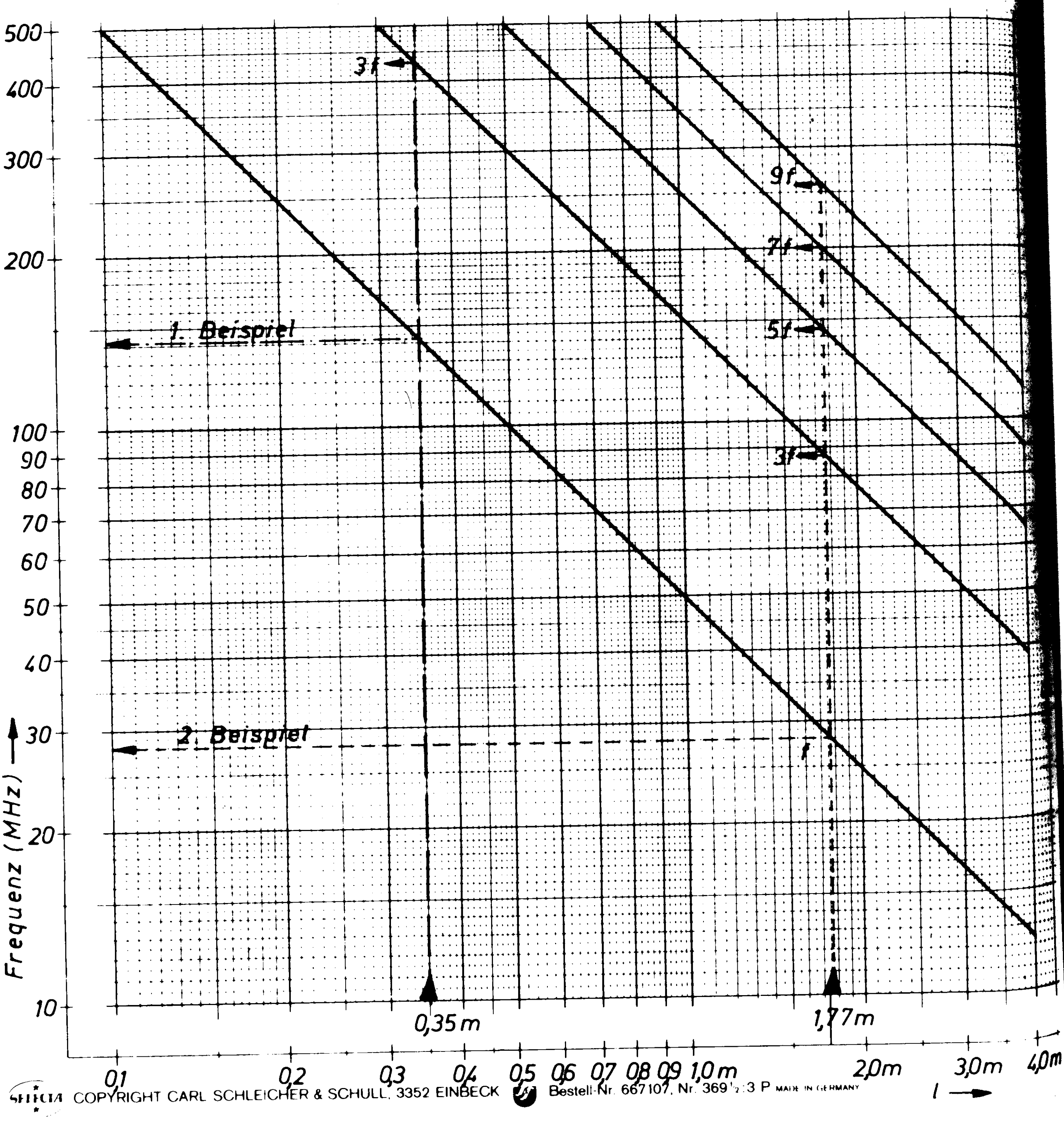
Figure 233
This figure indicates resonant frequencies at 3f, 5f and 7f. They are slightly below the calculated values due to additional shortening (ratio wavelength divided by diameter).
h: 30MHz/div.
v: 10 dB/div.

From the diagram we can directly read off the relation between stub length ($\lambda/4 \times v$) attenuated design frequency (f) and the position of the attenuated harmonics 3f, 5f etc.

First example: A 35 cm section of rf-cable (open end) designed for 144MHz will simultaneously attenuate 2m and 70cm signals.

Second example: If a 10m transmitter causes interference in a TV-set operating in the 6m band an arrangement according to C.5 or C.5.1 of length 1.77m connected to the VHF input will suffice. The resonances 3f, 5f etc. will cause additional attenuation in the UHF broadcast and UHF tv-bands.

Figure 234



C.5.2 Acceptor circuits in stripline technique

Acceptor circuits as described in Section C.5 are quite suitable for frequencies up to 432MHz. Beyond this band the cable dimensions are beginning to become critical and a design in stripline technique on epoxy based pc board should be superior. Now what sort of problems will such a device solve in the 23cm band?

Operation in the 23cm band is dominated by broad band systems. This holds true for the antenna, the preamplifier and the mixer. It is a fact that narrow band properties call for considerable effort at these high frequencies. Consequently, image frequency noise will be amplified and passed on to the mixer. This noise contribution will degrade the overall performance and is thus undesirable. Since this noise component should be attenuated by at least 3dB (preferably 6dB) some sort of filtering is required. If no radar interference has to be taken into consideration then the large mechanical effort of filters in heavy brass or aluminium technique as described in C.4.1 to C.4.4 is hardly justifiable and the following stripline circuit offers a neat solution to the problem.

Theory of operation: As shown in figure 235 short coupling lines (as discussed in section B.4 "directional couplers") are arranged in parallel with a stripline conductor (Z = 50 Ohm in this particular case). Open quarter wave length stubs are connected to these coupling lines at right angles. Upon tuning to the image frequency they will suck out signal components at that frequency. Since 6dB of image rejection is sufficient only light coupling is required. For higher values of attenuation 4mm wide copper strips (length L_K) are soldered to L_K and bent upwards such that they rest about 2mm above the 50Ohm stripline.

If the stubs are supposed to filter out the 1152 or 1268MHz components then 0.5 - 6pF trimmers must be added in positions "A" and connected to ground. This facilitates easy tuning to the image frequency. Placing these trimmers at the ends of the strips labelled "3" and bridging all the gaps will extend the range of frequencies that can be filtered out down to the 70cm band.

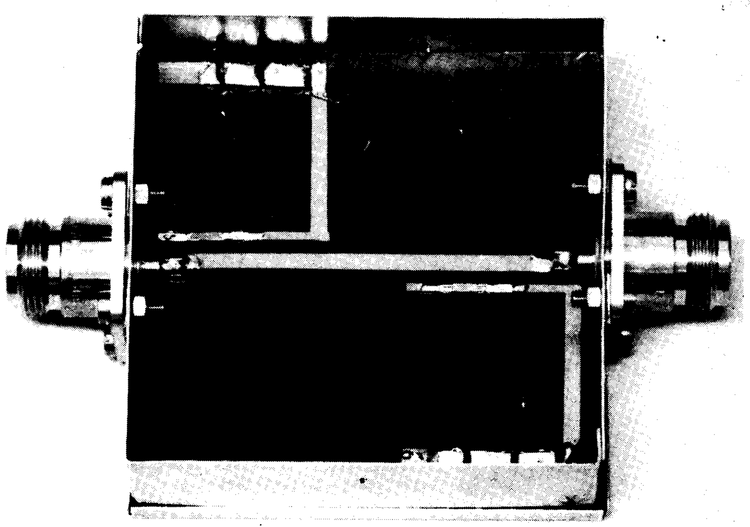


Figure 235 shows the stripline filter in its case 72x72x30mm

No additional components are fitted.

The width of the main stripline depends on the required characteristic impedance (i.e. 50Ohm line) and the properties of the pc board. The relevant theory is discussed in sections A.2.4 and G.3.2. For the 1.6mm epoxy board we decided on a strip width of 1.3mm for the second sample. All other lines should have a width of 2mm; but their dimensions are not critical.
The attenuation curves of the original sample are shown in figures 236 and 237. The stub interruptions have been bridged up to segments "2"; then a 5mm wide copper strip - connected to ground - was placed over segments 1 and 2. Tuning to the desired frequency is achieved by forcing the strip closer to or farther away from the stripline. Using a proper capacitor is certainly the neatest solution. These trimmer capacitors must - however - be of good quality since they greatly influence the overall Q-factor and thus the skirt selectivity. Trimmers with air dielectric are to be preferred.

Adjustment: If absolutely no test instruments are at one's disposal the filter module is placed in front of a broad band mixer (D.4.2). The fixed frequency generator (B.8) is set to deliver a 23cm band signal just above the noise level. Then the trimmers are adjusted for best readability. The 1dB (23cm) insertion loss may be neglected if the module is placed between the preamplifier and the mixer stages.

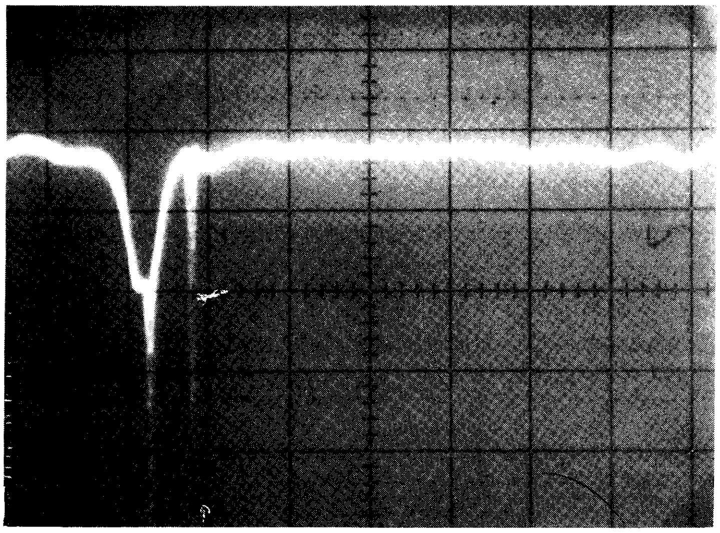


Figure 236
Stripline acceptor circuit tuned to 1240MHz. This corresponds to the image frequency when using a 28MHz if-strip and an injection frequency of 1268MHz.

h: 1 - 2 GHz
Markers: 1240 MHz
1296 MHz
v: 5 dB/division

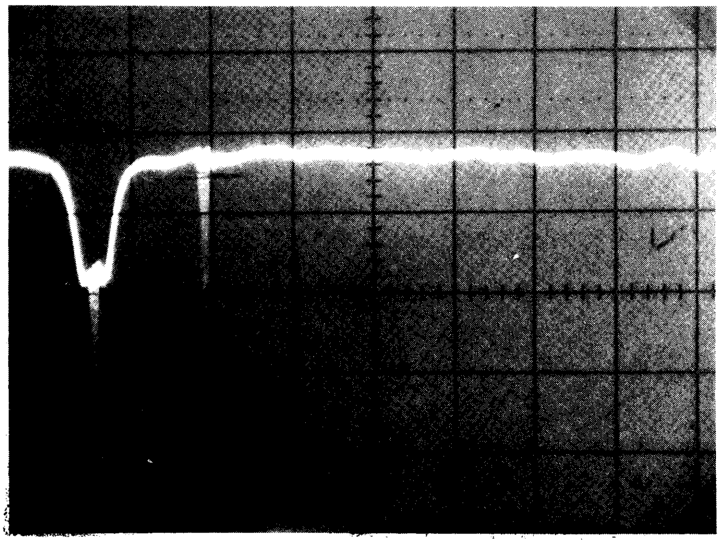


Figure 237
Stripline acceptor circuits tuned to 1152MHz to suppress the now unwanted f_0 signal.

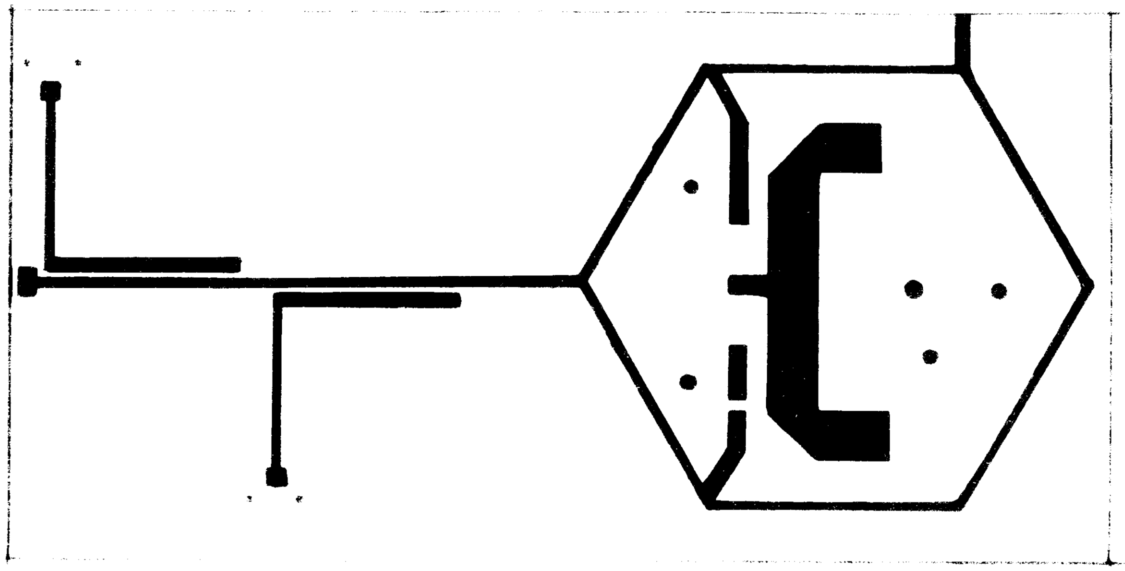
h: 1 - 2 GHz
Markers: 1152 MHz
1296 MHz
v: 5 dB/division

This filter arrangement lends itself to integration into a converter concept. In that case the pc board is simply extended by the filter dimensions.

There must surely be a large number of applications that are not necessarily limited to the 23cm band. The connectors should be chosen in accordance with the frequency range. In our case an N-type connector was selected.

If such a filter is supposed to operate on even higher frequencies it is strongly recommended to use teflon as substrate material. With increasing frequency the width of the stop band will increase as well. The same applies for tighter coupling. No trials were carried out in the 13cm band. In principle the same theory applies. Coupling lines and stubs should be shortened by about 10mm. Tuning should be performed by means of 0.2 - 3pF trimmers.

Figure 238 shows one way of integrating the filter arrangement into the converter concept described in sections A.5.2 and D.4. Here a case with dimensions 146x72x30mm (G.4) is used.
For stripline width and dimensions of ring segments the reader is referred to the respective chapters. The distance between the stubs should be quarter wavelength multiplied by the velocity factor of the substrate material.



C.6 The design of helical filters
(D.Reichel, DC9NL)

C.6.1 General considerations

The Helical resonator consists of a cavity containing a coiled inner conductor. One end is connected to the cavity - low transfer resistance being of great importance. The other end remains open or could be connected to a trimmer capacitor. Helical resonators are nothing else but quarter wavelength cavity resonators having coiled rather than straight inner conductors which results in significantly smaller dimensions. The cavity may have circular or rectangular cross-section. In either case the cavity surface must have excellent conductivity.

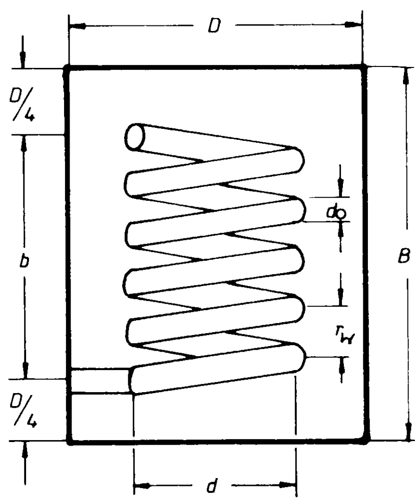
Coupling between consecutive helical resonators may be performed by means of capacitive, inductive or radiation coupling (through an iris).

The correlation between the no-load quality factor "Q", resonant frequency and cavity dimensions is shown overleaf. The hatched area indicates the range in which helical resonators of defined volume will have better Q-factors than other types of resonators. For higher frequencies and higher Q-factors a conventional coaxial cavity resonator is superior. If frequency and required Q-factor are below the lower boundary then conventional LC resonators are the best choice. These conditions are represented by the points "lower limit" and "upper limit" in the nomogram. The latter is accurate to within $\pm 10\%$ and was designed around the following assumptions:

The no-load Q-factor is achieved for single-layer low loss coils of copper wire operating in a seam-less cavity. A screened inductor designed according to the equations stated further down will show a corresponding Q-factor at resonant frequency (f_{res}).

C.6.1.1 Symbols:

Figure 239

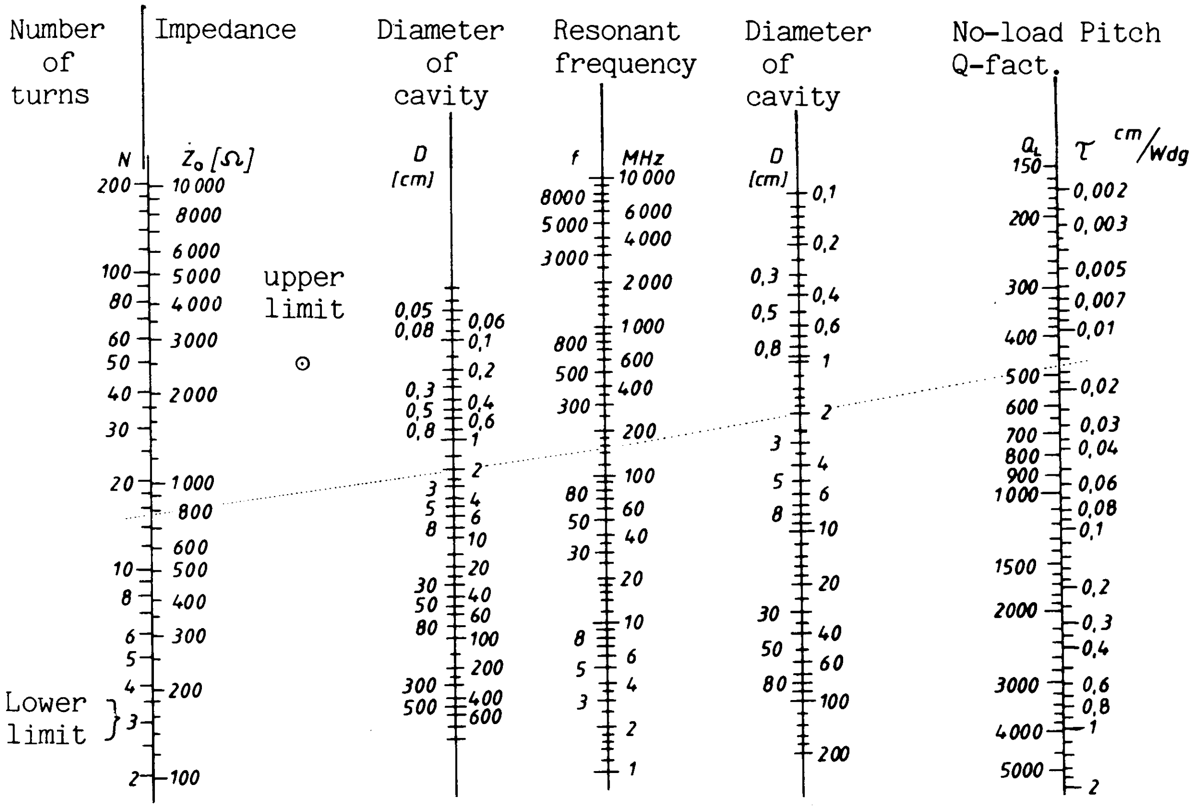


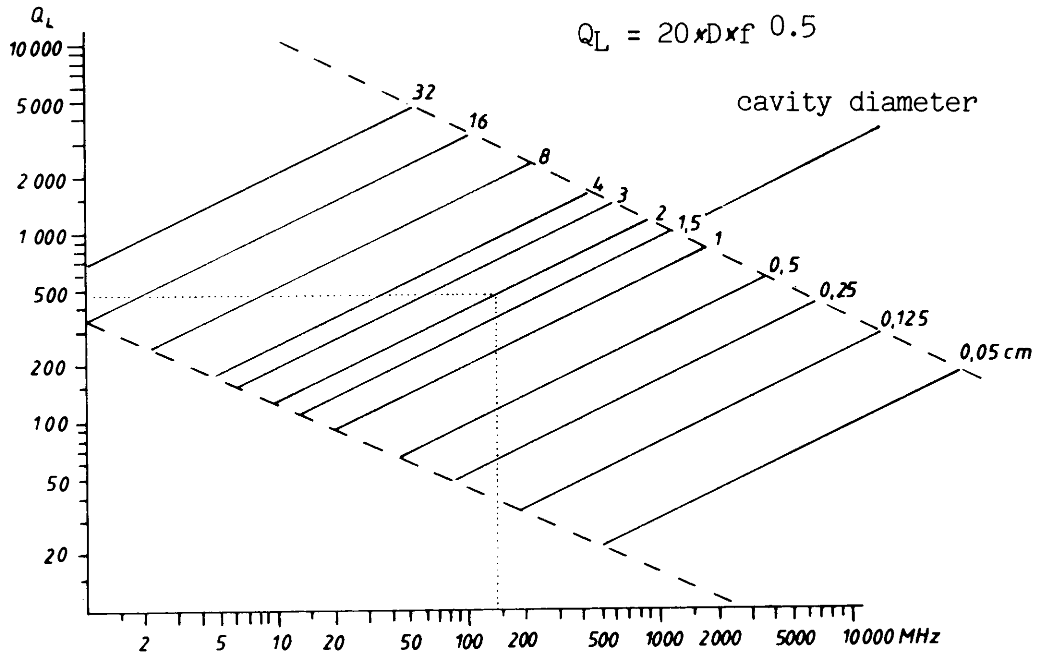
b = length of helix (inductor)
l = length of wire required for the helix
B = length of cavity
d = mean internal diameter of inductor
D = internal diameter of cavity
D_O= diameter of wire forming the inductor
f = resonant frequency in MHz
n = pitch in number of turns per cm
N = total number of turns
Q_L= no-load Q-factor
r_W= 1/n = spacing between turns
= penetration (skin effect: A.1.3)
Z_O= characteristic impedance in Ohm

All dimensions in cm

Figure 240

Nomogram for ../4 helical resonators





C.6.2 Equations and diagrams

The nomogram is applicable within the following limitations:

$0.45 < d/D < 0.6$ with $b/d > 1.0$
$0.45 < d_o/r_w < 0.6$ with $b/d = 1.5$
$0.5 < d_o/r_w < 0.7$ with $b/d = 0.4$
$(d_o > 5\delta)$

The maximum number of turns is calculated from $N = 4800/(f \times D)$ with $d/D = 0.55$ and $b/d > 1.0$

Spacing between turns and characteristic impedance:

$r_w = 1/n = f \times D^2 / 5840$ [cm per turn]

$Z_o = 25 \times 10^4 / (f \times D)$ [Ohm] with $d/D = 0.55$ and $b/d = 1.5$

The most important prerequisite for all considerations is:

$B \approx (b + D/2)$ and $r < d/2$

Empirical results point to the following relations:

$$L = \frac{(n^2 \times d^2)}{100} \times \left(1 - \left\{\frac{d}{D}\right\}^2\right)$$ [µH/cm axial inductor length]

$$C = 0.3 \times \left(\log \frac{D}{d}\right)^{-1}$$ [pF/cm axial inductor length]

$$v = \frac{f \times \lambda}{2.54} = 1000 \times (L \times C)^{-0.5}$$ [cm/usec]

$$l = 0.94\ \lambda/4 = 600 \times f^{-1} \times (LC)^{-0.5}$$ [cm]

$$Z_o = 1000 \times (L/C)^{0.5} = 92500 \times (b \times f \times C)^{-1}$$ [Ohm]

The following equation defines the quality factor Q_L as function of the cavity volume (V_K in cm^3).

$$Q_L = 21.5 \times f^{0.5} \times (V_K)^{1/3}$$

assuming that: $0.4 < d/D < 0.6$ and
$1.0 < b/d < 3.0$

Bibliography: Reference Data for Radio Engineers
Howard W.Sams & Co 1973

C.6.3 Examples of filters for the 2m and 70cm bands

C.6.3.1 Filters for the 2m band

A bandpass filter utilizing two helical resonators suitable to provide selectivity preceding or following an rf-amplifier is described.
The design offers both mechanical and electrical stability. It will accept considerable temperature variations without noticeable changes of its electrical properties. It is thus suited to operate either inside the receiver or directly at the antenna feed point.
To keep costs down the use of precision-type trim capacitors was avoided in particular as they would have further complicated the construction. Tuning is provided through brass bolts and lock-nuts. This results in an arrangement consisting of a box and a removable base plate. Both inductors and their coupling loops are installed on this base plate. Serviceability is thus optimized. Coupling between the two resonators is by radiation through a slot in the partitioning wall.
Brass stock of 4x20 mm was used for the construction as this thickness of material will pose no difficulties to drilling and tapping (M3).

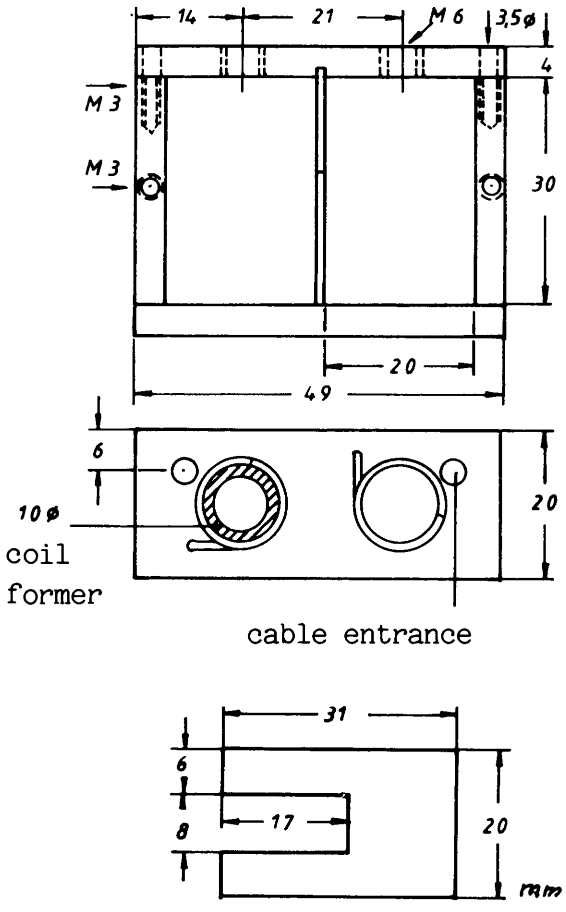


Figure 241
Construction of 2m filter

cavity frame.

Base plate

Partition and coupling iris

To start with, two each 30mm and 49mm long pieces are joined to form a rectangular frame. The partition with its coupling iris is held in a saw cut in one of the two long pieces. The two side plates are cut from 1.5mm brass sheet and screwed to the frame. This - already rather strong - assembly is now (preferably hard-) soldered. The base plate (long piece without saw cut) should - off course - be removed before soldering.

Now the top is drilled (4.8mm hole) and tapped (M6 thread) to accept the two tuning screws right in the centre of the two cavities.

Read off the required number of turns for a cavity diameter of 2cm and a frequency of 145MHz from the nomogram; the result: N = 15 turns.

Since the helix is supposed to be separated from the cavity ends by D/4 the effective inductor length is b = 2cm.

Length of wire $l = 0.94 \times \lambda/4 = 47$ cm.

With 15 turns d will be $47/(15\pi) = 1$ cm.

Spacing between turns $r_W = 1/7.5 = 0.133$ cm.

Since $b/d = 2$, d_O/r_W must be smaller than 0.6. This results in a wire diameter of 0.8 mm.

Such thin wire will exhibit unsatisfactory mechanical stability if no support is provided and a coil former must be used. It could consist of glass, PVC, nylon or any other plastic material of the correct diameter. Bear in mind that the relative dielectric constant E_r of the plastic material will shift the resonant frequency approximately 10MHz down.

The coil ends are inserted into holes in the base plate and soldered. It is advisable to insert long hollow brass rivets into the cable entrances. They are solder-joined to the base plate. Now it is quite easy to solder the screening braid to the thin rivet material. If no suitable high power soldering iron is available then the soldering - or at least the heating up - may be performed using the electric kitchen stove.

Afterwards the coil formers are inserted into the coils and glued. Next the coaxial cable is soldered to a tap 3/4 of a turn from the cold end and the screening braid is soldered to the hollow rivet.

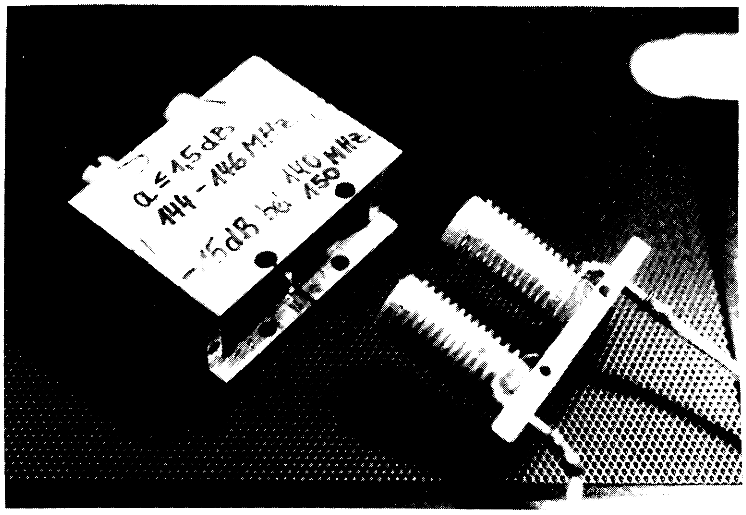


Figure 242
Construction of the filter. The case has been removed.

The photograph of the opened filter shows how the turns have been fixed to the coil former. Care should be taken to prevent any of the resin entering the space between the two inductors i.e the iris as this will certainly increase the insertion loss.
The screw mounting of the base plate may be done from the sides or to the frame. The only important aspect is a tight fit all around the cavity sides.
Upon completion of the assembly we may start to tune the filter.

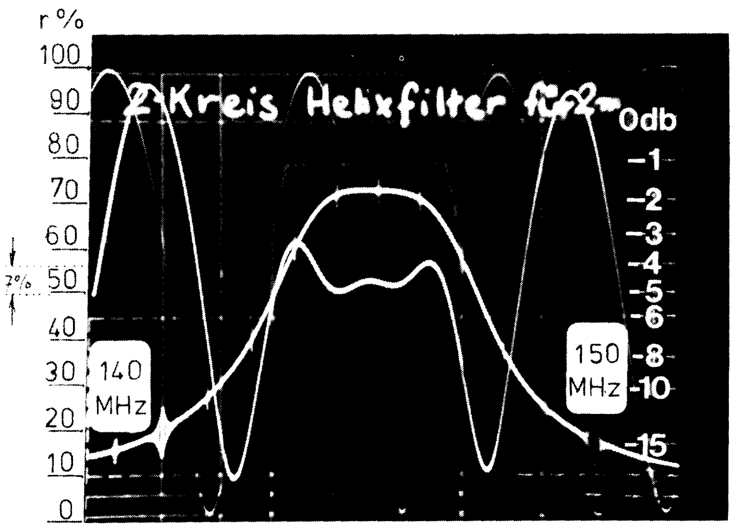


Figure 243
Frequency response and matching characteristics of the 2m filter.

Figure 243 represents the pass band of 145 ± 6 MHz. The numbers near the right edge of the photograph indicate the attenuation measured in dB. From that the insertion loss may be seen to be approximately 1.5dB.

The background indicates the 100% calibration of the incident voltage. This is taken as reference for the curve representing the matching characteristic (no-load and short circuit conditions of the signal generator).

The reflection coefficient r has a value of approximately 7% inside the pass band.

Tuning is achieved by turning both tuning screws until a receiver following the filter indicates maximum input voltage on 145MHz. Following that one should see whether the attenuation remains constant near the band limits. Should this be the case then alternate adjustments of the two screws should soon lead to the desired response. It is advisable to shift the point of least attenuation into the SSB segment of the band.

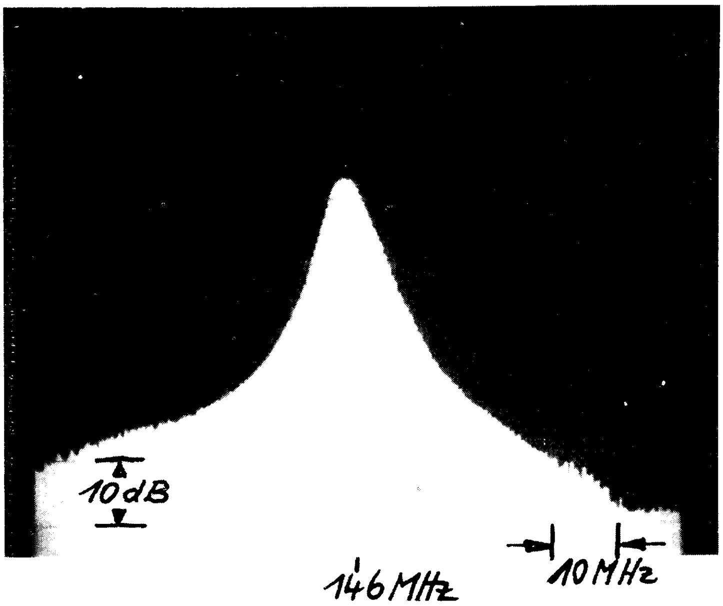


Figure 244
Attenuation properties of the filter over a frequency range of ± 50 MHz around resonance.

h: 10 MHz/division
v: 10 dB/division

C.6.3.2 Filter for the 70cm band

It was intended to design a filter that meets the requirements concerning stability and ease of construction.
For that reason the same type of construction as for the 2m filter was chosen.
The nomogram indicates the maximum number of turns corresponding to a cavity diameter of 2cm at a frequency of 435MHz.
N = 5 turns

Wire length $l = 0.94\ \lambda/4 = 16.5$ cm

r_w turns out to be 3 mm.

From $d_o/r_w < 0.6$ we obtain $d_o = 1.5$ mm

Inductor length $b = r_w \cdot N = 15$ mm.

Wire length and number of turns result in $d = 16.5 : 5 : 3.14 = 1.05$ cm.

Cavity height $B = b + (2\ D/4) = 21$ mm.

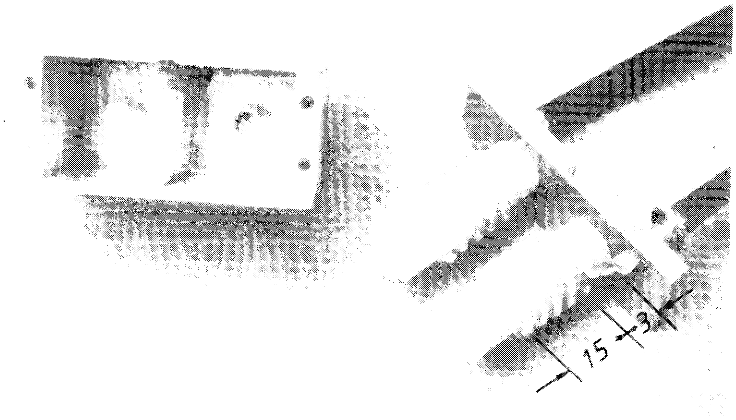


Figure 245
shows the helical filter prior to assembly. The taps are located 1/3 turn off the cold ends of each inductor. Note the coil winding sense.
(Coil formers not yet cut to length).

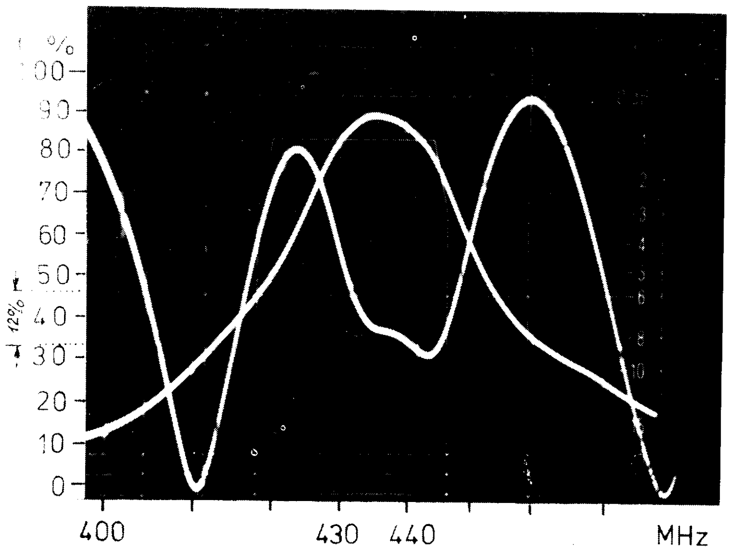
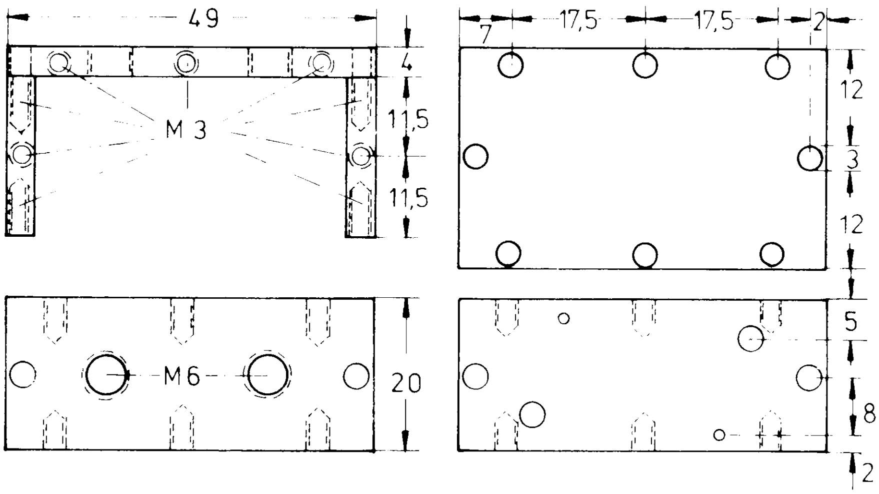


Figure 246
Insertion loss and return loss across the range
430 - 440MHz ± 35MHz
a = 0.5 dB
r = 12 %
resulting in
s = 1.3 dB

During the alignment operation it became apparent that the coupling iris had to be enlarged so that only a 1mm wide strip of the partitioning wall was left over (to achieve sufficiently low insertion loss). Leaving out the partition altogether turned out to be no disadvantage.

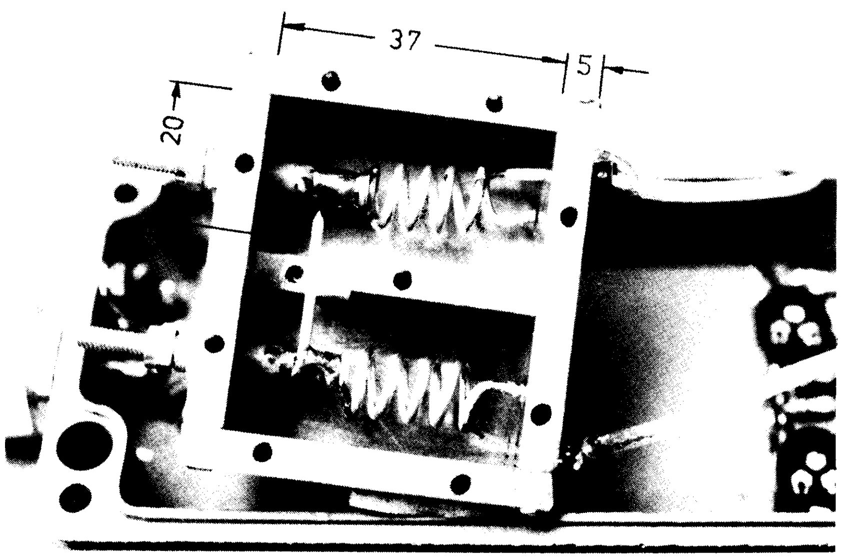
Figure 247
Construction of 70cm filter box



Should the reader use the case dimensions of the 2m helical filter then the iris should be enlarged to measure 13mm in width and 21mm in height.

Apart from radiation coupling capacitive coupling may be employed. This only requires a short length of wire - one end of which is soldered to a trim capacitor, a support, or directly to the helix. The precision-type capacitors shown in figure 248 do not lend themselves to direct mounting onto the thick walls. It is necessary to mill the wall thickness around the installation hole down to 2mm.

Figure 248
shows a helical filter employing capacitive coupling



C.7 Antenna multiplexer for broadcast, 2m and 70cm bands

(N.Schramm, DC9NI)

Use of the antenna multiplexer described below allows two transceivers (15 watt max.) for 2m and 70cm operation and the car radio (long, medium, short wave and UHF) to be connected to a 2m mobile antenna.

Theory of operation:

1. 2 m - branch:
   The 2m signal is passed on through the pi-filter C1, C2, L1 and the 2m series-tuned circuit (C3, L2) to the antenna terminal. Two 2m rejector circuits L6, C8 and L7, C9 as well as the 2m acceptor circuit (L10, C12) prevent 2m energy from reaching the car radio terminals.

2. 70 cm - branch:
   The 70cm signal is passed on through the pi-flter C4, C5, L3 and the 70cm series-tuned circuit (L4, C6) to the antenna. Once again two 70cm rejector circuits L5, C7 and L8, C10 as well as the 70cm acceptor circuit (ℓ9, C11) stop the 70cm energy from reaching the car radio.

3. Broadcast branch
   The frequency range from 3 MHz to 108 MHz (no measurements could be taken at even lower frequencies) may pass practically unhindered from terminals Ra to the antenna terminal.

Terminals U_V and U_r deliver dc voltages corresponding to forward - respectivly reflected - power to provide continuous monitoring of the antenna matching (and over-all function).

Measured attenuation:

2 m	--->	Ant	1.2 dB
70 cm	--->	Ant	0.9 dB
2 m	--->	Ra	52.0 dB
70 cm	--->	Ra	47.0 dB
2 m	--->	70 cm	38.0 dB
70 cm	--->	2 m	34.0 dB
Ant	--->	Ra	max. 3.5 dB (typ. 1.5 dB)

Alignment procedure:

1. - Connect milliwatt meter to Ra-BNC connector.
   - Connect power meter to antenna-BNC connector.
   - Connect 2m TX to 2m-BNC connector.
   - Connect 70cm TX to 70cm-BNC connector.

2. Turn on 2m TX (~1 Watt). Adjust C1, C2, C3 and C8 for maximum power meter reading. Adjust C12, C9 and C8 for P_{min} as indicated by the milliwatt meter.

3. Turn off 2m TX, turn on 70cm TX (~1 Watt). Adjust C4, C5, C6 and C7 for P_{max} as indicated by the watt meter.*

4. Repeat steps 2. and 3. at higher TX output power level (approximately 10 Watt).

5. Install multiplexer in the car. Connect 2m Antenna ../4 or 5/8.. to antenna terminal. Connect car radio to terminal Ra. Connect 2m and 70cm terminals to respective transceivers. This should be done via SWR meters which may be removed upon finishing all alignments.

6. Adjust C1 and C2 for best SWR when transmitting on 2m.

7. Adjust C4 and C5 for best SWR when transmitting on 70cm. Should this result in an unsatisfactorily poor SWR then the length of the coaxial cable connecting the antenna to the "Ant" terminals should be reduced or increased by 5 to 15 cm.

Figure 249 Circuit diagram of antenna multiplexer.
Trimmer capacitors in [pF]
Inductors 1mm CuAg on 6mm diam. mandrel.

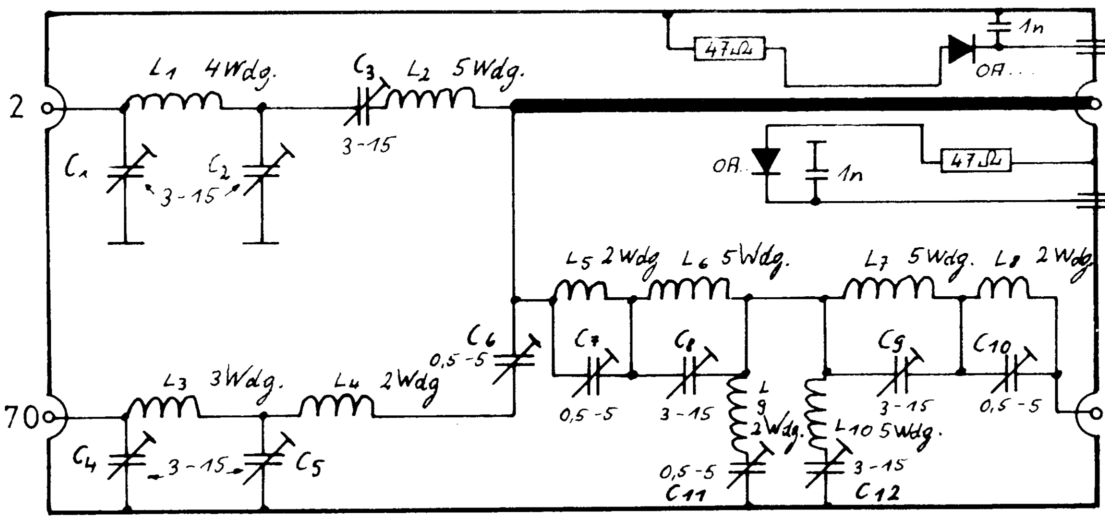


Figure 250

Printed circuit track suitable for a case measuring 146x72x30 mm

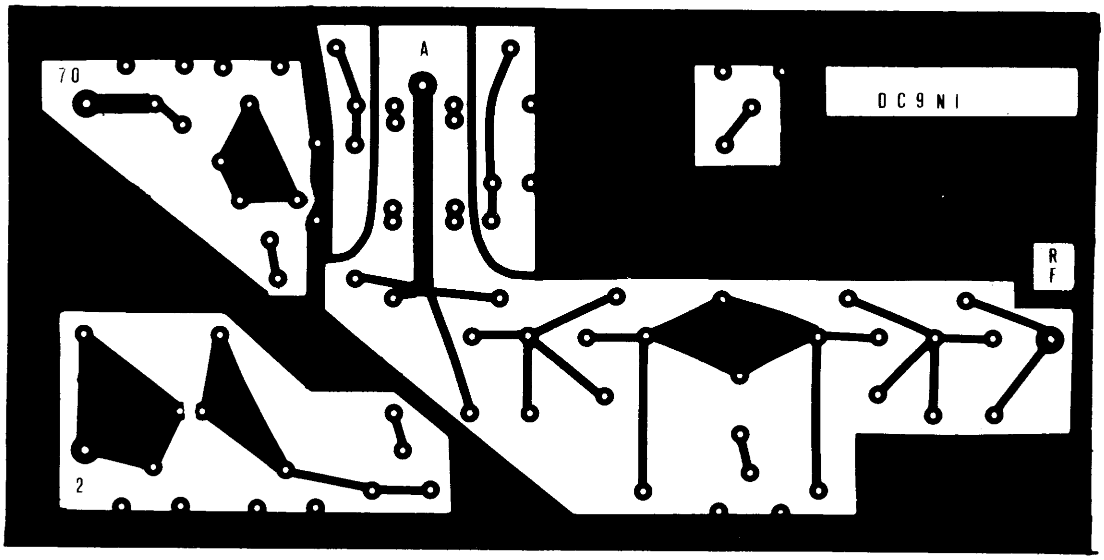
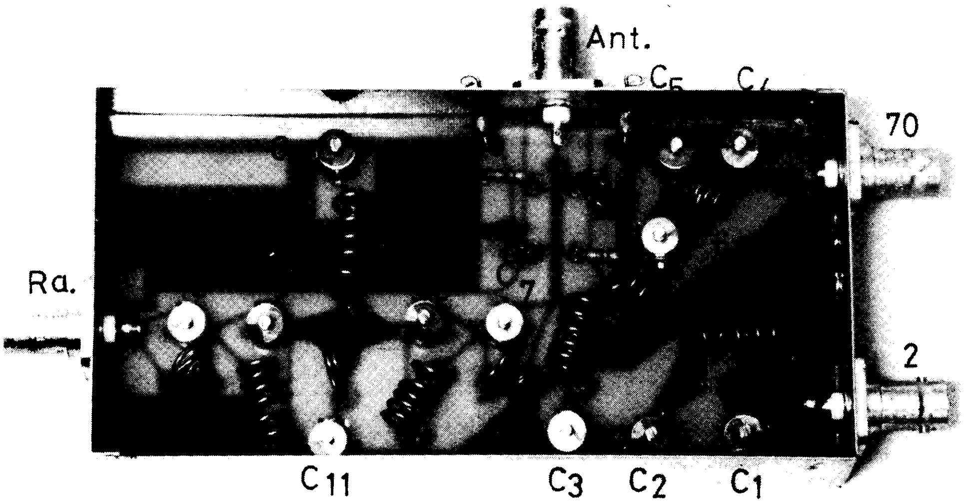


Figure 251

Component layout and identification.



D.1 UHF - Preamplifiers

The minimum signal detectable by a receiver is limited by the over-all noise power in a receiving system. This total noise is composed of predominantly three components:

- antenna noise (antenna temperature)
- noise contributed by the interconnecting cables (ohmic losses)
- noise generated by the receiver (essentially first stage).

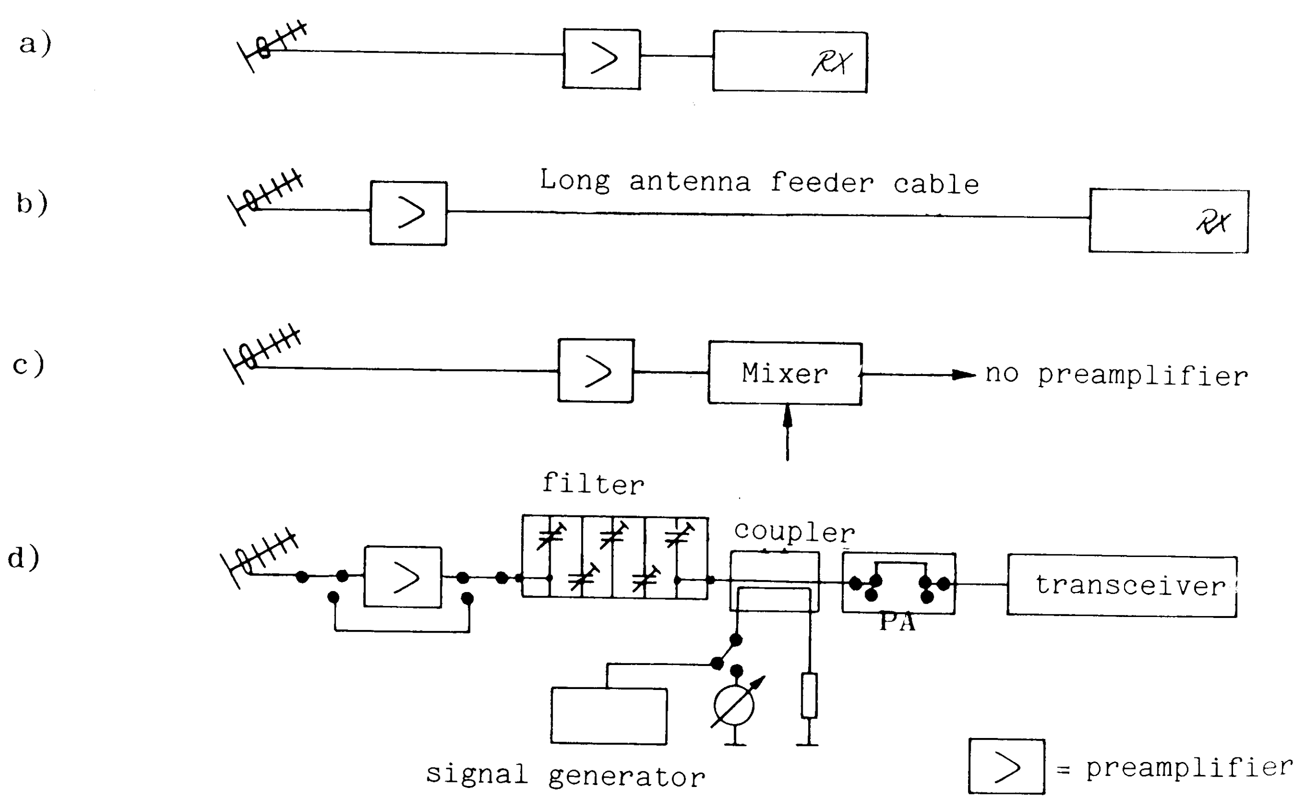
The amateur has little influence on the antenna noise. Cooling the antenna is too complicated and is usually limited to commercial application. The two remaining components may, however, be minimized through the use of high-performance low-noise preamplifiers.

Employment criteria, gain requirements, choice of transistors

Noise - respectively sensitivity - may not be improved indefinitely. Excessive gain will lead to increased de-sensitization (dynamic range, compression point) of the system (see A./4.2 - A.4.4). Less than optimum gain means throwing away the maximum possible sensitivity.
What then are the conditions making a preamplifier desirable? They are:

a. older receivers with poor sensitivity
b. long and lossy cables between antenna and receiver
c. insufficient gain preceding the mixer
d. additional attenuation due to relays, filters, directional couplers, distribution networks etc.

Figure 252



D.1.1 28 MHz IF-driver stages

Upon close examination of the station concepts according to D.7.1 and D.7.2 it is obvious that the different applications call for power splitters. In the above-mentioned examples the driver output power is split between two signal paths. One of them drives the transmit-mixer whereas the other one is used for monitoring and test purposes. This power splitting is tied to a power reduction of -6 dB in the mentioned examples. Should the driver output supply no more than 12 mW of total power then the drive power delivered to the transmit mixer will be insufficient. For that reason an additional stage should first raise the power level to approximately 100 mW. Each output of the power splitter according to B.1.3 will deliver a signal 6 dB down on the input which corresponds to 25 mW.
A very simple and reliable circuit is described below. A type P 8000 FET is well suited for this application. Preamplifier- type transistors may be used when employing this circuit in a receiving chain. The P 8000 happened to be in the scrap box. Thus, no other types were purchased and tried out.

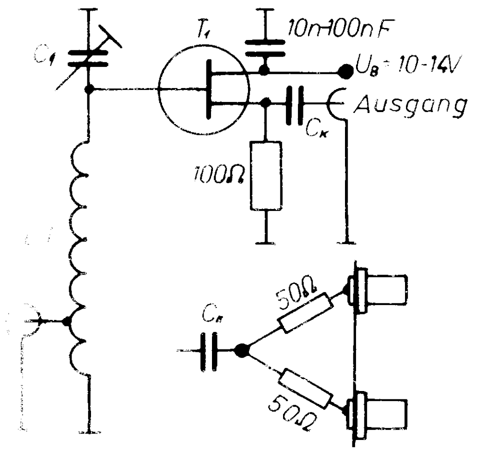


Figure 253
Circuit diagram of 28 MHz driver stage

L1 = 30 turns CuL 1 mm wound on 4 mm mandrel. Tapping point is 6 turns from cold end.

C1 = 4.5 to 70 pF foil trimmer

C_K = 680 pF to 1 nF

T_1 = FET P 8000

asured data for 100 mW output power

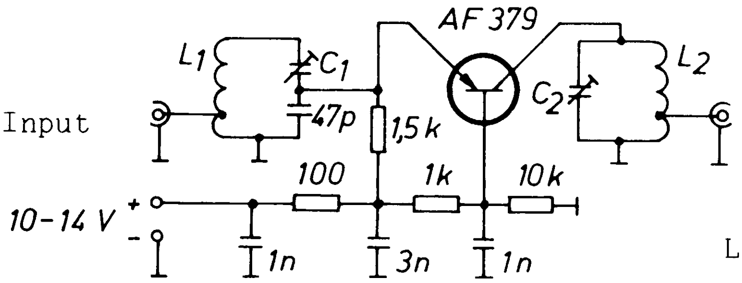
	$I_{quisc.}$	P_{in}	P_{out}	I_c	Gain
	40 mA	12 mW	100 mW	60 mA	9.2 dB

to 10 dB of gain will suffice for both the transmit and receive annels. Higher IF-gain in conjunction with strong input signals uld easily lead to overdriving the receive chain. When using a mixer XM D2T 10/70" according to D.4.1 in the transmit chain a power tput of 25 mW is quite sufficient. Employing this amplifier followed a power splitter will thus provide sufficient drive and at the same me deliver 25mW for monitoring purposes. The power splitter may be tegrated into the amplifier and the two power components could be uted to individual terminals. This would incidently allow two ters (70cm and 23cm) to be connected permanently to the driver. eration on two bands would be possible and band selection is [illegible]

D.1.2 2m preamplifier stages

Two circuits that were concieved many years ago are presented. One may replace the transistors against more modern types again and again since their pin arrangement usually remains unaltered. It is rarely necessary to change resistance values. These preamplifiers serve a good purpose when preceding ageing receivers of low sensitivity or as IF amplifiers following a 23cm converter. Instead of the 40673 any other dual-gate transistor suitable for 2m operation may be employed. The following figures show the details of preamplifiers utilizing either the AF 379 or the dual-gate transistor 40841. The BF 900 may be used as well; this does, however, result in a tendency for spurious oscillation. A type with lower transconductance is apparently quite satisfactory.

Figure 254
2m preamplifier employing AF 379 or AF 279/280

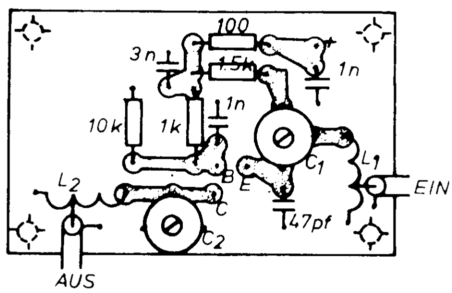


Output

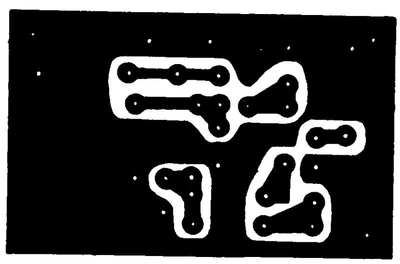
L1 6 turns 6mm diameter
1mm CuAg
tapped at 1 1/4 turns

L2 6 turns 6mm diameter
1mm CuAg
tapped at 1 1/4 turns

C1 / C2 foil trimmer
1.8 - 22 pF
Valvo or equivalent



Component location
(actual size)



Printed circuit
(actual size)

2m RF-preamplifier with dual-gate transistor 40673

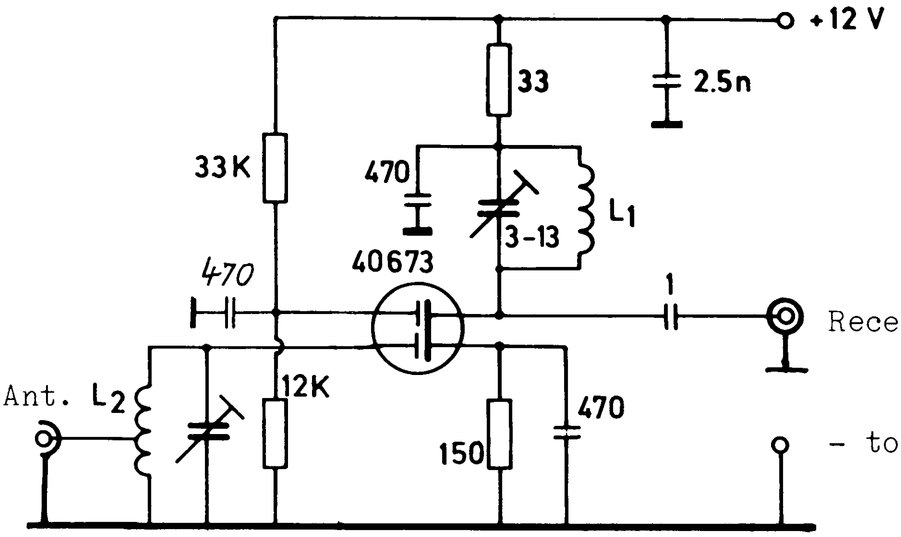
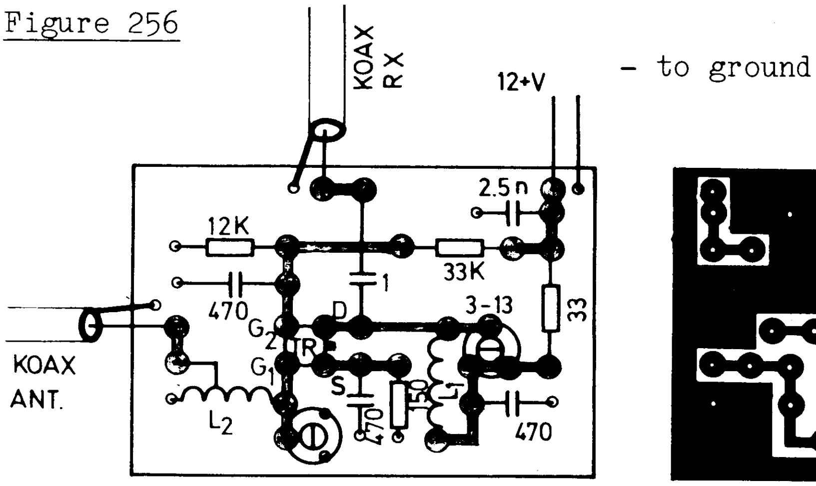


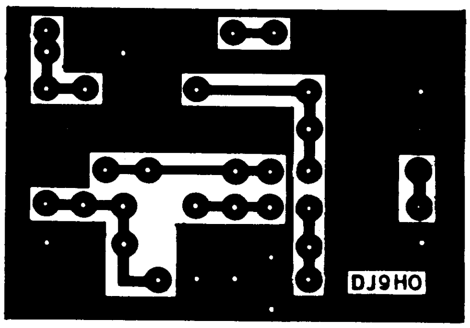
Figure 255

Circuit diagram

Figure 256



Component layout
(actual size)

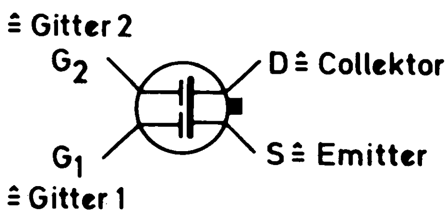


Printed circuit
(actual size)

Transistor 40673 or equiv.

L_1 6 turns
1mm CuAg on 6mm mandrel

L_2 6 turns
1mm CuAg on 6mm mandrel
tapped at 1 3/4 turns.



Dual-Gate-Transistor
(viewed from top)

The two photographs show the display of a spectrum analyzer. The 2m carrier has been wobbulated to aid recognition. Trace A depicts a 2m signal 6 and 5dB above noise level. Traces B represent the same signal after amplification by the respective preamplifiers. It may be seen that the preamplifier employing the AF 379 provides approximately 12dB (18dB minus 6dB) of gain. The preamplifier based on the type 40841 dual gate transistor provides approximately 11 dB of gain (16dB minus 5dB).

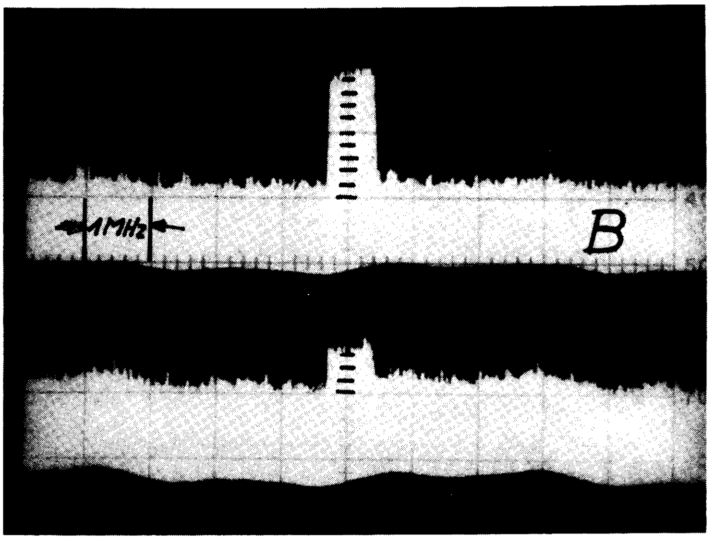


Figure 257

horizontal: 1MHz / div.
vertical: 10dB / div.

Signal behind the preamplifier with AF 379 (approx. 18 dB)

Signal without preamplifier (~ 6dB)

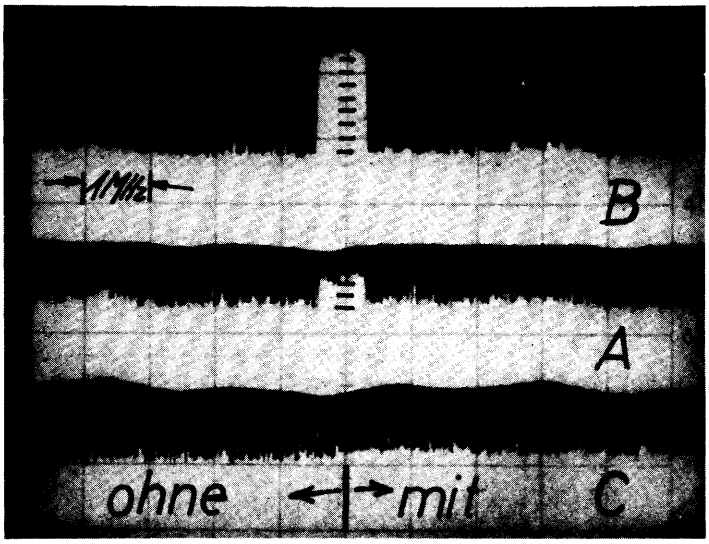


Figure 258

Signal behind the 40841

Signal without preamplifier

Noise (n_{rz}) of preamp. with AF 379 < 2dB

D.1.3 Preamplifier for 70cm employing BFR 34A/90/91

Building the single stage amplifier described below requires little effort. The r.f. signal is passed via a capacitor (~ 220pf) to the transistor base which is biased through resistor R_1 (50 to 160 Ohm was tried out). The lower value of this resistor will reduce - in particular when using the first two mentioned high-gain transistors - the risk of self-oscillation without seriously affecting the sensitivity. It furthermore represents some sort of termination for the coaxial input line over a wide range of frequencies.
It is possible to operate this preamplifier without it being preceeded by a filter - and this should be tried first. However, if there is a strong broadcast station in the vicinity or should one intend operation from an exposed "portable" location then preselection in the form of one of the multi- section filters according to C.3 should be provided. Preselectors and the converters described in this book will allow simultaneous operation on 2m and 70cm without risking desensitization of the 70cm receiver.
The preamplifier was tried out and its values measured when equipped with several transistor types. Quiescent currents lie in the range 3 to 10 mA. It is set to a value 1 mA higher than necessary for optimum gain. This setting promises stable operating conditions. In addition the module operates on a fixed supply voltage of 12 Volts as obtained from an IC stabilizer. The latter is a 78L12A (up to 30 mA at 12 Volts) which is commercially available at a very low price. Should this device fail then a series connected resistor of around 100 Ohm will protect the transistor against current overload and subsequent destruction.
The amplified r.f. signal is coupled to the following bandpass filter through a trapezoidal or disc-ceramic capacitor. The coupling capacitor (C_k) sets the bandwidth and in that context the gain. Values of between 0.5 and 2 pF are appropriate. The following figure shows the circuit diagram of the preamplifier. The case measures 72 x 72 x 30 mm and is fabricated from tin plate. For source of supply see section G.4.
Should one intend to stick more closely to the recommended operating conditions the supply voltage U_{CE} should be reduced even further (Zener diode and series resistor according to D.1.5) to the following values: BFR14A = 10 V; BFQ28 = 10 V; BFR34A = 6 V; BFR90 = 6 V; BFR91 = 8 V. Attention! Only in the case of the **BFR14A** is the beveled terminal the base connection!

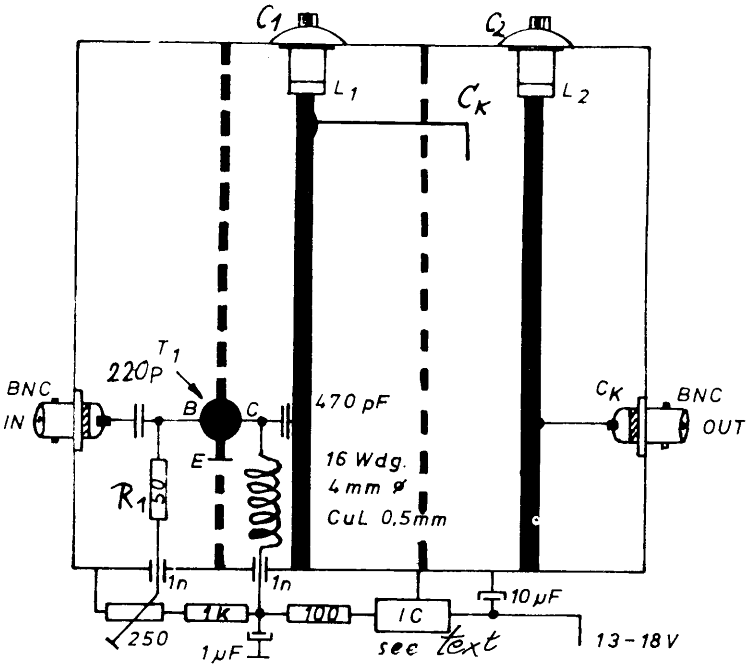


Figure 259
Circuit diagram of 70cm preamplifier

T_1: see text

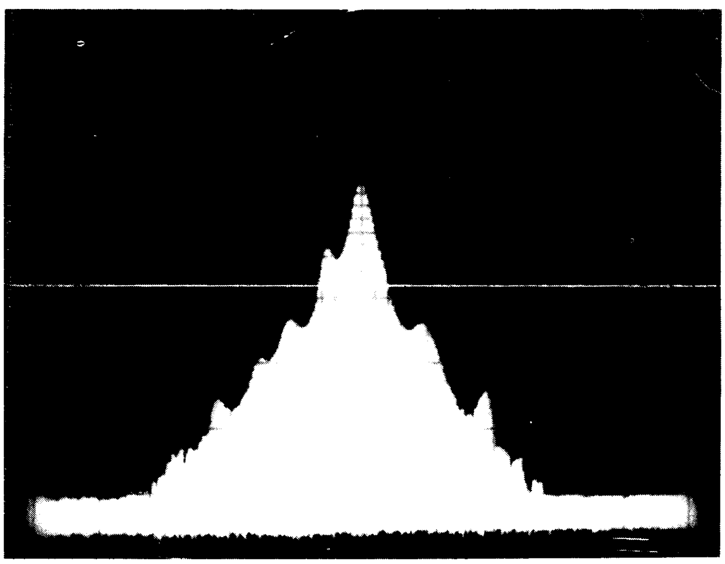
Voltage rating of electrolytic capacitors: approx. 30 Volt

C_1, C_2: Ceramic or glass trimmers ~ 1 to 10 pF

L_1, L_2: 2.5 to 6 mm diam. length: 55-60mm depending on C_1/C_2-dimensions

R_1: Low inductance composition or film resistor

To 70 cm waves solder is no longer a good conducter (see skin effect: A.1). To avoid lowering of the circuit quality factor and to improve the grounding conditions at the cold ends of L1 and L2 the brass lines are drilled, tapped (M4) and screwed down from outside by means of a brass bolt. The photograph shows a coil of 5 turns (wound over 4mm mandrel) installed as collector choke. C_k is a wire hook, soldered to L_1 or L_2, protruding through an opening in the partitioning wall and running parallel to the other line for about one centimeter. Slight variation of coupling (resp. gain) is achieved by bending this wire.



282 MHz 432 MHz 582 MHz

Figure 260
Passband with wire hook as C_k

h: 30 MHz/division
v: 10 dB /division

Bandwidth: (BFR34A)

- 3 dB	2 MHz
- 6 dB	6 MHz
- 10 dB	7.2 MHz
- 20 dB	16 MHz

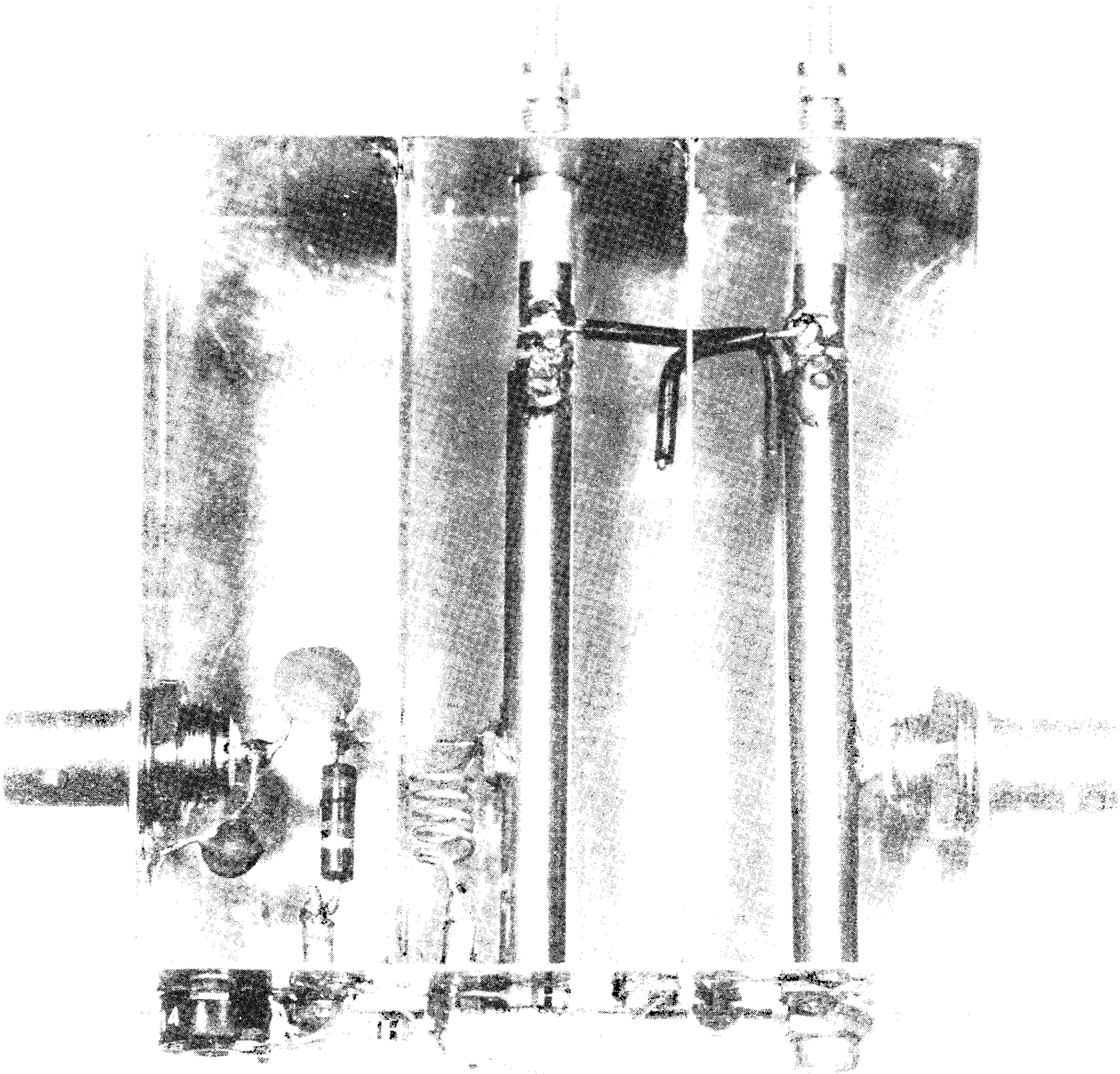
Transfer gain with transistors:

BFR 14 A = 21.8 dB
BFQ 28 = 20.7 dB
BFR 90 = 18.4 dB C_k=2pF
BFR 34 A = 17.3 dB
BFR 91 = 15.8 dB

Figure 261 Enlarged photograph of 70 cm preamplifier with bipolar transistors of types: BFR14A, BFQ28, BFR34A, BFR90, BFR91
Component specifications : See figure 259

The ceramic capacitor (input to ground) shown in the photograph is deleted. Powerful broadcast stations in the vicinity require selectivity in front of the preamplifier. Its transfer attenuation must be added to the noise factor of the amplifier and will always result in worsening the noise properties. An interim solution is the high-pass arrangement that already proved successful in the circuits of sections D.2.1 and D.2.2. If afterwards interference from out-of-band signals is still experienced then only a bandpass filter according to C.3 will solve the problem. All in all the circuit based on the BFR34A could be classed as a fair-priced no-problem arrangement.

Prior to first switching-on all components should have been installed and properly soldered (in particular stabilizers and electrolytics) since these transistors respond to voltage surges only once during their useful life...



D.1.3.1 70cm band GaAs-FET-amplifier with a 0.5dB noise factor

This circuit originates from Japan (JA1CZD). It had been reworked by DL7YC, was published in DUBUS-Info and may be regarded as a typical EME- preamplifier. Repeated measurements by OM Manfred (DL7YC) using automatic noise figure meters by both ATL and Hewlett Packard (HP) consistently revealed o.5 dB (± 0.1 dB measurement uncertainty). The preamplifier was also tested in connection with the EME antenna system of DL9KR and turned out to be 1 dB more sensitive than the one in use by DL9KR. At the time he was using a preamplifier based on the NEC244.

Suggested circuit: The whole arrangement is housed in a copper case to achieve favourable noise properties. All solder seams - except the partition - should be placed outside. The supply voltage is 5 Volt. Minimum noise was observed with a drain-source voltage of 2.5 Volt. The drain current is 10mA. Due to spread in component caracteristics it must be adjusted by means of R1 for the individual device used. R1 will thus have values of between 150 and 250 Ohm.

Figure 262 Circuit diagram according to JA1CZD and DL7YC.

The GaAs-FET is a type MGF-1400 by Mitsubishi.

Preceding and following the IC-stabilizer are 1.5uF electrolytics.

A diode guards the module against wrong polarity.

C_1 = 5pF, C_2/C_3 = 10pF

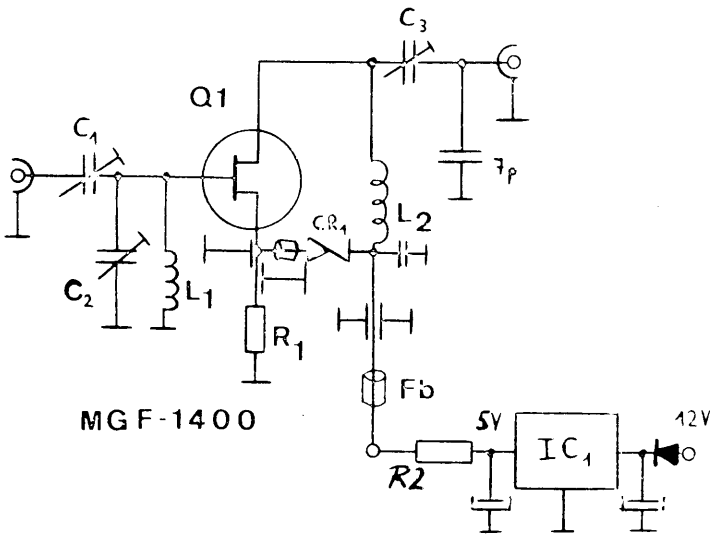


Figure 263
Gain and noise as function of capacitance C_1.

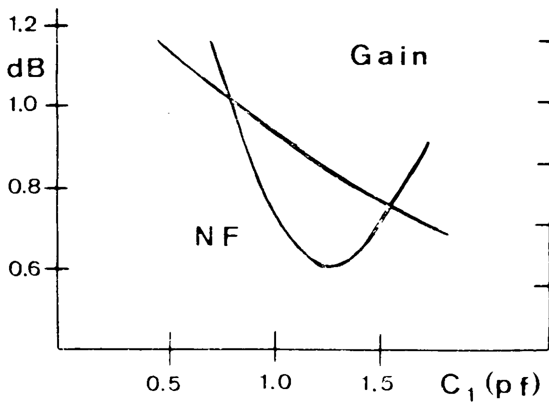


Figure 264
Gain and noise as function of I_D.

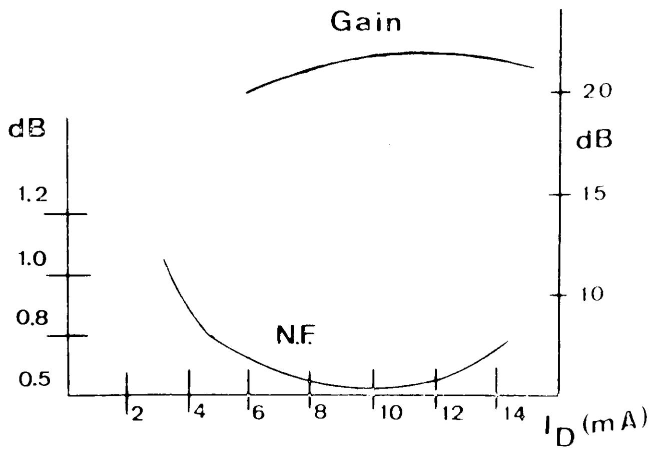
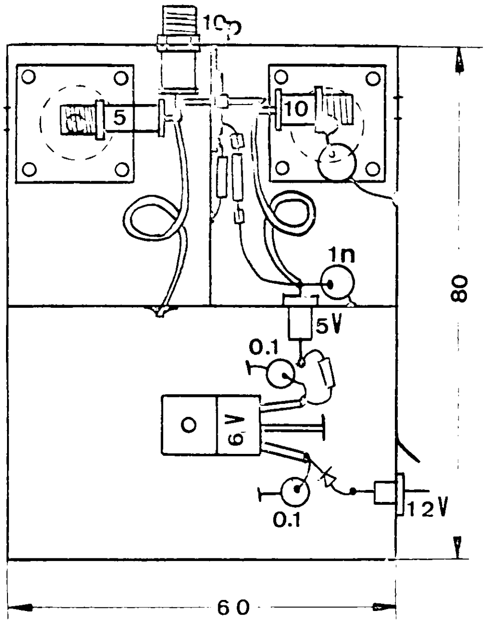


Figure 265
Construction and component lay-out as suggested by JA1CZD and DL7YC.
L_1 = 1 turn CuAg 1.2mm diam. wound on 5mm mandrel
L_2 = 1 turn CuAg 1.2mm diam. wound on 4mm mandrel shaped as shown in the photograph.

DL7YC



JA1CZD

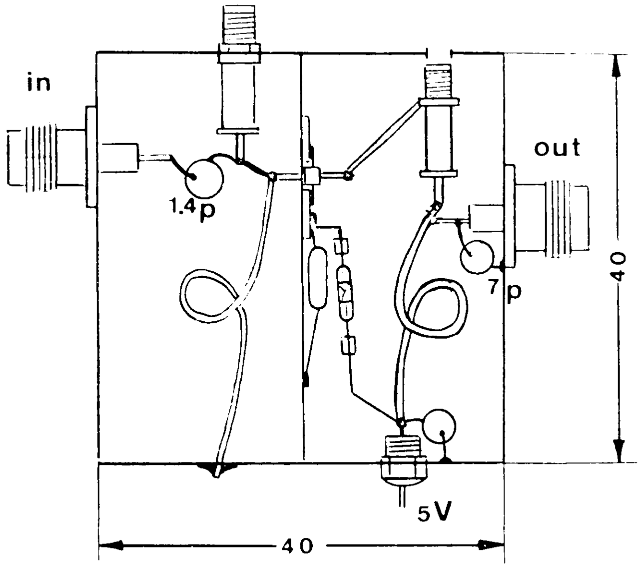


Figure 266
Photograph of preamplifier constructed according to DL7YC. The folded edge of the partition is drilled and tapped (M3) to increase the contact pressure of the cover plate. Both input and output (N-connectors) are mounted on the rear cover. C_1 and C_3 may be adjusted from outside through openings in the case.

Input Output

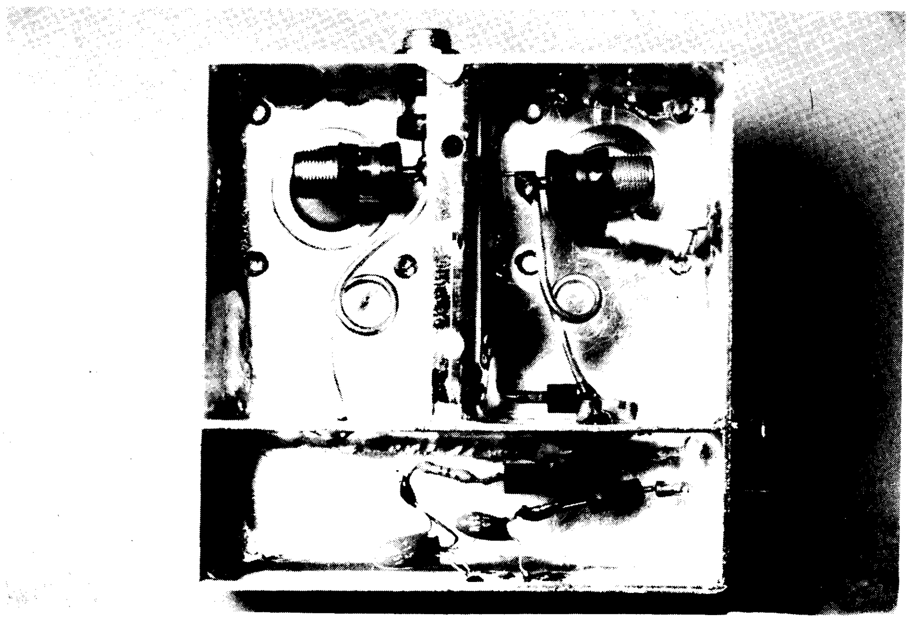
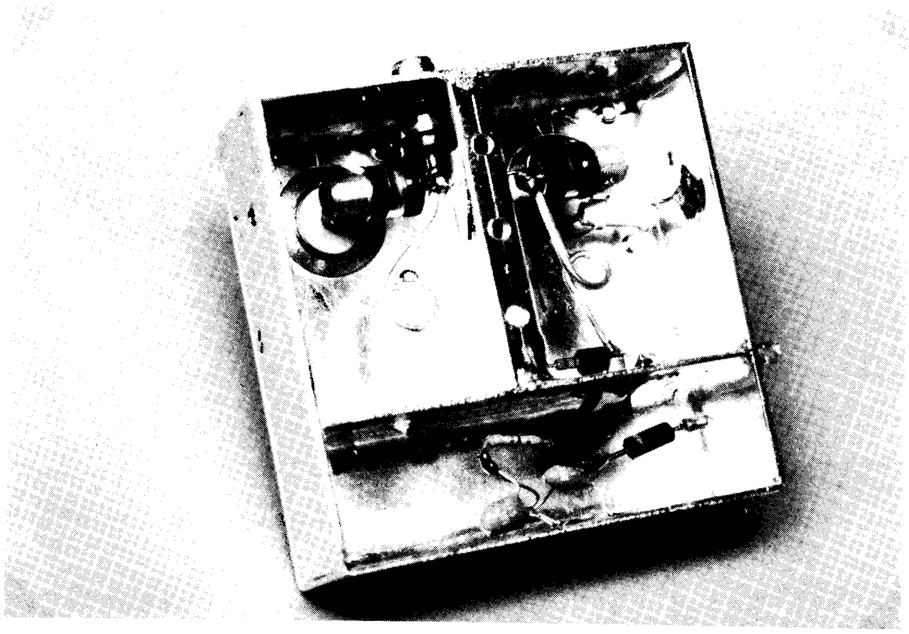


Figure 267
The input circuit is emphasized by this view of the module as constructed by DL7YC. The small openings through which the trimmers may be adjusted are clearly visible. Prior to final tuning all screws must be securely tightened. C_1 - C_3 should be high quality trimmers with glass dielectric (such as Johanson 5202).



Setting-up procedure for optimum operating conditions:
Initially R1 is set to 150 Ohm. To the left and to the right of the transistor mounting hole (4mm) both disc resp. trapezoid capacitors are soldered to the partitioning wall. Pull the mains plug of the soldering iron before soldering the transistor. Before touching any transistor terminal just touch the case briefly with the tip of the soldering iron to discharge any residual static electricity. The transistor may be held by means of pincers that are connected either directly to the case or via the hand holding the pincers by resting the hand on the case. The transistor will honour this treatment by staying alive....
R2 could initially be a 250 Ohm trimmer. Before connecting up R2 the correct functioning of the stabilizer should be verified. After installing R2 and upon application of the supply voltage 4 Volt should be measured at the cold end of L2 (R2 set to 100 Ohm). Next, the drain - source voltage is measured. A reading of 3 Volt should be obtained. The MGF-1400 exhibits optimum properties with a drain to source voltage of 3 Volt and with a drain current of 10mA.
With the particular transistor used by DJ9HO R1 turned out to be 160 Ohm and R2 was found to be 47 Ohm.

Figure 268
Sketch for the installation of te MGF 1400 to the partition wall.

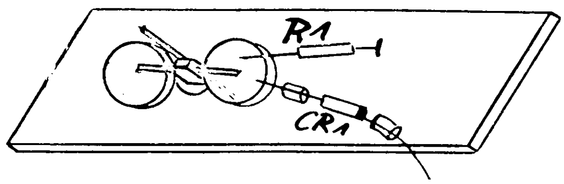
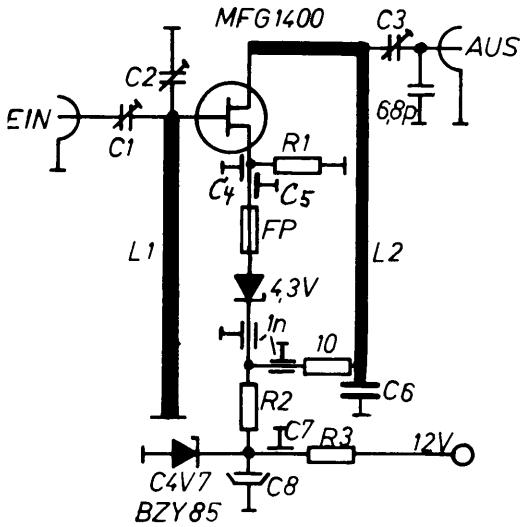


Figure 269
DJ9HO version of the preamplifier



C_1	=	0.5-6 pF
C_2	=	10 pF glass dielectric
C_3	=	2-22 pF
C_4-C_7	=	trapezoid.-C, 470 pF
C_8	=	15 uF, 35 Volt
R_1	=	160 Ohm
R_2	=	47 Ohm
R_3	=	330 Ohm
FP	=	Ferrite bead

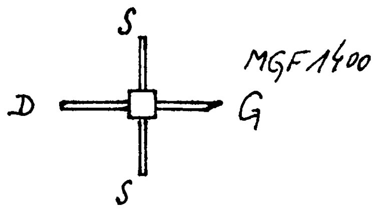
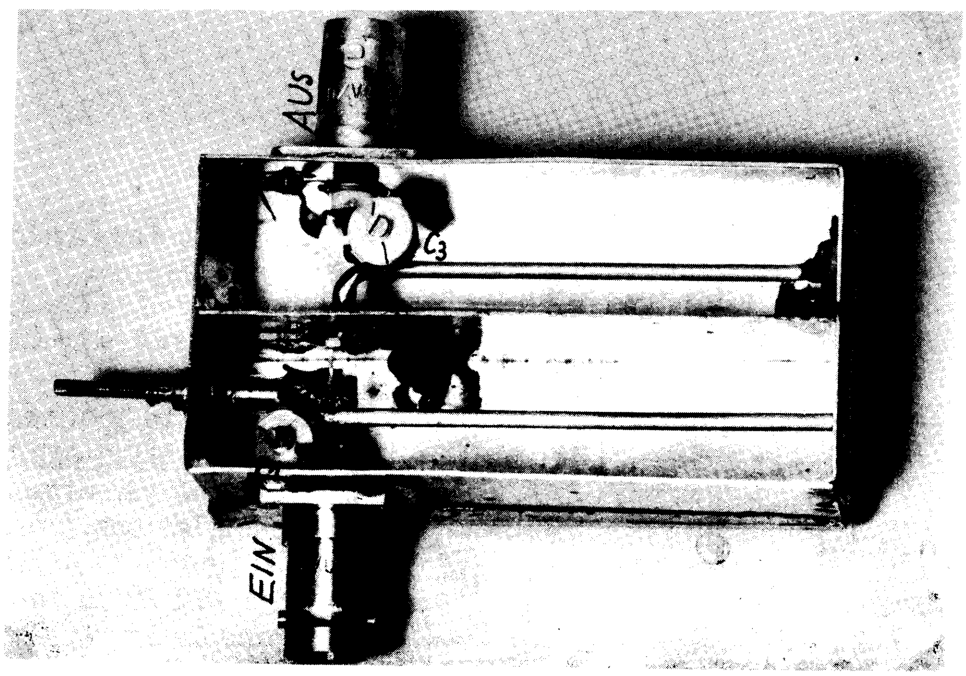


Figure 270
Assembled preamplifier in brass case 72 x 36 x 30 mm. All components from R2 up to the 12 Volt supply voltage terminal are mounted outside on the rear cover. At least C_1 and C_2 should have air or glass dielectric.



D.1.4.1 23 cm single-stage preamplifier

If required to construct a preamplifier in minimum time one will usually resort to the classical wiring technique. Since the coupling may be varied easily and since the losses may be lower when compared against the printed stripline technique this type of preamplifier is still popular. Stage gain depends on the type of transistor chosen. If one does not aim for the ultimate dB through optimum input matching the construction shown in the photograph will yield good results. The preamplifier is built into a commercially available case (tinplate, 72x36x30mm).

Figure 271 shows the preamplifier. The rear cover has been removed.

C1 is a 15pF fixed capacitor.

C7/C8 are disc-type or trapeze-type capacitors of approximately 70pF.

C6 = abt. 220pF

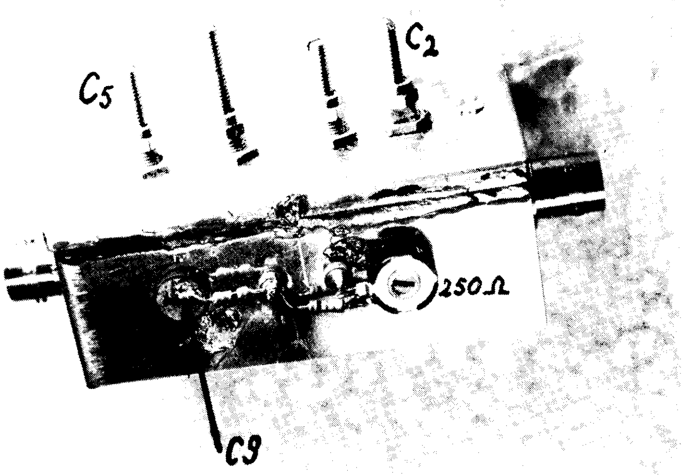


Figure 272 Side view of preamp. It may be necessary to add C9 to the stabilizer circuit which depends on the operating voltage of the chosen transistor.

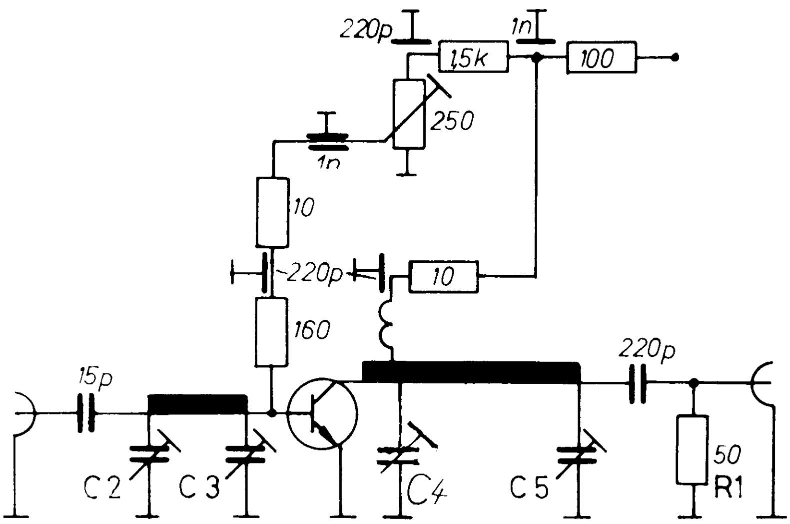


Figure 273 Circuit diagram of preamplifier.

C2-C5: 0.5 to ~ 6 pF.

R1: 50 to 160 Ohm. To be installed only in case of parasitic oscillations.

T1: BFQ 28 in this example

Stage gain: ~ 14 dB.

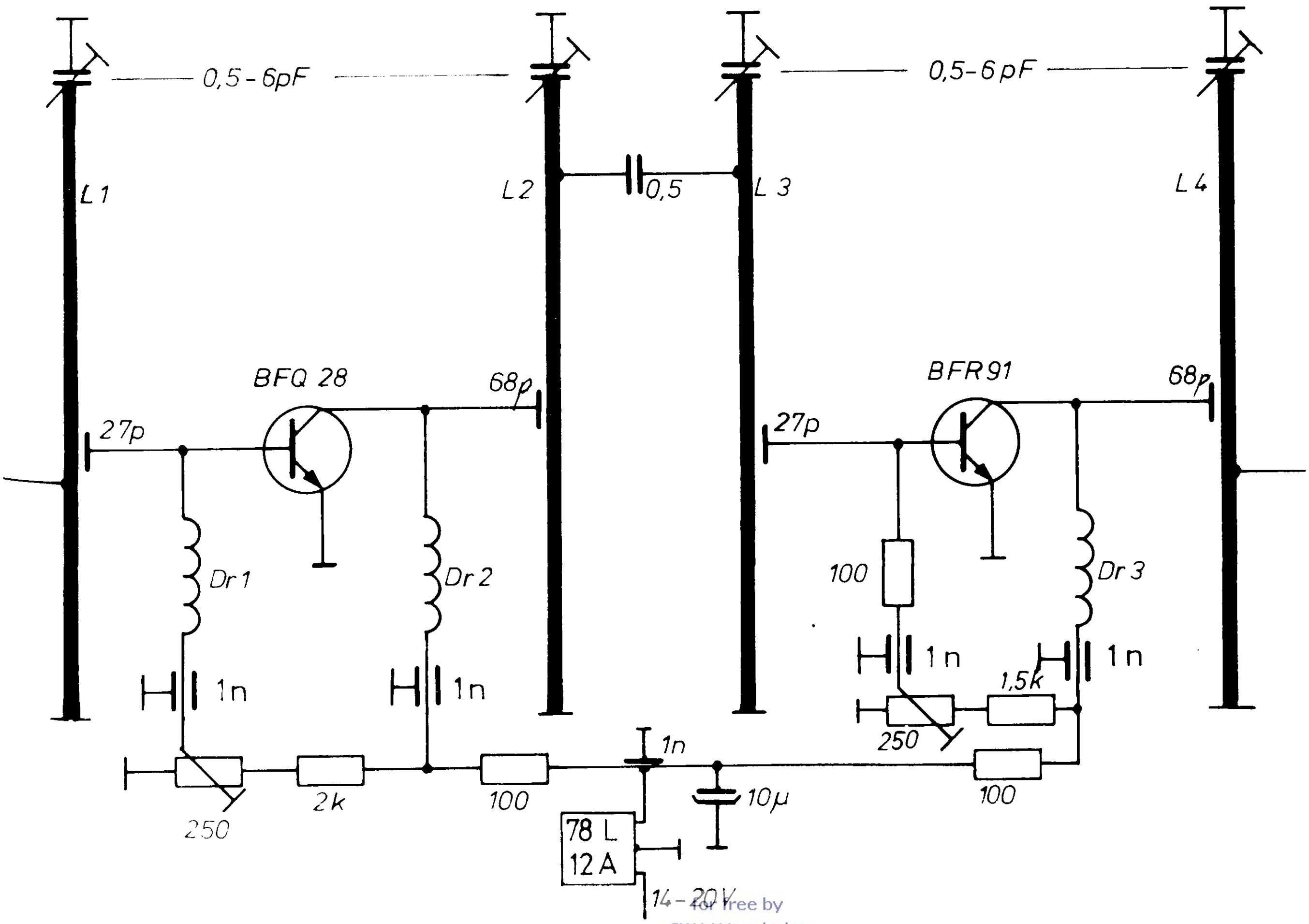
D.1.4.2 23 cm two-stage preamplifier

It has been explained in section D.1 that a mixer should be preceded by 20dB of preamplification. On 23cm this magnitude of single-stage gain can be provided by only one rather expensive transistor. The amateur is thus obliged to construct a two-stage arrangement. In that case a low priced transistor may at least be employed in the second stage.

Choice of transistor: The shown arrangement makes use of one each transistor BFQ28 (Siemens) and BFR91. The attainable gain is 20.8dB. It is also possible to decide for two type BFR34A transistors resulting in 14dB of gain.
As a rule the transistor having lower noise and higher gain should be chosen for the first stage as the latter is responsible for the overall noise properties and thus the sensitivity of the receiving system.

Voltage stabilization: Every amateur can testify from own experience that electrically stable operation and smooth amplification are highly dependent on stable operating voltages. Supply voltage fluctuations can lead to reduced gain. For that reason a voltage stabilizer (IC: 78L12A) has been provided. A stable operating voltage of 12 Volts is available for supply voltages of 14 to 20 Volts. Many types of transistors have their optimum noise properties at even lower voltages. In that case the IC may be replaced by a Zener diode (of for instance 9.1 or 6 Volts) and a suitable dropping resistor (see circuit in D.1.5). The series resistors of approximately 100 Ohm shown in the following circuit diagram must, however, be retained even after the Zener diode.

Figure 274 Circuit diagram of two-stage amplifier in conventional wiring technique.



Construction of the preamplifier:

The preamplifier is built into a TEKO case measuring 80x48x25mm. L1-L4 are made from silver plated copper wire of 1.5mm diameter. The length of these inductors depends on the dimensions of the trimmer capacitors and is 27mm in this particular case. Input and output are coupled to taps 8mm off the cold ends. Coupling capacitors (input 27pF, output 68pF) are of the trapezoid or disc-ceramic type. Chokes Dr1/Dr2 are made up from 6mm of 0.5mm diam. copper enamel wire wound on a mandrel of 2 to 3mm diameter.

Two partitioning walls are solder-mounted inside the case. Unfortunately the openings in these walls are too large. Upon installation of the two transistors these openings must be closed up such that only holes of 3mm diameter remain (passages for the transistors).

Tuning is provided through ceramic trimmers of 0.5 to 12pF that happened to be around. Trimmers of 0.5 to 6pF are quite sufficient. Should - due to different trimmer dimensions (trimmers fully extended at f_{res} = 1296MHz) - the inductors be found to be too long they may be produced from 4mm material. This will reduce their inductance and tuning for resonance is once again possible.

Again the trough line inductors are inserted into tight fitting holes and soldered from outside. The least possible quantity of solder should be used inside the trough.

Coupling from L2 to L3 is provided by a 0.5pF capacitor. In this case the partitioning wall was omitted. With extensive test facilities at one's disposal coupling could be arranged to match the chosen transistors and the desired passband. The d.c. network shown in the circuit diagram is mounted outside the case. Suppression of surge voltages as a result of man-made short circuits is performed by the 10uF electrolytic. Its value (10 to 100uF) is not critical. Should a transistor pass away into eternity the capacity of this condensor should be checked. Depending on the types the first transistor should be adjusted for 3 to 5mA and the second for 5 to 10mA. Prior to first switching-on the 250 Ohm potentiometers must be set to ground.

After applying the operating voltage and careful adjustment of the quiescent currents tuning for optimum signal properties may be carried out. The cover should be installed before attempting fine adjustment. This amplifier may be used both in receiving and transmitting chains (following the mixer).

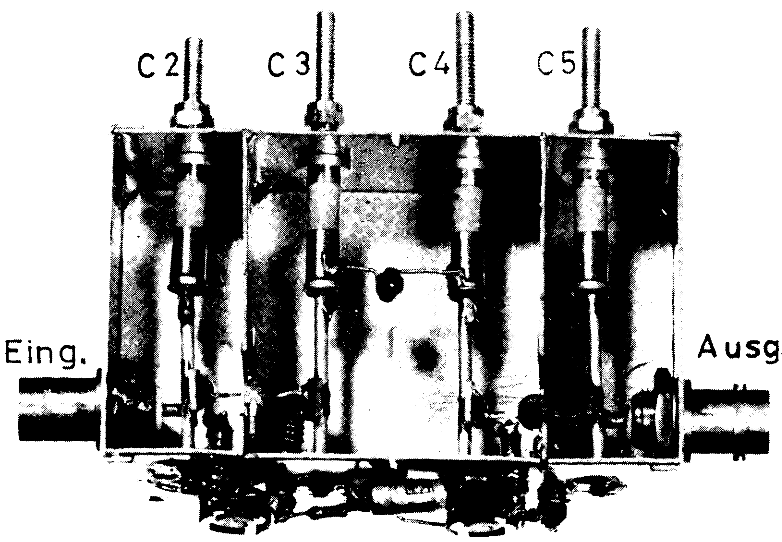


Figure 275

Construction of two-stage preamplifier for the 23cm band. No dimensions are given since the photograph is approximately actual size. Depending on the setting-up a noise figure of 5 to 8dB is to be expected.

D.1.5 Low noise preamplifier for the 23cm Meteosat band
(Josef Grimm, DJ6PI)

Low noise preamplifier circuits have been developed to bring the noise figure of input stages in 23cm receiving systems down to 2 to 3dB. These amplifiers are built in stripline technique on glass fibre reinforced epoxy boards.
There is little doubt that multi-compartment coaxial construction or the use of PTFE (Teflon) and RT-Duroid p.c.board will not lead to even better performance. This advantage was given up intentionally since glass fibre reinforced epoxy board is easily available and reasonably priced. In view of the low effort and cost the slight reduction in performance seems acceptable.

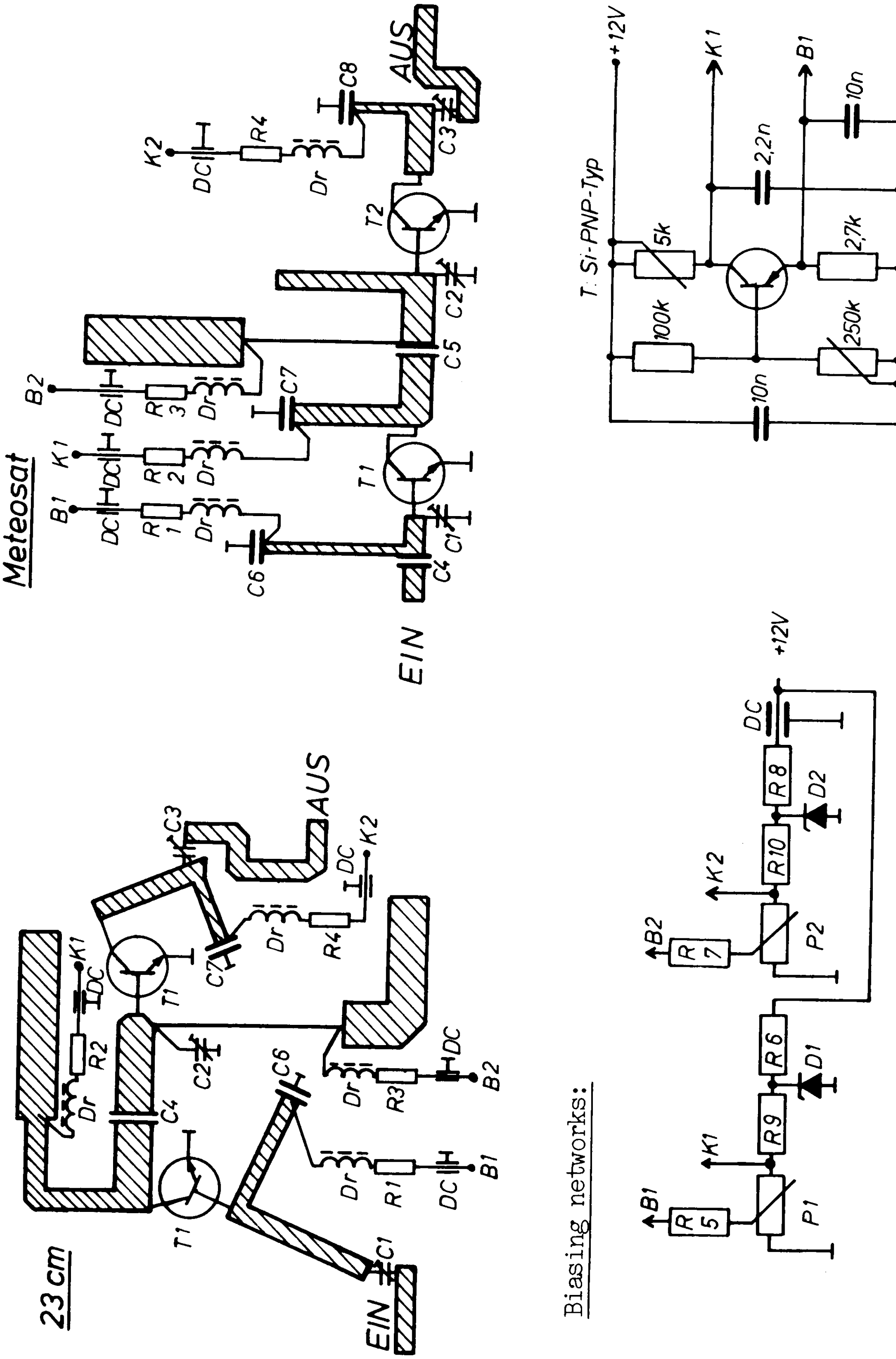
As described in (1) the amplifiers are of the two-stage design to ensure sufficient gain in front of the mixer.
The first stage is optimized for lowest possible noise figure and makes use of the rather expensive NE transistor NE 645 35. The second stage is designed for maximum gain and employs the cheaper type NE 578 35 transistor.

Because of the high dielectric constant of glass reinforced epoxy ($_r$ = 4.8) and shorter Smith-chart-derived transformation sections the amplifier dimensions could be reduced significantly.
The matching networks (in conjunction with printed transformation lines, inductances and capacitances) ensure noise - respectively power - matching to the transistors. Three ceramic tubular trimmers are provided to allow compensation of component and line tolerances.
Due to the low quality factor of the printed tuned circuits the tuning peaks are rather broad.

D.1.5.2 Construction of the stages

a) Drilling and sawing
Holes of 3.2mm diameter for the feedthrough capacitors are drilled into the print boards at the positions marked by soldering rings. Using a jig saw, slots are cut along those short edges of the emitter contact surfaces that face one another. Again using a jig saw somewhat broader slots are cut into the board for the installation of the capacitors C6, C7 (and C8 in the case of the Meteosat band).
The trimmer capacitors C1 (Meteosat only) and C2 should have the shortest possible connections to ground. Therefore narrow slots are cut into the boards in the positions marked by lines.
High quality N-type connectors should be used to avoid ruining preamplifier performance through low-grade connections. Holes to suit the shroud flanges are cut into the board at both input and output.

Figure 276
Circuit diagrams of two different amplifier versions and biasing networks.

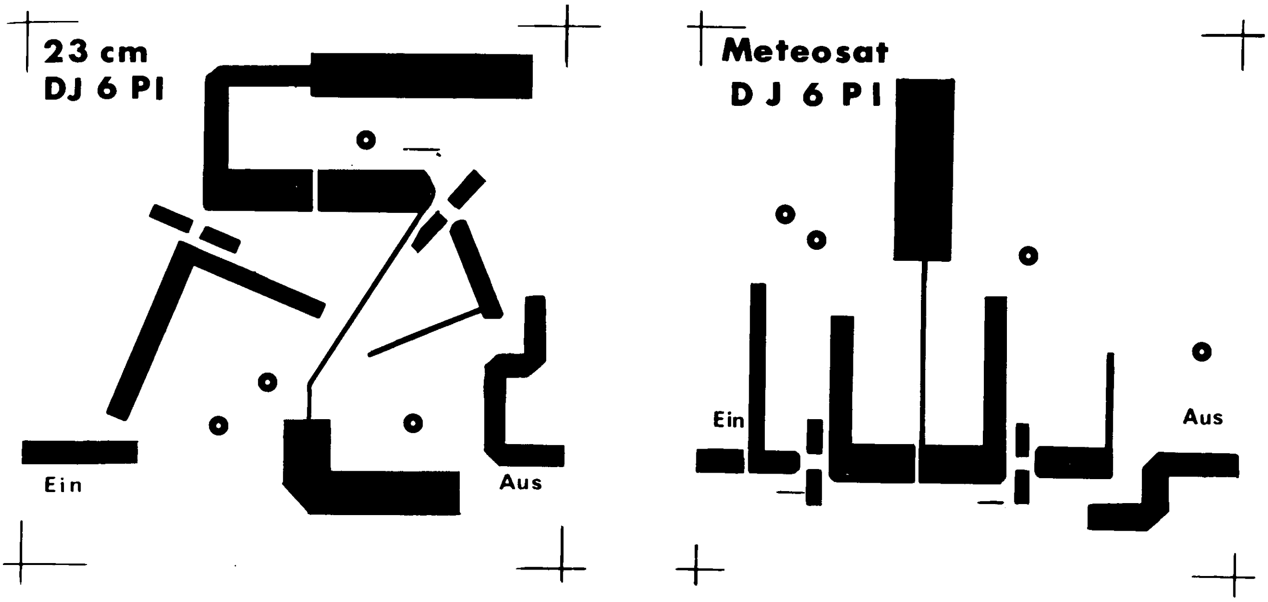


Simple ciricuit for both amplifier versions

More elaborate circuit for first stage.
5K potentiometer for U_{CE} adjustment
250K potentiometer for I_C adjustment.

Figure 277 Etching template for p.c.board of 23cm and Meteosat band preamplifier. Use standard glassfibre reinforced epoxy board of 1.6mm thickness, both sides copper clad.

Full-scale drawing



Components:

T1 = NE 645 35 (marking Kz)

T2 = NE 578 35 (marking Cg)

D1/D2 = Z-diode ZPD 9.1

Dr = Ferrite bead threaded on resistor lead

C1-C3 = 0.5-3 pF miniatur tubular trimmer for p.c.board mounting

C4-C5 = trapezoidal or disc capacitor 30..100pF (skip C5 in 23cm version)

C6-C8 = trapezoidal capacitor 470..1000pF (skip C8 in 23cm version)

DC = Feed-through capacitor 470...1000pF

R1,R2,R3,R4 = 10 Ohm

R5 = 18 K

R6 = 220 Ohm

R7 = 12 K

R8,R9 = 150 Ohm

R10 = 100 Ohm

P1: 25K potentiometer

P2: 10K potentiometer

Input and output: N-type (or BNC) connector, flange version

Case: Schubert tin plate no. 5 (72x72x30mm)

All non-standard components are available through SSB- Electronic or Verlag UKW-Berichte, Baiersdorf.

Figure 278
Enlarged connecting diagram of transistors.

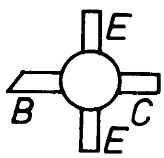


Figure 279

Construction of 23cm preamplifier (first version). There are slight deviations from the final circuit and template in the output stage due to a small modification after taking the photograph.

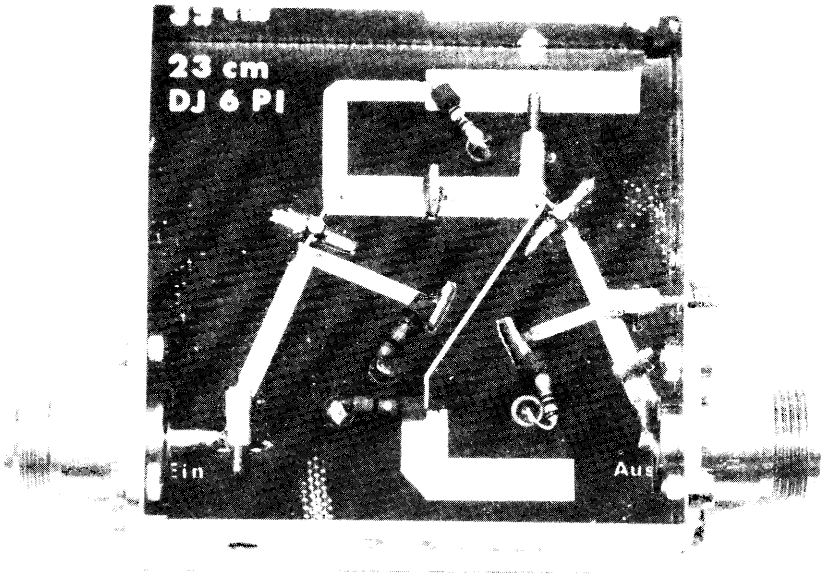


Figure 280

Rear side of the above mentioned amplifier. The photograph clearly shows the dropping resistors, Zener diode and trim potentiometers.

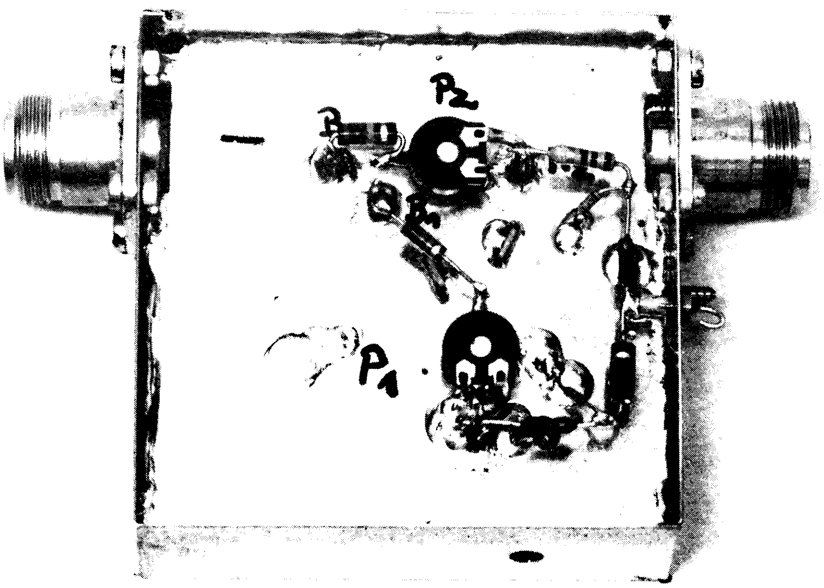


Figure 281
Construction of the Meteosat-version for 1700 MHz. Once again there is a slight deviation in the output coupling that could not be incorporated in the photograph in time.

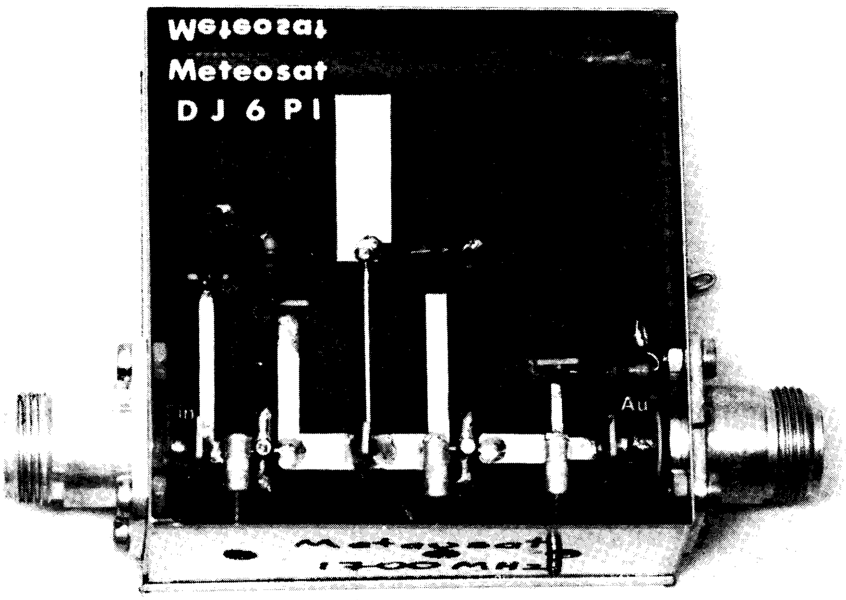
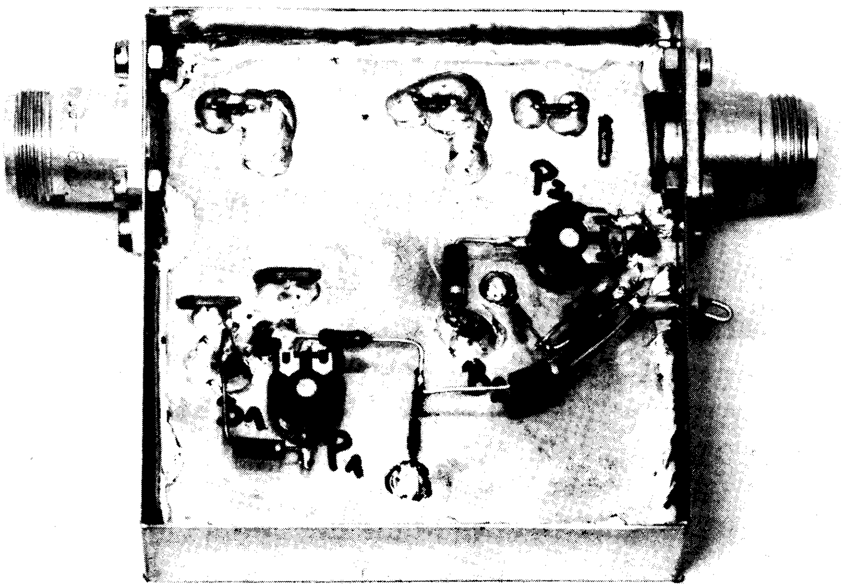


Figure 282
Rear side of the above mentioned amplifier. Again the biasing network for the two transistors is clearly visible.



b) Soldering tasks:
The trapezoidal isolating capacitors C4, C5 (C5 only in Meteosat version) are solder-mounted standing up at the interuptions of the striplines. The leads of the blocking capacitors C6, C7 (C8 only in Meteosat version) are inserted into the slots. One foil side is solder-joined to the stripline and the other foil side to the ground foil. The foil at the stripline must not make contact with the ground foil!! The four feed-through capacitors are inserted into the respective holes and the capacitance coating is solder-joined to the ground foil. Ferrite beads are threaded over one each of the resistor leads R1-R4, the leads are clipped next to the beads and soldered to the following points:
1) Capacitance coating of C6, C7 (and C8 for Meteosat) facing the stripline.
2) Intersection of the narrow and the wide choke line (in 23cm version: Collector supply of T1 and base supply of T2; in Meteosat version: Base supply of T2).
The remaining leads of the resistors are soldered to the top terminals of the feed-through capacitors. 2mm wide strips of copper foil are pushed through the four slots at the ends of the emitter contact surfaces and are soldered to both the ground foil and the emitter surface. Afterwards the transistors are soldered to the striplines and emitter surfaces - with particular attention to the correct pin sequence and shortest possible soldering time.
Next, the ground leads of C1 (Meteosat only) and C2 are threaded through the p.c.board slots and soldered to the ground foil. The ground leads of C1 (23cm version only) and C3 are bent and soldered to the input stripline.

c) Installation inside case: Holes for input/output connectors and for the trim capacitors are drilled into the case frame. The spigots of these connectors are clipped such that they will fit the stripline ends. A hole (3.2mm diam.) to accept the feed-through capacitor for the common voltage supply is drilled into the frame in the vicinity of the output connector such that it will face the ground surface of the p.c.board. Next the r.f.connectors are screwed to the frame, the spigots are soldered to the stripline ends and the ground surface is solder-joined to the frame. The components of the biasing network are installed and wired inside the cavity facing the ground surface.

D.1.5.3 Adjusting the preamplifiers

The same circuit as in (1) is used for setting and stabilizing the working point. Prior to first switching-on the wipers of both potentiometers P1 and P2 are turned towards the ground stops. A mA-meter is inserted into the connection between the d.c. biasing network and the feed-through capacitor K1. By careful adjustment of P1 a current of I_C = 7mA is set. The meter is then replaced by a jumper. The same procedure is repeated for K2. In this case P2 is used to set I_C to 10mA.
Utilizing the signal supplied by a signal generator, a fixed frequency generator according to B.8 or a QSO the tuning capacitors are adjusted for optimum results by means of a non-metallic screwdriver. C1 must be set to give best listening results or minimum noise by means of a noise figure test set - if available.

D.1.5.4 Noise figure optimization

Should the reader possess a noise figure test set or should he at least have access to such a set then other measures apart from adjusting C1 could lead to further noise figure improvements:
Data sheets specify the operating point of the transistor for lowest noise figure (for NE 645 35: U_{CE} = 8Volts, I_C = 7mA).
Individual samples may deviate from this at different frequencies. The biasing network according to (1) will allow merely I_C to be varied; U_{CE} will change accordingly. A network for reasonably independent control of I_C and U_{CE} is published in (2).
It is quite possible that minimum noise figure will occur at for instance I_C = 5mA and U_{CE} = 6.2 Volt. Amateurs having access to a noise figure test set are therefore adviced to duplicate the more ambitious stabilizing circuit for the first stage.

D.1.5.5 Measured data and supplementary equipment

As a result of component spread and p.c. board tolerances a noise figure of 1.8 to 3dB and a gain of 20 to 25dB is to be expected. The image frequency noise contribution must be reduced by means of selective filters (see C.4) in the case of wide band mixers. A suitable filter must thus be inserted between preamplifier and mixer. Printed stripline filters lead to high insertion loss which - when employed in front of the preamplifier - has to be added to the noise figure. If such a filter is added after the preamplifier the overall gain has to be reduced by that loss figure. For that reason only coaxial or interdigital filters should be selected (see C.4).

Notice: The printed circuit may be copied in quantity if the call sign is included.
Should the reader find it impossible to construct this preamplifier he is invited to contact the originator DJ6PI.

References:

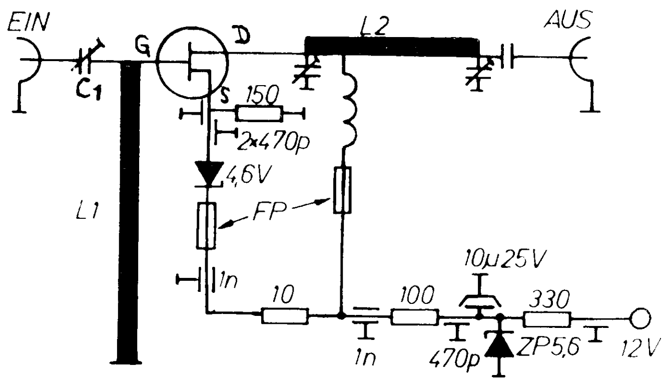
(1) Grimm J.
Twin stage low noise preamplifiers for the 24cm to 12cm bands.
(UKW-Berichte vol. 3/1979 pages 130 - 141)

(2) N.N.
A low noise 4 GHz transistor amplifier using the HXTR - 6101 slicon bipolar transistor.
(Hewlett Packard Application Note 967)

D.1.6 GaAs-FET preamplifier for the 23cm band (G=19.8dB, NF=1dB) using the N-channel Schottky-barrier gate low noise GaAS-FET MGF 1400

Having achieved good performance figures with the GaAs-FET MGF 1400 by Mitsubishi on 70cm we felt compelled to try it on 23cm as well. Expectations were raised even further by the fact that the makers claim a maximum oscillator frequency of around 50 GHz for this type! Since we intended to include the circuit in this book which at the time was about to go to the printers we had no time to waste. OM Manfred (DL7YC) managed to aquire five of these transistors. OM Josef (DJ6PI) scrounged and passed on the data sheets as well as giving us some advice. These specifications indicate a gain of 15 to 9dB over the frequency range of 4 to 10GHz with noise figures of 2dB (4GHz) and 3.5dB (10GHz) respectively. This hints towards applications up to 10GHz.

Circuit description: A single-stage arrangement was chosen since there is plenty of gain available. The d.c. biasing network was taken from the proven circuit in section D.1.3.1. Input matching is achieved through inductance L1 in conjunction with the coupling trimmer which is soldered to a point roughly 10mm below the hot end of L1. We experimented with a foil-type trimmer 0.5-6pF but it is strongly recommended to choose a trimmer with glass (or comparable) dielectric (0.1-3pF). Output matching is achieved by a half-wavelength line. At this point the quality factor is not quite so important and trimmer capacitors with ceramic insulation (0.5-6pF) were selected. They double as soldering posts for the inductor L2. Coupling out was initially provided through a variable capacitor (3.5-70pF) which was finally replaced by a fixed capacitor of 40pF.



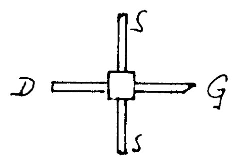


Figure 283
Circuit diagram of the 23cm preamplifier

For optimum adjustment a drain-source voltage of 3 Volts and a current of 10 to 15 mA should be set.

Maximum gain occurs at a V_{DS} of 3 to 4 Volts.

A 4.3 or 4.6 Volt Zener diode is acceptable for stabilization if the 100 Ohm dropping resistor is reduced to approximately 50 Ohm.

FP = single-hole ferrite bead.

Figure 284 23cm GaAs preamplifier with particular emphasis on the input arrangement. A trimmer with glass dielectric should be chosen instead of the 6pF foil trimmer shown. Should no such device be available then a trapezoidal capacitor of 22pF should be installed between input and gate. This will reduce the gain by about 1.5dB without affecting the noise properties.

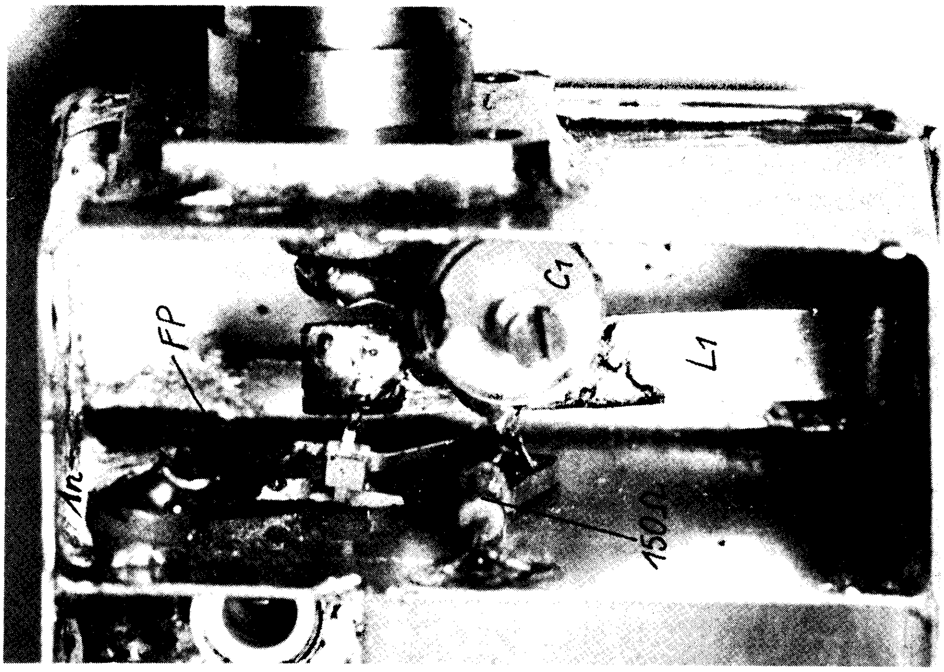
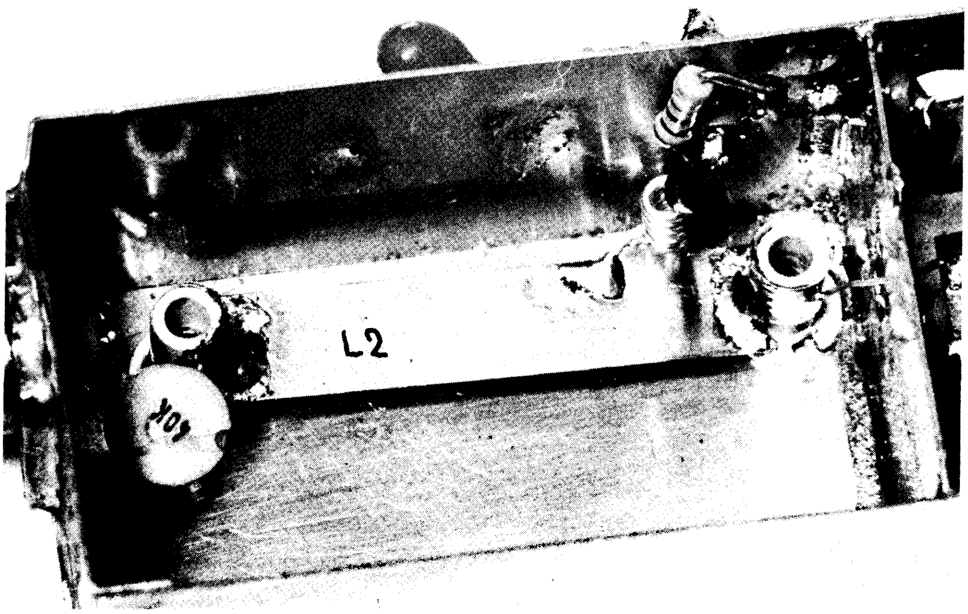


Figure 285 Half-wave linear circuit of preamplifier. The spacing between L2 and ground surface is approximately 2 to 3mm. The trim capacitors are set to about mid-range. L1 is made approximately 5mm longer and its lower end is angled-off by that amount. This end is tinned, pressed into the correct position and soldered to the inside of the case by placing the soldering iron against the outside of the case. The inside contact area may be tinned beforehand - using the smallest possible amount of solder.



D.2 Converter circuits for UHF (General remarks)

It is common practise to employ converters at UHF. It is thus possible to utilize the fairly expensive but technically mature HF and VHF equipment on UHF, too. As apparent throughout section D the modular design is the preferred technique. Whoever has had a chance to align a converter by means of a spectrum analyzer and simultaneously observe the spurious and harmonic products of the oscillator with respect to the power of the wanted signal will confirm the necessity of this approach. Injection signal generation should thus never be performed without placing bandpass filters between the various stages of the injection chain. It is a great advantage to install the stages in individual compartments, separate boxes or to spread them out and use selective circuits to filter out the wanted signals. Placing the stages too close to one another could easily result in crosscoupling even between the first and the last stage. It is quite impossible to realize all the stages of the injection circuit plus the mixer stage on one board without the compartment-type of construction. Under these circumstances all of the multiplied oscillator signals will mix with all input signals and all mixer products. One should thus not be surprised to hear broadcast and t.v. transmissions inside the amateur bands. To save the reader money and effort a number of proven rules governing the construction of receivers and converters follow.

1. Only the injection frequency should "see" the mixer.
2. Install injection signal generator in separate case.
3. Thorough screening between r.f. and i.f..
4. Stabilization of each modules operating values.
5. In general never place more than two amplifying stages in one case to avoid spurious oscillation.
6. Duplicate only such circuits that have been evaluated sufficiently.

The choice of i.f. depends greatly on the available or expected receiver following the converter. An i.f. outside the amateur bands would be beneficial to prevent strong stations from radiating directly into the circuit behind the converter head - thus suggesting activity inside the selected band. The choice of such an i.f. could demand more elaborate equipment and one is usually stuck with whatever happens to be available. A 10m receiver turns out to be the best i.f. strip for 70cm converters as any experienced amateur will testify. The 2m band is thus spared altogether and undisturbed multi-band operation becomes a possibility (during contests etc.). The geographical situation is the governing factor for the 23cm amateur. Again 10m seems the best choice in densely populated areas or exposed locations; furthermore the 2m rig will remain unoccupied and is free for parallel operation as is occasionally necessary. Should SSB-CW-portable stations be available (such as IC202, IC402 etc.) then the 2m set will be suitable for 23cm and the 70cm set will serve all higher bands for both receiving and transmitting purposes (See also D.7).

D.2.1 Simple high-performance 70cm converter
(two transistors AF279S, i.f. 28-30 MHz)

The suggested circuit shows that a newcomer can get started on 70cm without too much effort. If required this concept could be extended by an external preamplifier (D.1.3) and - depending on geographical conditions - modules to improve selectivity (C.3.3). This would result in a highly selective and sensitive receiving system.

Preamplifier and mixer are equipped with low-priced Germanium PNP transistors type AF279S (similar: AF239S, 279, 280, 379 etc.). The mixer is described in A.6.1. and was merely supplemented by a preamplifier.

The r.f.signal is routed via a highpass filter to the input tuned circuit and from there through 220pF to the emitter of T1. One leg of the trapezoidal or disc-type capacitor (470pF) is connected to the case, the other leg is soldered to the base of T2 and the two resistors (1K, 10K). The amplified input signal is passed on via L3/L4 and the coupling capacitor (~50pF) to the mixer transistor. The oscillator input is terminated in 50 Ohms to improve electric stability. The correct injection level (approx. 1-2mW) is set by means of a coupling trimmer. This signal is generated by a circuit according to D.6.1 and measured according to B.2.1.

Adjustment is achieved by setting the trimmer capacitors of the i.f.filter for optimum noise on the receiver and by setting the r.f.tuning elements for maximum signal level. Excessive signals (as during transmitter hunting) may be attenuated by approximately 40dB by switching-off the preamplifier stage. The overall bandwidth is essentially governed by the i.f.filter in this sort of concept. This appears logical since only single tuned circuits are employed at the input and behind the first stage.

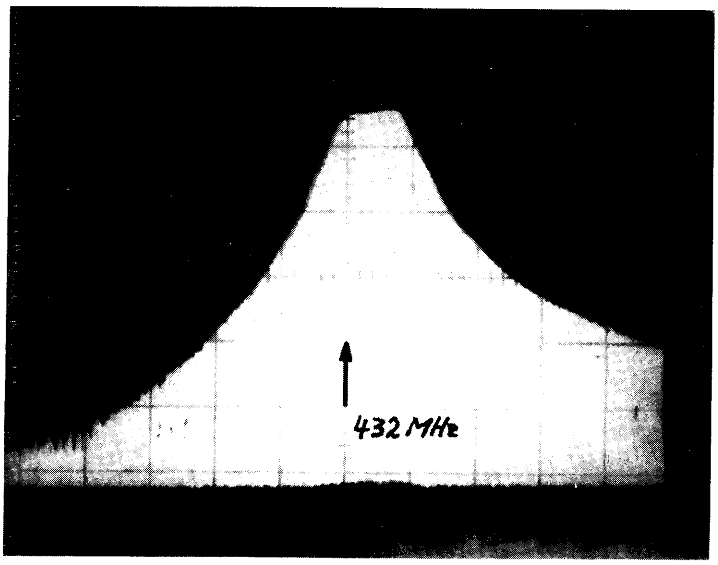


Figure 286
Overall passband of the converter, which is predominantly ruled by the i.f.filter - repectively the coupling capacitor (3.3p.f)

h: 3 MHz/division
v: 10 dB/division
f_{centre}: 432.0 MHz
Bandwidth: 3 MHz at -3 dB

Figure 287 Circuit diagram of simple 70cm converter

L1 = 3 turns CuAg 1mm diam. on 4mm mandrel

L2/3 = Cu 2mm diam., length is function of trimmer dimensions

L4 = total length 25mm connecting leads of 50pF capacitor

L5/6 = 25 turns Cu enamel 0.5mm diam. on 6mm mandrel

L7 = 5.5 turns Cu 0.5mm diameter insulated, wound over cold end of L6 in equal winding sense

i.f.filter trimmers: 4.5 - 70 pF (VALVO or equiv)

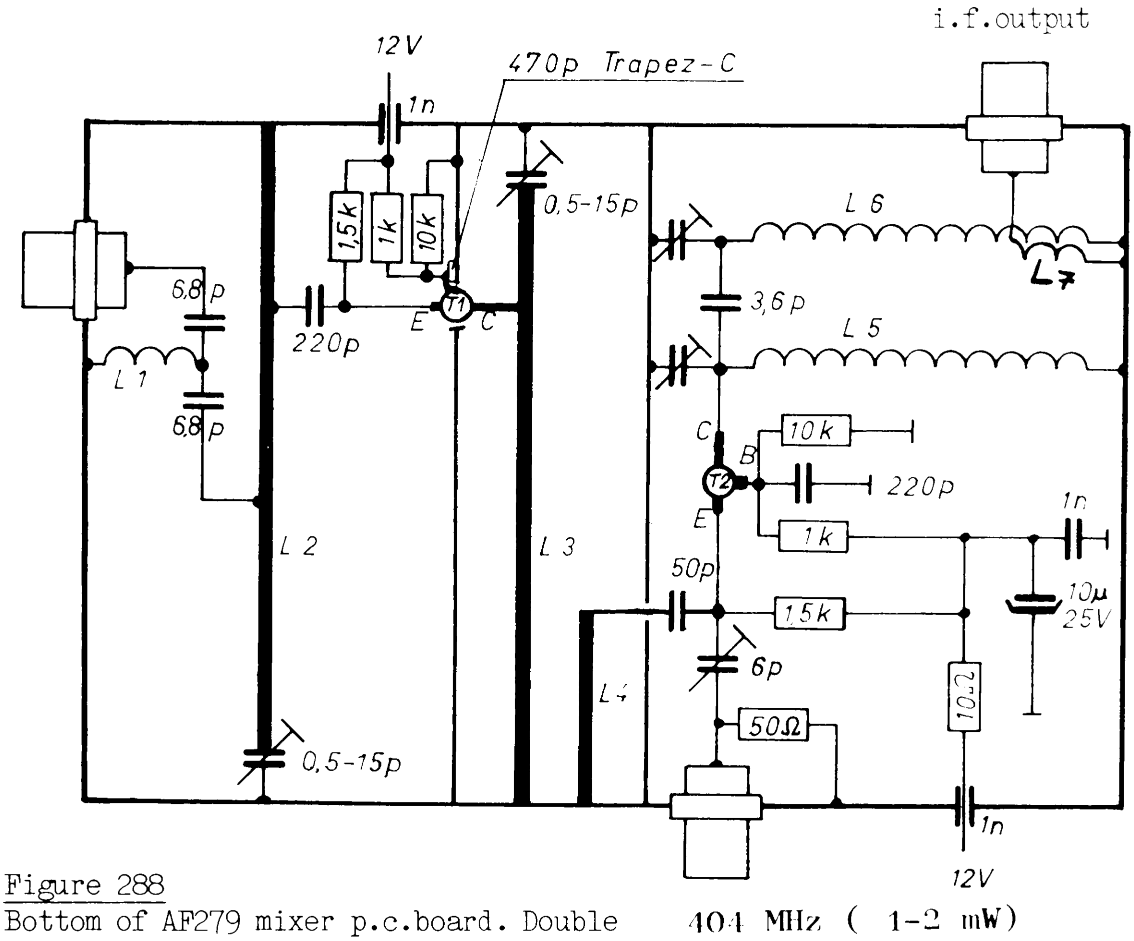


Figure 288
Bottom of AF279 mixer p.c.board. Double sided copper clad epoxy board was used. After drilling the holes the copper cladding around the holes should be removed by means of a drill bit (3- 4mm) to prevent ground contact of the components after installation and soldering.
The input stage components are mounted straight onto the case without the use of a p.c.board as clearly shown by the photographs.

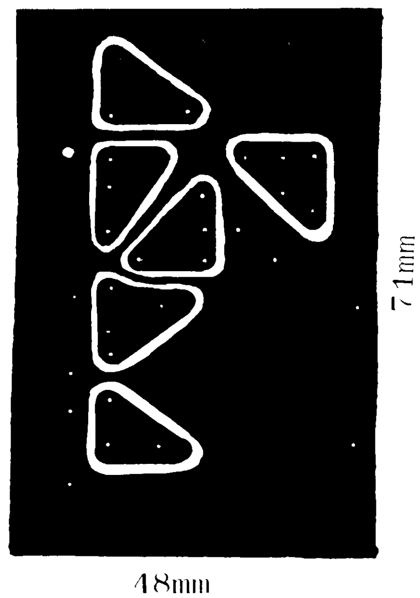


Figure 289 Construction and component layout of the simple 70cm converter

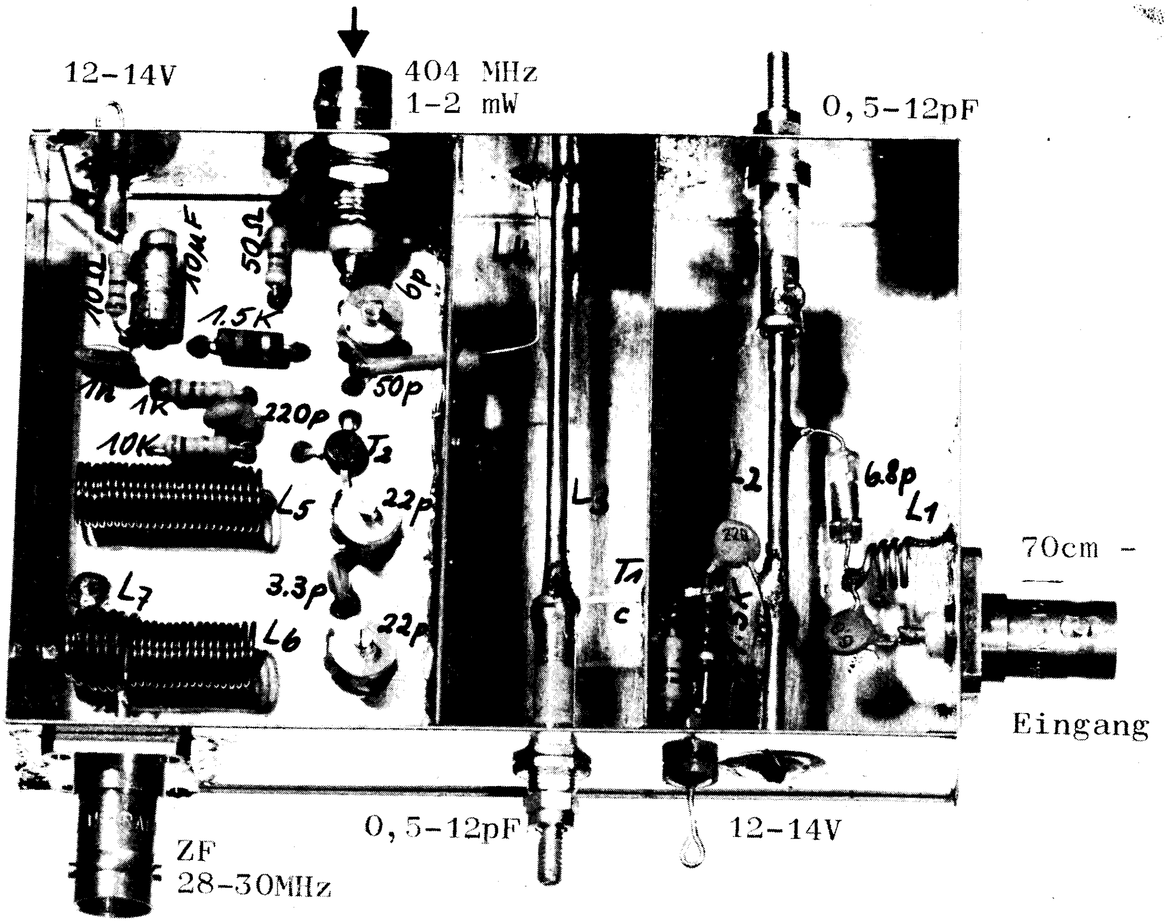
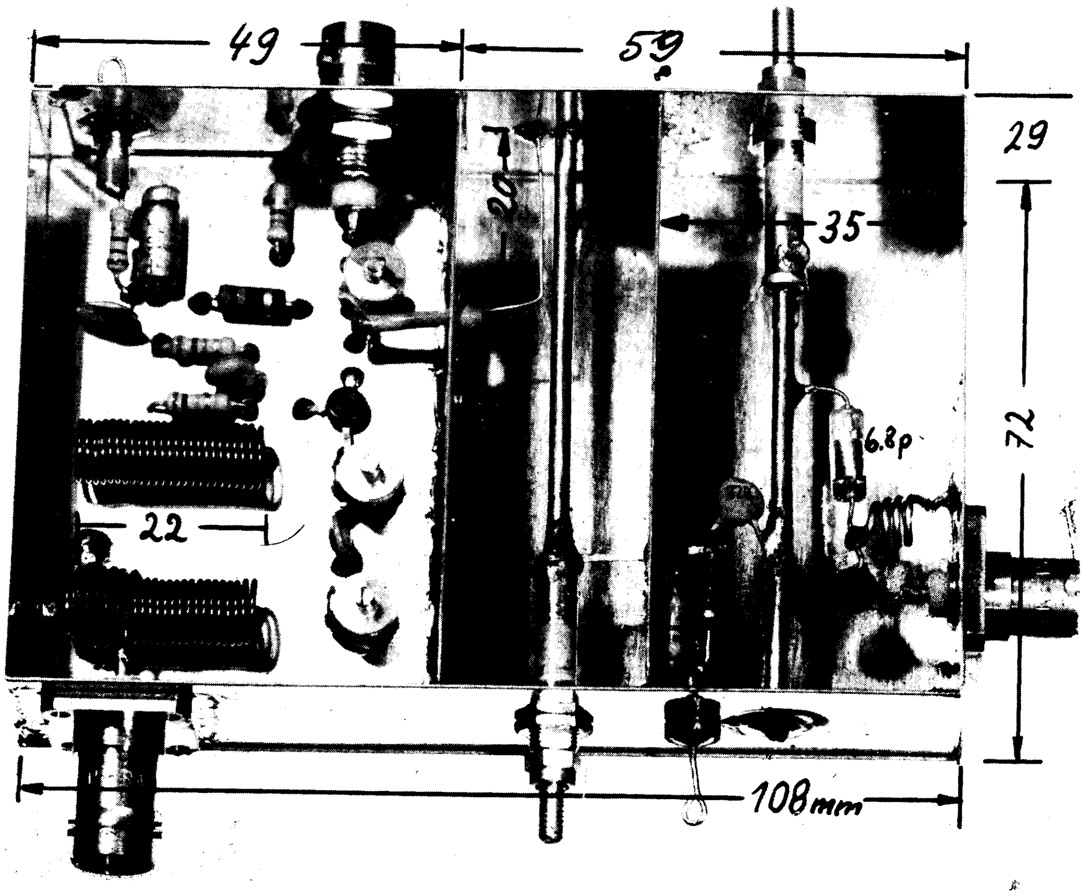


Figure 290 Dimensions of converter. The case is comercially available.



D.2.2 Selective converter for the 70cm band, suited for SSB/ATV

The proposed circuit may be constructed using either cheap or expensive transistors. The achievable selectivity will be very good in either case. Any bandwidth between 2MHz and 10MHz can be realized by merely bending the coupling strip. Selectivity is provided by a triple-tuned-circuit filter according to C.3.3 and the i.f.bandpass. The concept is thus equally well suited for SSB and ATV.
The r.f.signal is routed through a highpass filter and to the base of the preamplifier. The gain of a transistor stage in grounded emitter circuit is rather sensitive to voltage changes. Thus, an integrated stabilizer (type 78L12A or equivqlent) provides a constant operating voltage of 12 volts for the input stage out of supply voltages ranging from 13 to 20 volts. T1 could be any one of the listed types. However, reasonably priced are BFR34A and BFR90. Bipolar low-noise NPN UHF transistors by NEC, HP, VALVO etc. will also do a fine job. Transistor gain should - however - be around 16 to 20dB at this frequency (see D.1.1).
The r.f.passband is set by bending the silver-plated copper strip (Ag not necessary). The selectively amplified r.f. is then coupled to T2 via a capacitor (40 to 60pF) - one of its connecting leads being arranged to act as coupling loop. A foil-type trimmer (grey, ~6pF) is used to inject the 404MHz signal into the module where it is terminated in 50 ohms to improve stability. An injection level of 4mW is quite sufficient. Optimum sensitivity does not only depend on careful tuning and correct adjustment of quiescent currents but also on the correct injection level. Quiescent currents should initially be set to 5mA per stage.

Alignment procedure:

1) Connect operating voltage to mixer. Adjust the 250 ohm potentiometer so that a mA-meter inserted into the circuit will indicate a current of 5mA more than the basic current (which is given by the resistor chain 100+1500+250 ohm).
2) Connect the injection signal (max. 5mW) and set the 6 pF capacitor so that the mixer current increases by another 3 to 5 mA.
3) Upon connecting the receiver (28-30MHz) both trim capacitors of the i.f. bandpass filter are adjusted for maximum noise level. By now it should be possible to hear the harmonics radiated by the switched-on 2m set.
4) Next step is to connect the stabilizer to the supply and to set the preamplifier current to 3-10 mA according to step 1.
5) The harmonics of a 2m set, the signal generator according to B.8 or some other signal generator is used for fine-adjustment of the three cavity resonators. The coupling strip is shaped as shown on the photograph. The spacing betweenL2 and L4 controls the bandwidth. The converter is closed by means of a second cover plate. Alignment is finished by readjusting L3-L5 and T1 current. In case of extremely strong t.v. or broadcast transmissions in the neighbourhood an additional filter according to C.3.3 will help.

Figure 291 Selective 70cm converter in conventional wiring technique. The tinplate case (146 x 72 x 30mm) is a commercial product (G.4). In contrast to the circuit diagram this converterhead has been fitted with a tuned input circuit. The latter attenuates unwanted signals above 450MHz. The r.f. signal is passed to the base of T1 through a 15pF capacitor. Simpler to construct and with supperior noise properties is the arrangement according to the circuit diagram. A filter according to C.3.3 will provide protection in cases of interference. The photograph was taken prior to installing the partitioning walls to provide a clearer picture. The coupling strips L_K are bent upwards approximately 3mm short of the trimmer foils of L2 and L4 and cut off after 5mm and thus provide the coupling capacitances of the triple filter. The mixer is constructed on a piece of p.c. board which is then installed inside the case.

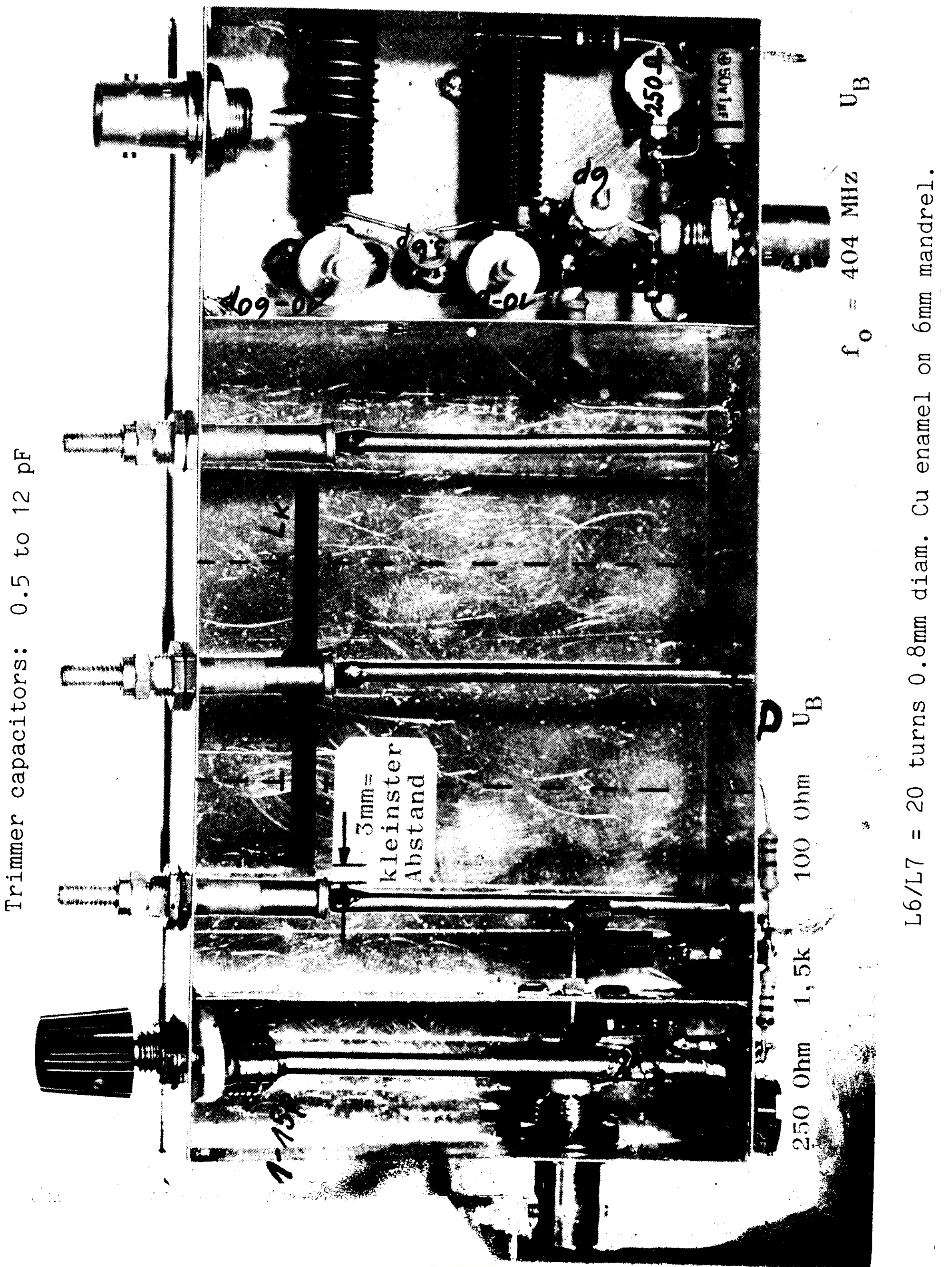
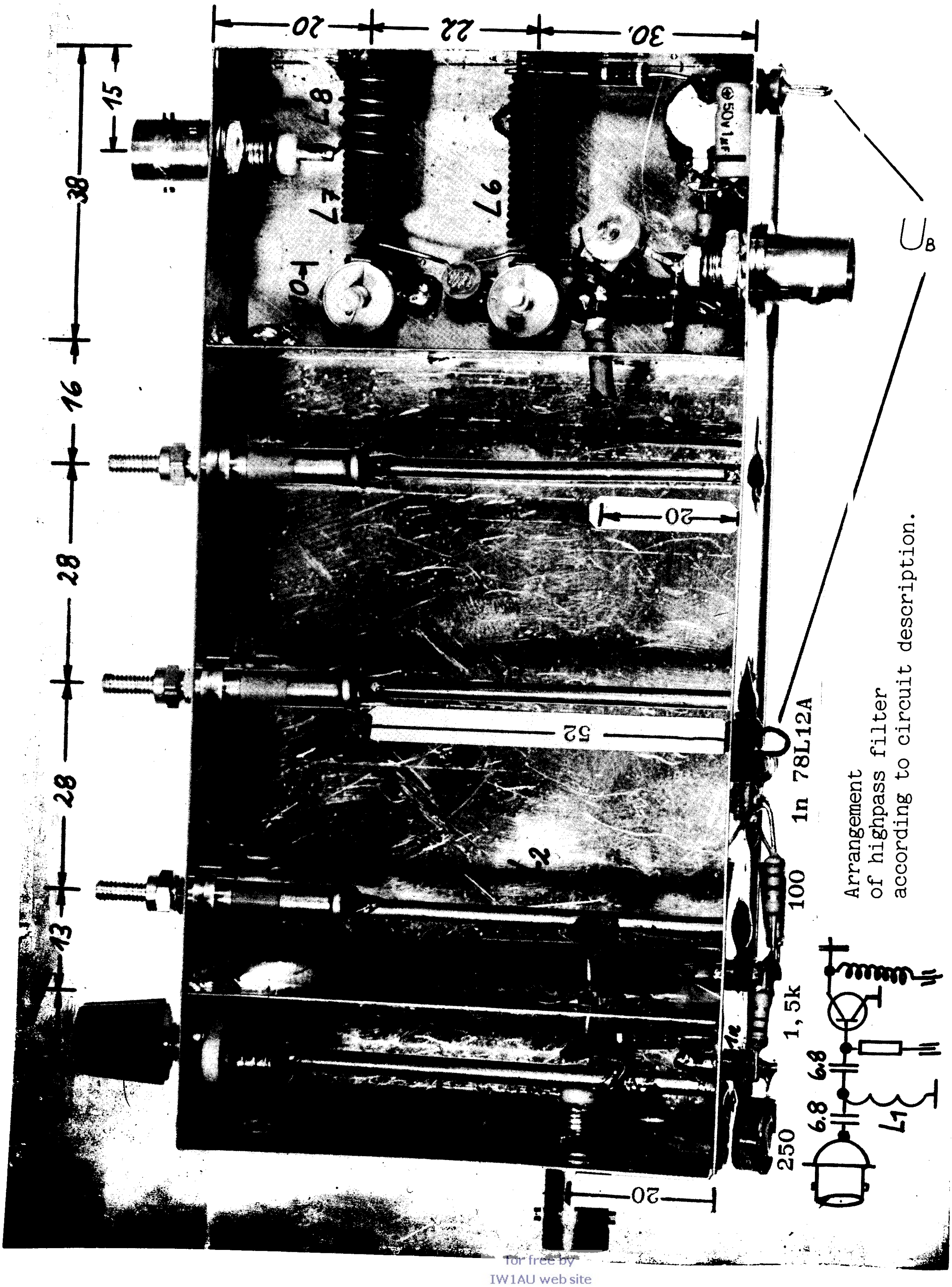
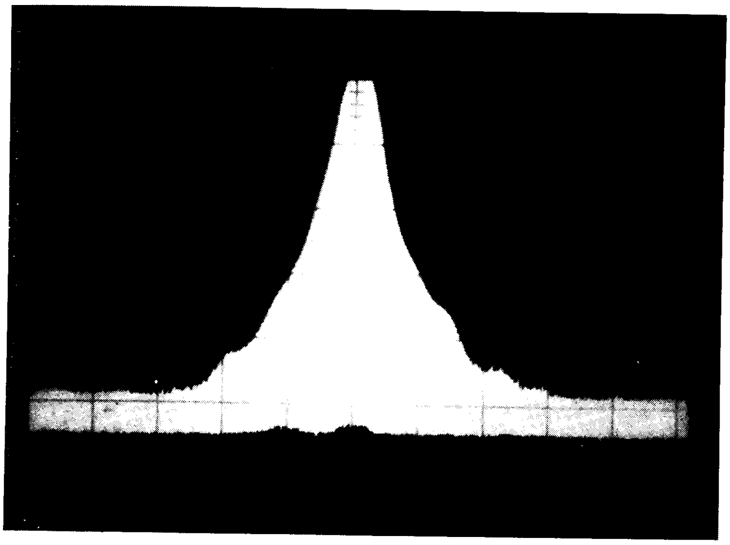


Figure 292 Dimensions (mm) of selective 70cm converter.
L2 - L4 are forced through holes that are drilled into the sides along their centerlines with diameters to provide a tight fit. Solder from outside only! The properly shaped connecting lead of the coupling capacitor constitutes L5 and is positioned next to L4 as shown. The partioning walls indicated in the previous photograph are absolutely essential to obtain a passband according to figure 293. A 10 ohm resistor was inserted between the 250 ohm potentiometer and the base biasing resistor for convenience.





283 433 583 MHz

Figure 293 r.f. passband of selective converter. It is governed by the filter composed of the three tuned circuits.

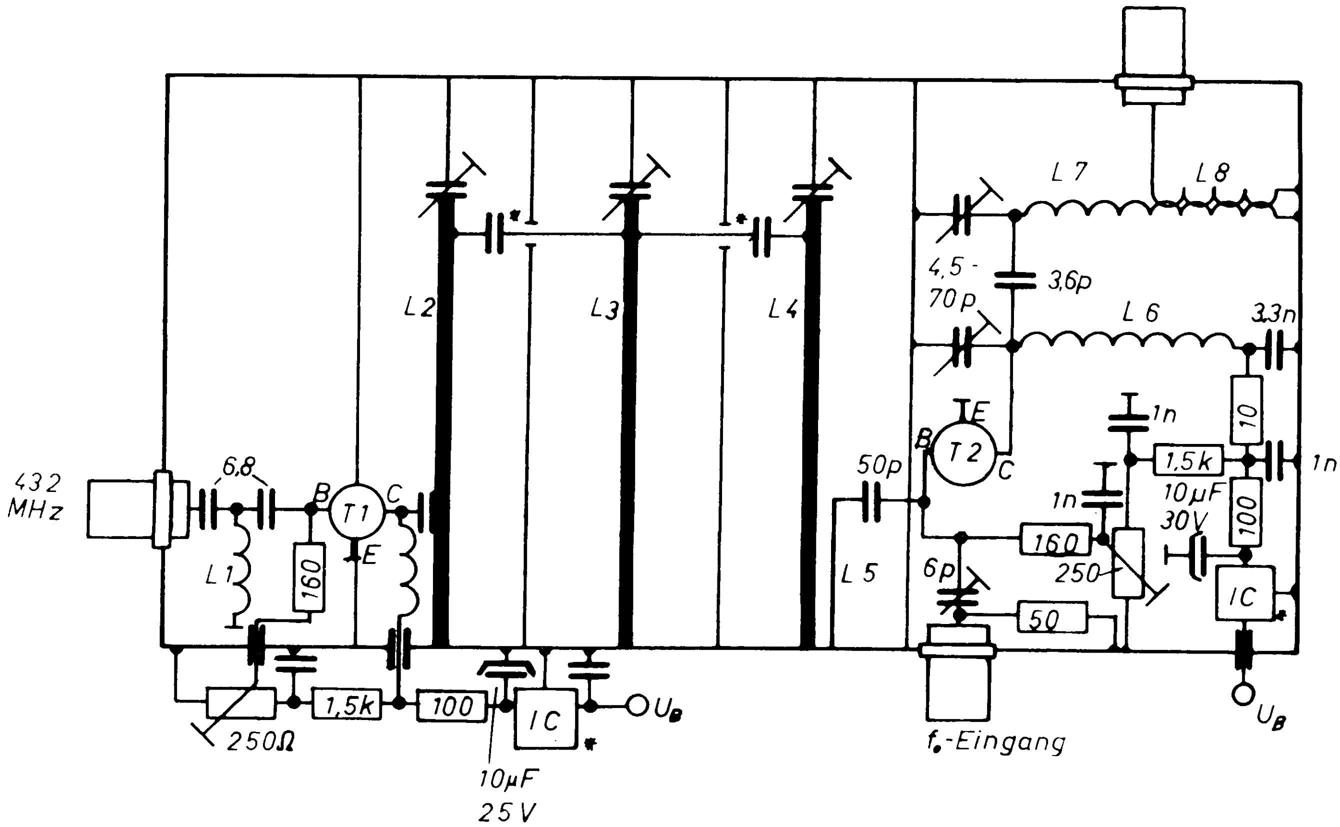
h: 30 MHz/division
v: 10 dB/division

f_{centre}: 433.0 MHz.

The width of the passband may be varied from less than 2MHz to more than 10MHz by bending the ends of the coupling strip closer to or farther away from L2 and L4. If no sweep generator with display facilities is available then start with approximately 3mm spacing between the coupling strip and the capacitor foils of L2 and L4. The lines are tuned for maximum output using a fixed frequency generator (eg. B.8). Following that, the ends are bent so as to increase the spacing to 4mm and the tuning is repeated. If the output level is now significantly lower, then the optimum spacing is already passed (see D.2.1 for alignment procedure). Final adjustment is performed using a beacon signal or the signal of a local station.

Figure 294 Circuit diagram of 70cm converter.

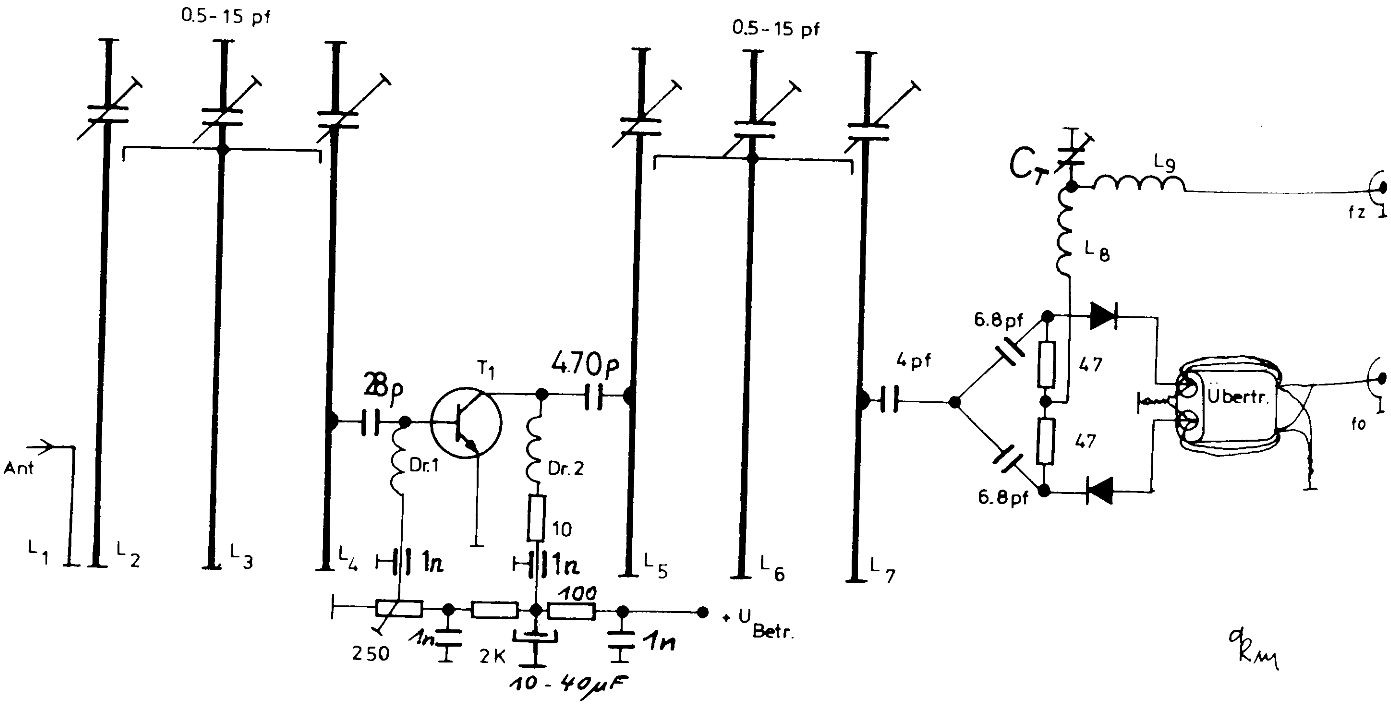
L1 = 2 turns 1mm CuAg on 4mm mandrel.



D.2.3 Selective 70cm converter with excellent strong signal handling capabilities

The following building project is a typical example for the possibility of combining various modules described in this book. It is evident from the circuit diagram that this converter is the combination of triple tuned circuit filters according to C.3.3, a preamplifier according to D.1 and a push - pull mixer according to A.5.2. Should one expect strong signals (at either the signal frequency or at other ranges) this - or a similar - concept should be chosen. The details are discussed in the respective sections and are not repeated else where. The chosen transistor should provide approximately 20dB of gain. This will allow loose coupling between L5 to L7 resulting in a slight reduction of the overall gain and a significant narrowing of the r.f. passband. Reducing the coupling from L1 to L4 will have a detrimental effect on sensitivity. If no transistor with this sort of gain is available (by HP, NEC, Siemens, Valvo to mention just a few) then a BFR34A could be employed with only slight reduction in performance. An additional stage of preamplification in conjunction with an additional triple filter could bring up the gain. The strong signal handling capability would, however, suffer. As stated before, the first stage should be equipped with a low noise high transconductance transistor and the second stage with a transistor having higher collector power rating. Since the pre-mixer gain of 20dB is quite sufficient the choice of the second transistor is not ruled by gain considerations but rather by the ability to handle higher signal levels.

Figure 295 Circuit diagram of 70cm converter. Further details are specified in sections C.3.3, D.1 and A.5.2 or may be obtained from the photographs.



L2-L7 = dimensions depend on parameters of trimmer capacitors (In this example: L2-L7 = 60mm long, CuAg, 2mm diam.)

L_K = copper strip, silver plated, total length 35mm, width 4mm, last 5 to 7mm of each end bent according to photograph. The openings in the partitioning walls are 6mm diam.

L8/L9 = 30 turns of 0.2mm diam. Cu-enamel on 5mm mandrel

C_T = 10 to 60 pf foil type or ceramic trimmer capacitor.

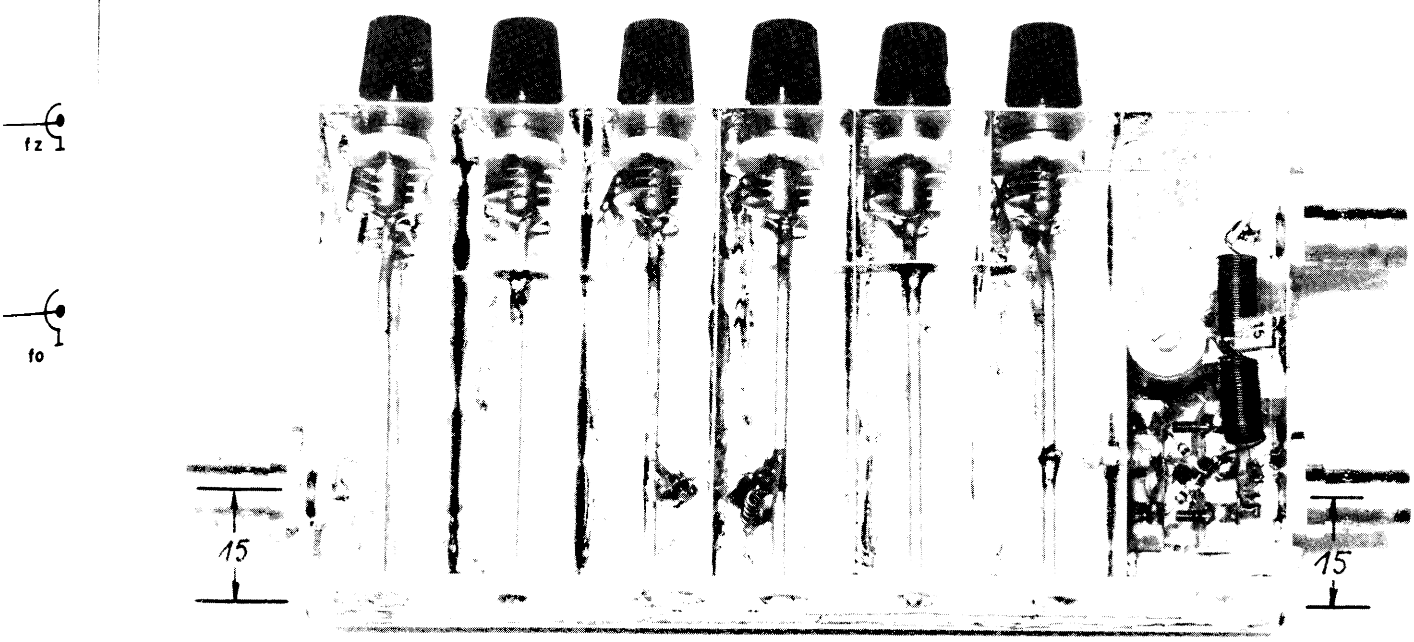
Dr1 = quarter wave length of 0.2mm diam. Cu-enamel on 2mm diameter coil former.

Choosing a different i.f. necessitates modifications to f_o, L8, C_T and L9.

The mixer itself is constructed on posts in the right hand compartment of the case. This makes etching of a p.c. board unnecessary. Since the lines are nearer to the partitioning walls than to the base plate the ground reference is shifted towards these walls. It then follows that near-optimum tuning can be achieved even with the cover plate removed.

Choke Dr.2 is formed by winding five turns of the lengthened connecting lead (of the 10 Ohm resistor) on a 2mm diam. mandrel. The value of this resistor depends on the transconductance resp. the oscillation properties of T1. Coupling to and from the transistor is through trapeze-ceramic capacitors (approximately 30 and 470pf). Both capacitors are soldered straight to the trough lines - approximately 18mm off the cold end. The stabilizing circuits and the voltage dividers are mounted onto the rear cover. It is highly recommended to stabilize the operating voltages at 12 Volts by means of an integrated circuit such as 78L12A.
Supply voltage variations between 13 and 18 Volts will thus cause no gain irregularities.

Figure 296 Filter trough lines and mixer/i.f.compartment of converter.



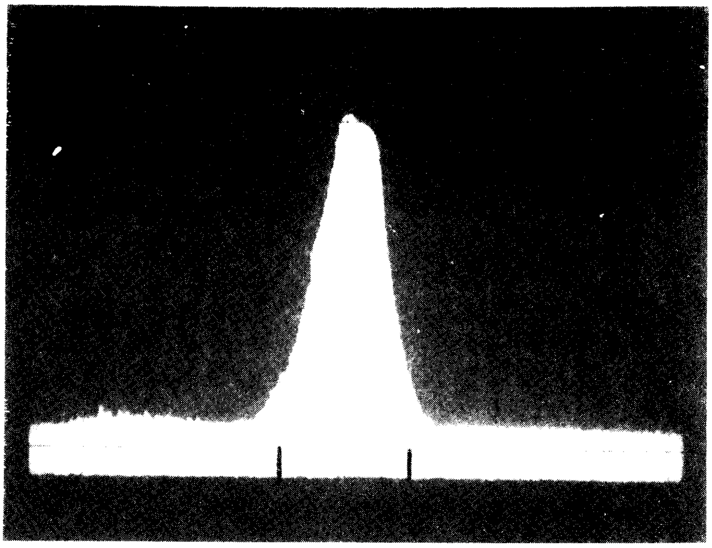


Figure 297
Overall pass band of converter over the range 382 to 482 MHz.

h: 10 MHz/div.
v: 10 dB /div.

Centre 432 MHz

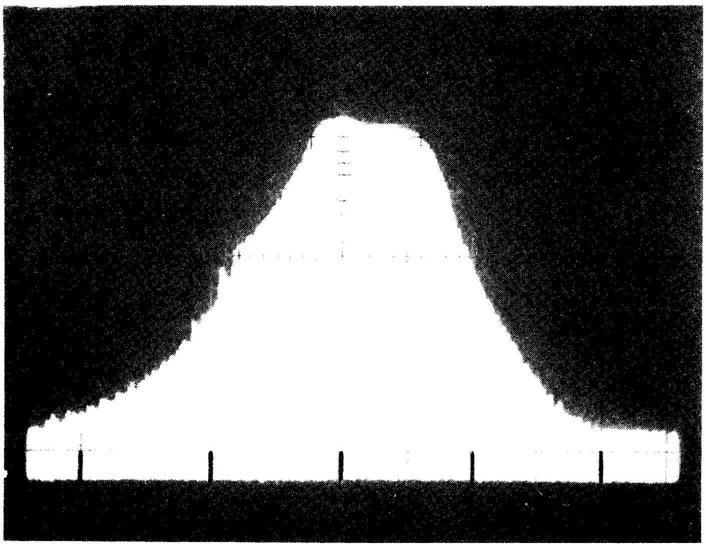


Figure 298
Stretched presentation of frequency band to show 3 dB bandwidth.

Δf_{3dB} = 5.5 MHz

The overall bandwidth may be made wider or narrower by making the coupling between the resonant circuits looser or tighter. The tuning capacitors for L2 to L4 should be low loss air or glass dielectric. Ceramic capacitors for L5 to L7 will merely reduce the stage gain.

Alignment:

Starting from ground the 250 ohm potentiometer is turned in until I_{T1} reads about 5 mA. Harmonics of a 2m transmitter are then utilized for course adjustment while the cover is still removed. Final tuning is performed with the cover installed and by using a real 70cm signal. This concludes the alignment process.

D.2.4 Converter for the 23cm band in stripline technique

The advantages of modular design were explained in the introduction (D.2) of the converter section. These arguments are even more valid for the 23cm band since screening between the individual stages could become even more of a problem. Furthermore, technology - in particular as far as transistors are concerned - progresses so rapidly that the possibility of replacing a complete stage should be an important design consideration. Since all other modules remain operational and only the stage concerned needs to be updated swift modification is guaranteed.

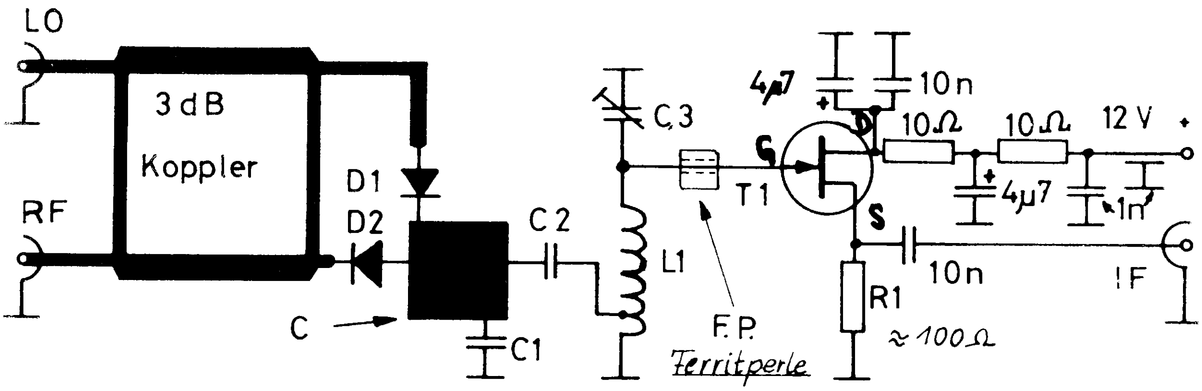
It is thus advisable to install just the mixer - combined with one i.f. stage at the most - in a separate case. As mentioned elsewhere in in this book (D.1) 20 dB of preamplification should be provided. In conjunction with a preamplifier according to D.1.4.2 - D.1.6 a very sensitive concept may be realized. Employing a GaAs-FET preamplifier as described in D.1.6 should result in the optimum solution presently achievable.

The above-mentioned preamplifier reaches its 1 dB compression point at a level of -20 dBm (i.e. 0.01 mW input level). Because of the 20 dB gain this will result in an output power level of 1 mW. In other words: The preamplifier is capable of handling input signals of up to 0.01 mW without distortion.

These levels have been taken into consideration in the following converter description. In conjunction with the high performance i.f. stage (FET: P 8000) the converter head described hereafter will reach its 1 dB compression point (A.4.3) at exactly 1 mW antenna input power level. If preceded by the GaAS-FET preamplifier the 1 dB compression point will be at -20 dBm.

The converter may be designed for intermediate frequencies of between 28 and 144 MHz.

Figure 299 Circuit diagram of the converter in stripline technique.



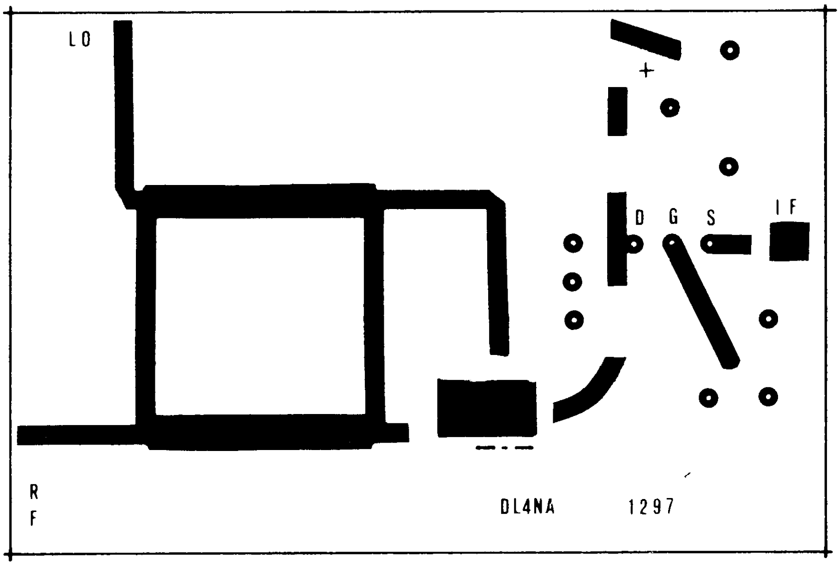
Component values for intermediate frequencies of either 28 or 144 MHz

	28 MHz	144 MHz
L1 =	25 turns 0.6mm on 4mm mandrel tapping at 5th winding	5 turns 1mm CuAg on 6mm mandrel tapping at first winding
C1 =	30 pF disc capacitor	10 pF disc capacitor
C3 =	3.5 - 70 pF + 33 pF(fixed)	5 - 25 pF
C2 =	10nF	1nF

T1 = P 8000, D1/D2 = HP 2800 or better

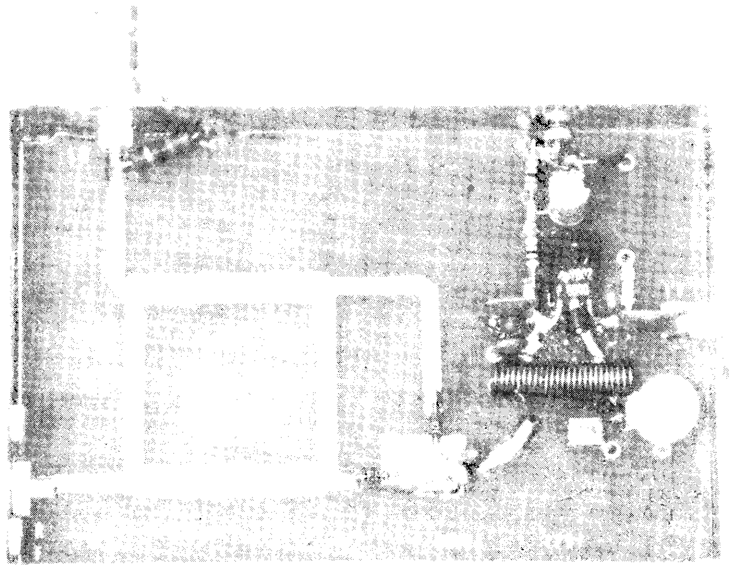
Set I_D to approximately 30 mA by means of R1

Figure 300 Converter head printed board. The 3dB coupling line was calculated by DL4NA according to (1). The i.f.stage is a proven circuit as designed by DC9NI and DJ9HO. The theoretical background for calculating these 3dB couplers is treated in sections A.2.4 and A.5.3.1. Double sided copper clad epoxy based material of thickness 1.5mm was used.



An injection frequency of 1152MHz should be selected for an i.f of 144MHz. The corresponding value for an i.f. of 28MHz is 1268MHz. A circuit is suggested in D.6.2. An injection signal power of 2 to 5mW is required.

Figure 301 converter as built by DC9NI. The i.f.stage using the P8000 is clearly visible to the right. The local oscillator input is terminated in a 50 ohm film or composition type resistor (low inductance). The only alignment necessary is to adjust C3 for signal maximum.



Reference: (1) Unger/Harth "Hochfrequenz-Halbleiterelektronik"
S. Hirzel Verlag Stuttgart

D.2.5 Converter for the Meteosat band in stripline technique

The circuit diagram corresponds to the one of the previous section. Merely the frequency dependent 3dB coupler was redesigned for the new frequency. The i.f of approximately 137MHz is obtainable with the component values of the 144MHz version (increase capacitance of C3). Again, copper clad epoxy board of 1.5mm thickness is used.

Figure 302 Printed board of converter head for the Meteosat band. An injection signal level of 2 to 5mW is sufficient.

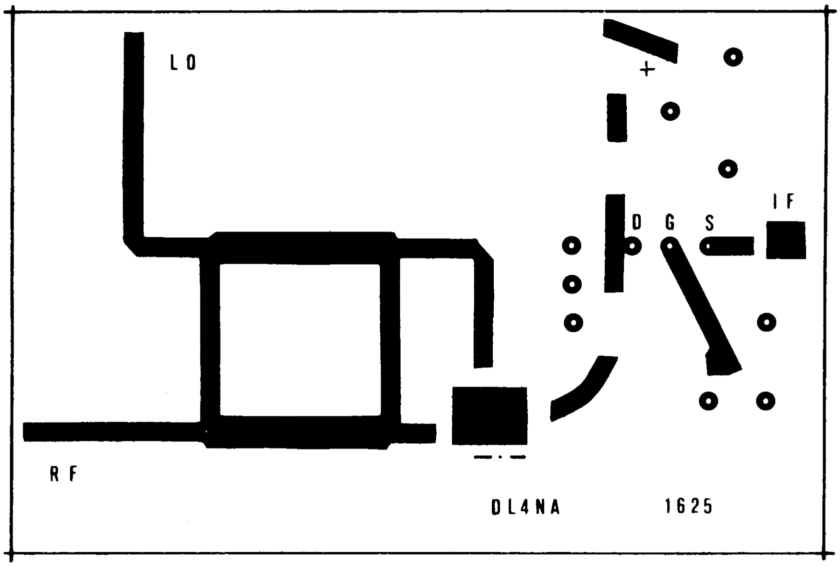
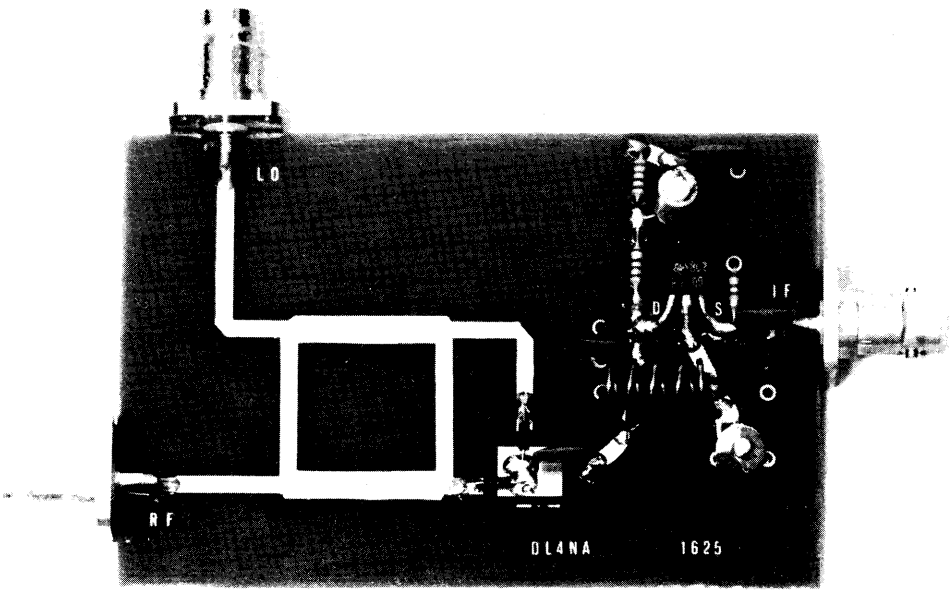


Figure 303 Assembled converter head without case. C3 is replaced by a foil trimmer 5-25pF. After installation in a tinplate case (110x72x30mm) the rear ground foil of the p.c.board is soldered to the case. Again, all the alignment required is to tune C3 for maximum signal. A BF245 may be used should no P8000 be available. This does, however lead to a slight reduction of gain, large signal handling and noise properties.



D.3 Multiplier stages
(N. Schramm, DC9NI)

A frequent requirement is the generation of frequency stable signals at low to medium power level at the higher frequency bands - for instance
- injection signals of 0.5-10mW for converters in the 2m - 3cm bands
- injection signals of 10mW - 1 watt for transverters in the UHF band,
- driver signals of 50mW - 10 W for FM and CW transmitters.

There are two fundamentally different types of multipliers:

1. The active multiplier
based on transistors or tubes. A signal is fed to the input of a component and is distorted by its non-linear characteristics. A resonant circuit tuned to the desired harmonic is then utilized to select the desired multiple of the input signal. The active multiplier exhibits gain.

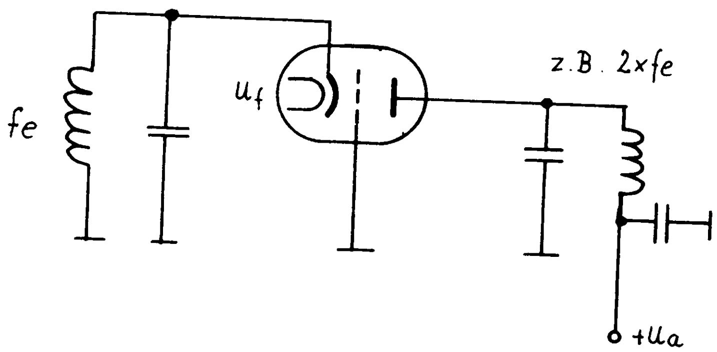


Figure 304

2. The passive multiplier based on varactor diodes.
There are two different types of varactor diodes:
   a) Utilization of the highly voltage-dependent junction capacitance (like BB105);
   b) Utilization of the charge storage properties of storage varactors. Here the sudden current reversal due to the blocking action is employed.
   The passive multiplier gives rise to transfer attenuation.

Designing multipliers requires attention to the following aspects

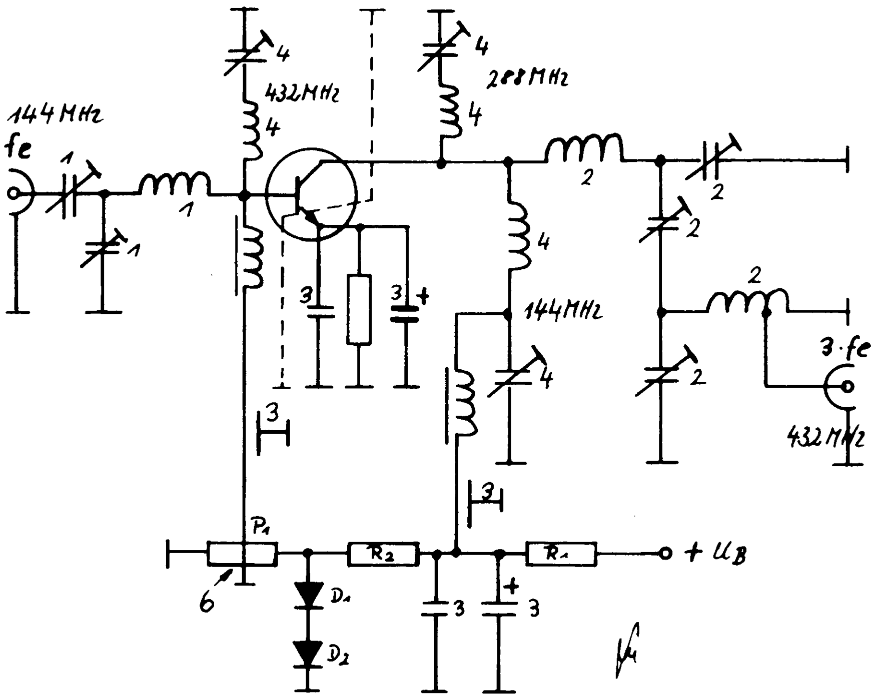
1. Perfect matching between driver and multiplier stages;
2. Optimum matching at output and selectivity;
3. Careful blocking of all frequencies that exist in the stage and additional low frequency blocking of supply voltages of keyed systems (CW).

4. If large even-harmonic power components are generated in systems with odd multiplication factors then idler circuits should be included to short-circuit the undesired components - thus improving the efficiency of the varactor.
5. It is obvious that all components should have the appropriate power rating.
6. The working points for doubling, tripling or quadrupling require different settings for optimum results.

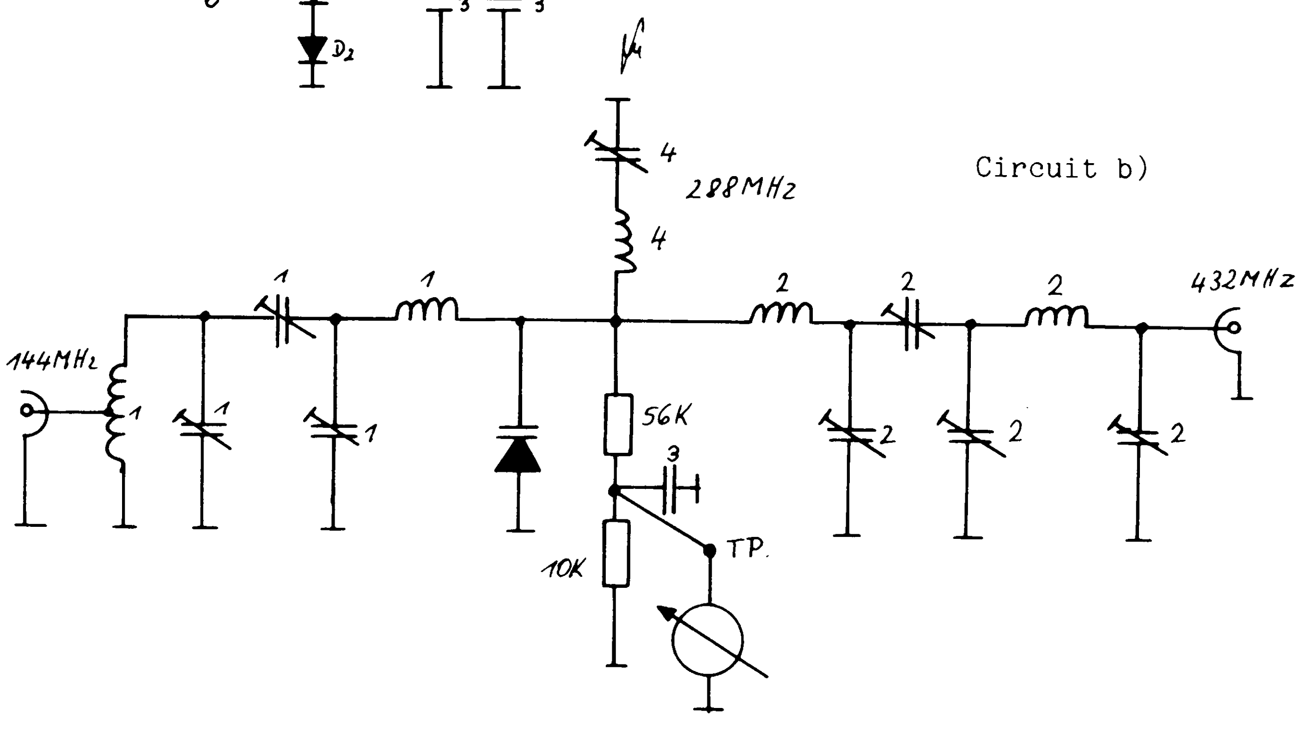
<u>Figure 305</u> a) Example of active multiplier
b) Example of passive multiplier

The numbers (1 to 6) next to the components indicate by which of the mentioned aspects the respective component is governed.

Circuit a)



Circuit b)

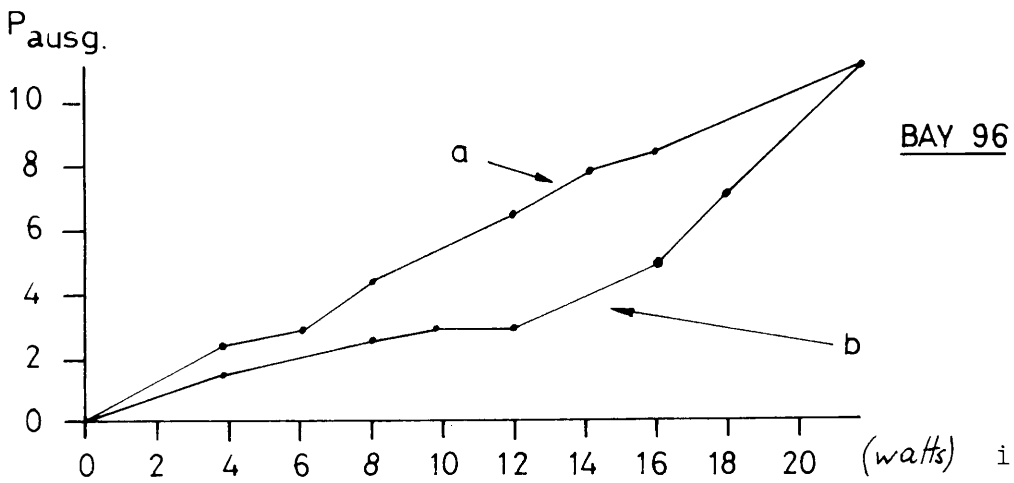


Varactor diodes exhibit significant impedance changes as the applied power is altered. It is thus necessary to re-tune varactor multipliers if the input power level is changed.

The following diagram displays the interrelation between input power and output power of a passive varactor multiplier using a type BAY 96 diode according to the afore-mentioned circuit variant b).

Figure 306

output power



input power

If all tuning elements (trimmer capacitors) are re-set to optimum corresponding to the respective input power levels then the best efficiency of ciricuit b) can be obtained. The output power achieved is depicted by curve a.

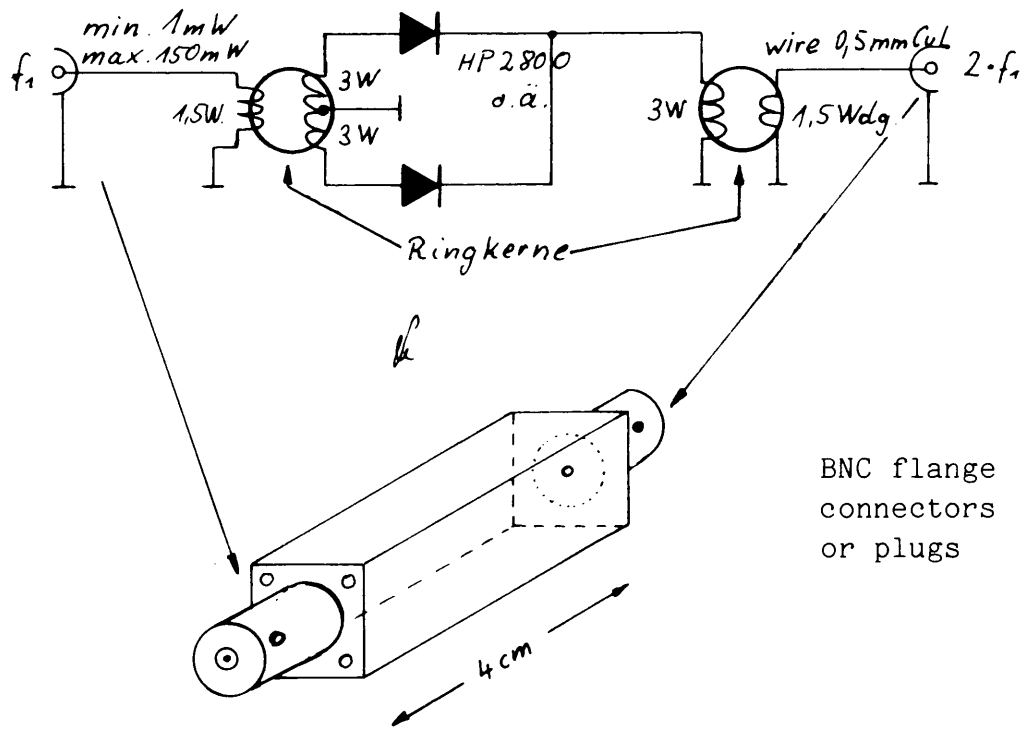
If the tuning is optimized for maximum input power and the power is then reduced without re-tuning then the less favourable power transfer as shown by curve b is achieved.

As shown by these two examples it is recommended to optimize the varactor multiplier for the actual input power level. If pre-stage modulation (FM of CW) is employed then this is a simple process. For the AM case, however, one should not merely look for favourable output power but also for low distortion and best readability. This requires monitoring through the opposite station or by a separate receiver.

D.3.1 Active and passive 2m/70cm multipliers

Broadband doubler for up to approximately 150 mW input power:
The output signal contains strong even harmonic components. A filter on the desired frequency is absolutely essential.

Figure 307 Frequency doubler using Schottky diodes

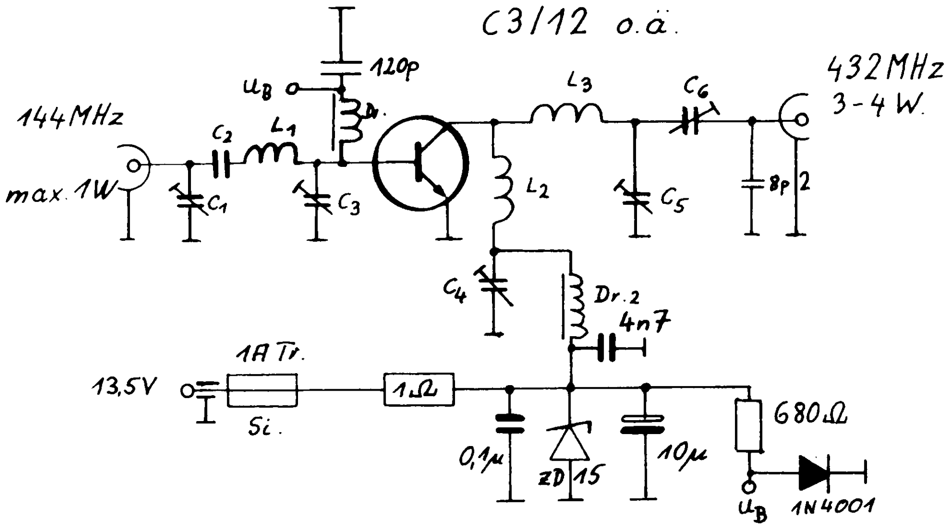


List of materials:

1 BNC flange connector
1 BNC flange plug
1 U-shaped case (tin plate), sides measure 18mm, length 40mm, with cover
2 toroidal cores (UHF version) or 2 double-hole cores (for instance salvaged from t.v. input circuit).
2 Schottky diodes - such as HP 2900/338
copper enamel wire, 0.5mm diam.

Important notice:
The specified numbers of turns are highly dependent on the core material and the operating frequency. The listed values allow operation up to approximately 500 MHz.

Figure 308 144/132 MHz transistorized tripler



C1 = 3.0-30 pF
C2 = 1 nF
C3 = 3.0-30 pF
C4 = 0.5-15 pF
C5 = 0.5-5 pF*
C6 = 0.5-15 pF
C7 = 0.5-15 pF
*) if necessary

	turns	diam.	wire
L1 =	3.5	on 6mm,	1mm CuAg
L2 =	2	on 6mm,	1mm CuAg
L3 =	1	on 8mm,	2mm CuAg
Dr =	Valvo, six holes		

Aligment:

Required equipment:

2m-TX 1 Watt, SWR-meter, dummy load with power meter, amperemeter, power supply 13.5 Volt.

Procedure:

Connect 2m-TX via SWR-set to input, connect dummy load to output and insert amperemeter into positive lead of power supply.

1. Adjust C1 and C3 for max. collector current and good 2m SWR.
2. Adjust C5, C6, C4 for maximum output power.
3. Repeat this cycle several times.

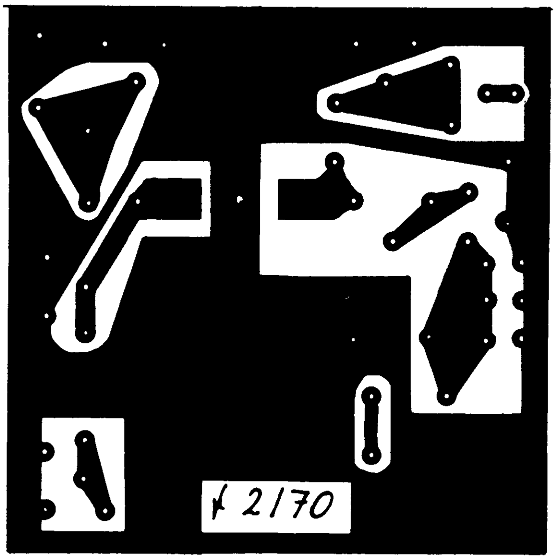


Figure 309
Printed board of multiplier. Solder posts and components are mounted on the other side of the p.c.board.

Figure 310 Assembled active tripler

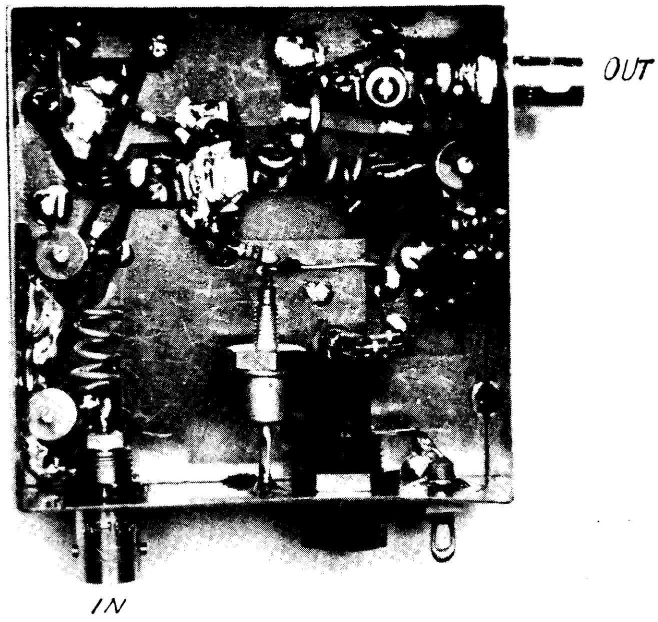
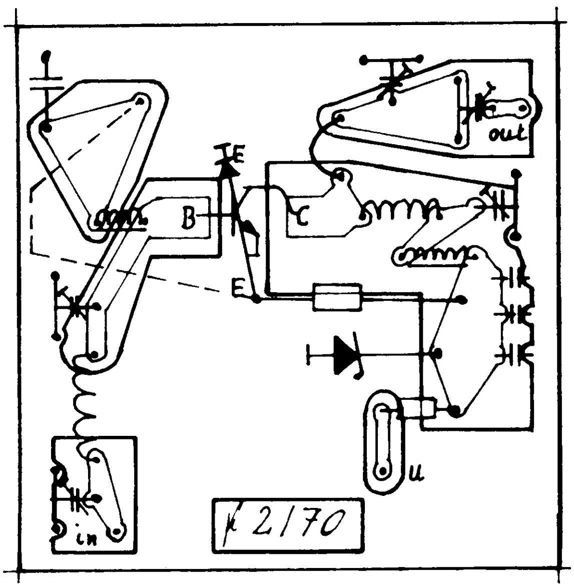
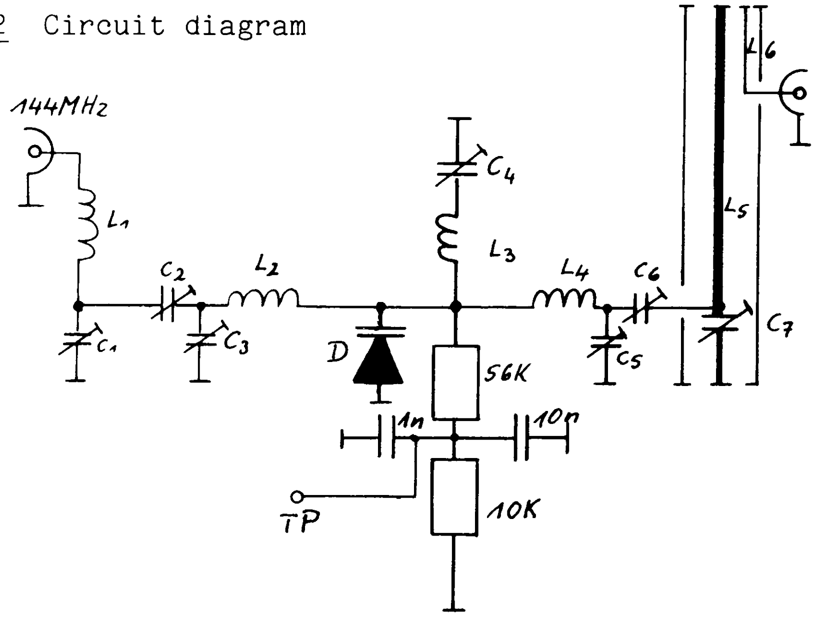


Figure 311 Component location. The module was installed in a case measuring 72 x 72 x 30 mm. The maximum input power of 1 Watt must not be exceeded for the type C3-12 transistor without risking its destruction.



144/432 MHz varactor diode tripler

Figure 312 Circuit diagram



	turns	wire diam.	coil diam.
L1	5.5	1mm	6mm
L2	5.5	1mm	6mm
L3	3.5	1mm	6mm
L4	2	1mm	6mm

L5 tubing, brass or copper, 11mm diam., 13cm long in trough 27x27x145mm
L6 3 cm long, positioned at cold end of L5
C1...C5 3 - 13 pF
C6 metal vane, 5mm wide
C7 3 - 13 pF
D BAY 96 or equivalent. (Very well suited - but with varying results - are the base-collector junctions of r.f.power transistors, such as 2N3375, C3/12, C12/12, 2N3553)

Alignment of 144/432 varactor tripler

Required equipment:
2m transmitter, SWR-meter for 2m, dummy load for 70cm with power meter (at least relative), voltmeter.

Warning! The output power of the 2m transmitter must not exceed the maximum allowable power rating of the diode!

Procedure:
1. 1/10 of the power for which the diode is rated is fed into the input via the SWR meter. The dummy load is connected to the output. The voltmeter is connected to TP.
2. Adjust C1, C2, C3 for maximum voltage reading with simultaneous optimum SWR between 2m-TX and tripler. Repeat this sequence several times.
3. Adjust C5, C6, C7 for max. output power (voltage reading will drop).
4. Adjust C4 for minimum voltage reading.
5. Increase input power to full value. Repeat steps 2, 3 and 4. All adjustments are highly dependent on input power!

Figure 313 Construction of passive multiplier 144/432 MHz in case 146 x 72 x 30 mm.

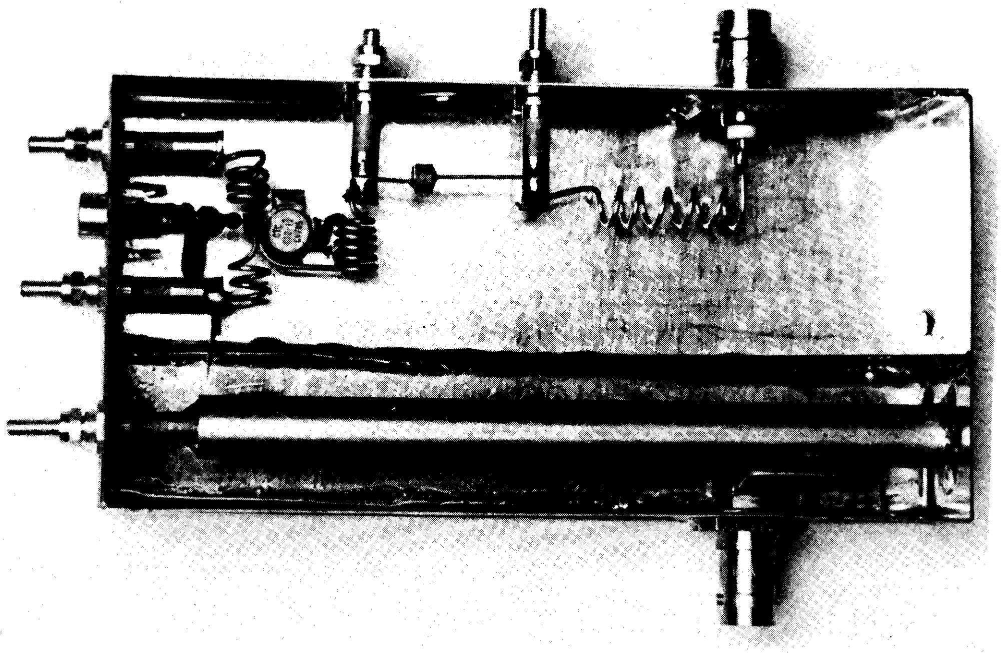
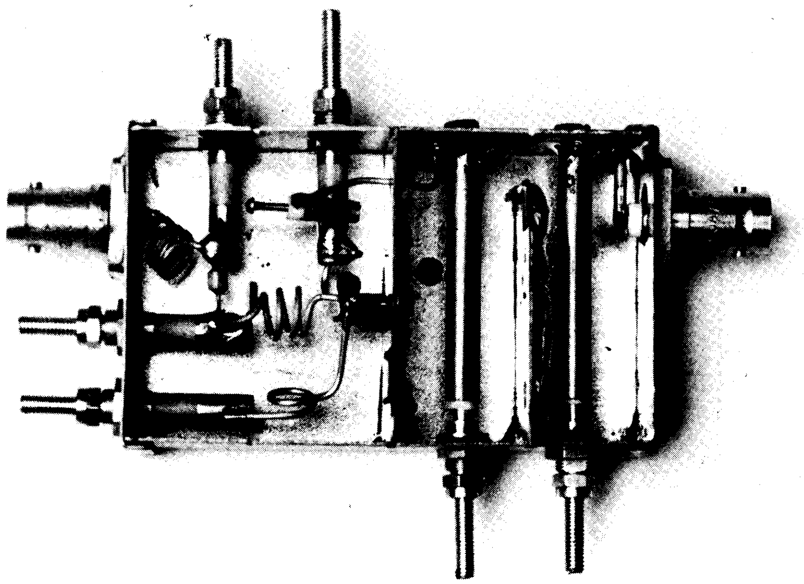


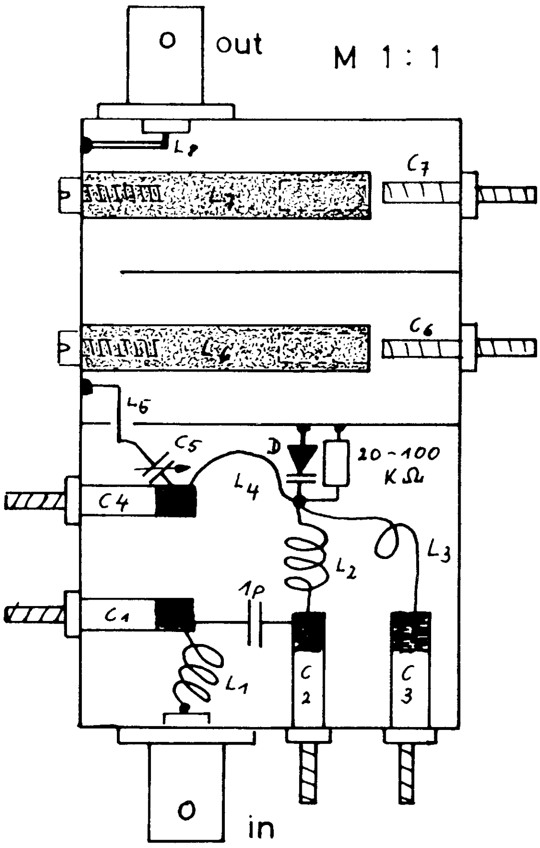
Figure 314 Construction of varactor tripler 432/1296 MHz in TEKO case (dimensions 81 x 49 x 25 mm).
The circuit is described on the following page.



D.3.2 70cm/23cm varactor tripler for low to medium power in TEKO case

Peculiarity of the circuit

Two different versions have turned out to be successful. Depending on the selected diode they will show differing conversion efficiencies (P_{in}/P_{out}) - see list.



Type	Pin	Pout	Version
BB 105	1 watt	200 mW	B
BA 149	1 watt	230 mW	A
BYX 19	12 watt	6.3 W	B
1N4148	1 watt	180 mW	A

Figure 315 Version A.
The diodes BA 149 and 1N4148 have been tested. As indicated by the list, diodes may be used that are normally not suited for this frequency range. It is thus worthwhile to test diodes from the scrap box.

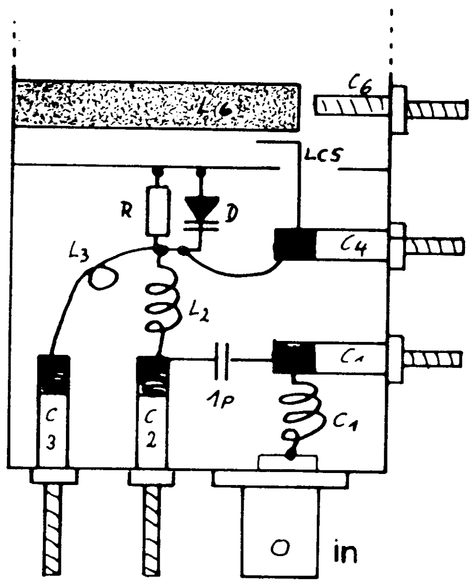
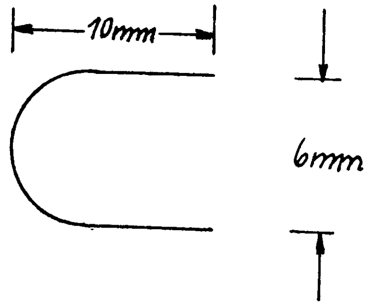


Figure 316 Version B

Both diagrams full size.

List of material:

1 TEKO case 81 x 49 x 25 mm
2 BNC flange connectors
1 Resistor 20 to 100 ohm, low inductance
2 screws M3 x 10
C1...C4 3 to 13 pF
C5 0.3 to 3 pF
C6, C7 tuning spindles (ceramic tubing of trimmers removed)
L1, L2 3 turns of 1mm CuAg on 6mm mandrel
L3 1 turn of 1mm CuAg on 6mm mandrel
L4 U-shaped loop of 1mm CuAg



L5, L8 see figure 315
LC 5 see figure 316
L6, L7 brass rod, 6mm diam., 38mm long, drilled and tapped M3 to fit tuning spindles (if necessary, internally covered with Teflon foil).

The diode must be chosen according to power requirements.

Alignment:

1. Connect 70cm TX via SWR-meter to connector P_{in}.
2. Connect power meter to P_{out}.
3. Switch on 70cm signal (Warning: Do not exceed power rating of varactor diode!). Adjust C1 and C3 for minimum SWR.
4. Adjust C6 and C7 for maximum power. Adjust C5 resp. bend LC5 for optimum P_{out}.
5. Adjust C3 for maximum P_{out}.
6. Set C1 and C2 for maximum power and simultaneously good SWR.
7. Repeat steps 4 to 6.

Notes on construction:

a) The partition between L6 and L7 has the same height as the case but is 5mm smaller than the internal case width (5mm coupling iris).
b) The partition between input network with diode and L6 may be made up from brass sheet or tin plate for low power levels. In the case of higher power levels - i.e. if using the BXY 14 HG - it should be made from copper sheet.

D.4.1 10m/70cm transmit converter (TXM D2T 10/70)

A transmit converter for UHF should present the state of the art and should be suited for a variety of applications. To the designer this implies:

1. Integrated means to provide selectivity (filter to obtain high oscillator and image rejection.
2. Choice of modern transistors at moderate prices.
3. Variable transistor complement to suit different output power requirements resp. to fit the levels planed throughout the station concept.

A balanced diode mixer according to A.5.2 was chosen as converter. Higher input levels may thus be handled and the driver level is less critical when compared against a transistorized mixer. This is advantageous in so far as only two high-quality transistors are required to provide the necessary gain for 1.3 watts output power.

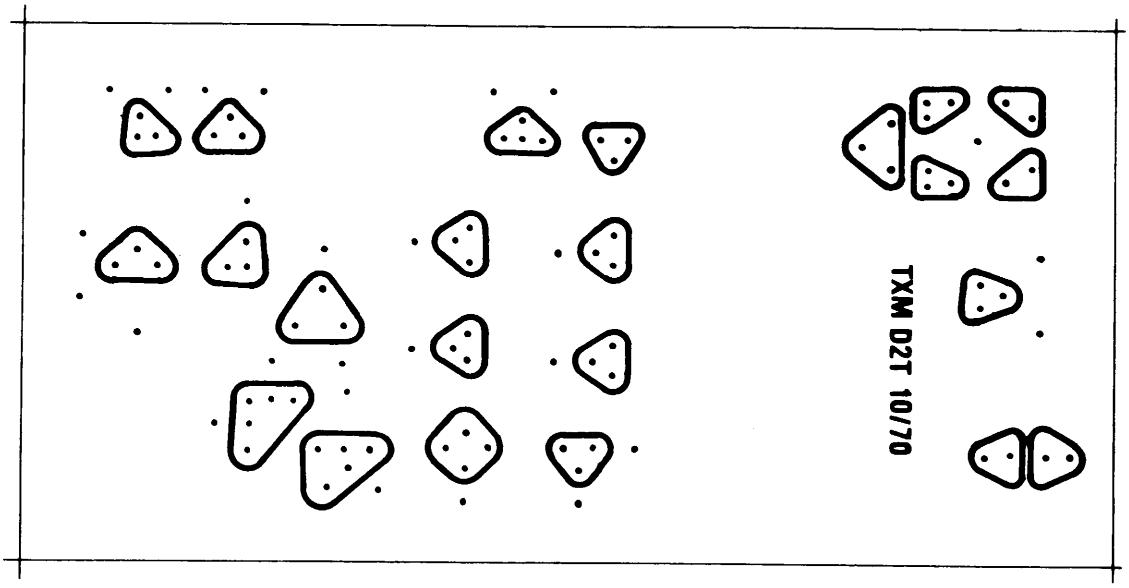
The required injection signal of approximately 20 mW at 404 MHz is available from the module X3T 404 of D.6.1 and the transverter output terminals of modern short wave transceivers can usually supply the necessary 5 to 20 mW.

The mixer is followed by a bandpass filter for the desired signal and a trap (L6 + 6pF) for the now unwanted 404 MHz signal. The design of the two transistor stages is fairly straightforward. T1 and T2 are current stabilized by resistors. The values listed in the table below are thus for continuous operation. Different transistors were measured according to the required output levels. The following list is meant to indicate the possible power levels and the corresponding transistor complements. Additional circuits for higher power levels are treated elsewhere in this book (D.5.2).

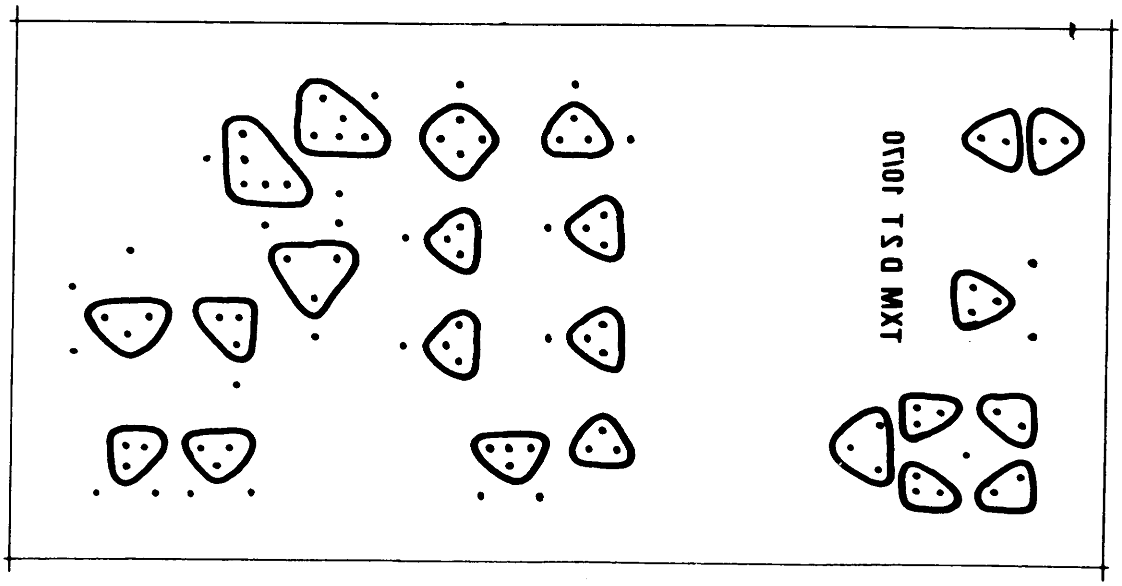
T1:	T2:		output power:		
BFR34A	CHEO	by CTC	700mW		U_B = 12.5 V
BFR 96	C1-12	by CTC	130mW		
BFT65	C1-12		180mW		
BFR90	C1-12		200mW	input level at:	
BFR91	C1-12		240mW	28 MHz = 10 mW	
BFR34A	C1-12		380mW	404 MHz = 20 mW	
BFQ28	C1-12		550mW		
BFR14A	C1-12		610mW		
BFR34A	BFT98	by Siemens	920mW		
BFQ28	BFT98		1000mW		
BFR14A	BFT98		1150mW		
BFR34A	C1-12		500mW	28 MHz = 10 mW, 404 MHz = 70 mW	U_B = 12.5 V
BFR34A	C1-12		600mW	" "	U_B = 14.0 V
BFR14A	C1-12		750mW	" "	
BFR14A	BFT98		1350mW	" "	

The combination BFR34A/CHEO turned out to be quite cost effective. With a supply voltage of 14 volt an output power of nearly 0.9 watt is available. For all other combinations the 28 MHz level may be raised slightly. More than two stages in one case will give rise to spurious oscillation since the power radiated by the second stage into the base of the first transistor is likely to be higher than the drive power delivered by the mixer.
The following figures show the printed board, construction, component location and the spectrum.

Figure 317 a) Printed board underside (photographic exposure side)
b) Top side drill template; mirror image to aid assembly in conjunction with the photograph of the component side.

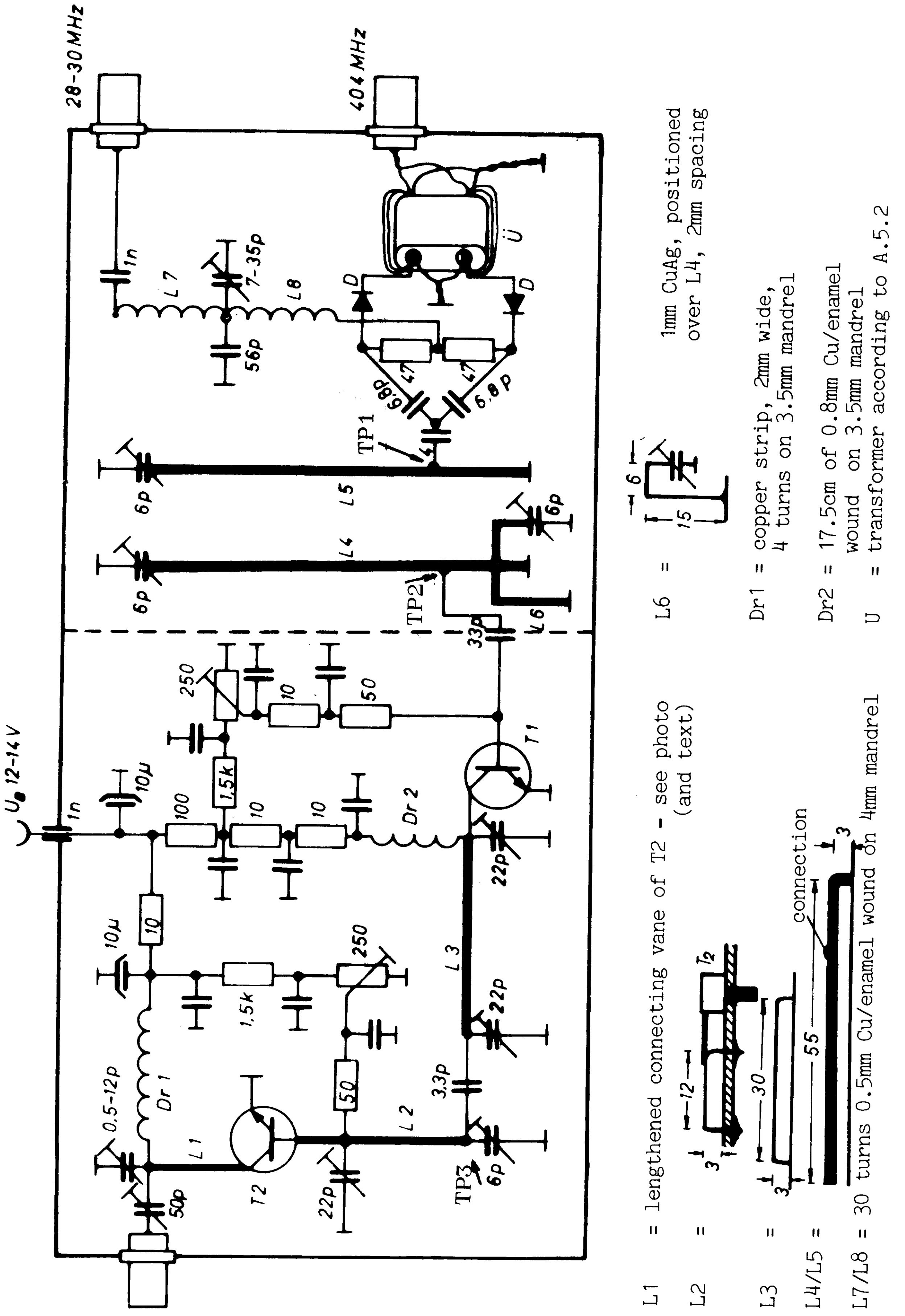


a)



b)

Figure 318 Circuit diagram of transmitter mixer (TXM D2T 10/70)
All resistors in ohm unless specified otherwise. Diodes D are Schottky diodes, in this case HP 5082 or 5082 2995.
2805



Assembly notes: L4 and L5 are mounted inside individual cavities and are thus separated from one another. Coupling is achieved by means of a capacitor (which defines the degree of coupling). A capacitor of 1.8pF was used in this particular case. The capacitor is threaded through a hole of 6mm diameter and both leads are soldered to the hot ends of the resonators.
If the CHEO is employed then
L1 = 2 turns of 1mm CuAg wire wound on a 3.5mm diam. mandrel.
Dr1 = 5.5 turns of 1mm CuAg wire on 3.5mm mandrel (not critical).
L6 is - just like L4 - pushed through a hole drilled to provide a tight fit and soldered from outside. It runs for 15mm parallel with L4 (3mm spacing) towards TP1, crosses L4 at TP1 and is routed back. The length is governed by the dimensions of the trimmer which should be a foil or tubular capacitor. It is either soldered to the side wall or - in the case of the tubular trimmer - bolted to the side wall. The shown third version of the module this ciricuit was skipped since the remaining 30dB of attenuation of the image frequency (404MHz) was found to be acceptable and since an additional filter according to C.3.3 was planned to be inserted between this module and the power amplifier LV2/432 (D.5.2.2). With this combination absolutely nothing of either the image frequency nor the injection signal could be detected at the output of LV2/432 (!).
The collector lead of T2 as well as L1 was soldered to a ceramic mounting post. The trimmer 0.5-6pF (0.5-12pF) provides mechanical rigidity to the other end of L1.
The combination L7/L8 in conjunction with the two capacitors is tuned to the i.f. = 28.5 MHz (maximum output level). Should a trimmer with larger tuning range (4-70pF) be available then the fixed capacitor could be deleted (see photograph). Should you intend to mix 144 + 288MHz to obtain 432MHz then only L7/L8 and the trimmer need to be replaced according to A.5.2. The broadband input circuit (f_0) requires no modification up to 1GHZ.

Figure 319 Overall view of the TXM D2T 10/70 in its case 146x72x30mm. This case is fabricated from tin plate and is commercially available (see G.4).

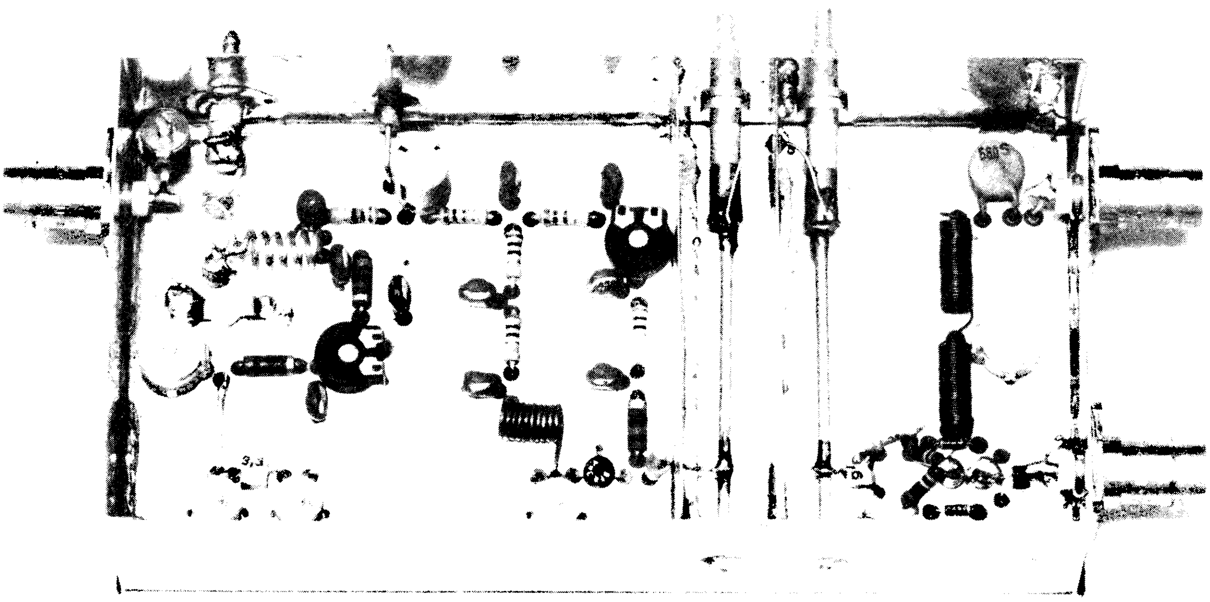
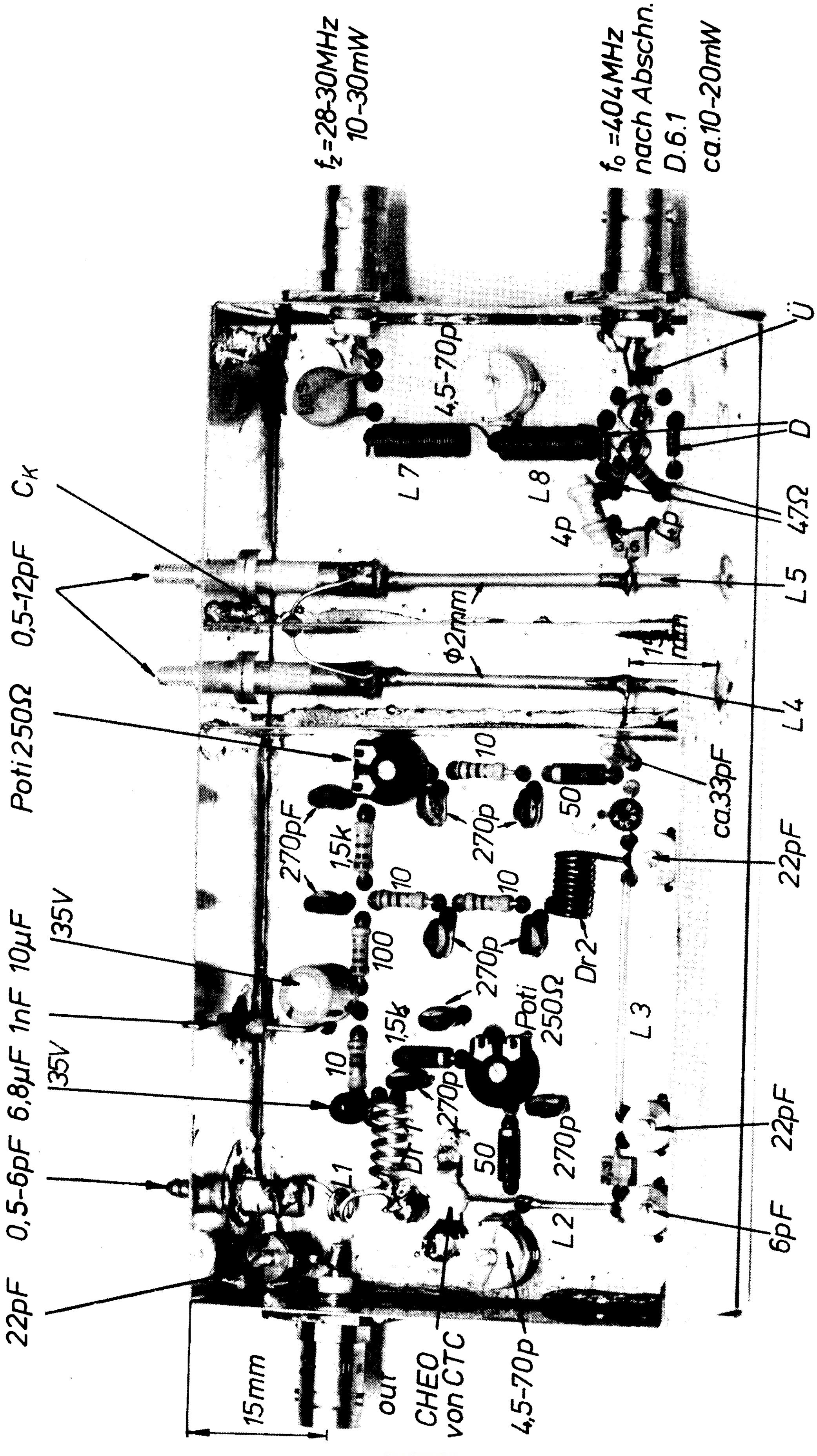


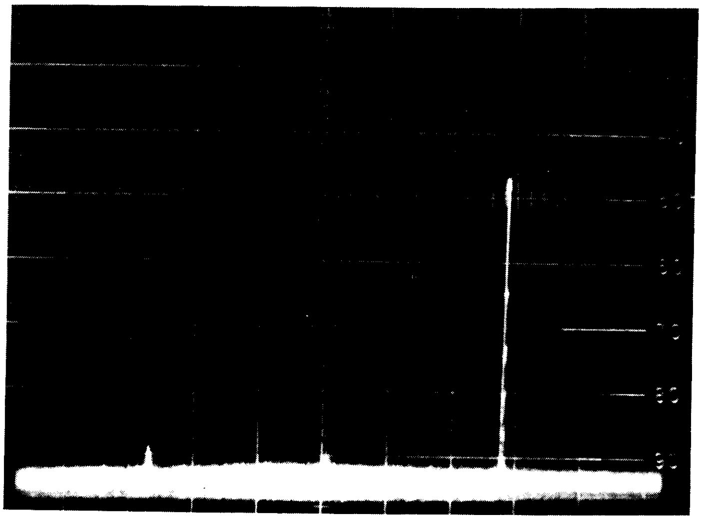
Figure 320 Component location of transmitter mixer TXM D2T 10/70



Alignment:
Initially all trimmers are set as indicated by the photograph. The potentiometer wipers are then turned towards ground. After applying the supply voltage the quiescent currents of T1 and T2 are set to approximately 20mA and 40mA respectively. Next, the two input signals are fed to the module; upon turning the 6pF capacitor to maximum capacitance both the heterodyned and the image signal must be apparent at test point TP1 on L6. Using a coupling loop according to B.1.6 and a mW-meter according to B.2.1 or other means the trimmers belonging to L7/L8 and L4/L5 are adjusted for maximum signal level. A filter - set previously to 432MHz and connected to TP2 - ensures in conjunction with the following mW-meter the correct selection of the desired signal. This could also be achieved by means of an UHF-dipper (see B.1.5). Should none of this be available then the aerila should be connected directly to TP2. Another local station may then report via 2m any field strength variations. This procedure could be repeated with TP3 and the output terminal. Afterwards all tuning elements should be readjusted with the cover in position (through openings provided).

Upon amplification by additional stages it could happen that the (suppressed) injection signal is still present 43dB below the desired signal (no drive signal on 28MHz). The 6pF capacitor and L6 are in that case tuned to give minimum level of this 404MHz signal.

The UHF spectrum of this module is presented in figure 321. It can be seen that both f_O (404MHz) and the image frequency f_O-28MHz (376MHz) are at least 43dB down on the desired signal. This is quite sufficient for normal requirements. Filters according to C.3 should otherwise be employed to remove these products from the spectrum. All other products are attenuated by at least 50dB.



376 MHz 404 MHz 432 MHz

Figure 321

UHF spectrum of the transmitter mixer TXM D2T 10/70 at an output level of 380mW.

f_O = 404 MHz
IF = 28 MHz
f_{sp} = 376 MHz (404 - 2 IF = f_{sp}

h: 10 MHz per division
v: 10 dB per division

Example: If only wideband amplification (bandwidth more than 28MHz) would lift the output to 100 watt then the injection signal component (f_O = 404MHz) would be 5mW. The following selective valve based power amplifier according to D.5 will however suppress f_O and f_{sp} even further.

D.4.2 10m/23cm resp. 2m/23cm transmitter mixer in conventional wiring and stripline technique

It must be pointed out that the mixer described hereafter is equally well suited for both transmitting and receiving purposes. Because of power handling properties and simple construction the choice fell on balanced diode mixers.
Depending on ones possibilities construction could be in either conventional wiring or stripline technique. The conventionally wired version was described in A.5.2 for both 70cm and 23cm waves.

Stripline technique version

Injection of both signals f_o and f_z into the mixer is performed best by a quadrature or ring hybrid arrangement. The difference between the two versions was discussed in A.5.4. The mixer described below was derived from (1) and is realized as $6\lambda/4$ ring hybrid. Since the striplines are frequency dependent they can be optimized for one frequency only. The mixer is, however, fed with two signal frequencies (f_o, f_z) to obtain the desired signal frequency of 1296MHz. The ring hybrid is therefore designed for $1296-1/2f_z$, i.e. 1224MHz for an i.f. of 144MHz. The bandwidth of this arrangement must in this case be 144MHz (12% of 1296MHz). Additional errors from dielectric constant, dimensions etc. may creep in and give rise to incorrect values which should be taken into account by appropriate choice of the bandwidth.

Epoxy was selected as base material since it is cheaper than Teflon while still keeping losses at an acceptable level. The choice of i.f. is governed by:

- the existing receiver
- the image rejection (employing filters) and
- the unwanted mixing products (such as ghost signals as result of wrong choice of frequencies in the injection signal generator).

From the bandwidth point of view a f_z of 28 MHz would be an attractive choice. The image rejection would, however, be inadequate. A good compromise for the 23cm band is an intermediate frequency of 144MHz. The crystal frequency is then 46.666667 MHz (1152MHz devided by 27). Starting with 48MHz (x24) would also result in an injection frequency of 1152Mhz; 48 MHz x 3 will, however, give rise to a strong ghost signal at the lower band limit of the following 144MHz receiver and is thus not recommended. A concept for injection signal generation suitable for transmitting as well as receiving purposes is described in D.6.2.

But now back to the mixer: The hybrid ring is of hexagonal shape after measurements in the USA have shown that this has no disadvantages compared with the circular version. Line impedance is, however, critical. The stripline width of 1.3mm on 1.5mm thick glassfibre/epoxy board (double sided copper cladding 36um) should therefore be carefully maintained see also A.2.4).

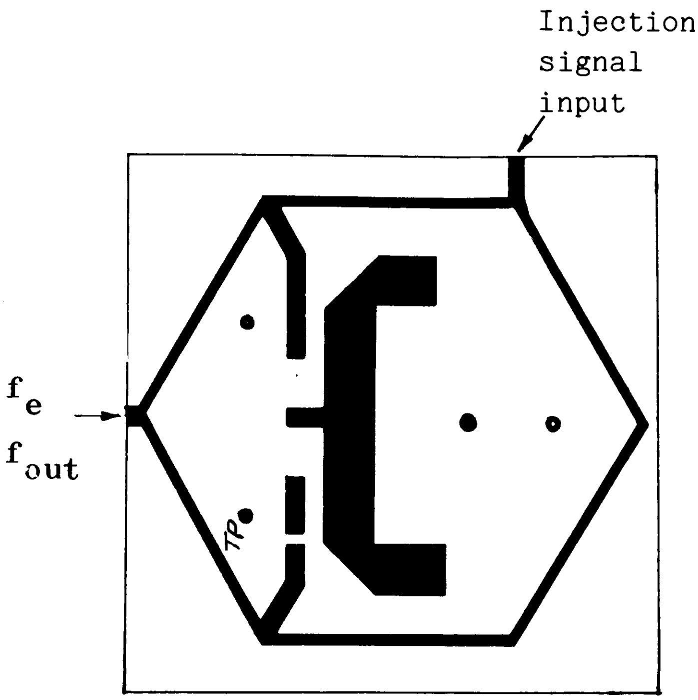
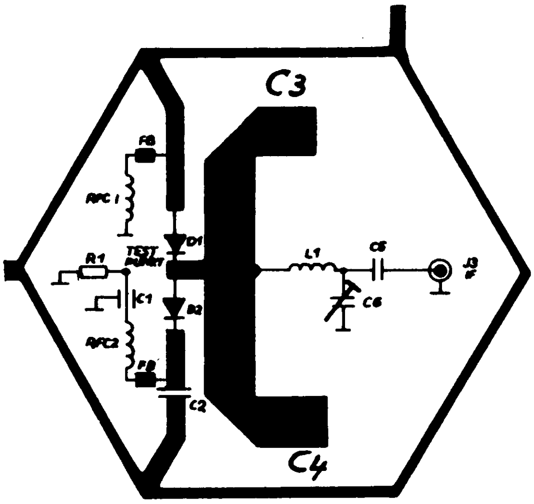


Figure 322
Printed board of ring hybrid mixer. The underside is used as ground foil and must not be etched away.

P_{fo} = 5 to 10mW for both TX and RX applications.

The board may also be used without the internal circuit arrangements to split the power between two outputs. In that case the lines to the diodes should be relocated towards the side walls and should end in coaxial connectors. This possibility is computed for the applicable frequency in B.1.3 and D.5.9 and is discussed in conjunction with building instructions and applications.

Figure 323 shows the circuit diagram superimposed on the printed board.



C1 = 1nF feed through-C

C2 = trapezoidal capacitor 100 to 220 pF

C3 = λ/4 stripline (25 ohm) (open) for 1296 MHz

C4 = λ/4 stripline (25 ohm) (open for 1152 MHz) see C.5 for theory

C5 = 1nF ceramic C

C6 = trimmer 0.5-6pF for F_z = 144 MHz

R1 = 10 ohm

D1/D2 = Schottky diodes for UHF such as: HP 82 2995 034 or (1) HP 5082-2817

L1 = 5 turns CuAg 1mm wound on 4mm mandrel

The 50 ohm termination of the i.f could also be achieved by a fixed attenuator; this would, however, lead to a significant increase of mixer losses. C3 /C4 represent open ended quarter wavelength lines and lead to further decoupling of the various radio frequencies - i.e between f_e resp. f_o and f_z.

Mixer isolation was not measured in this particular case - according to (1) it is around 20dB between 1050MHz and 1300MHz - respectively 33dB at 1152MHz.

The rectified injection signal is available across R1. Depending on the applied level and the selected diodes voltages in the mV range (approximately 10-40mV) may be drawn for test purposes. R1 is soldered at the board underside.

The choke inductance is 0.22nH (should be 0.33nH according to ref.1). FB = ferite bead for UHF (single hole).

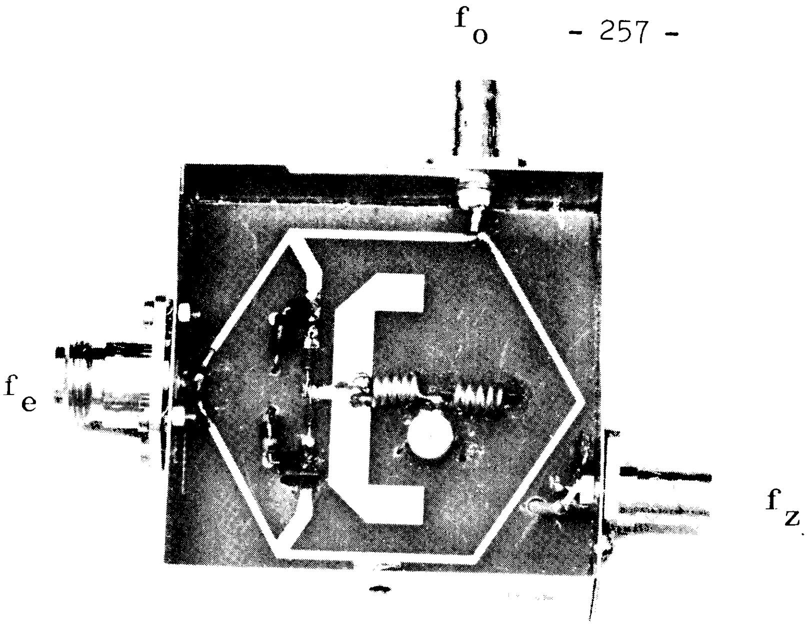


Figure 324 shows the mixer installed in its case.

Case Dimensions: 72x72x30mm

The board underside acts as ground foil and is soldered to the case.

$f_z \triangleq IF$

The coupling capacitor C5 was moved to the i.f.connector and an additional inductor L2 was installed. This gave superior matching. Should the frequency plan of the station call for a different signal generation then only the combination L1/L2 and C6 need to be redesigned for the new i.f. The following values apply to an i.f. of 28MHz: L1/L2 - 30 turns of 0.5mm Cu/enamel wound on a 4mm diameter mandrel. C6 should cover the range 4.5-70pF.

Mixer alignment

is achieved by merely adjusting C6 for maximum signal.
The mixer bandwidth covers the range 950 to 1400MHz.
Pf_o is 5 to 10mW. The mixing loss is -7dB.

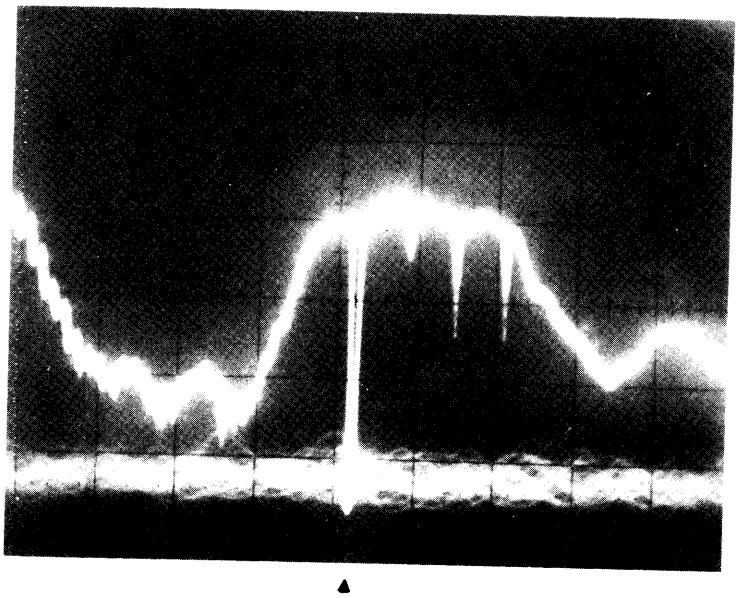


Figure 325 indicates the mixer frequency response.

All negative spikes are frequency markers.

Displayed range:
200 MHz to 2 GHz

Vertical scale:
10 dB per division

Markers: 1000 MHz (centre)
1150 MHz
1250 MHz
1350 MHz

The photograph indicates a fairly high signal transfer at frequencies below 200MHz. This is of no significance since only the 23cm band is entered when receiving. The frequencies for the transmitting case are defined anyhow. Contrary to the mixer in A.5.2 the image frequency (1152-144 = 1008 MHz) is not suppressed in this module. An additional filter behind the mixer is thus compulsory. It is therefore recommended to construct the mixer in conventional wiring with an integrated filter (A.5.2) for transmitting purposes.

Reference: (1) rat-race balanced mixer for 1296 MHz, "hr" July/77

D.4.3 Transverter for the 23cm band (i.f. 144 or 28MHz)

This compact transverter module by DK1FI is installed in a case measuring 146x72x30mm. It includes the transmitter channel, the receiver channel and the double balanced mixer.

The mixer is used for both transmitting and receiving purposes. By switching the supply voltage between the two feed-through capacitors the mixer is connected to either the receiving chain or the transmitting chain.

The output power is 0.5 to 0.7 watt depending on the chosen transistor. The first receiver stage is equipped with a BFQ28. A BB105 is soldered to the collector of the third stage to prevent power from leaking away into the last resonant circuit of the receiver when transmitting. Upon removal of the supply voltage from the receiver the diode capacitance will change. This had been included in the tuning process and will now shift the circuit off resonance. In spite of tight coupling no mixer energy will thus be absorbed by the final resonant circuit of the receiver.

Because of the high gain of the three stages (~25dB) and the corresponding tendency for spurious oscillation it is necessary to secure the cover to the case by several screws. It is recommended to activate and to optimize the stages of both the receiver and the transmitter channels one after the other.

Needless to say this concept may be combined with other modules described in this publication. To give an example, the first and second stage may be skipped and a preamplifier according to D.1.4 could be integrated.

It is furthermore possible to install the first and second stage of the transmitter in a case 72x72x30mm and to utilize them as 200mW driver in a different context. Construction and many details are presented by a series of photographs.

Component list:

Case made from tinplate or brass, dimensions 146x72x30mm with two cover plates.

T1:	BFQ28	R1:	270 ohm	R11,R13,R14:	22 ohm
T2:	BFR34A	R2,R18:	2.7 K	R12:	680 ohm
T3:	BFT65	R3:	470 ohm	R15:	39 ohm
T4:	BFR34A	R4:	180 ohm	R16:	910 ohm
T5:	BFR96	R5,R8	1 K	R17:	100 ohm
T6:	BFT98	R6,R9	330 ohm	R19:	220 ohm
		R7:	220 ohm	P1:	500 ohm
		R10:	5.1 ohm	P2:	100 ohm

C1-C14: 1 nF CHIP — C12-C24: 1 nF feed through capacitor
C25-C26: 27 pF — Trimmer capacitors: 0.5-5 pF VALVO or equivalent
C28-C29: 16 pF CHIP — D: diode 1N4148
mixer diodes: Schottky HP 2900 or better
transformer: Ring-type twin hole core, useful up to 200 MHz
3 times 3 turns of 0.3mm diam. Cu/enamel
All chokes: 3 turns of 0.3mm Cu/enamel wound on 2mm mandrel.

The mixer has already been discussed in section D.4.2 so there is no need for a detailed description. Quiescent currents are adjusted to 30mA (T5) and 60-70mA (T6). An i.f. power level of 11 mW yields 0.55 watt continuous output power with P_{fo} = 60 mW and U_B = 13 V. Higher quality Schottky diodes lead to increased output power.

<u>Figure 326</u> Circuit diagram of 23cm transverter

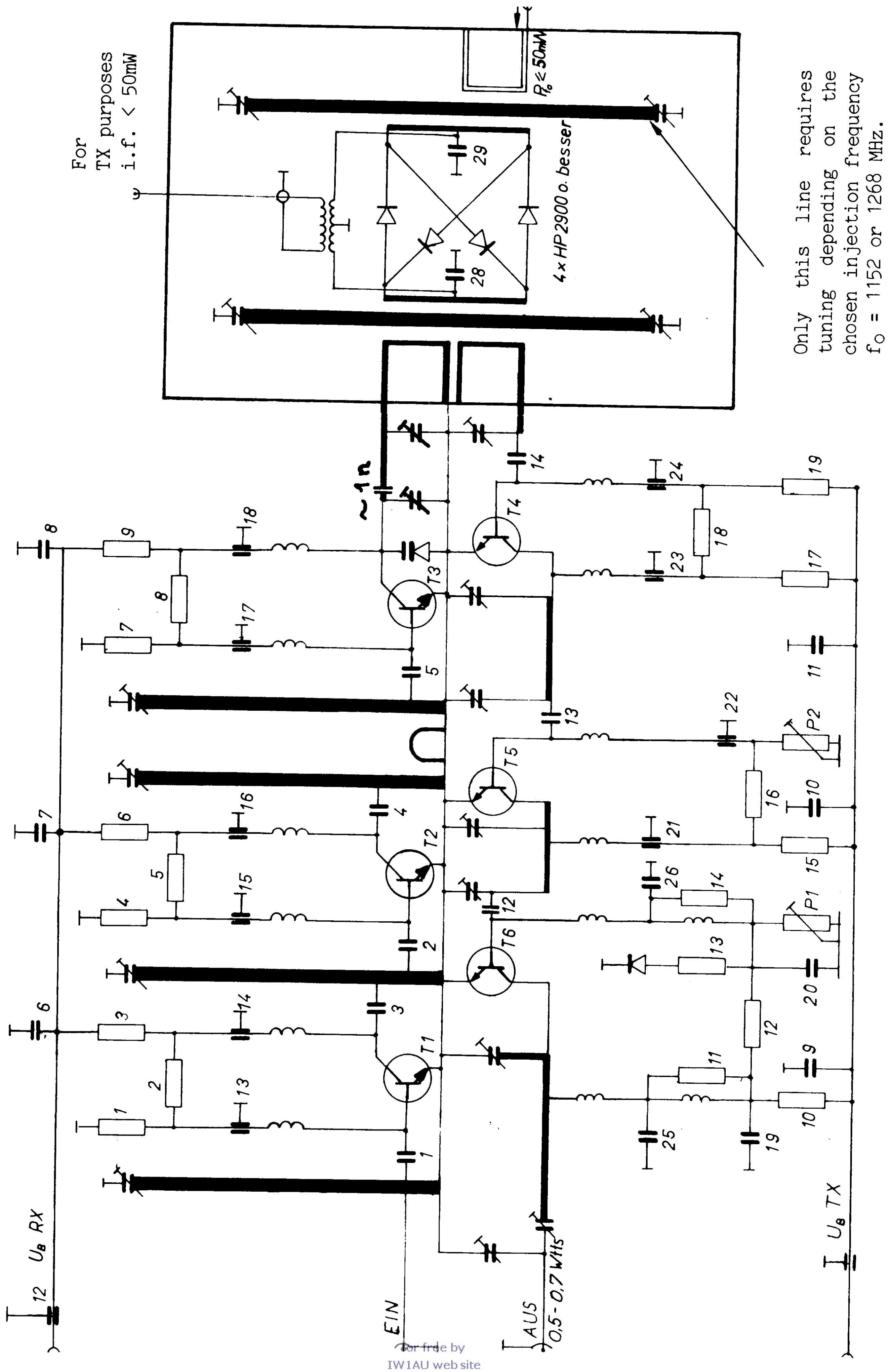


Figure 327 Enlarged presentation of transverter by DK1FI

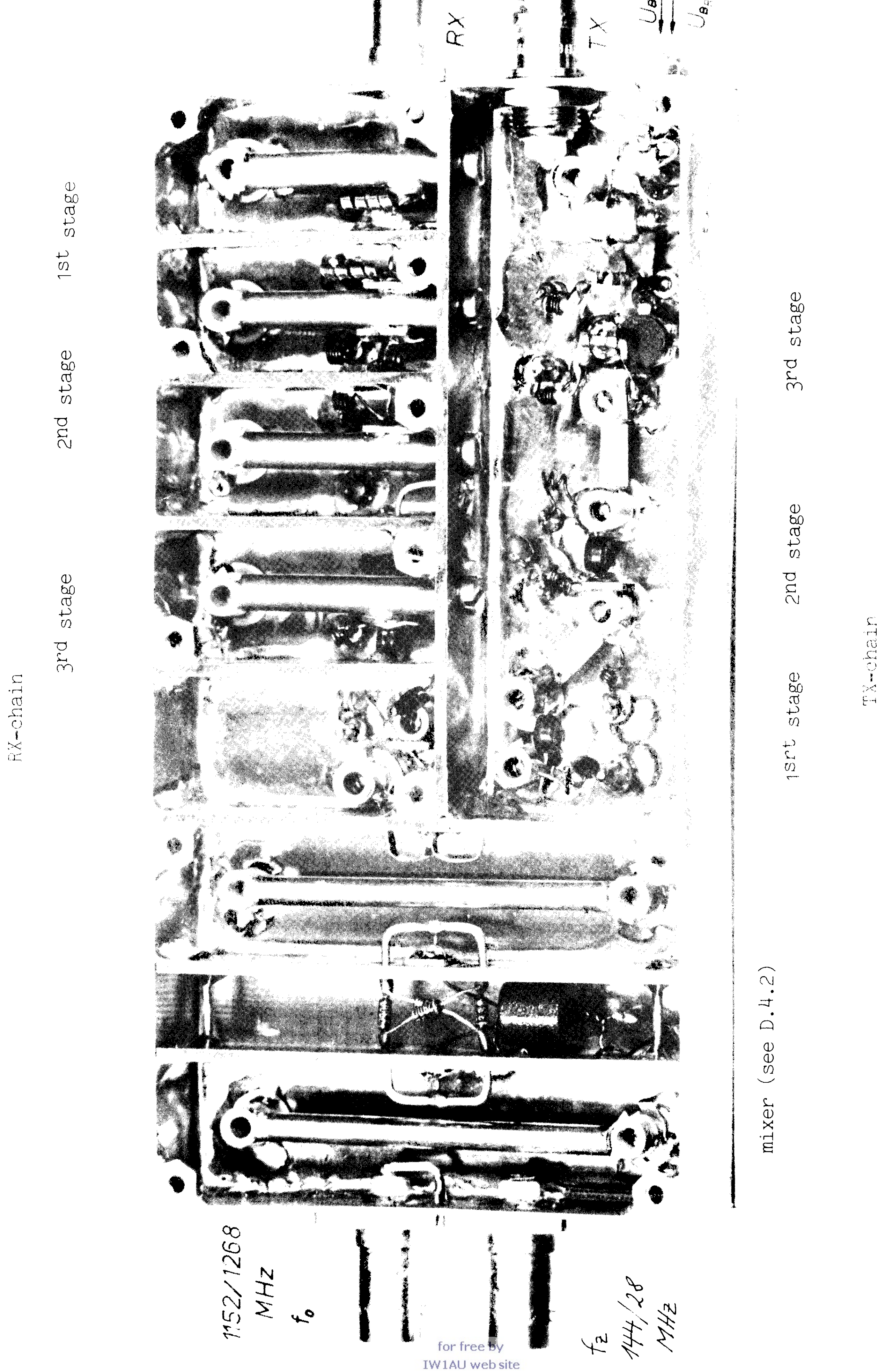


Figure 328 Enlarged partial presentation of mixer and final stages of RX and TX chains.

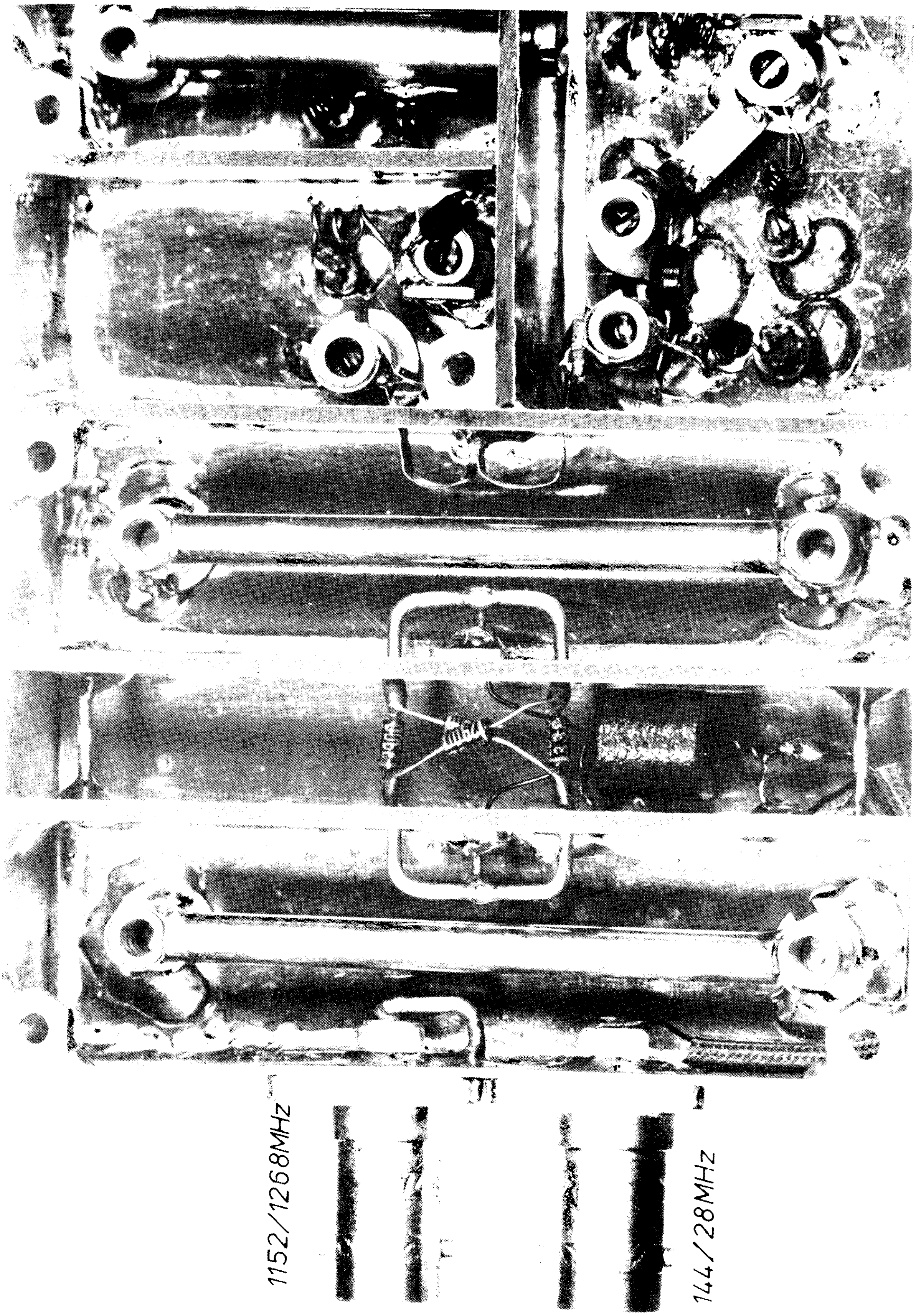


Figure 329 Transverter detail

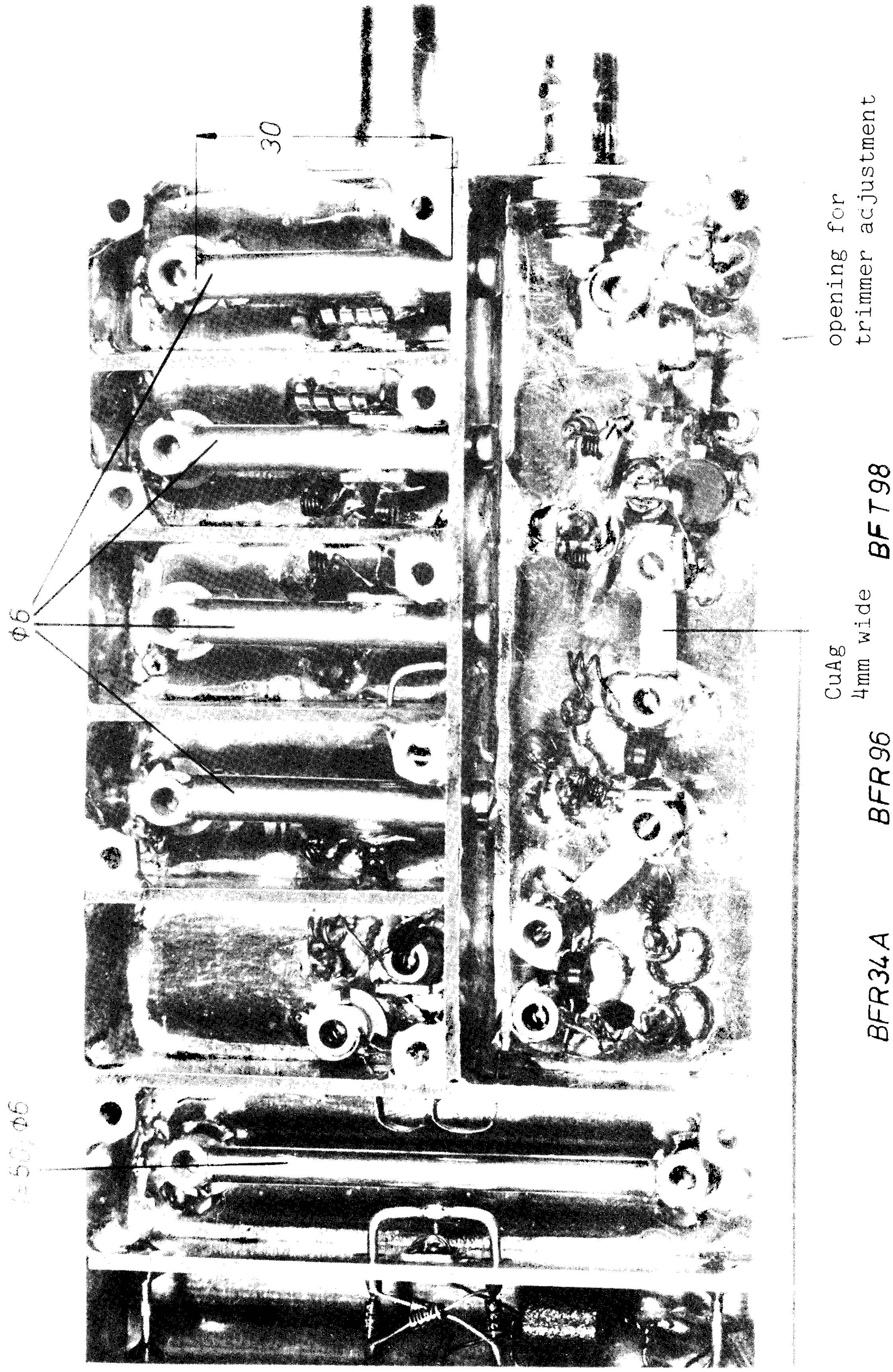
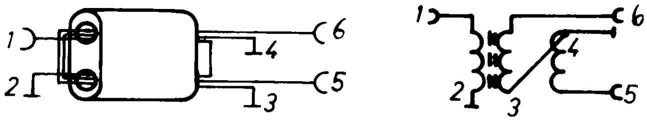


Figure 3 Winding sense of transformer windings. It was eventually discovered that diodes marked HP82 2995 034 in red casing give better results.
For test purposes one should install whatever diodes happen to be around.



3 turns each 2x0.3mm Cu/enamel
1x0.3mm Cu/enamel

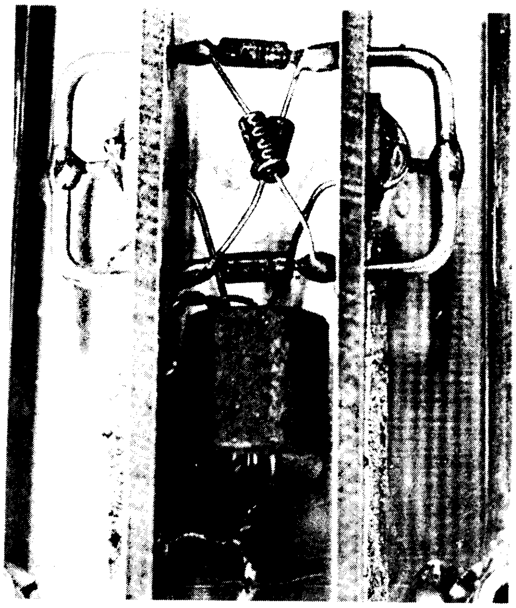
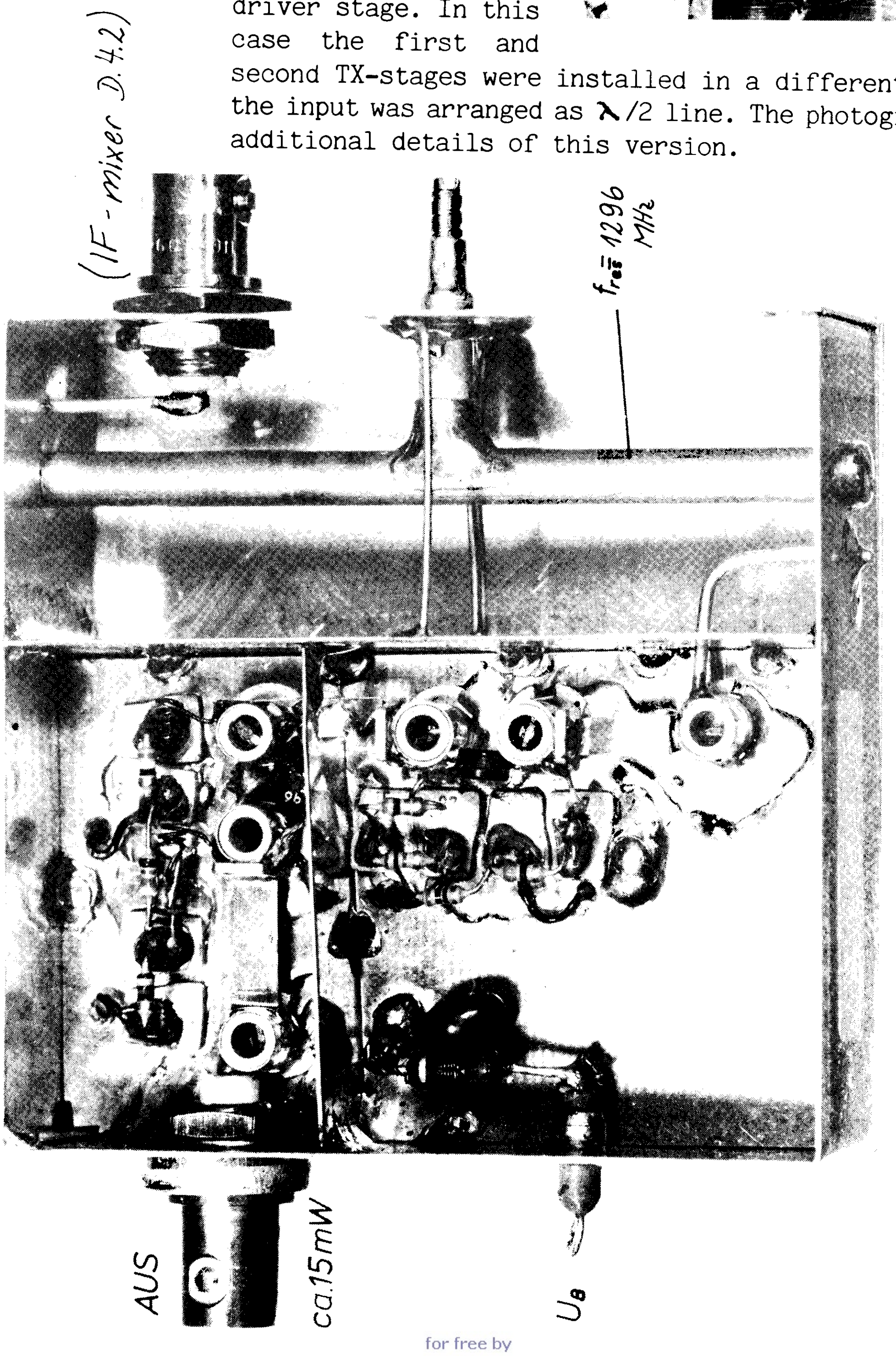


Figure 331 Example of driver stage. In this case the first and second TX-stages were installed in a different case and the input was arranged as $\lambda/2$ line. The photograph shows additional details of this version.



D.5.1 Low power amplifier for 70cm (3 watt, 10 dB gain)
(J. Grimm, DJ6PI)

1) Introduction:
Starting with the transistor impedances Z_{in} = 3-j5 ohm and Z_{out} = 17-j8 ohm a power amplifier for low power output based on the C3-12 was designed. Frequency relevant circuits were realized as printed striplines (as in the more powerful stage by DJ9HO described in the next chapter) to simplify the construction. Since the amplifier had to fit into a 72x72x30mm case all matching networks were designed using the Smith chart. Input matching is achieved through a 12mm long stripline (characteristic impedance Z_o = 33 ohm) and a variable capacitor 4-20 pF. The transistor output impedance is transformed to 50 ohm using a stripline of length 36mm (Z_o = 33 ohm) and a variable capacitor 3-12 pF. Chokes Dr1 and Dr2 were not realized in stripline technique due to space considerations. Details concerning the biasing network and protection against low frequency oscillations were taken from a circuit by SSB-Elektronik (Iserlohn).

Drilling and soldering:
A hole of 9.5mm diameter is drilled into the printed board between the two striplines. A 1mm diameter hole is drilled next to the live end of trimmer potentiometer P1. The position is marked "x" on the component location diagram. The hole is countersunk slightly at the board underside by means of a 6mm drill. The round edges of the emitter contact surfaces of the 9.5mm hole are filed straight and the top and bottom ground surfaces are interconnected by a 6mm wide copper strip.

Figure 332 The copper strips are soldered to both the top and the bottom ground foils.

The connecting leads of all resistors and blocking capacitors are bent and cut as short as possible and soldered in position. The 10nH ready-made chokes are shunted across the 22 ohm resistors. The trapezoidal insulating capacitors are soldered across the gaps in the striplines so that they stand up. The connection vanes of the transistors are cut to 5mm length - after first marking the case at the collector terminal. The C3-12 is then inserted from the top and its connections are quickly soldered to the striplines. The p.c. board is solder connected so far down that the metal block of the transistor rests on the lower cover. The mounting stud protrudes through a hole (4.5mm diameter) in the bottom cover plate and is bolted to the latter. The cover plate thus doubles as heatsink. The board is then inserted into the case and soldered along the edges. The connection leads of the feed-through capacitors are then soldered to the copper tracks on the p.c.board. Next, the bottom cover is once again dismantled and the ground foil soldered to the case. D1 is soldered to the ground surface near the transistor with which it is brought into contact using heat conducting paste. The other end of D2 is inserted into the 1mm hole and soldered to the printed track next to P1 (warning: No short circuit to ground!). The bottom cover plate is re-installed and secured by bolts.

Alignment: Move P1 wiper to ground. Apply 12 volts. 1.4 volts should be present at the "live" end of P1. If not - check polarity of the diodes. Insert a mA - meter into the supply lead and set a current of approximately 50 mA by means of P1. Apply low drive power - gradually going up to a maximum value of 300mW - and adjust C1 and C3 for maximum output power.

Measured data of our sample:

P_{in} [mW]	P_{out} [W]	I_C [mA]
50	0.600	
100	1.5	
200	2.5	
300	3.5	~ 500
500	4.3	~ 700

A C3-12 as used by DJ9HO delivered an output of 2.4 watt with 200mW drive and an input current of 50mA. The gain is thus slightly over 10dB up to 3.5 watt output power. The drive power should not exceed 300mW. The transverter according to D.4.1 is thus sufficiently powerful and actually provides power to spare. In case of higher drive power being available one of the other amplifiers described on the following pages could be used.

Component list:

T	=	C3-12 by CTC
D1,D2	=	1N918, 1N4148
C1,C4	=	trapezoidal C, 100..1000pF
C2	=	4-20pF air, foil, ceramic
C3	=	3-12pF air, foil, ceramic
C5,C9	=	1nF
C6,C10	=	10nF
C7,C11	=	100nF
Dr2	=	4 turns CuAg 0.8mm dam., wound on 4mm mandrel.
Dr3,Dr4	=	10uH ready-made choke, value not critical
C8	=	100pF
DC	=	feed-through-C, 270...1000pF
P1	=	trimmer pot. 50ohm
R1,R2	=	22 ohm
R3	=	390 ohm
Dr1	=	6 turns Cu/enamel or CuAg 0.5mm diam. wound on 4mm mandrel

Figure 333 Circuit diagram of 3.5 watt amplifier with C3-12 (CTC)

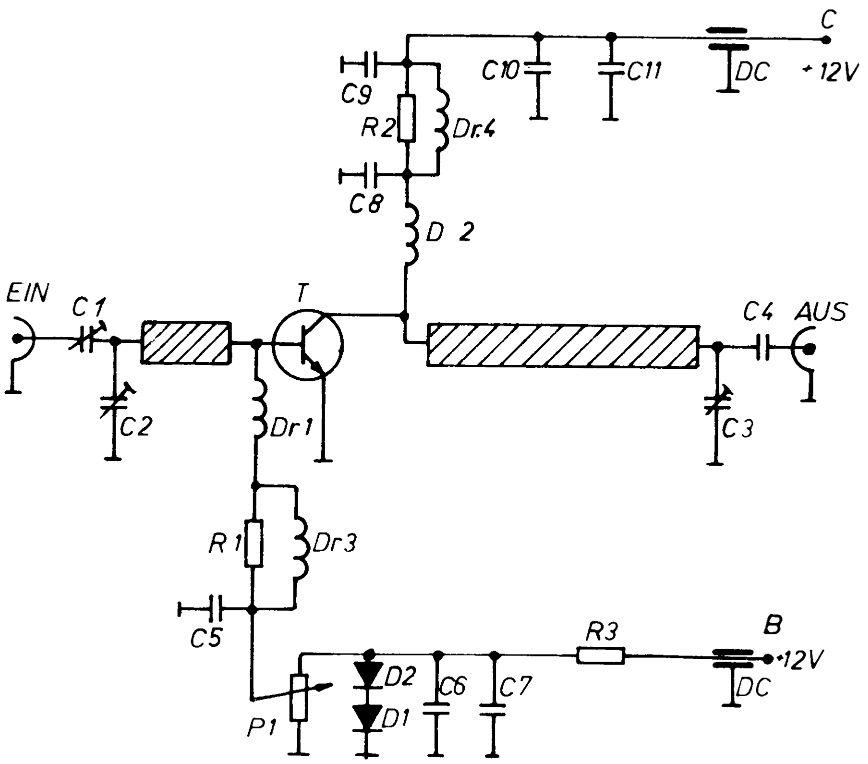
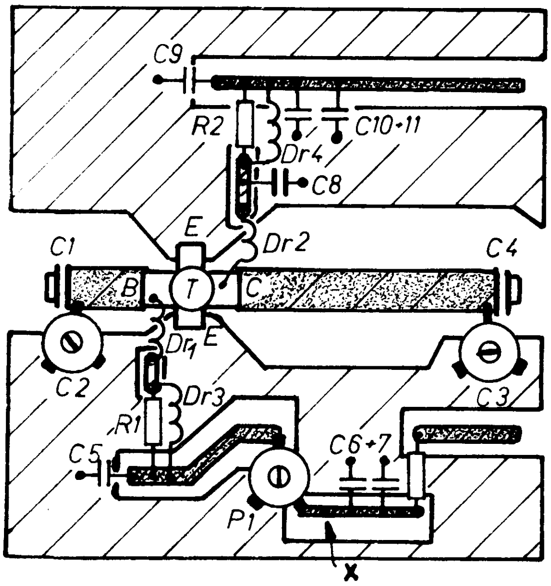


Figure 334
Layout of 3.5 watt amplifier for 70 cm using the C3-12 transistor by CTC. The second connecting lead of D2 is inserted from below and soldered to the copper track at position "x".



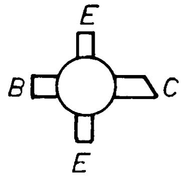
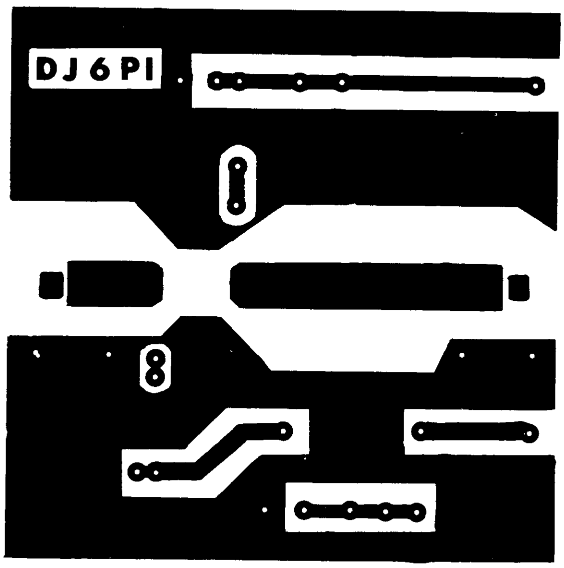


Figure 335
Top side of p.c.board. The underside of the double sided copper clad epoxy board remains as ground surface for the circuits. Apart from D2 and D1 all components are soldered to the top side and are not threaded through the board. The white dots merely indicate the positions to which the components are to be installed. The board is not drilled.



Scale:

Actual size

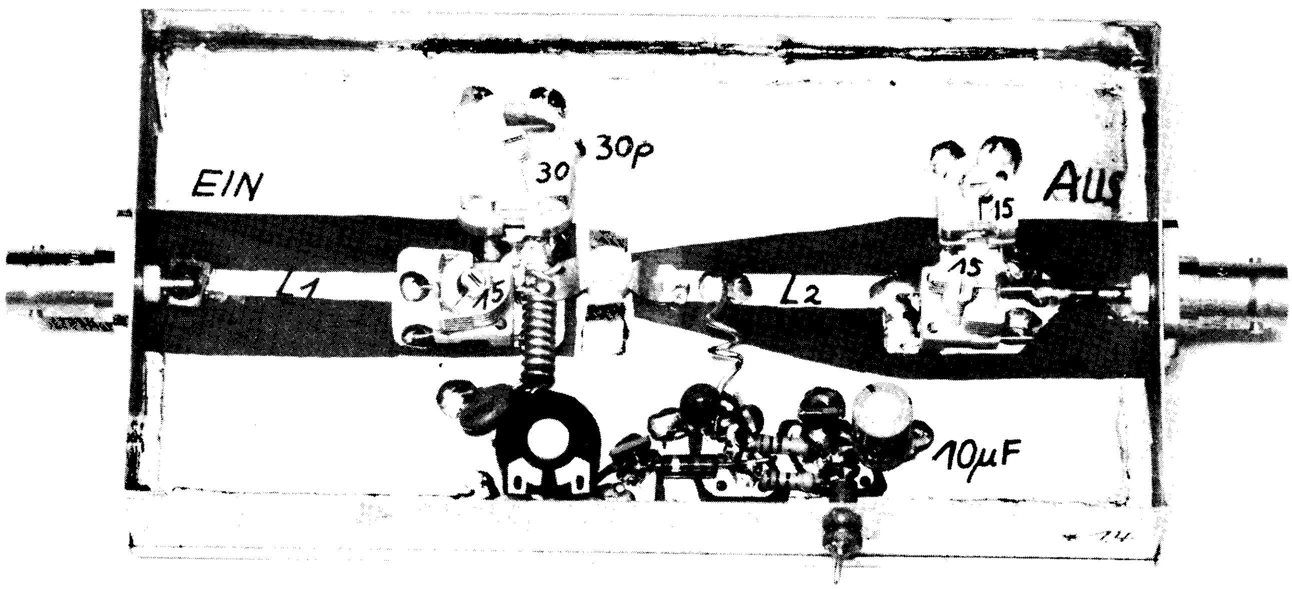
D.5.2.1 Transistorized driver LV 1/432 (printed line circuits)

There is a wide spectrum of applications for transistorized single stage power amplifiers. The broad band circit described hereafter may be equipped with different transistor types to suit the available driver and the desired output power. The r.f. is routed to the base of T1 via a series tuned circuit. The two trimmer capacitors (15pF, 30pF) allow good matching. If the 30pF trimmer is replaced by one of smaller physical dimensions it should have a maximum capacitance of approximately 50pF. The 250 ohm potentiometer is used to set a quiescent current of 50 mA at 12 volts. As shown by the photograph, all solder lugs and components are installed at the foil side of the single sided copper clad p.c.board. Series resistor (R_V) and electrolytics are incorporated to protect against voltage surges resulting from short circuits (destruction of T1). The total resistance (R_{Vges}) should be 10 ohms at 10 watts output and 5 ohms at a desired output level of 3.5 Watts. This measure does not provide maximum but rather reliable output (approx. 0.8 dB reduction). Anyone with plenty of funds is invited to investigate the performance limits of the transistors installed in the unit (hi)...

The prototype (photograph) was equipped with a transistor C5-12 by CTC.

Adequate cooling is provided by a 1mm thick and 30mm wide brass strip which is installed across the underside of the p.c.board and soldered to the sides of the case. The heat generated by the transistor which is bolted to this brass strip is thus transferred to the case.

Figure 336 Construction and component layout of the LV1/432 in its case measuring 146x72x30mm.



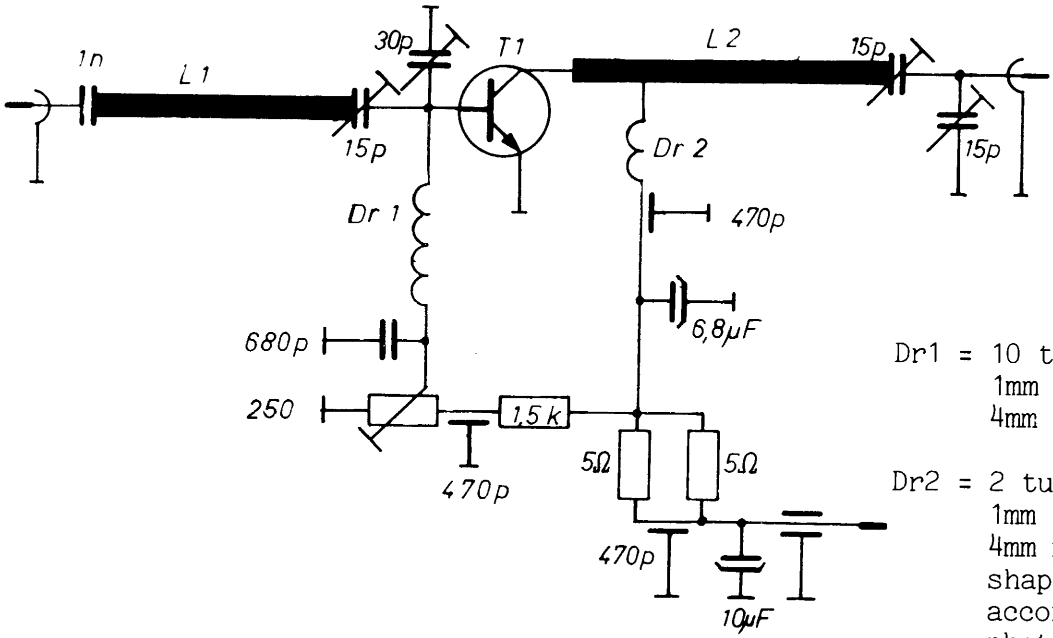
Measured data of power amplifier LV1/432 in single-tone operation

Supply voltage = 13.5 volt, quiescent current = 50 mA

P_{in} [watt]	P_{out} with R_V [watt]	P_{out} w/o R_V [watt]	gain [dB]	I_{max} [ampere]
0.5	2.5	2.8	7 / 7.5	0.36 / 0.4
1	3.5	4.2	5.45 / 6.25	0.45 / 0.5
2	5.0	6.5	4 / 5.15	0.75 / 0.9

The 1dB compression point (see section A.4.3) lies at 4.0 watt output power - corresponding to 0.9 watt input power. The drive power should not exceed 1 watt. Should only 0.5 watts be available then the C3-12 could be employed. Underrating (insufficient power handling capability) will, however, lead to distortion and should be avoided in the interest of spectral purity of the signal. Increased power is supplied by the LV2/432 according to D.5.2.2.

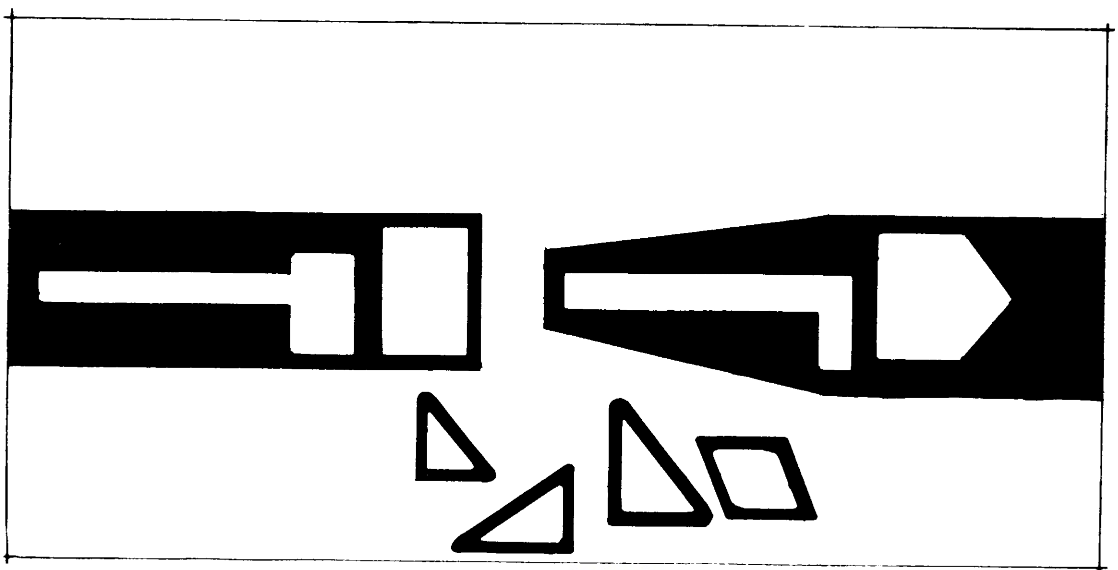
Figure 337 Circuit diagram of LV1/432. L1/L2 are defined by the p.c.board.



Dr1 = 10 turns of 1mm CuAg on 4mm mandrel

Dr2 = 2 turns of 1mm CuAg on 4mm mandrel shaped according to photograph.

Figure 338 Printed board of LV1/432. All components and soldering posts are installed on the p.c.board. Black areas to be removed by etching. P.c.board dimensions 145x71mm.



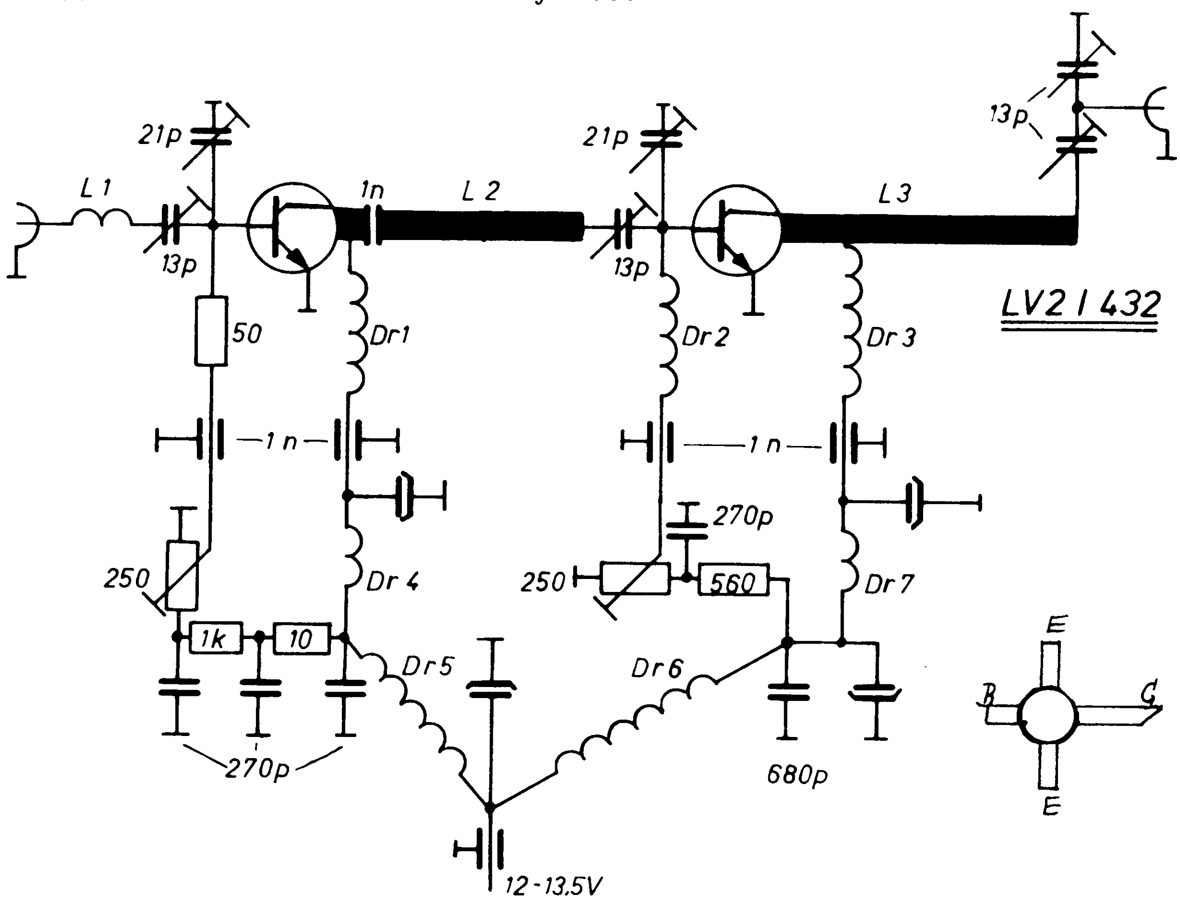
D.5.2.2 Transistorized power amplifier LV2/432

Following the advice "never put more than two stages in one case" an amplifier was designed which

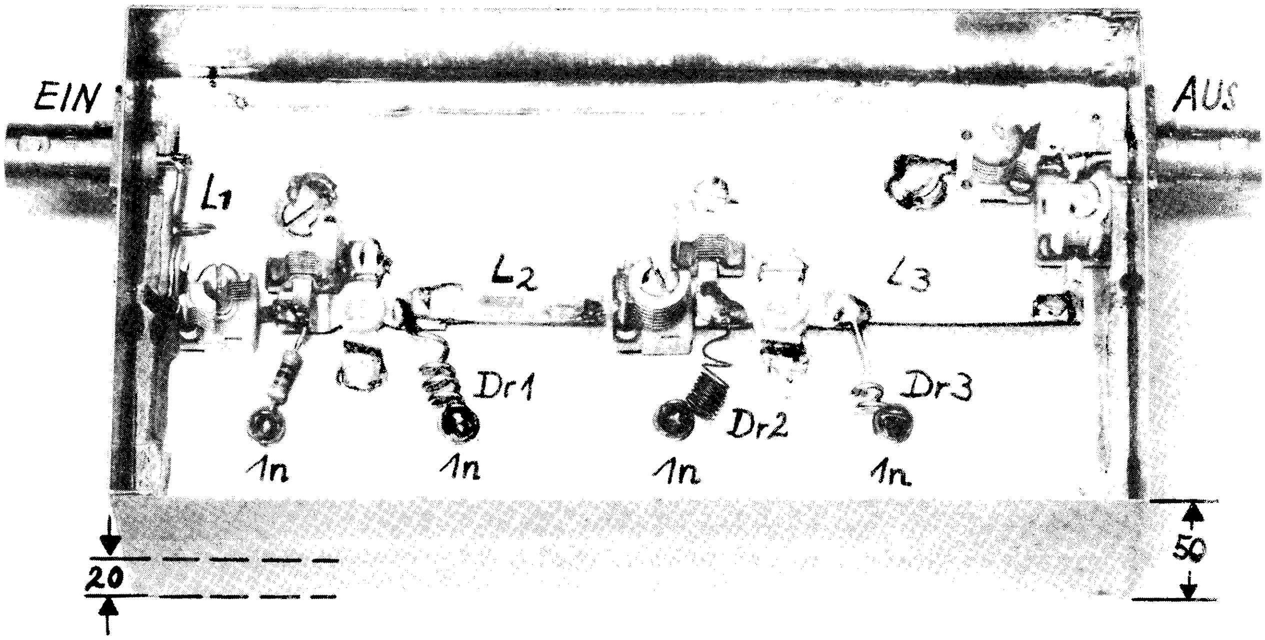
1. could be used behind the TXM D2T 10/70 (section D.4.1) and
2. reliably deliver 8 to 10 watts (1dB compression point),
3. could be used to drive a p.a. with 2 x 2C39,
4. is simple to build while still operating reliably and
5. can be built using standard and fair-priced transistors.

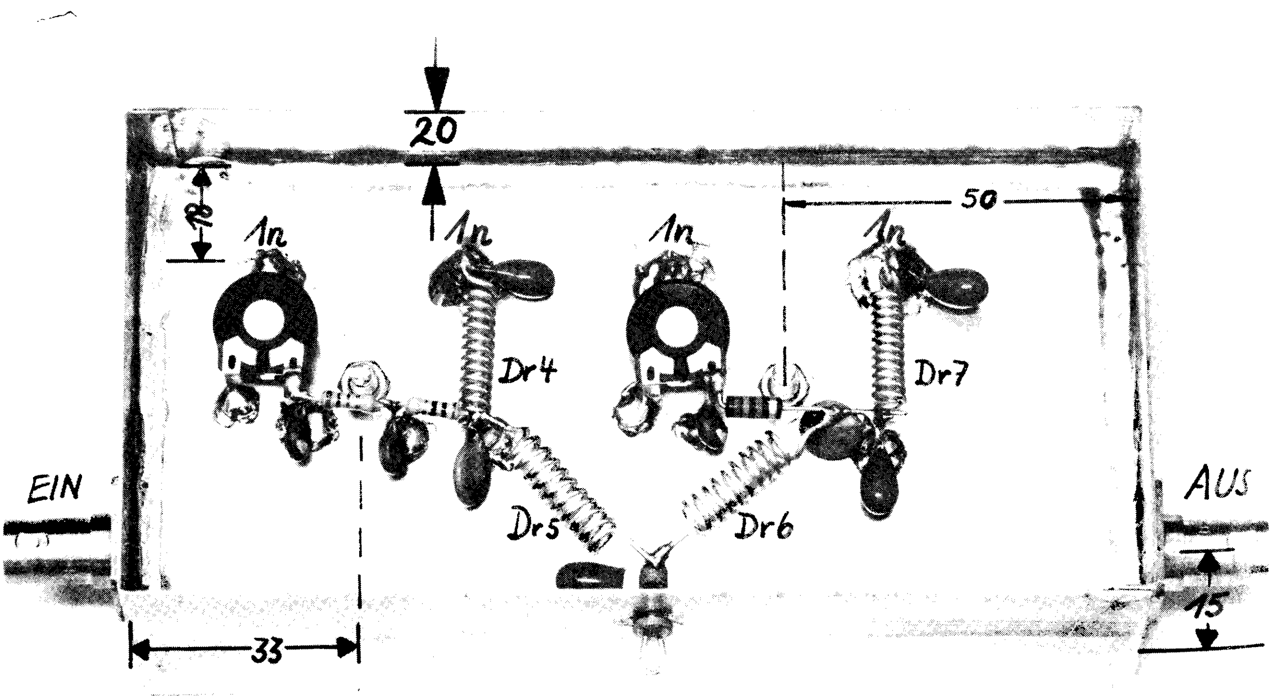
The conventional wiring technique was chosen in view of simple construction and optimum power utilization. The intermediate floor (brass sheet, dimensions 145x71x1mm) is not just a stable ground surface (which is always necessary for stable operation of a circuit) but also a heatsink that absorbs the heat energy of the transistors mounted to it on studs and transfers this heat to the case. This measure results in significantly increased cooling surface. After a test run of 30 minutes in continuous operation at 8 watts the transistor quiescent currents (I_R) had not changed significantly. No further measures were thus taken to stabilize against thermal runaway of I_R. The case was merely warm to the touch.

Figure 339
Circuit diagram of transistorized power amplifier LV2/432. All electrolytics are 6.8uF (not critical) at 35 volts. Our sample is equipped with tantalum-electrolytics.



Figures 340/341 Top side and underside of power amplifier LV2/432 in its case measuring 146x72x50mm. The clearly visible positions of the variable capacitors refer to f_{res} = 432 MHz.





Component data:
L1 = 1 turn CuAg 1mm wound on 5mm mandrel, length 20mm
L2 = CuAg strip 20 x 4 x 0.5 mm, 5mm above ground surface
L3 = CuAg strip 35 x 4 x 0.5 mm, 5mm above ground surface
Dr1 = 4 turns CuAg 1mm wound on 4mm mandrel, shaped as shown on the photograph.
Dr3 = 3 turns CuAg 1mm wound on 4mm mandrel, length 10mm
Dr4/7 = 12 turns CuAg 1mm wound on 4mm mandrel, length 25mm
Dr5/6 = 10 turns CuAg 1mm wound on 5mm mandrel, length 30mm
Dr2 = 17.5 cm Cu/enamel wound on 4mm mandrel, the first two turns are stretched - see photograph.

Circuit description: The r.f.signal is routed via L1 and the capacitive voltage divider to the base of the C5-12, amplified by T1 and passed on to the final stage via yet another series tuned circuit. Matching of T2 is achieved similarly. The series tuned circuit followed by the capacitive voltage divider provides optimum matching to the 50 ohm load. The 50 ohm resistor at the base of T1 does not only provide correct termination for all spurious signals but will also reduce feedback to the base of T1 (feedback of the amplified signal from the output of T2).
This is one reason. The V-shaped arrangement of the chokes (Dr1-3) the second - and the installation of only two stages in the case, the third reason for the complete lack of spurious oscillation tendencies of the power amplifier. Only air insulated variable capacitors should be employed for optimum power utilization.

Alignment:
1. Adjust all trimmer capacitors to the positions shown on the photograph.
2. Move all potentiometer wipers towards ground potential.
3. Apply supply voltage and carefully set the quiescent currents (I_R); T1 = 50mA, T2 = approx. 60mA.
4. Apply P_{in} of approximately 100mW and optimize output.
5. Increase P_{in} to a maximum value of 1 watt and perform fine adjustment.

The quiescent currents must not be altered any more, stick to these values!

Measured data: At U_B = 13.5 volts, I_R = 150 mA overall value including base bias voltage divider, single-tone operation.

P_{in}	P_{out}	gain	I_{max}
0.5 W	9.5 W	12.75 dB	1.4 A
1.0 W	13.0 W	11.15 dB	2.0 A
2.0 W	14.1 W	8.5 dB	2.3 A

Do not exceed 0.8 watts input power (1 dB compression point) to avoid distortion of the otherwise narrow signal with this type of transistor which was designed for FM and is basically non- linear.

D.5.9 Increased power from power amplifier combinations

An output power of 40 to 60 watts may be obtained from a transverter with a TX chain consisting of a driver and a 2C39 final. This power level is quite adequate for DX from favourable locations. Once this power is mastered (no TVI etc.) one tends to develop a certain urge for higher levels. Satisfying this urge usually implies a significant effort in material (power supply, voltages, valves etc.). The following arrangement proves that combining simple-to-build 2C39 power amplifiers can also lead to respectable power levels. Since two voltages (U_A and heater) and two 2C39 power amplifiers are all that is required to generate 100 watts this might well be the perfect solution. The mentioned power output will suffice for all applications apart from EME.

Description: The arrangement is based on the four-port hybrid - the theory of which was discussed in section B.1.3. An incoming signal of - say - 10 watts is divided evenly between the two power amplifiers by means of the power splitter (0^o coupling). Each power amplifier is thus driven by 5 watts and will deliver 50 watts output power. The two output components are then combined by another four-port hybrid composed of coaxial cable sections to add up to the total output of 100 watts. Should one of the power amplifiers fail then the power components at output C (termination) will no longer cancel. Instead, half the power of one stage (i.e. 25 watts in this case) will be fed to the 50 ohm resistor. The latter should therefore have a rating corresponding to 50% of the output of one amplifier. If an arrangement according to OM Reichel (DC9NL) is connected to the output of one power amplifier and if the other amplifier is disconnected then a $\lambda/2$ coaxial line (short circuited at the end) will act as an acceptor circuit. The power delivered by the second valve is consequently completely absorbed by the cable network. Apart from that the material-saving arrangements are well suited.

Figure 387 shows the combination of two power amplifiers. The modern amplifier of section D.5.3.7 should be employed on 70cm because of its inherent advantages (improved cooling and input coupling). See D.8 for a 12 volt mobile power supply.

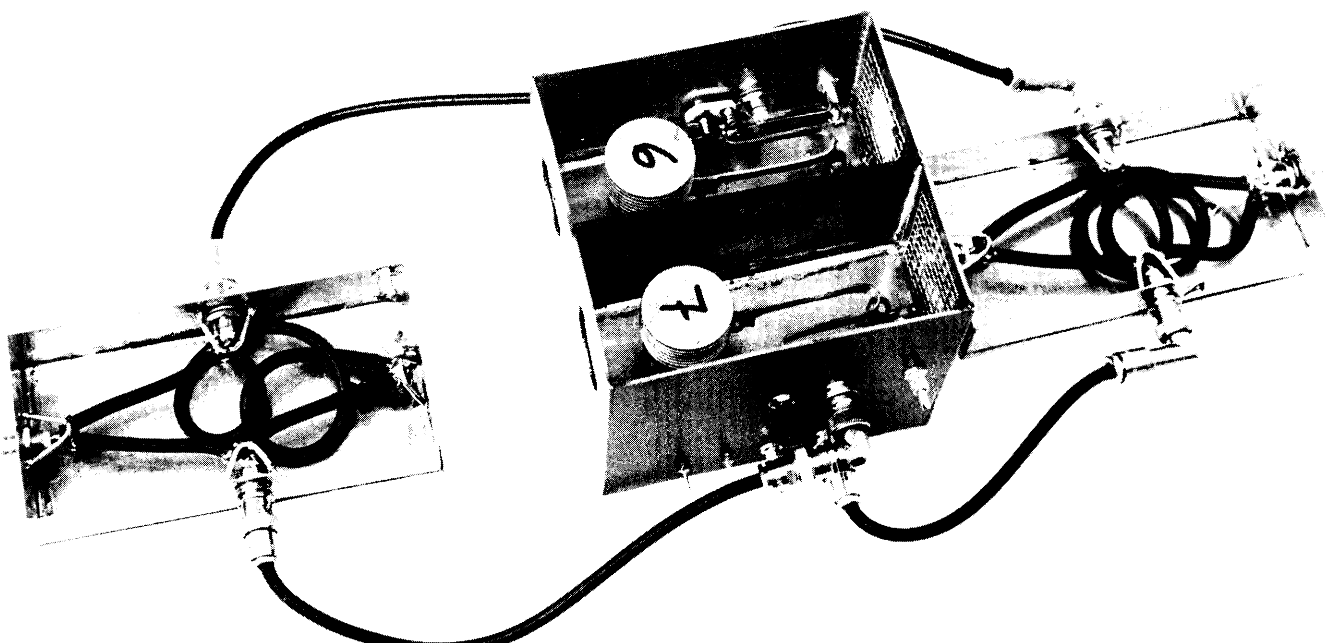


Figure 388 Combination of four amplifiers. Overall gain [illegible] The final hybrid ring should be made from thicker coaxial cable, such as 12mm diameter, 75 ohm impedance t.v. quality.

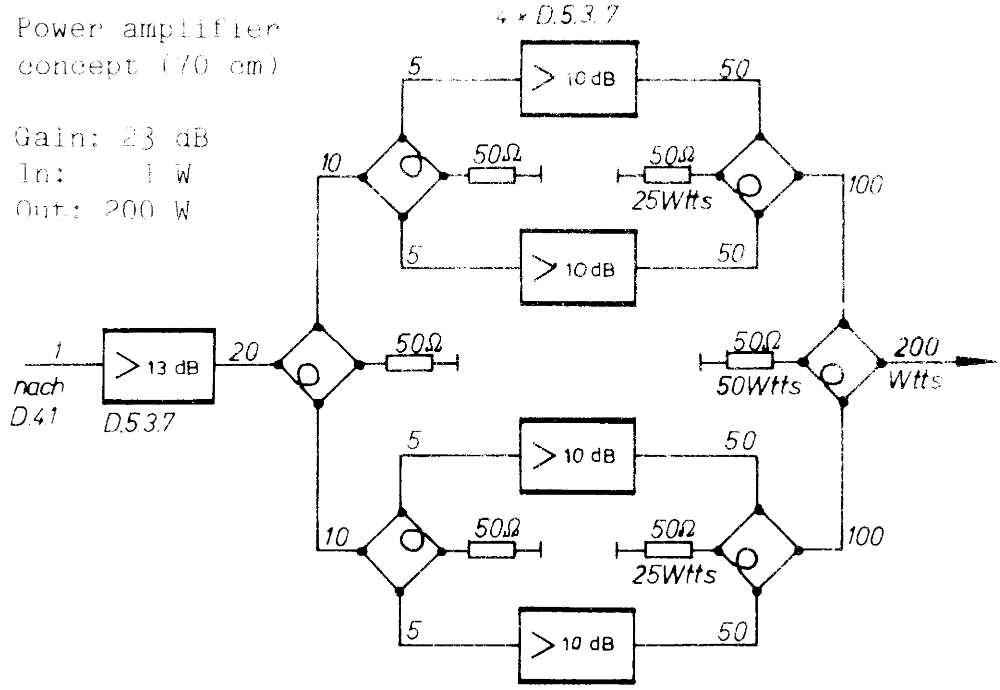
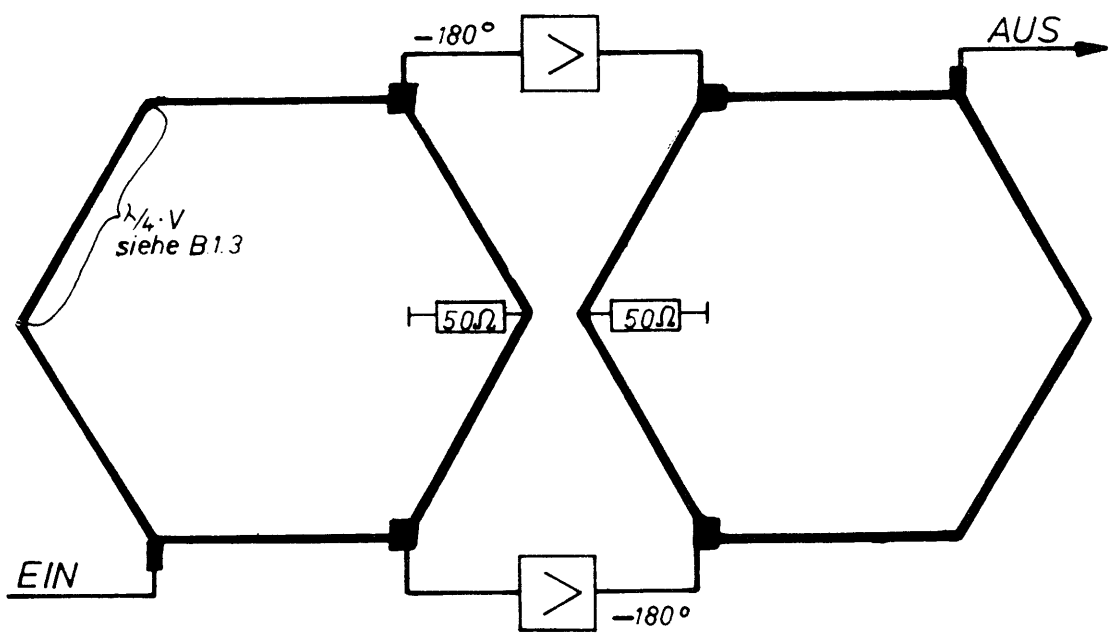


Figure 389 As mentioned before, power adders should not be composed of coaxial cable sections on 23cm and above. The printed version (180° delay) is well suited for higher frequencies. On 23cm the $\lambda/4$ sections are approximately 30mm long for epoxy material of 15mm thickness. See B.1.3 and A.2.4 for theory of operation. It is thus possible to construct a portable transistorized station of considerable power. Up to twice 20 watt may be combined by this arrangement. All-metal versions (E.3.1) should be employed at higher power levels.

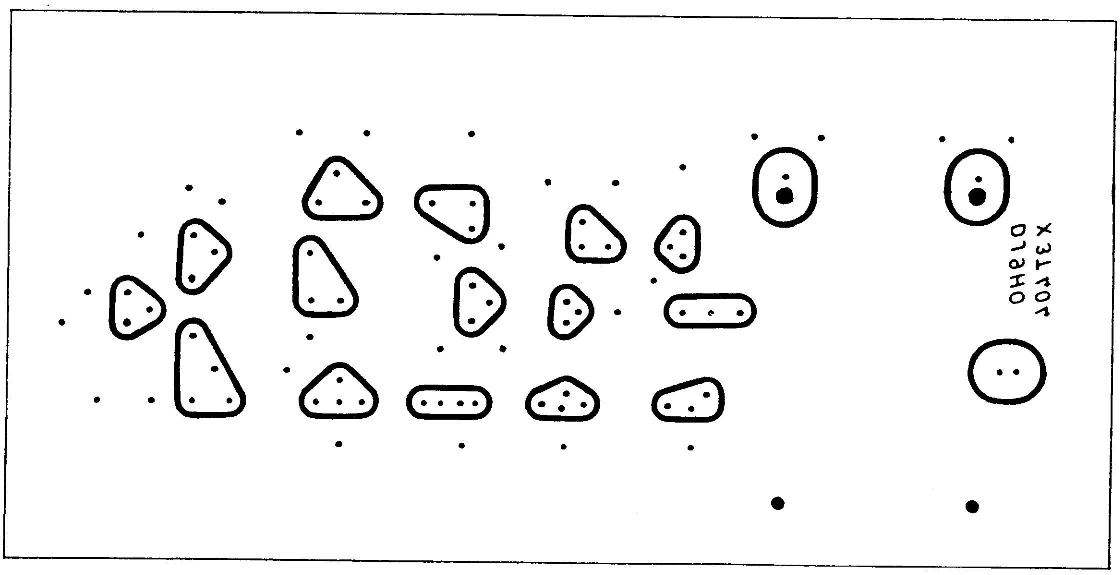


D.6.1 A 404-MHz injection signal generator (X3T 404)

This easy-to-build and reliably operating module is equipped with three modern, high transconductance NPN-silicon-planar-transistors SIEMENS BFW92 (f_T = 1.6GHz). The crystal oscillator (at 44.888889MHz) is followed by two triplers that multiply and amplify the oscillator signal to 404MHz and 20mW. Bandfilter coupling supresses the ever present harmonics of the crystal frequency (f_Q) by approximately 40dB. This means that for instance the 808MHz signal - being the dominating harmonic - is attenuated by factor 10000, i.e. 2µW. This value (40dB) is often specified for carrier suppression in SSB apparatus. A 2.5pF capacitor delivers 3mW to a second connector for supplying a converter with the injection signal. The module may be employed as a beacon transmitter (48 - 144 - 432MHz) without further modifications. It may be employed for receiver tuning in conjunction with an attenuator (see B.1). Varactor multiplication leads to further applications since a BFT65 in the final stage is capable of delivering 40mW.

The publication of a circuit diagram was not judged to be necessary since the two large photographs will answer any questions concerning component location. The p.c.board diagram must be viewed from top. In conjunction with the two photographs it will aid the construction. The board dimensions (30x72x146mm) were standardized to fit a commercial case, made from tin plate, with upper and lower covers (G.4).

Figure 390



Theory of operation:

T1 and the crystal constitute the oscillator. The signal is coupled from C2/L1 to C3/L2; C1 - in parallel with R (≈150 ohm) - is placed in the emitter circuit and controls both the conditions for oscillation and to some extent the frequency. C1 should be a foil trimmer capacitor (~3-22pF). Using a frequency counter the injection signal may be adjusted to exactly 404.000MHz. C4 transfers the signal to the first tripler (T2). The 134.667MHz signal is then coupled via L3/C8 and L4/C10 as well as C11 to the base of T3. The 404MHz signal appearing in the collector circuit is filtered out by L5/C15 and L6/C16 such that all spurious products are at least 40dB down. Figure 391 shows the spectrum of the module between 1 and 1000MHz. It was obtained and optimized using a spectrum analyzer. Power is supplied via a 20 ohm resistor from 10 to 13 volt. The sole purpose of the Zenerdiode (13-16V) is to protect the module against voltage surges resulting from short circuits during installation etc. since the transistors do not thrive on that sort of treatment! The current consumption is 50mA. Should the meter indicate more than 100mA you can be certain that at least one transistor has sadly passed away.

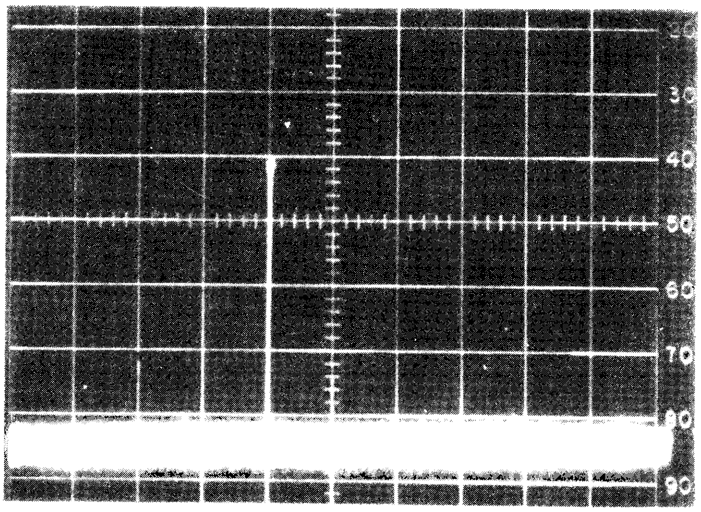


Figure 391
r.f. spectrum
1 - 1000 MHz

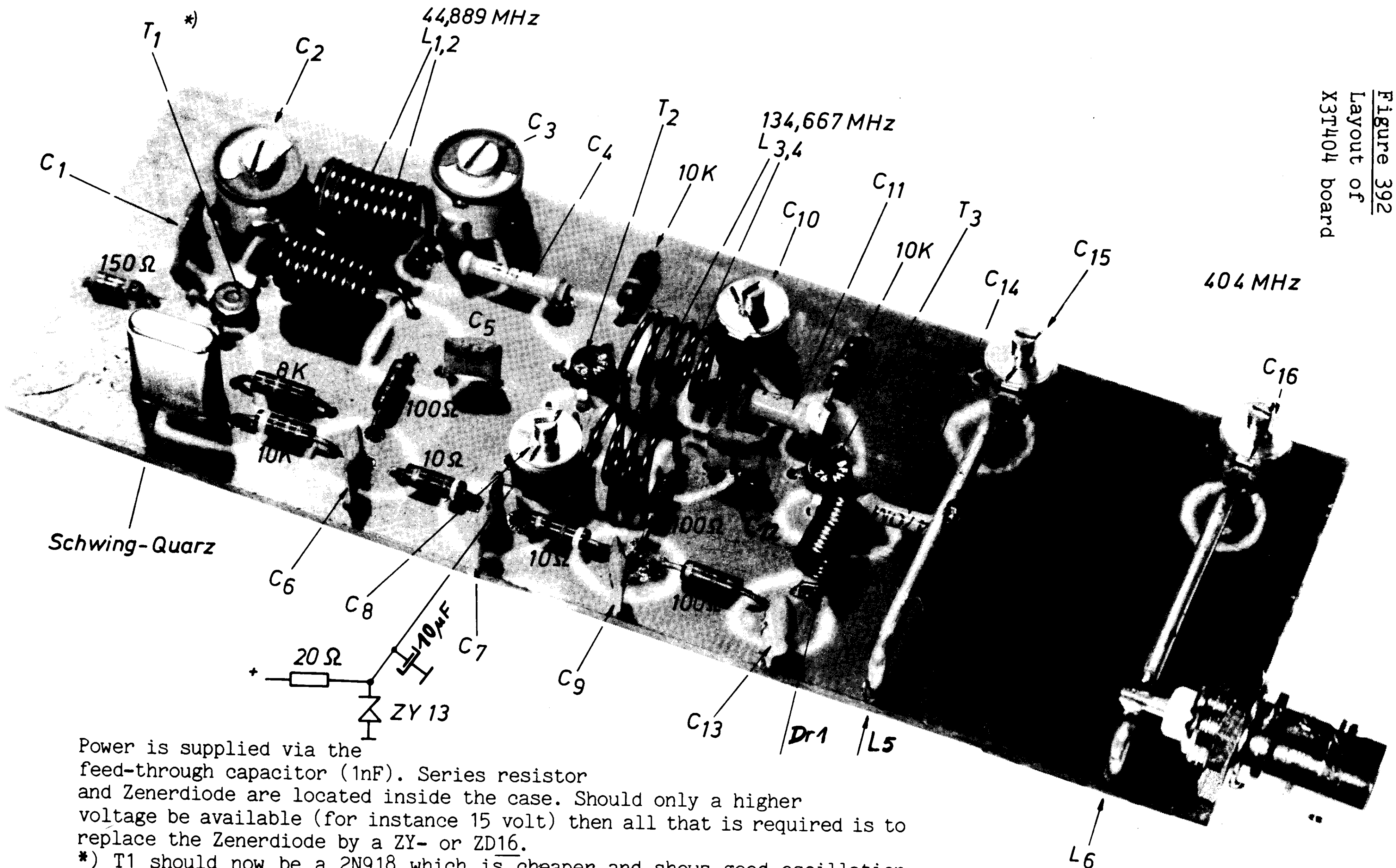
horizontal scale:
100 MHz per division

vertical scale:
10 dB per division

Components:

Crystal	= 44.888889 MHz, series resonance, type HC 43/U QT- C-1024 by Quarz Muller, D-5568 DAUN Postfach 1205
C 1	= 15pF or trimmer 3-22pF, Dr1 = 17.5cm on 5mm diam.
C2, C3	= ~30-80pF or 40pF fixed, 30pF trimmer.
L1, L2	= 10 turns, 1mm Cu/enamel on 6mm diam.
L3, L4	= 8 turns, 5mm diam.
L5, L6	= 2mm Cu, spaced 6mm from board
C4	= 35 tubular capacitor
C5, C6, C7	= 3.3nF; C9, C12, C13 = 220pF; C14 = 470pF
C8, C10	= 3-22pF foil trimmer; C15, C16 = 2-6pF foil; C11 = 15pF

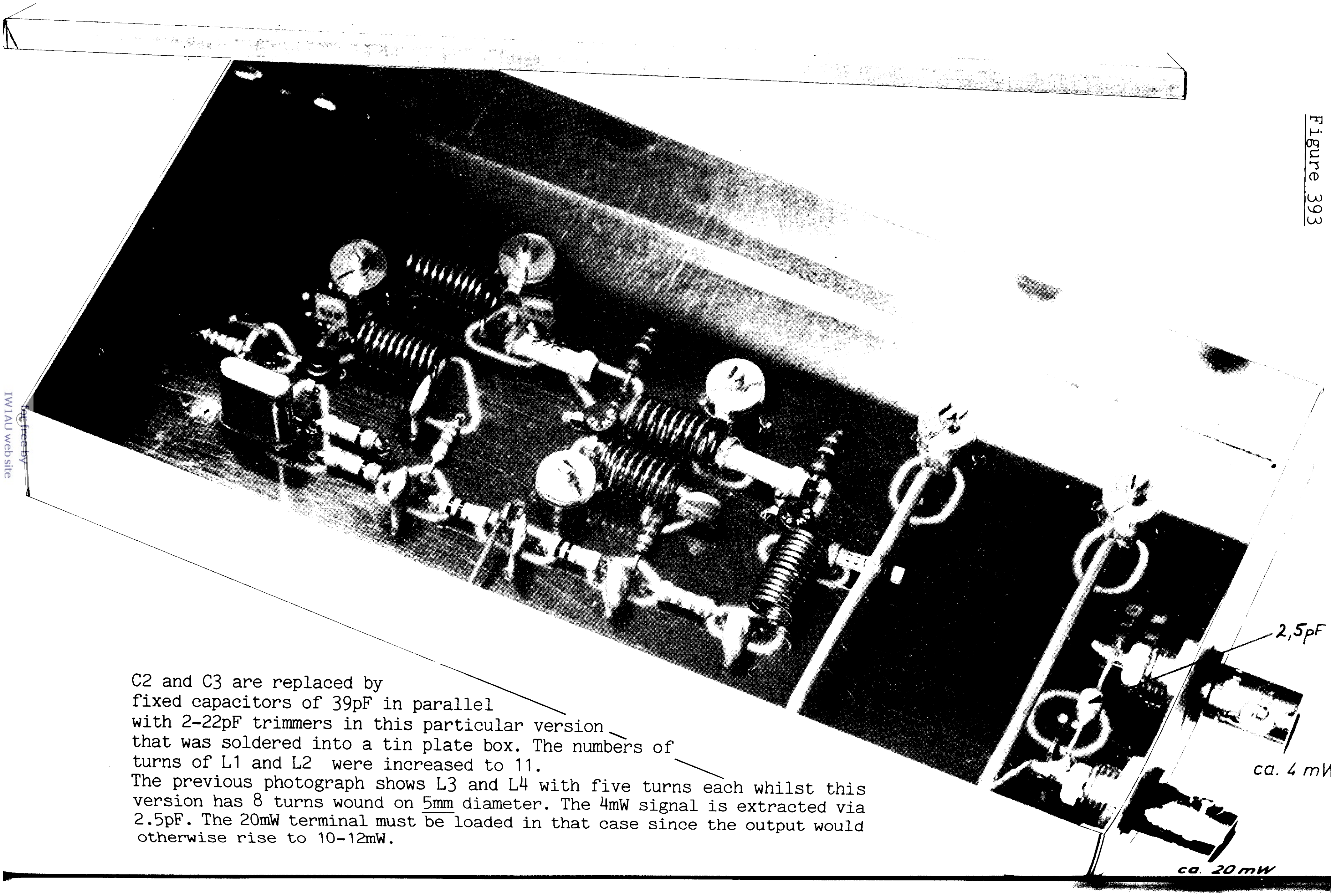
Figure 392
Layout of
X3T404 board



Power is supplied via the feed-through capacitor (1nF). Series resistor and Zenerdiode are located inside the case. Should only a higher voltage be available (for instance 15 volt) then all that is required is to replace the Zenerdiode by a ZY- or ZD16.

*) T1 should now be a 2N918 which is cheaper and shows good oscillation properties. The emitter resistor (150 ohm) is then replaced by one of value 2 Kohm. C1 - if substituted by a 25pF trimmer - allows slight frequency correction.

Figure 393



C2 and C3 are replaced by fixed capacitors of 39pF in parallel with 2-22pF trimmers in this particular version that was soldered into a tin plate box. The numbers of turns of L1 and L2 were increased to 11.
The previous photograph shows L3 and L4 with five turns each whilst this version has 8 turns wound on 5mm diameter. The 4mW signal is extracted via 2.5pF. The 20mW terminal must be loaded in that case since the output would otherwise rise to 10-12mW.

D.6.2 Injection signal generator for 1152/1268 MHz

ca. 20mW

The design of this injection signal generator was directed by the following requirements:

1. Suitability for 23cm power mixers, i.e. P > 25mW.
2. Application as beacon or TX-hunting transmitter.
3. Low mechanical effort, resp. construction in standard box.
4. High attenuation of all spurious and harmonic products.

The suitability of an oscillator board for high level mixing demands a power level of 25 to 50 mW. The design was based on the proven circuit of section D.6.1 - the only difference being a different transistor in the oscillator stage.
A transistor in the final tripler is a rather costly affair if selected for an output power of 25 to 50mW (on 23cm). This power level which required quite a financial effort, can - however - not be utilized on 70cm. Simpler and more cost effective is the amplification on 70cm and consecutive tripling by means of a varactor. In that case the increased power is available on both bands simultaneously (as beacon and for transmitter hunting etc.). The terminal frequency of 1152 resp. 1268MHz is obtained by f_Q x 27 (f_Q x 3 x 3 x 3) with the ninth harmonic amplified by means of a BFT12 or a transistor called CHEO (by CTC) from 20mW (behind T3) to 100 resp. 200mW to provide some reserve power. Upon installation of the 30pF capacitor and tripling to the final frequency more than 25 resp. 50mW are obtainable from the two output terminals simultaneously.

Bandfilter coupling throughout all tripler stages ensures high suppression of all unwanted frequencies. A diode in the power supply connection safeguards against destruction through inverted polarity. All stages are stabilized through series resistors. Whereas the two cover plates will suffice as ground surfaces for L6 and L7 it is necessary to instal a copper plate (copper or brass, 1mm thick, 35x69mm, spaced 3mm off the base towards the inside) in the final frequency area. This measure will improve both mechanical and electrical stability of the module. The cold ends of L6 and L7 are drilled, taped (M3) and bolted from outside. This is also recommended for L8 and L10. We did, however, mount L8 - L10 on to the partition and soldered from inside the L6/L7 cavity. Soldering inside a cavity such that r.f. has to cross the solder joint should be avoided if at all possible (i.e. no soldering of L8 - L10 from inside the L8 - L10 cavity in this particular case) as the poor conductivity of solder would reduce the circuit quality factor (see A.1.3). This would, in turn, lead to power loss and instabilities.
The two trimmer capacitors (6pF) near the varactor diode BA149 are set to near-minimum (see photograph). Capacitive coupling between L8 and L9 is provided by a 4mm wide copper strip which is brought to within 1mm of the hot end of L9. The tuning assemblies for L8 to L10 were salvaged from trimmer capacitors of the type employed for L6 and L7. Brass bolts (M3) and locknuts could be used instead. The photograph should answer any additional questions.

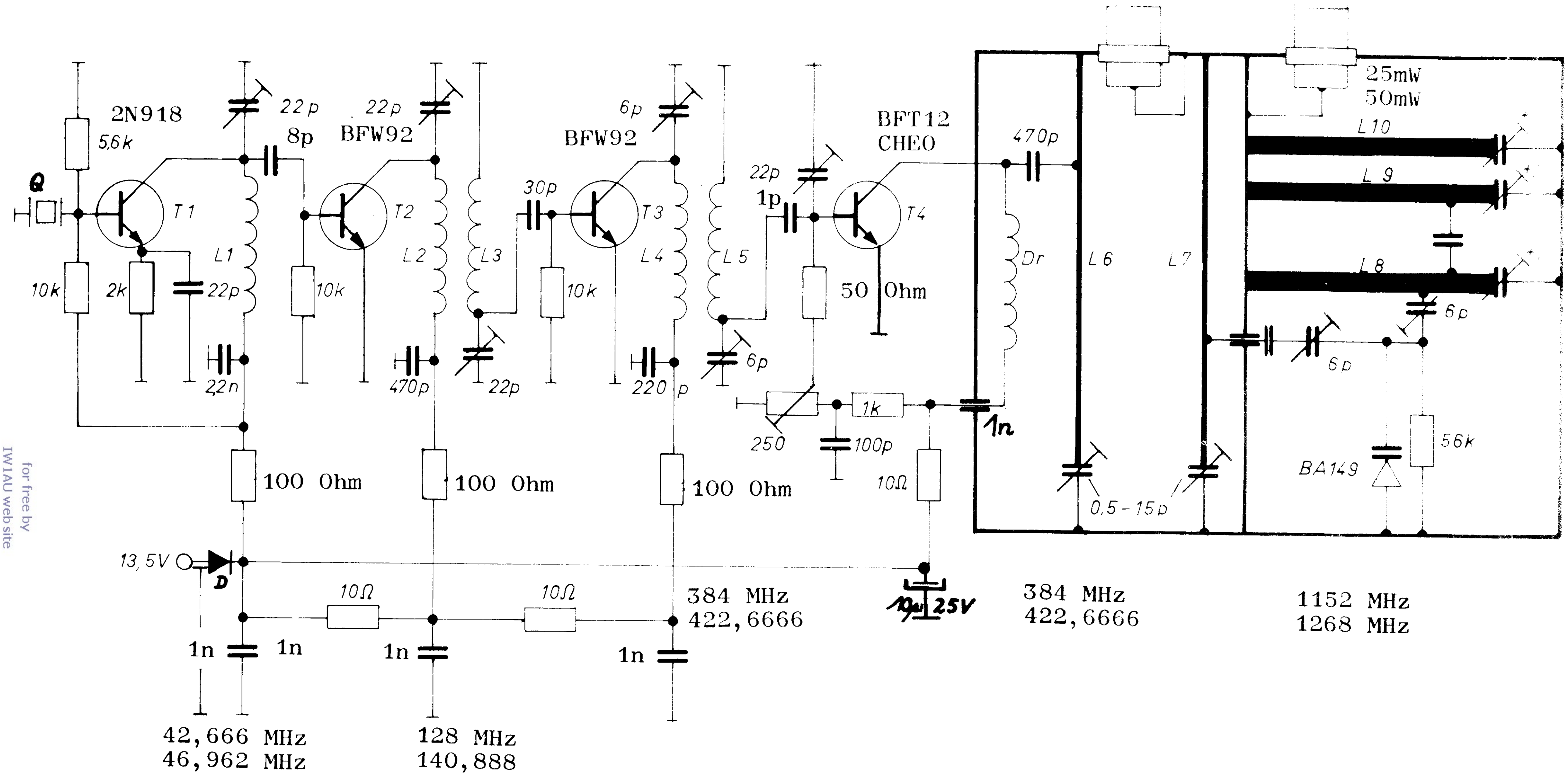


Figure 394 Circuit diagram of the 23cm injection signal generator. Blocking capacitor values are not critical. A brass or copper base plate should be installed inside the 1152 resp. 1268 MHz compartment. It could be extended to the 384 MHz circuit. This measure results in improved stability of the module. Tuning is provided through spindles that protrude into the inside of L8 - L10 and were salvaged from trimmer capacitors. The BA149 turned out to be the best varactor diode and is frequently found in old TV sets.

L1 = 22 turns of 0.8mm Cu/enamel wound on 6mm diam.
L4/5 = 2-3 turns of 1mm Cu/enamel wound on 4mm diam.
L8/9/10 = brass tubing, 6mm diam., 30mm long, soldered from compartment outside.

L2/3 = 7-9 turns 1mm Cu/enamel on 4mm diam.
L6/7 = length depends on trimmers, [illegible]6mm diam
Dr = 5 turns of 1mm Cu wound on 4mm diam.

L1 could be manufactured according to either the circuit description or the photograph - the difference being in the number of turns and the internal diameter. Both solutions lead to success.

Alignment: A dip meter is sufficient for the oscillator and f_Q x 3 circuits. The setting-in of oscillations is indicated by a current rise in T1 and T2. Since bandfilter coupling is used throughout the module it is necessary to proceed stage by stage from oscillator to output. This is the only way to end up with the correct output frequency. A test receiver or the instruments mentioned in sections B.1.4 and B.1.5 enable one to stick to the frequency plan. Alignment with the case still open is possible up to L5. Coarse adjustment requires the base cover and fine adjustment both covers to be installed before aligning the remaining stages. The quiescent current of T4 is set to 20-40mA in either case; this requires the excitation to be removed (i.e. the oscillator must be made inoperative). As L2/L3 and L4/L5 are tuned to resonance the current drawn by the subsequent stage will rise. Fitting of an idler circuit to the BA149 made no difference worth mentioning.

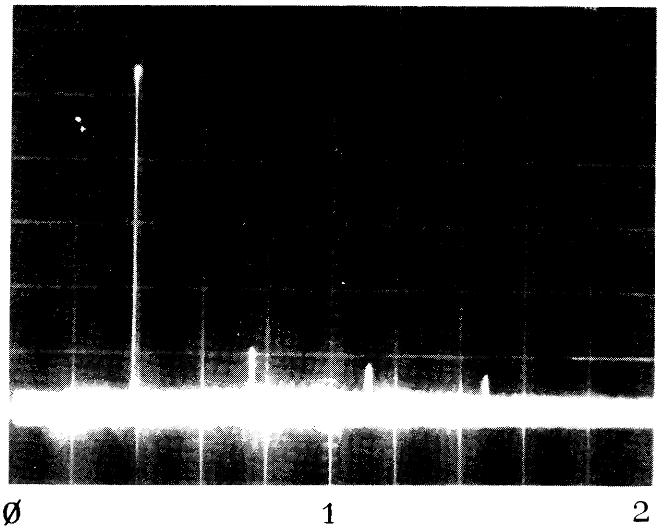


Figure 395
Frequency spectrum at f_Q x 9

In this case: 384 MHz

h: 200 MHz/division
v: 20 dB /division

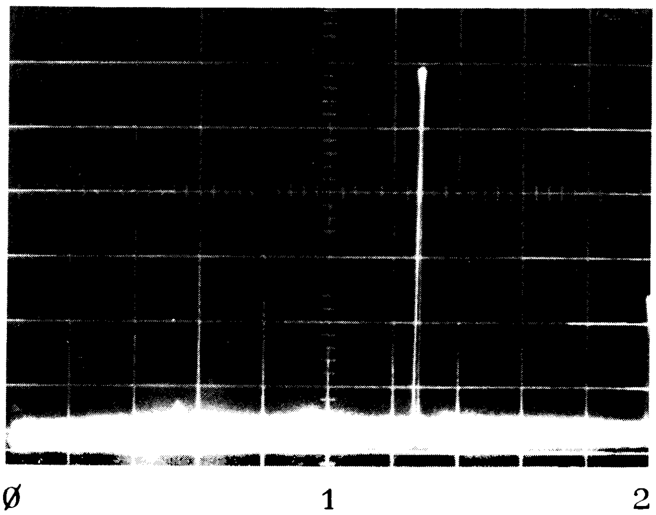
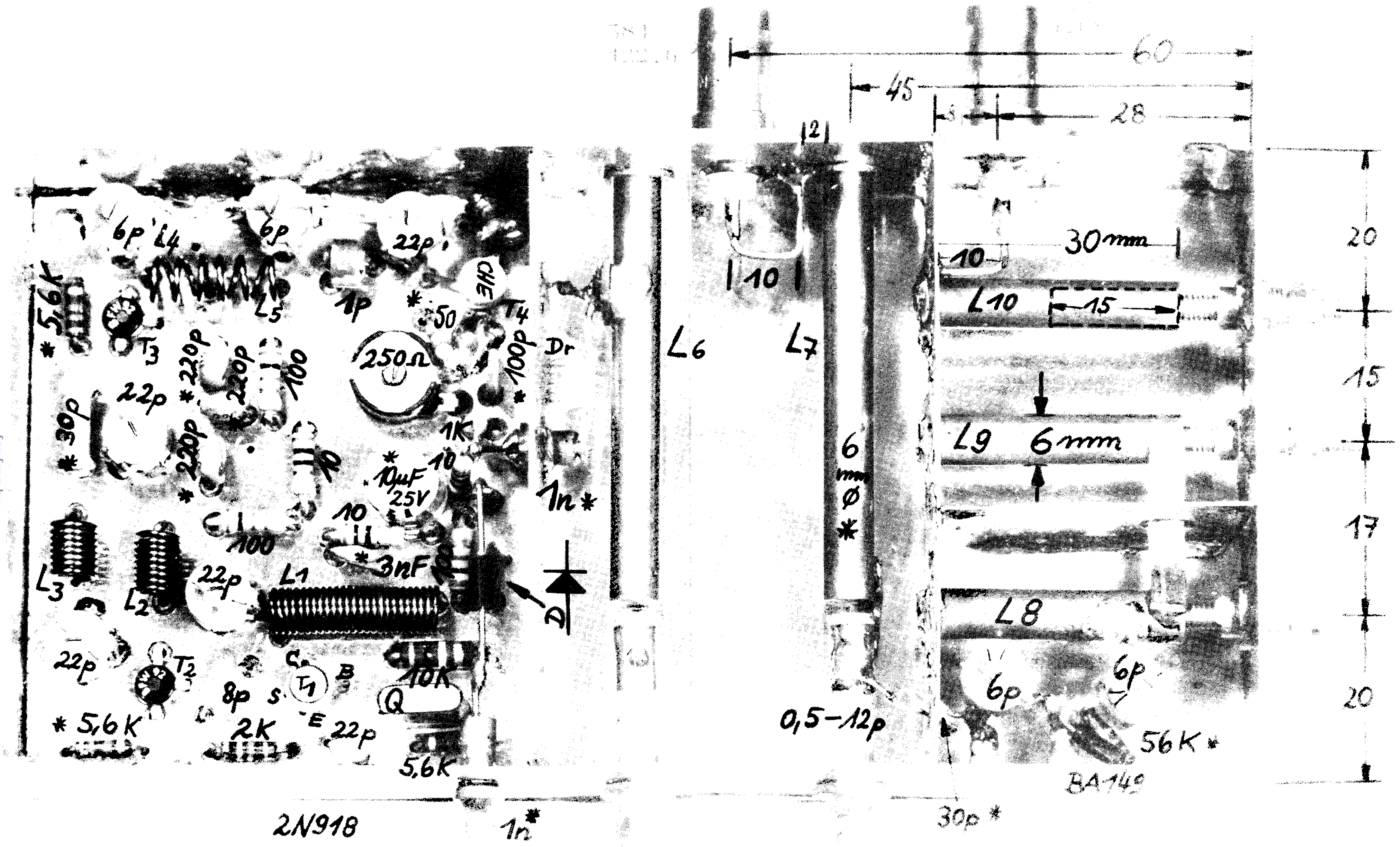


Figure 396
Frequency spectrum at f_Q x 27

In this case: 1268 MHz

h: 200 MHz/division
v: 10 dB/division

Figure 39: Layout of injection signal generator. All [illegible] be read off the photograph. Case dimensions: 140x72x30mm



* = value not critical

mounting posts for improved mechanical stability.

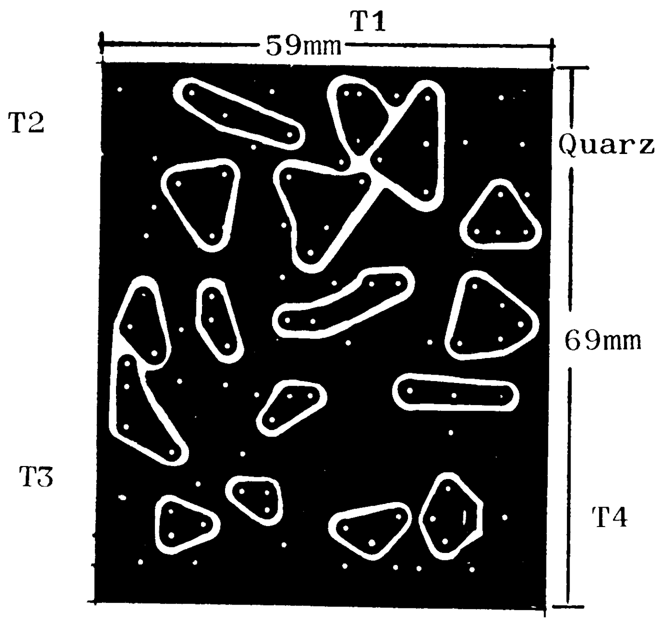


Figure 398
Underside of double sided copper clad board. This board carries all components up to the input of T4. The BFT12 is soldered directly into the partition next to L6. The base connecting lead must, in that case, be lengthened by a piece of CuAg wire to reach the soldering point near the 22pF capacitor. Holes for all crystal types listed further down are provided.

After drilling the holes the copper on the top side around them must be removed by means of a drill (3 to 4mm diameter) to prevent short circuits resp. unintentional contacts to ground.
Crystals specifically produced for this oscillator may be obtained from:

Quarz Müller, D-5568 Daun / Eifel, Postfach 1205

Other crystals may lead to significant differences in output frequency.

The following types by Quarz-Müller have been tested: HC 6/U, HC 18/U, HC 25/U and HC 43/U.

Ordering information: HC 43/U QT-C-1024 series resonance.

Crystal frequencies: 42.66667 MHz for injection frequency 1152 MHz
46.96296 MHz for injection frequency 1268 MHz.

Figure 399
Crystal types HC 6/U, HC 25/U and HC 18/U. The type HC 43/U corresponds roughly to the HC 18/U.

If temperature stability is considered to be of importance then the type HC 6/U is to be recommended.

The shown crystals are of different makes and frequencies.

D.7 Station concept for the 70cm and 23cm bands

A large proportion of UHF amateurs prefers transverter or converter operation. The short wave transceiver (28 MHz band) acts as exciter and receiver for both the 70cm and 23cm bands. Whereas 28MHz represents an almost perfect i.f. for the 70cm band the situation is somewhat critical for 23cm operation. This is because injection frequency (f_o) and image frequency (f_{sp}) are rather close to the desired signal frequency. The following broadband transistorized amplifiers provide no selectivity and unwanted signal components are going to be amplified as well. Even though the mixer will attenuate the injection signal by 20 to 30 dB this is not sufficient to stop this signal from reaching the antenna. Furthermore, the image frequency signal will usually be present with a power level comparable or significantly higher than the desired signal (frequency f_n). The use of filters is thus required. Suitable filters are described in section C.
Many amateurs only possess 2m transceivers but would still like to operate on 70cm. This can not be recommended. While it is possible to solve the receiving problems it is virtually impossible to get the transmitting problems under control. These arise from the considerable effort required to filter out the tripled frequency which is furthermore not suppressed by the mixer. This leaves us with the necessity for double conversion. When considering the efforts discussed in various sections of this book to suppress spurious and harmonic products one will get an idea of what is involved to build all the compartments and to arrange all the screening in order to separate wanted and unwanted signal components. The statement that "these products are outside the amateur band" is not acceptable. The only solution could be the purchasing of one of the fair-priced 27MHz SSB stations that are quite simple to convert to 28MHz operation. This would leave the 2m station free for application on - for instance - the higher bands.

Figures 400 and 401 clarify the problem on 23cm with intermediate frequencies of either 28 or 144 MHz. The block diagrams are shown to the left, the spectral situations are shown to the right - with and without filtering.

Figure 400 Method with 1268 MHz injection frequency and 28 MHz exciter frequency. Notice that the image frequency f_{sp} is attenuated by only 15 dB in spite of the effort. There is practically no attenuation for the injection signal f_o.

Block diagram of signal generation

Spectral display
in front of the filter after the filter
(triple filter according to C.4.2)

f_{sp} 15dB below f_n

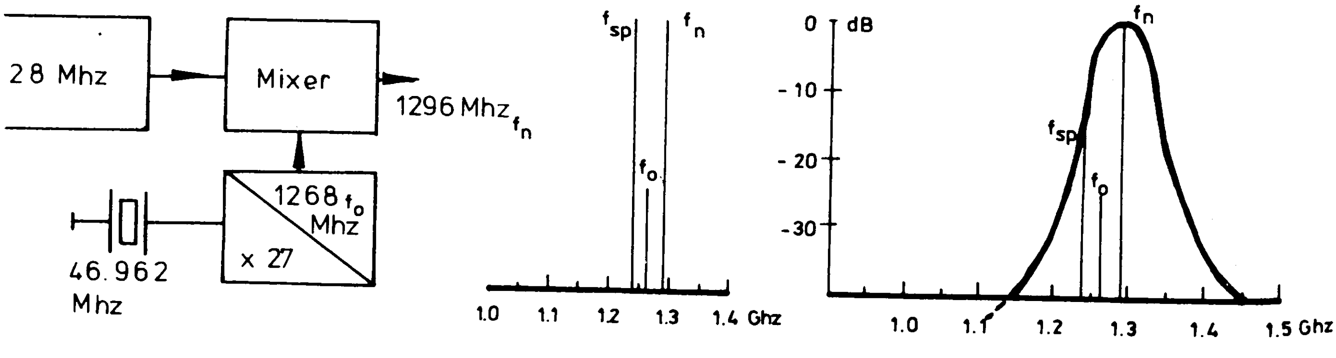


Figure 401 Method with 1152 MHz injection frequency and 144 MHz exciter frequency. Compared with the previous figure the larger frequency separation is noticeable. f_{sp} is already attenuated by 58 dB (can not be shown in this diagram since the curve extends below the base line). f_o is also attenuated by 15 dB. Both products are thus at least 40 dB below the desired signal.

Block diagram of signal generation

Spectral display
in front of the filter after the filter
(triple filter according to C.4.2)

f_{sp} 58dB below f_n

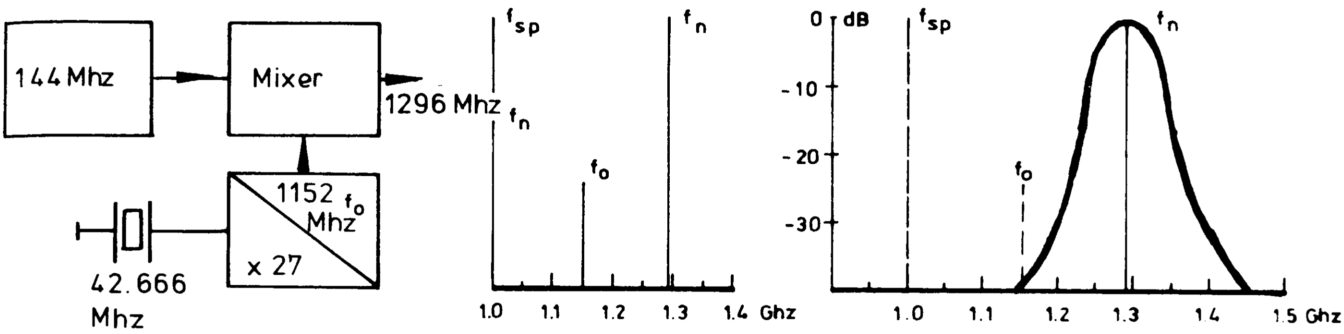
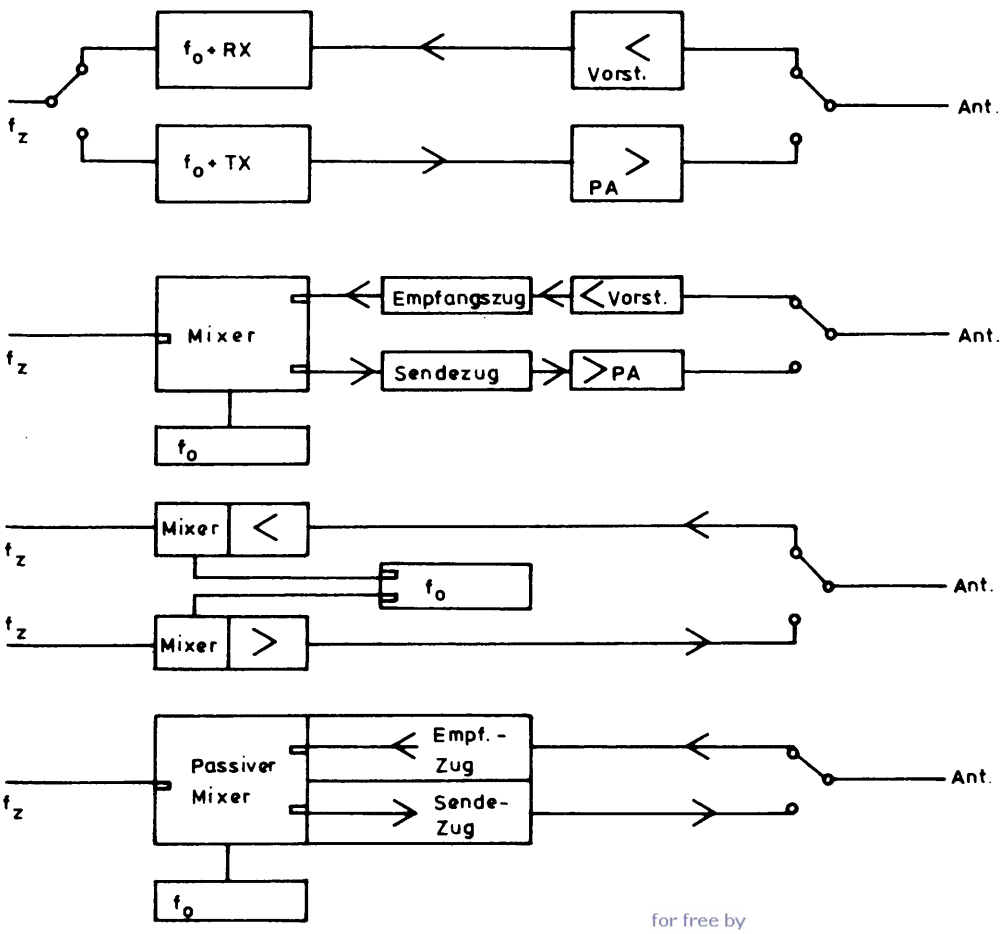


Figure 402 surveys various common concepts. The first concept is outdated and has the disadvantage that separate oscillators in the RX and TX chains could lead to frequency deviations when switching between transmit and receive modes. All other concepts may be composed from modules described in this book.

Station concepts for the 70cm and 23cm bands



D.7.1 A modern 70cm concept

Before going qrv in SSB or CW on 70cm several aspects should be considered. A monoband 70cm transceiver is usually as expensive as a good 2m set and can not normally be afforded in addition to the 2m and shortwave equipment already purchased. A common approach which is both cost effective and quite modern is the use of converters resp. transverters. Starting with a 2m signal - i.e. 144MHz - a 288MHz signal must be added to obtain 432MHz. The third harmonic of 144MHz is also 432MHz; this product is, however, not readable on 432MHz and cannot be removed after the mixer. The rejection of this component depends therefore on the ability of the mixer to suppress the input components. When in receive mode it is, on the other hand, possible that a 2m signal at 50 to 60dB at the receiver location will pass the 70cm converter, reach the 2m receiver and will appear as a 70cm signal. It is furthermore possible that a reasonably strong signal will be picked up by the coaxial cable running from the 70cm converter to the 2m receiver. A cable with double shielding will give a significant improvement without, however, solving the problem as such. It is also impossible to operate duplex on 2m and 70cm with this concept.

An almost perfect solution is to start with 28MHz. An injection signal of 404MHz yields 432MHz, the quality of which is essentially governed by the expensive and technically advanced short wave equipment capable of operating in SSB, CW and (occasionally) AM. Since the 2m band is completely by-passed simultaneous operation on both bands (contests etc.) is possible. For good reasons professional equipment is increasingly designed to a modular concept, i.e. screened individual modules, and the amateur is well advised to adopt this up-to-date technique.

The following concept is based on these aspects and has a number of important advantages:

- Exchangeability of all modules. As new techniques are developed they may be incorporated and the complete station may be kept up to date.

- The ability to insert filters, directional couplers, amplifiers and measuring devices between the modules. This allows - for instance - the reduction of the interference caused by the transmit chain without reducing the receiver sensitivity through lossy filters, or to place an additional converter up front to reach the 12cm band with an i.f. of 432MHz.

- Only one coaxial relay is required for r.f. switching.

- The effects of cable losses in the receive chain may be avoided by placing the converter and the coaxial relay near the aerial fed point. This results in a significant improvement of the sensitivity.

Figure 403 gives further details of the concept. The specified filters may be varied according to requirements (see D.3). Incorporating, for instance, an additional coaxial relay after the power amplifier would allow the change over to a dummy load (see B.2). It is then possible to tune the final prior to going on the air or to measure the output power at any time without great effort.
This concept may be realized by combining various modules. Injection signal generation (404MHz) is performed according to D.6.1. Depending on one's financial resources and geographic conditions there are three different converter circuits to choose from (section D.2). If necessary, additional selectivity may be obtained as described in section C.
The type TMX D2T 10/70 mixer from D.4.1 is employed in the transmit chain. The power is then raised to the required drive level of the final by modules described in section D.5. A power amplifier comprising a valve from the 2C39 family (D.5.3) or 4X150 family amplifies the signal to the specified values. It is worth mentioning that an output power of 50 to 150 watts is quite sufficient and constitutes a good compromise between power and effort.

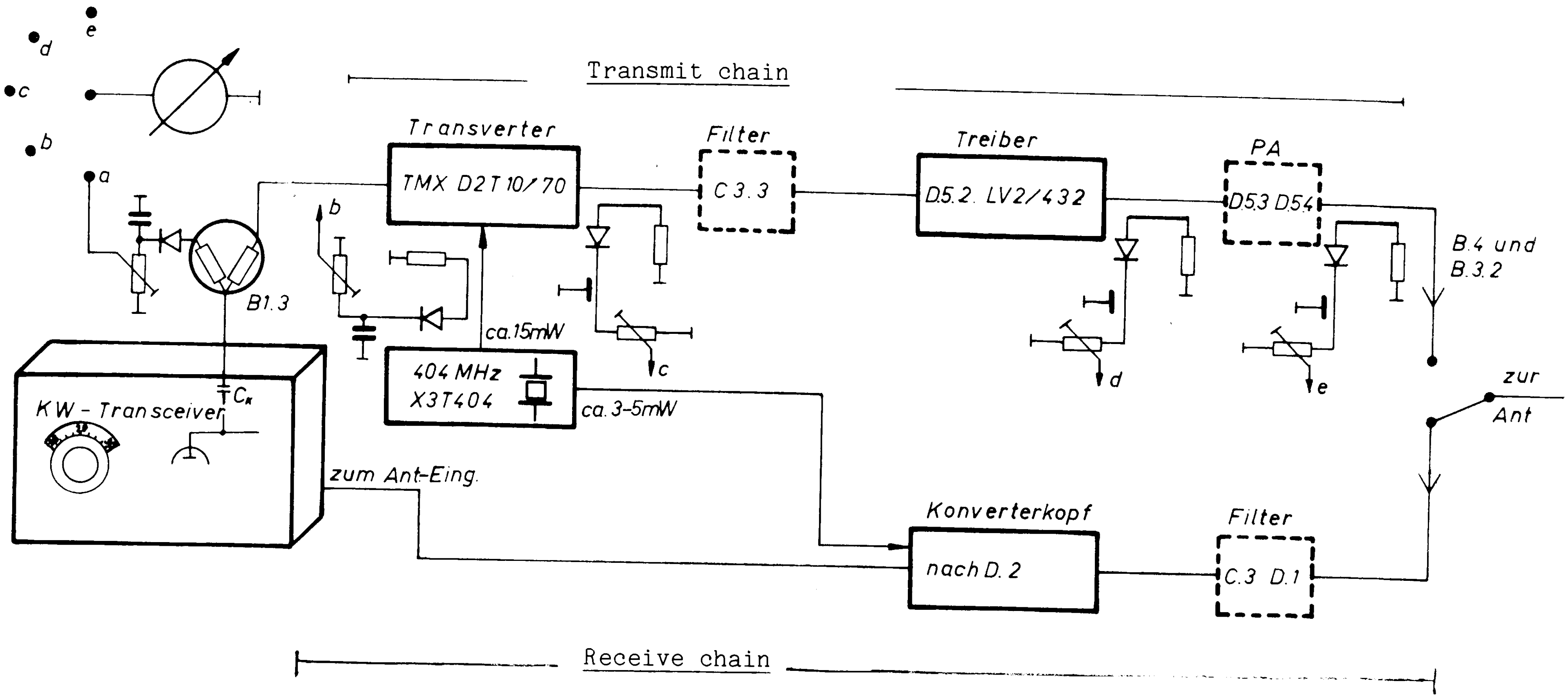


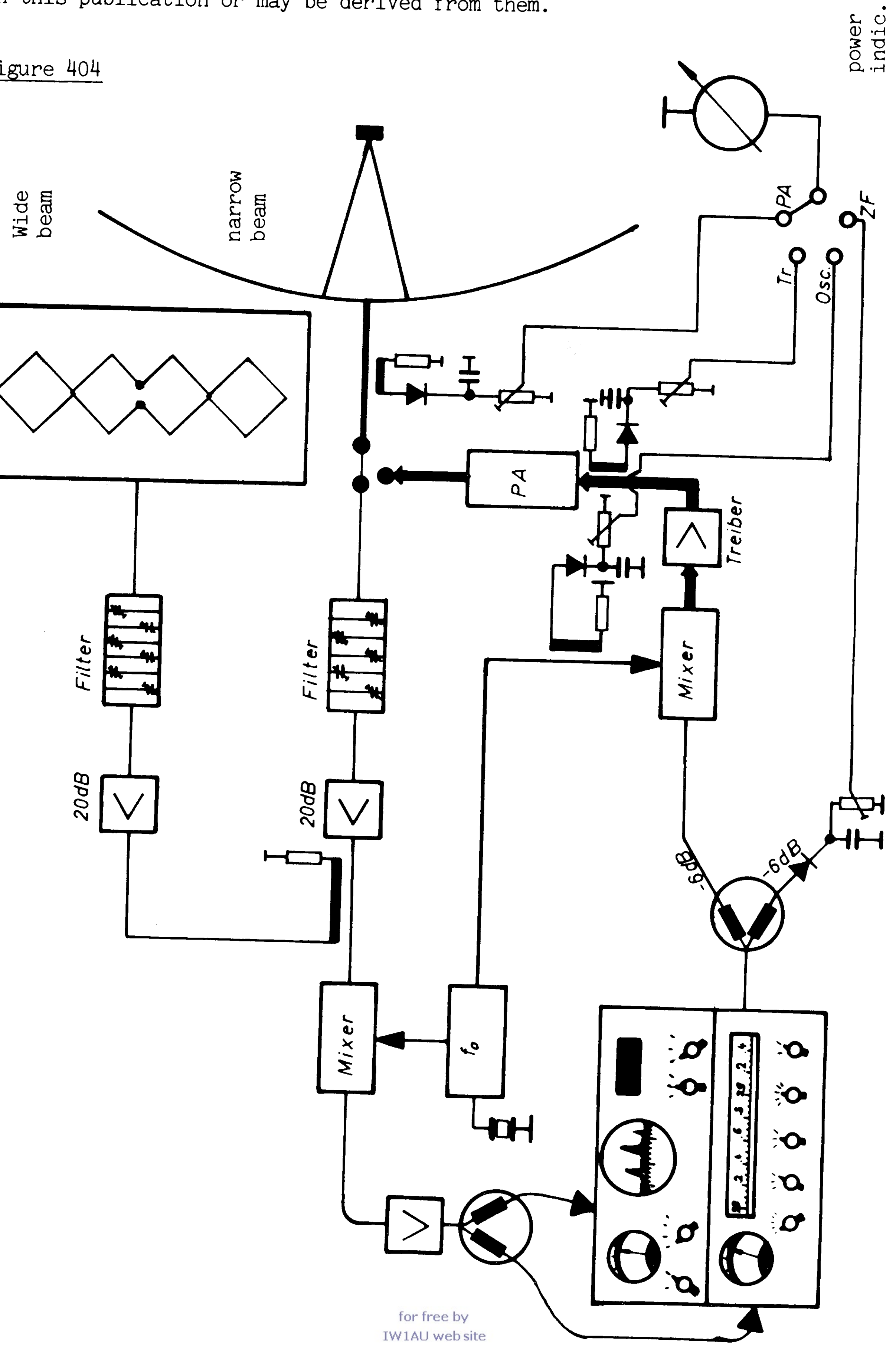
Figure 403 Block diagram of 70cm concept

The block diagram shows the interconnection of the various modules. If not already provided by the maker, a 5pF capacitor (500V) is soldered to the anode of the driver valve and brought out to a BNC connector via a screened cable. This coupling capacitor (C_k) may have to be altered depending on the different power levels met in various transceivers. A mW-meter according to B.2.1 may come in handy. During transmission or reception only the transverter resp. converter is supplied with power. The 404MHz module is continuously connected to the power supply. The power supply leads of the short-wave final- amplifier heaters are routed via a switch. In transverter mode it is rectified (single phase) and smoothed by a 10,000 to 15,000uF capacitor and constitutes the low voltage power supply in a concept without an operative short wave final amplifier.

D.7.2 A modern combination concept for 70cm resp. 23cm

In contrast to D.7.1 a more elaborate combination is presented. It offers the following possibilities: Combined or independant operation with wide and/or narrow antenna beam width. Simultaneous monitoring of activities on the entire band in the instantaneous antenna direction. Supervision of all power modules by switchable power indicator. Easy replacement of all stages because of modularization. Descriptions of the necessary modules are contained in this publication or may be derived from them.

Figure 404

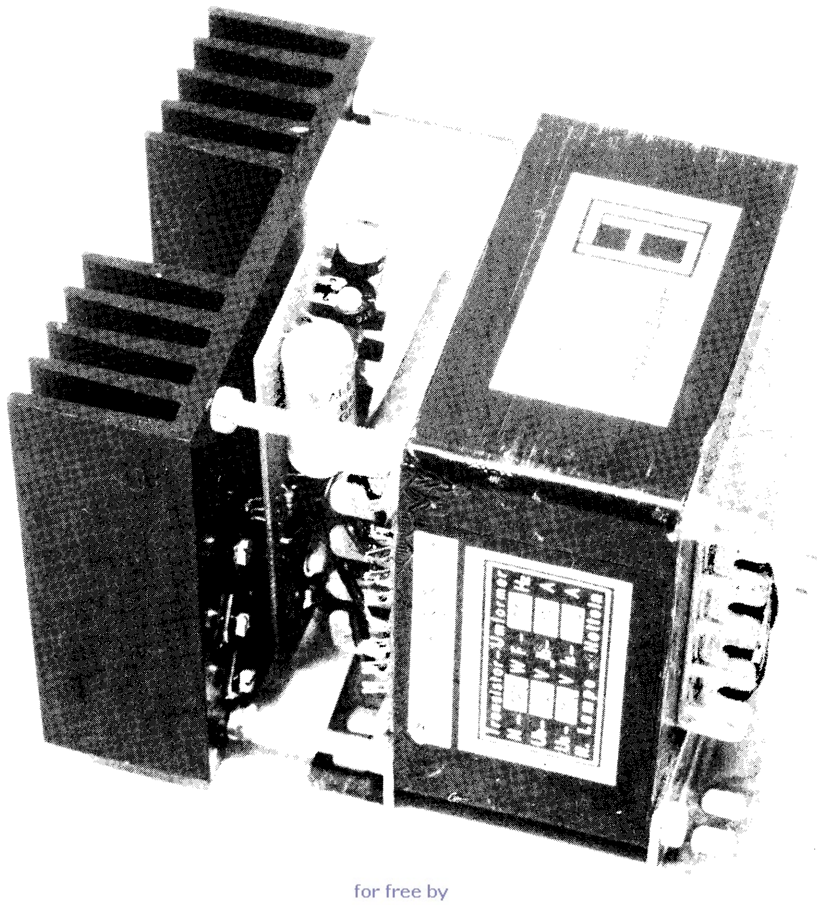


Station concept for 70 cm or 23cm

D.8 12 Volt mobile power supply

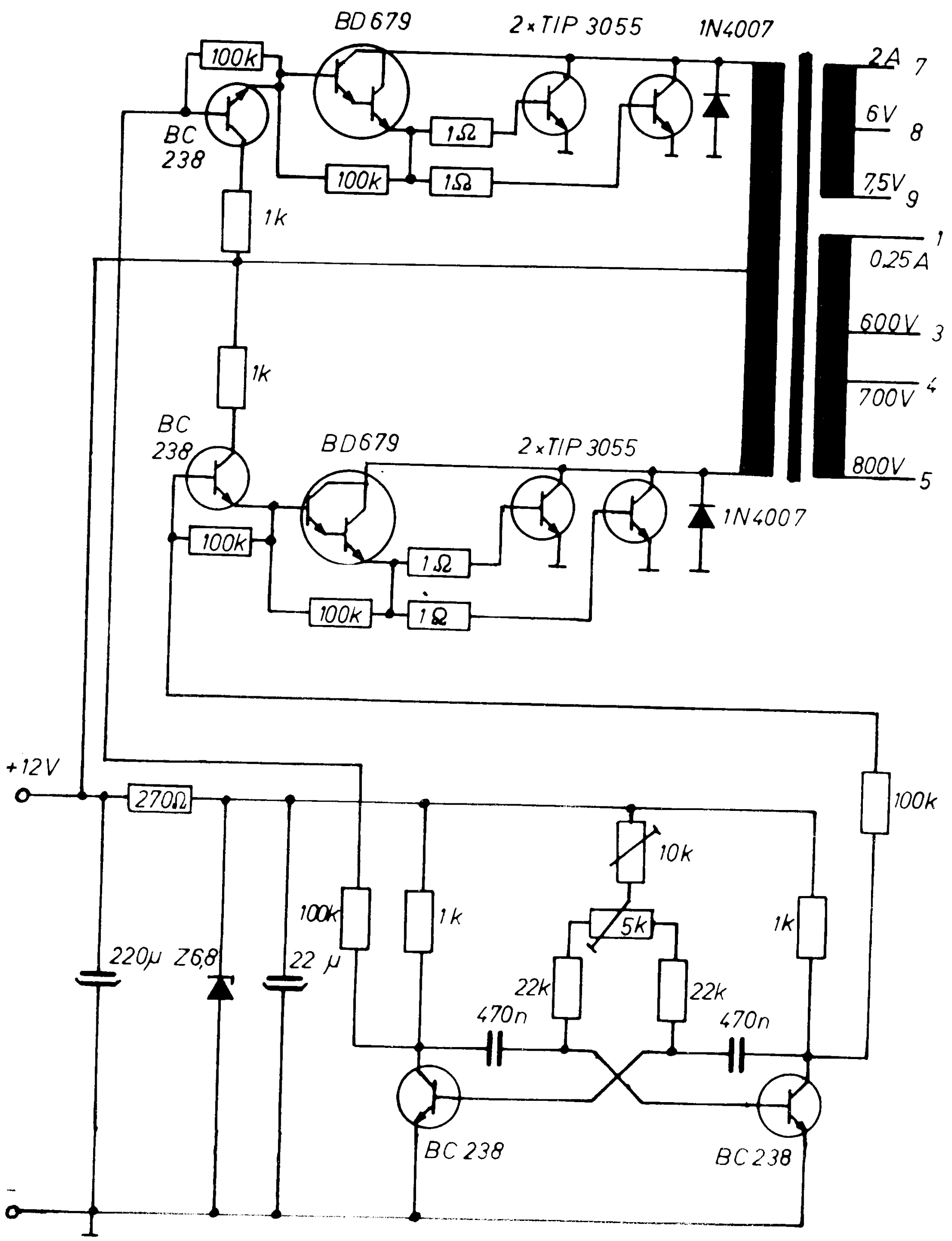
Many friends of the 70cm band have aquired one of the many handy SSB/CW transceivers offered on the market. In cases of favourable propagation conditions or during contests they are thus quickly qrv with low power from an exposed location. It is then not unusual that they receive a station with S2 to S3 - i.e. good readability - without having their own call answered. There is no point in advising the other chap to donate his insensitive receiver to the museum. The more likely explanation could be that he is transmitting at 100 watt against the 3 watt of the portable set. The difference (see G.3.1) is about 15dB - i.e. three S-increments. The antennas need not be considered since they are effective for both reception and transmission. It is thus not necessary to calculate the effective radiated power.
A simple to construct 2C39-p.a. could raise the 3 watt signal by 10 to 13dB (i.e. 30 to 60 watt) depending on the state of the valve. The net difference against the 100 watt of the received station would now be only approximately 3dB. Should the other chap run a fixed station with 300 watt or more output power he is likely to be an UHF-specialist that is complementing his high output power with an extremely low noise receiver.
A 2C39 p.a. obviously calls for a power supply to deliver the necessary heater and anode voltages. The module described below converts the 12 volt d.c. input to the required voltages (after rectification and smoothing of U_A, see circuit in figure 386).

Figure 405 150 watt converter for the supply of 2C39 power amplifiers. Input voltage: 12 volt d.c. The square wave output voltages are inscribed.



The circuit diagram shows two TIP 3055 per branch. This number is absolutely necessary for an output power of 150 watts. The cross section of the supply leads must be in accordance with the primary current. A fuse of 15 to 20 ampere should be placed in the connection between the car battery and the converter module. There is no point in building this converter just as there is little point in building short wave transceivers. The circuit is nevertheless published in case some one should like to duplicate the module. Winding data and winding sense are a bit difficult to describe. About 860 turns (0.3mm diam.) on a M85a core are required for the 600 volt winding, the corresponding figures for the heater are 9 turns for 6 volt. Additional questions and inquiries concerning prices should be directed to DJ8MO (Lampe company).

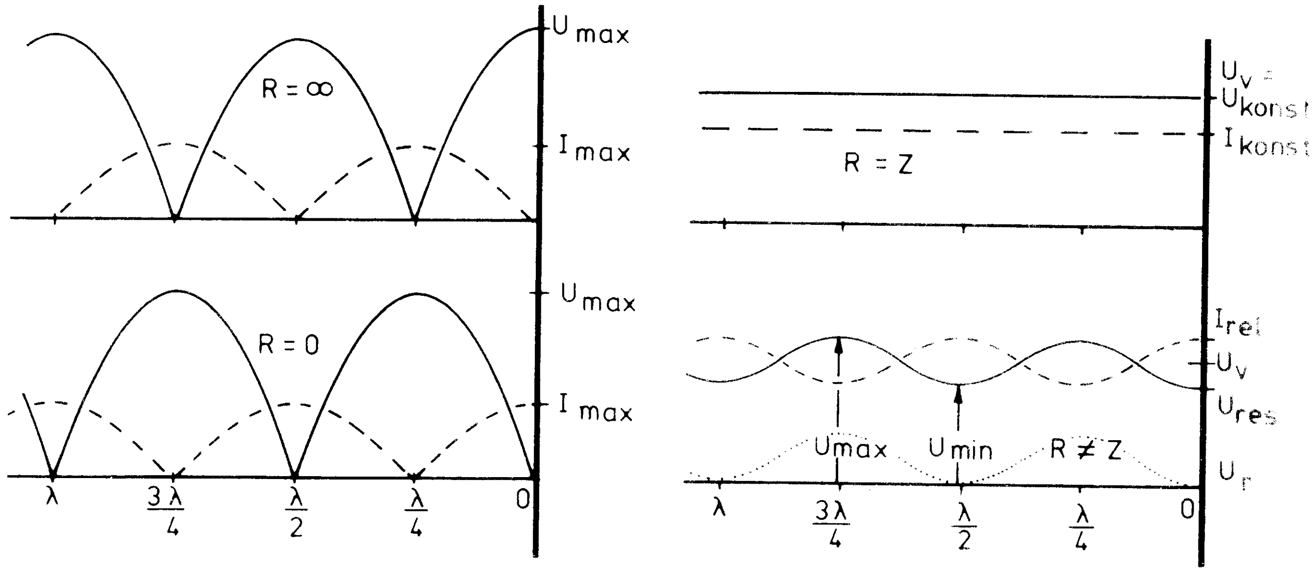
Figure 406 Circuit diagram of 150 watt converter for 2C39 p.a.



E.3 Matching and balancing

Matching denotes the process whereby a load is connected to a generator in such a way that generator internal resistance and load resistance as seen by the generator are equal. To the amateur this means in general that transmitters and receivers that are connected to an (artificial) antenna should have input resp. output resistances equal to the cable characteristic impedance and the antenna resistance. Matching is necessary to prevent the useful power from being reflected. Power losses and undesired radiation are thus avoided.
A test line allows the measurement of the voltage and current distributions. The following diagrams show the situations on open circuited, short circuited and correctly terminated lines (R = Z).

Figure 407



$U_r - U_r = U_{res}$

$Z_L = U_{max} / I_{max} = U_{konst} / I_{konst} = U_{min} / I_{min}$

There are various ways for the mathematical treatment of matching. Amateurs usually employ the VSWR (Voltage Standing Wave Ratio) which corresponds to the ripple factor s.

$$s = \frac{1}{m} = \frac{U_v + U_r}{U_v - U_r} = \frac{U_{max}}{U_{min}}$$

U_v = forward voltage

U_r = reverse voltage

The presented range starts with matched state at "1" and ends with open circuit or short circuit at "infinity".
The matching factor m, however, starts at "1" for the matched condition and "0" for open resp. short circuit.

$$m = \frac{1}{s} = \frac{1 - r}{1 + r}$$

r = reflection coefficient

(see also B.3)

The reflection coefficient is the percentage of forward voltage reflected back to the generator.

$$r = \frac{U_r}{U_v} \times 100\% = \frac{s - 1}{s + 1} = \frac{R - Z}{R + Z}$$

Z = charact. impedance
R = mismatched termination

Another commonly used expression is the return loss. This is the logarithm of 1/r (see also note in E.4.6)

$$a_r = 20 \log U_v \,/\, U_r \text{ [dB]}$$

Because of the international adoption of the 50 ohm coaxial standard for the interconnection of r.f. apparatus (world wide standard in amateur circles anyhow) there should not arise any problems with cables and plugs designed to this standard. The use of even the sometimes very cheap discarded coaxial and Flexwell cables (commercial applications in 60 ohm standard) in conjunction with 50 ohm material does not unduly worsen the matching conditions.

$$\text{Reflection coefficient } r = \frac{R - Z}{R + Z} = \frac{60 - 50}{60 + 50} = \frac{10}{110} = 0.09 = 9\ \%$$

This corresponds to a VSWR of 1.3

Clearly, this "super scrap" is still quite valuable for our purposes as long as different impedances are not mixed in one cable run. One otherwise risks establishing resonant sections resulting in unintentional impedance transformation.

The aspect of impedance matching is on its own, however, not sufficient - in particular when dealing with antenna systems. Most radiators are composed of symmetrical elements which cannot be fed from non-symmetrical coaxial cables straight away. It is therefore necessary to employ suitable balancing devices. They ensure even current and voltage distribution, optimum radiation of the transmitter energy and prevent waves from travelling along the outside of the coaxial cable - which is often the cause of BCI and TVI (particularly in densely populated areas).

When facing a balancing problem it is necessary to take the possible need for impedance transformation into consideration. The half wavelength by-pass line could provide the solution in such a case. It consists of a piece of coaxial cable corresponding to half a wavelength shortened by the velocity factor of the cable. The outer braidings of the three ends are connected to one radiator and one each of the inner conductors to the two radiators as shown.

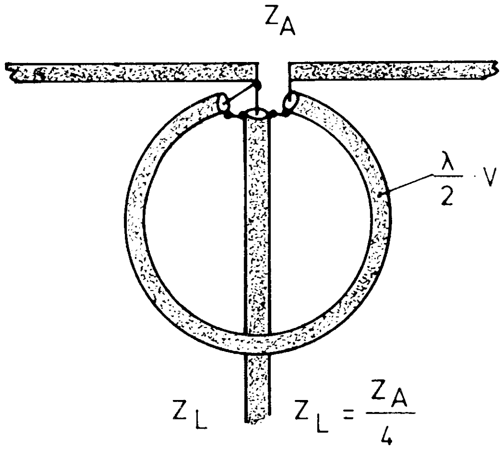


Figure 408
Coaxial balun with 4:1 impedance step-up.

Make sure that there is no ohmic connection between inner and outer conductor at the right radiator. This balance-to-unbalance transformer (balun) provides a 4:1 impedance step-up. It has reasonable broadband properties and can handle the usual transmitter power levels. The construction of balun transformers from coaxial cable sections is possible up to the 23cm band. Beyond that frequency the sections become too short for precise dimensioning and the cable losses are rising sharply. Balancing without impedance transformation is provided by a device known as EMI - loop, quarter - wave open balun or Pawsley stub. This device may also be arranged as coaxial sleeve balun.

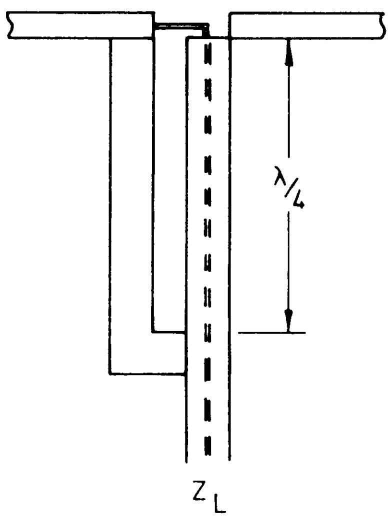
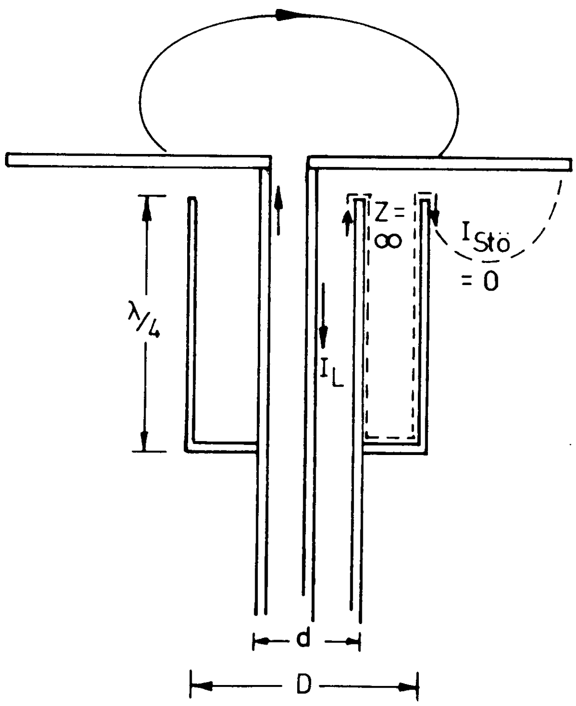


Figure 409
EMI - loop,
quarter wave open balun
or
Pawsey stub

The Pawsey stub was tested on 70cm aerials. The sheath waves existing prior to balancing could be reduced but were not entirely removed. This could be due to the fact that the balancing action is on one side of the cable only (because of the parallel wire arrangement) rather than being effective all around the cable.
Optimum results were achieved with a quarter wavelength coaxial sleeve balun. Three different variants were tried, two of them being shortened by the introduction of some dielectric and the remaining version employing air-dielectric which again turned out to be the least complicated to construct.
The coaxial cavity with its high blocking impedance (which it has at the antenna feed point - being a quarter-wavelength line with one end short circuited) ensures that the energy is not distributed over the two dipole sections at random (giving rise to sheath waves). Rather, a homogeneous field is built up for instance between the radiator connected to the outer conductor and the radiator connected to the inner conductor - thus ensuring symmetrical radiation.

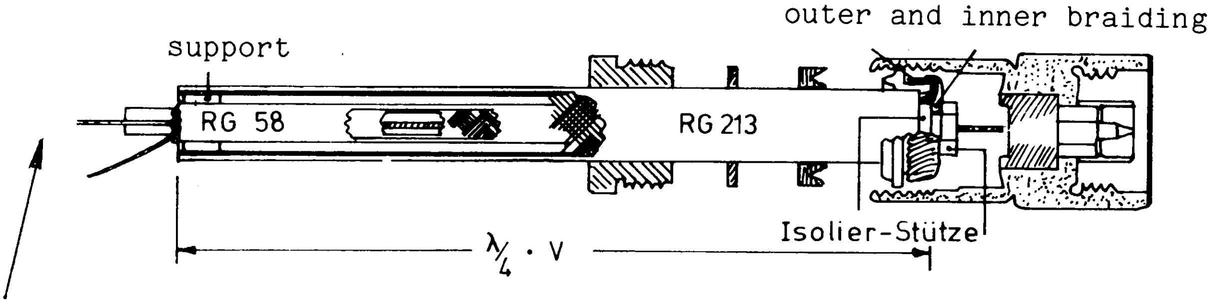
Figure 410
Coaxial sleeve balun



The length of the coaxial cavity is approximately 0.93 x $\lambda/4$. The ratio D/d should be around 2.5 to 4. The radiators should be connected as close as possible to the high - impedance end of the cavity if feeding by coaxial cables. Because of the shortening effect of the dielectric both the centering plate for the inner cable and the cover plate should be no longer than absolutely necessary.

Occasionally space limitations etc. do not allow a stiff assembly. For these cases a coaxial sleeve balun composed of RG58 coaxial cable and parts of the RG213/U cable (external insulation and screening mesh) was fabricated. It is a simple matter to remove inner conductor and dielectric from RG213/U or German video cable (designated 1,1/6,6 75 Ohm). The complete assembly may be mounted to a plug connector. The result is a flexible coaxial balun that may be screwed onto the antenna feeder.

Figure 411 Flexible coaxial sleeve balun



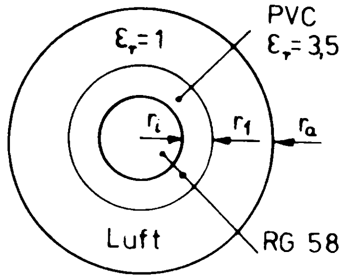
To the dipole

To make such a balun, start with a 14cm long piece of RG213/U from which inner conductor and dielectric have been removed. A 19cm long piece of RG58/U is threaded into this sleeve and wrapped with insulating tape 1cm off the end facing the plug until it fits tightly inside the sleeve. Attach the plug according to the procedure for RG214 cable. The critical part is the slim insulation of the internal conductor of the RG58 which should be supported by means of a 3mm thick disc of PE (internal insulation of RG214) prior to soldering to the spigot of the plug. The hole in the disc should be opened up with the aid of a 4mm drill.
The sleeve of the remaining 4 cm of the RG58/U is removed at the exit. The inner conductor with its insulation is pushed through the coaxial screening. This end is centered by means of insulating tubing and sealed against moisture by means of shrink-on sleeve. Shrink-on sleeve and cable sleeve should first be covered with contact adhesive. The adhesive should have dried before applying open fire to the thermo-shrinkable tubing to avoid any risk of fire.

The physical length of the balun sleeve is calculated according to the following equation:

Figure 412

$$\varepsilon_{rw} = \frac{\ln \frac{r_a}{r_i}}{\frac{\ln r_1/r_i}{\varepsilon_{r1}} + \frac{\ln r_a/r_1}{\varepsilon_{r2}}} = \frac{0{,}7}{0{,}09 + 1{,}38} = 1{,}49$$



This diagram shows a cross section of the various dielectric layers of the quarter wavelength coaxial sleeve balun.

Screening	RG 58	= r_i	= 3.6	mm
	RG 214	= r_a	= 7.25	mm
Sleeve	RG 58	= r_1	= 4.95	mm

Approximately 3cm of RG 58 are wrapped with tape for centering purposes (ε_r = 3.5).
The velocity factor V = $1/\sqrt{\varepsilon}$ for PVC is 0.53
The remainder of $\lambda/4$ (electrical) = 17.5cm - 3cm (tape) = 14.5cm has a V = $1/\sqrt{\varepsilon_{rw}}$ = $1/\sqrt{1.49}$ = 0.82

0.53 x 3 cm	=	1.59 cm
0.82 x 14.5 cm	=	11.88 cm

Total length	=	13.47 cm
		========

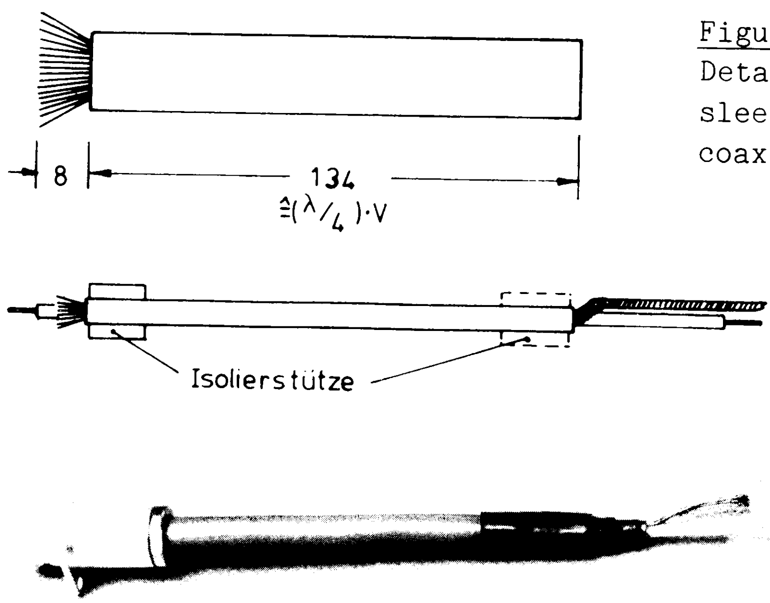
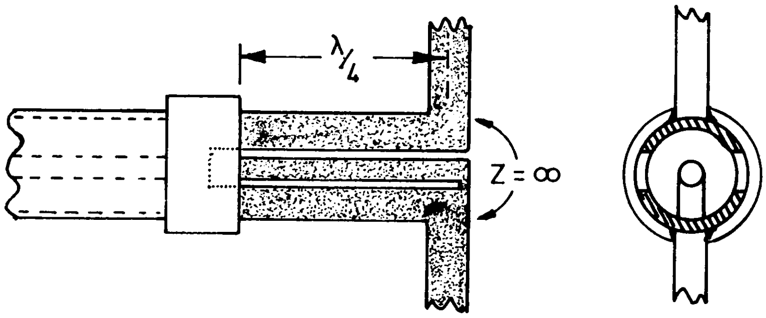


Figure 413
Details of the coaxial sleeve balun made from coaxial cable

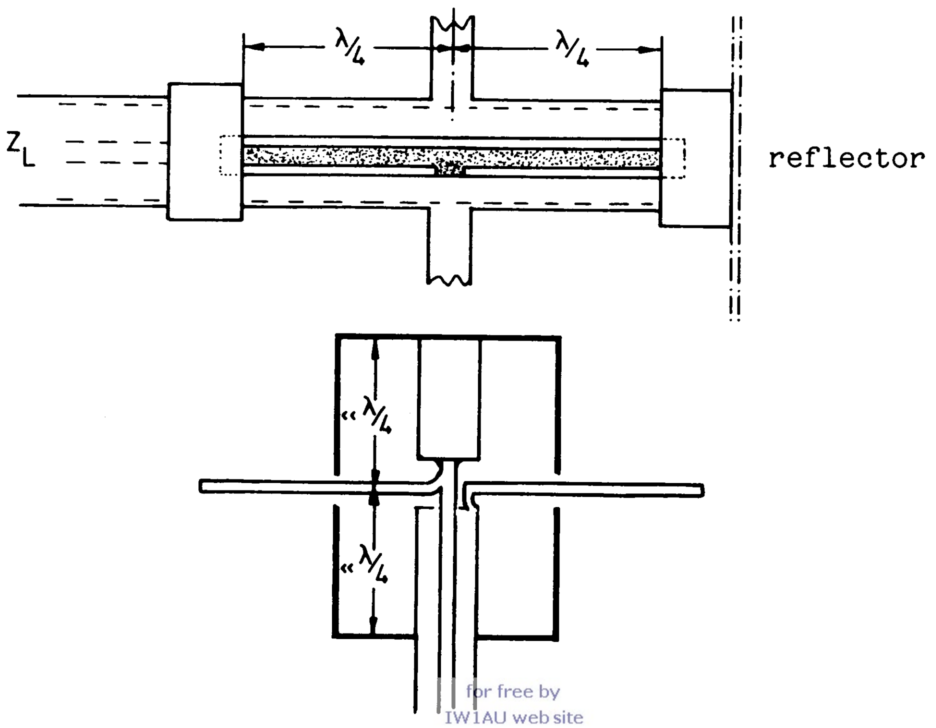
A special form of the Pawsey stub is the balancing by means of a $\lambda/4$ slot in the coaxial cable at the feed point of a dipole illuminating a parabolic reflector. Each half of the dipole is connected to one of the arms formed by the slotting. The inner conductor is also connected to one of the radiators. Assuming that the energy will at the beginning of the slot spread between the inner conductor and that arm of the outer conductor which is not connected to the inner conductor, it is easy to understand that a field between the two arms will build up near the dipole feed and that the field lines will extend from the radiator connected to the inner conductor to the radiator connected to the cable sleeve. Symmetrical radiation in spite of the coaxial feeder is therefore achieved.

Figure 414 Simple slotted coaxial sleeve balun



Balancing of the feeder-to-feed transit could be improved upon by employing a second balancing assembly or slot. An additional advantage of this balancing cavity may be utilized if the dipole is employed to illuminate a parabolic reflector. A small parasitic reflector is installed at the end of the second $\lambda/4$ cavity that reflects any energy (radiated forward) back into the reflector. Each cavity may require shortening in spite of air dielectric due to the capacitance between the insides of the two cavities.

Figure 415 Double slotted coaxial sleeve balun

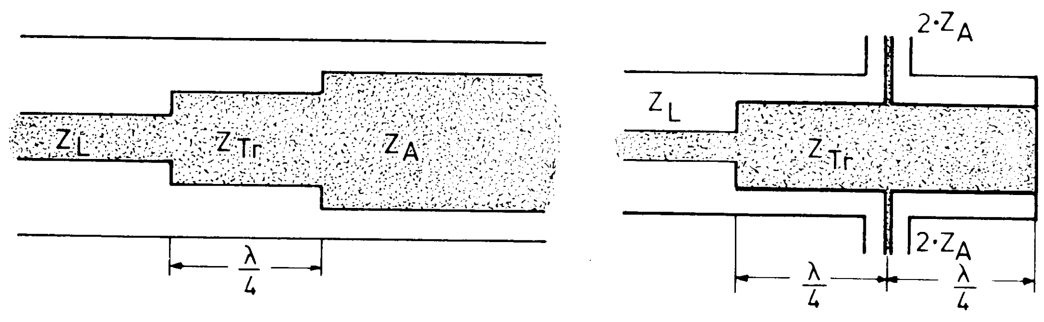


E.3.1 Stacking of aerials

Stacking of aerials requires some means of impedance transformation to retain the characteristic impedance of the line. These transformation devices are usually composed of quarter wavelength cable sections that may require parallel connection to obtain a specific impedance. It is thus frequently a disadvantage that the required impedances cannot be produced with sufficient accuracy.

$$\sqrt{Z_{line} \times Z_{aerial}} = Z_{transformation\ line}$$

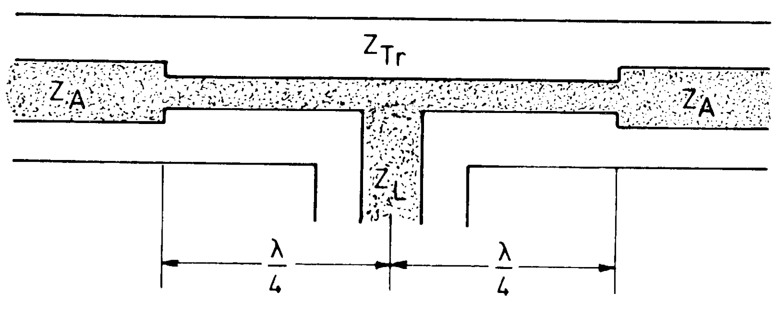
Figure 416



Rather than using one transformation line to match - for instance - two aerials the problem may be solved by employing two transformation lines behind a T - junction (i.e. one per aerial).

Figure 417

T-junction



It is thus obvious that a clear situation exists if four aerials are stacked. In that case the $\lambda/4$ sections may be cut from the same cable as used for the feeder (same characteristic impedance). If no commercial T-junctions are available they may be constructed from copper T-junction pipe fittings. They are mechanically strong, r.f.tight and should be sealed by filling up with resin. A VSWR of around 1.2 is to be expected.

Figure 418 T-junction made from pipe fittings. It is a copper T-junction for oil lines.

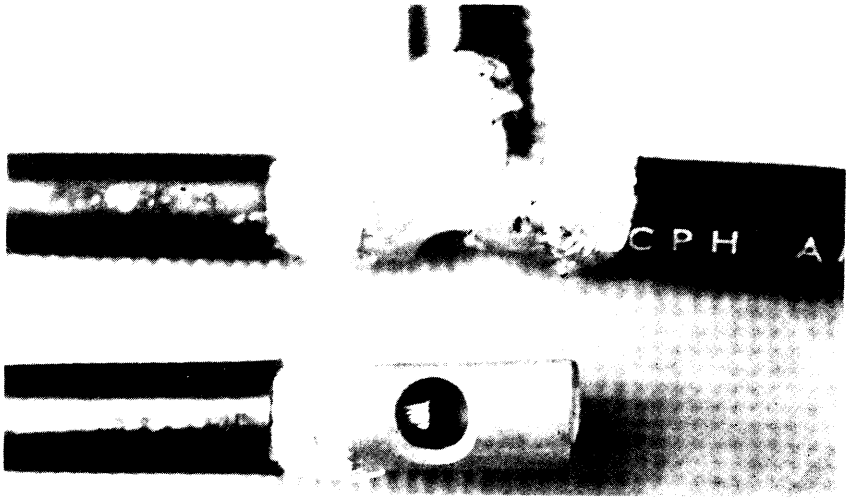
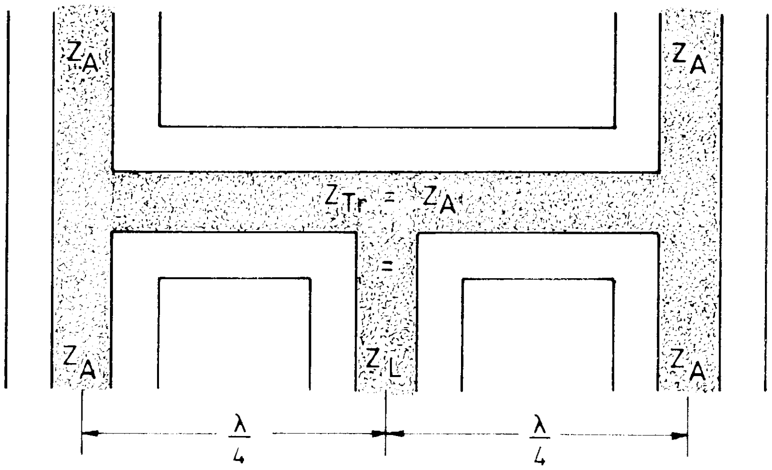


Figure 419 Theoretical arrangement for the stacking of four aerials



The following description shows how well-performing power splitters for feeding two aerials may be constructed using aluminium shapes and aluminium sheet. The U-shape may be obtained from department stores or hobby shops. The inner conductor consists of copper or brass tubing, the diameter of which defines the impedance (A.2.3; G.3.2). For outdoor use the ends should be covered by means of a piece of U-shaped stock. The edges are then sealed using silicon-based sealant. Look out for a brand that does not release acetic acid while curing.

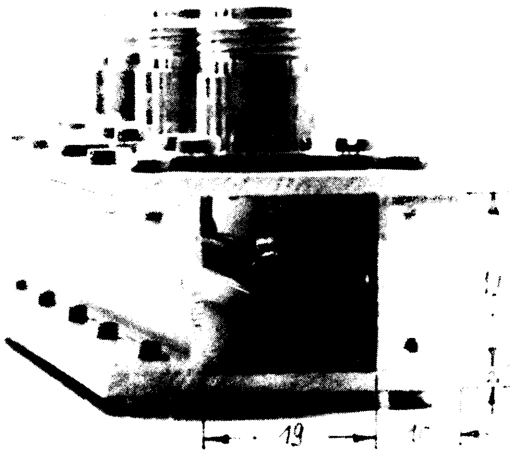


Figure 420

List of materials for 23 cm:

3 N-connectors UG 58/AU
2 U-shaped aluminium profile 10 x 19mm, 150mm long
2 aluminium sheet 2mm, 150 x 39mm
1 copper/brass tubing 6mm diam., 120mm long
2 U-shaped aluminium profile 10 x 19mm, 19mm long

The following photograph illustrates the construction. In contrast to the prototype the connectors should be turned through 45° to avoid interference between the flange mounting holes and the aluminium profiles.

Figure 421 shows the disassembled 23cm version with all relevant dimensions.

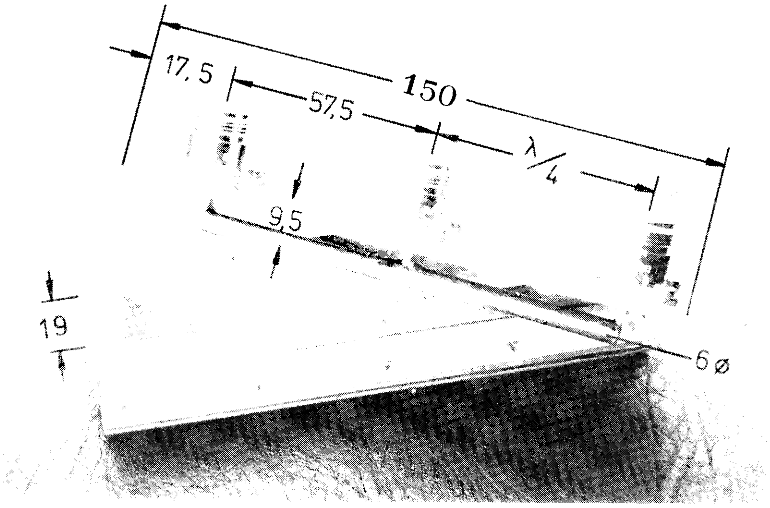
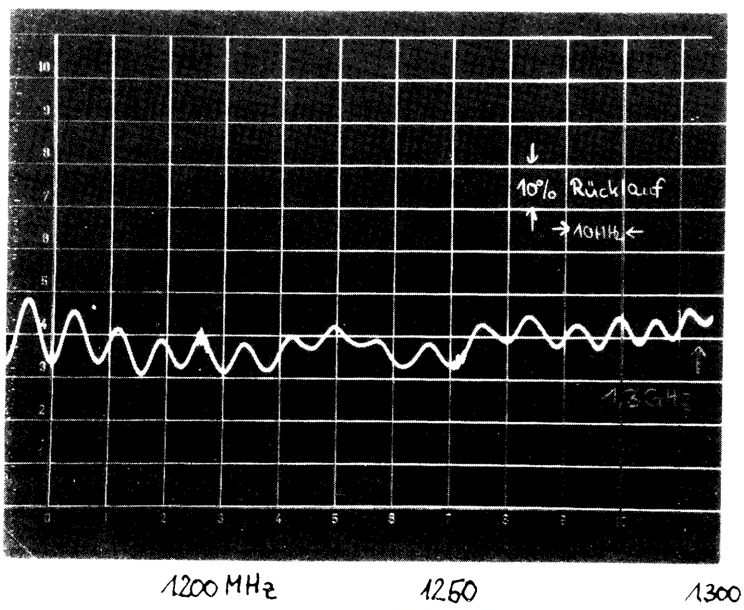


Figure 422 shows the ripple factor over the range 1060 to 1300MHz. The output connectors to the left and to the right were terminated in 50 ohm. r is about 7% which corresponds to a VSWR of approximately 1.15

h = 10 MHz/division
v = 10 % reflection



No problems are encountered when using this type of device in conjunction with 70cm aerials.

Material for 70cm version:

3 N-connectors UG 58 A/U
2 Aluminium U-profile
19 x 10, 380mm long
2 Aluminium sheet 2mm
39 x 380mm
1 Brass/Cu tubing
6mm diam., 350mm long
2 Aluminium U-profile
10 x 19mm, 19mm long
36 screws M3 x 6, cyl. head

Figure 423

Leistungsteiler für das
23 cm Band

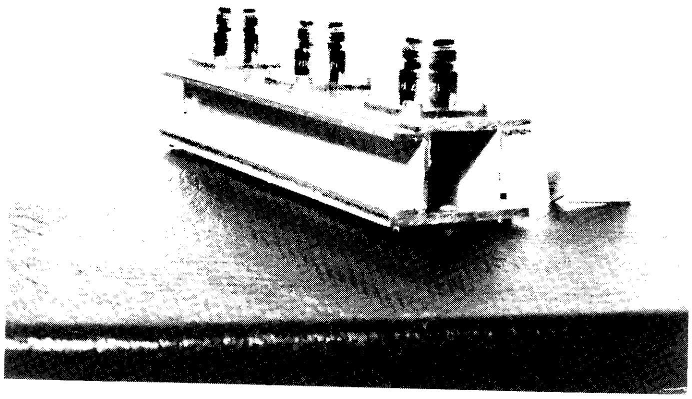


Figure 424

Matching properties described by reflection coefficient r.

r = 7%, i.e. VSWR = 1.15

(Suitable for D.5.9)

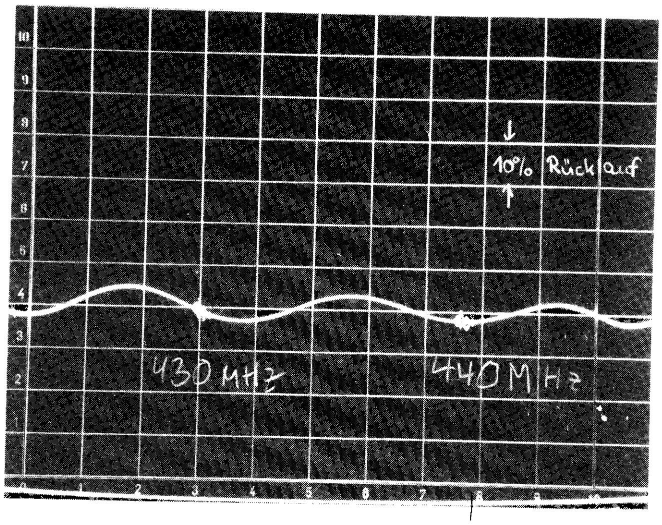
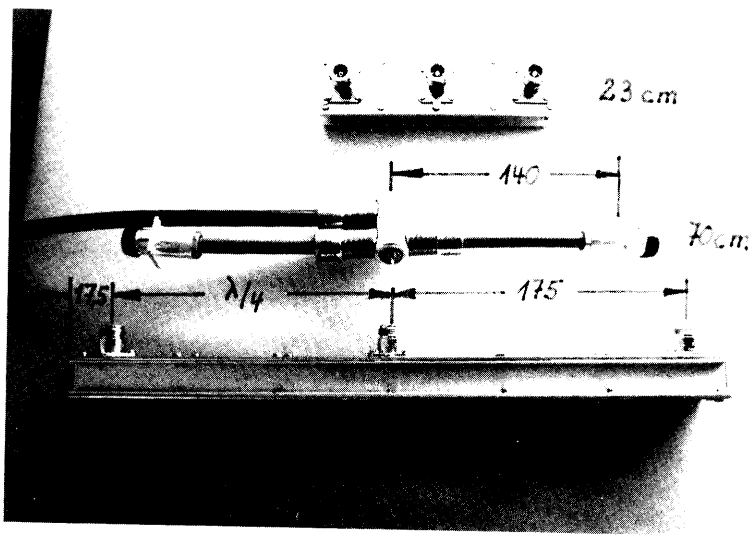


Figure 425 shows the two power splitters constructed from aluminium and a third version based on commercial grade coaxial fittings.

The length of 140mm results from cable impedances that differ from those according to the equation

$$Z_{T1} = \sqrt{100 \times 50} = 71 \text{ ohm}$$



Adjustable $\lambda/4$ transformation line

A power splitter for 70cm aerials is described which allows matching of different feed point impedances to the feeder. It must, however, be pointed out that its duplication requires significant mechanical work involving, for instance, a lathe and a blow lamp for soldering.

The body consists of a brass square section of dimensions 25 x 25mm with a wall thickness of 1mm. The inner conductor is designed to give a characteristic impedance of 20 ohm. Only half of it consists of 16mm diameter brass bar. A piece of brass sheet 78 x 16mm is braised to the underside of the brass bar. The flat section is required to give a sizeable area opposite two tuning discs. These will introduce impedance discontinuities whereby impedance transformation may be achieved in steps. The tuning discs should be covered with Teflon sheet to prevent short circuiting with the inner conductor. The tuning spindles are secured by means of locknuts. The square shaped body is drilled according to the chosen type of connectors and the cover plate is prepared.

The shown prototype employs an unusual type of 50 ohm connector which really stems from the old days of 60 ohm standard. Centering of the inner conductor prior to final assembly is achieved by cutting recesses at 90° intervals into the inner conductor into which the four spigots of the connectors will slide when inserting the inner conductor from the far end.

Figure 426

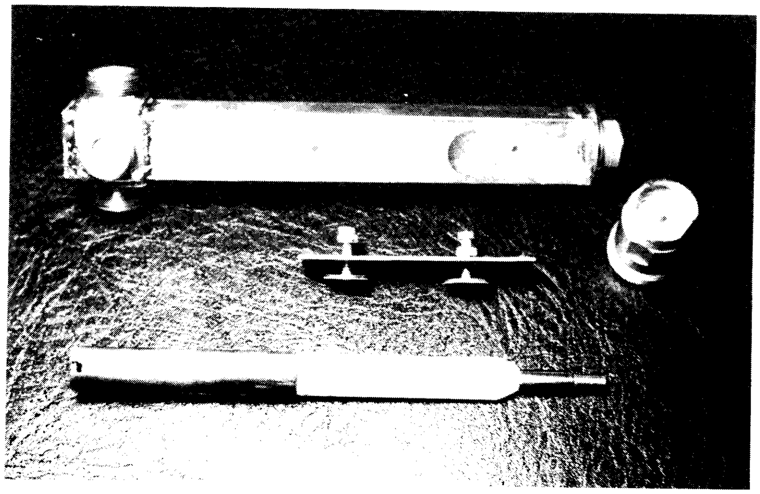
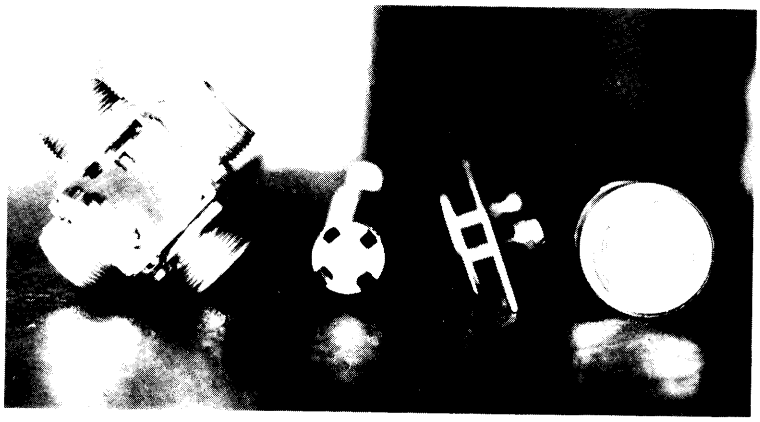


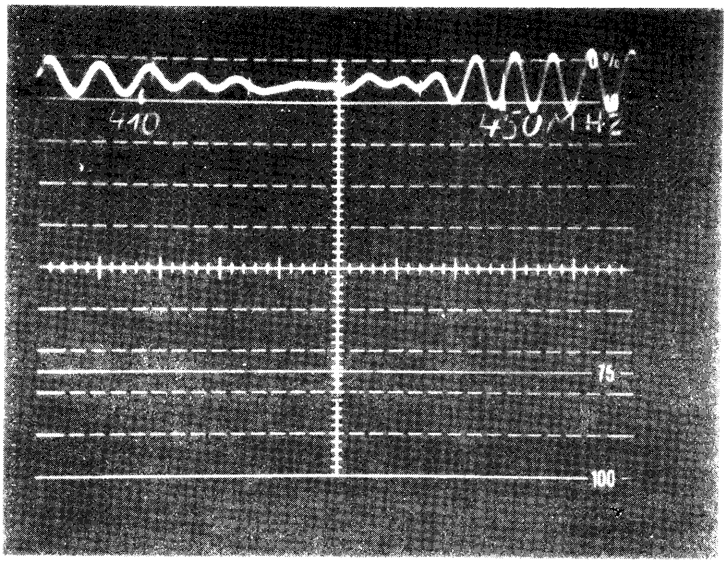
Figure 427

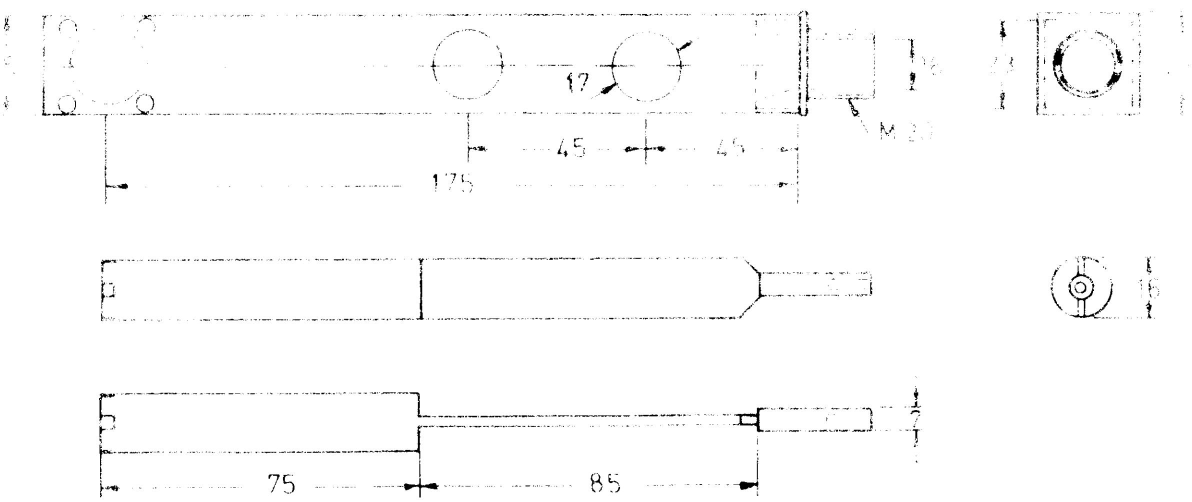


The part which constitutes the transition towards the input connector is turned from stock material and braised to the flat part of the inner conductor. Its free end is drilled and tapped to provide a press fit with the connector spigot. Finally, the two tuning assemblies with their base plate are installed. The inner conductor may now be soldered to the four output connectors and one cover plate only is fixed at the top.

The result justifies the effort insofar as any mismatch between a coaxial feeder of 50 ohms characteristic impedance and any number of aerials (from one to four) connected to the splitter may be compensated to yield a VSWR of 1.05 to 1.2.

This design has a narrow passband, lower losses and provides improved matching when compared with the coaxial cable line transformers. Silver plating of all brass components is recommended to reduce the skin effect.



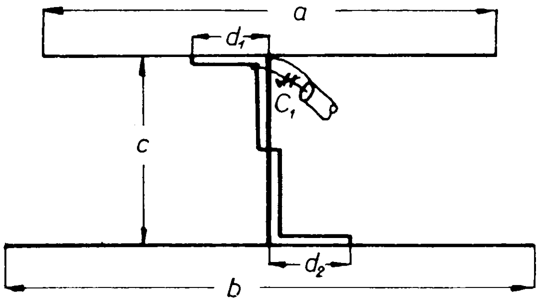


E.4 Construction of proven aerials

E.4.1 The "HB9CV" aerial for 70cm

Particularly during contests and periods of favourable propagation conditions there has been a marked increase in 70cm activities from exposed locations since portable transceivers for SSB and CW appeared on the market. The light-weight HB9CV aerial has gained considerable importance in this context.
When constructing such an aerial a BNC connector (or equivalent) is installed at the feed point. Using a rigid interconnecting cable with plug connectors installed at either end it is possible to mount this aerial directly onto the transceiver. While standing on a rock or tree stump the transceiver will act as base. The aerial shown in the photograph was constructed from 4mm diam. brass rod. After soldering a layer of protective varnish is applied. Alignment requires a VSWR (B.3.2) meter to be inserted into the feeder line and C1 is then adjusted for best signal strength (which corresponds roughly to best VSWR).

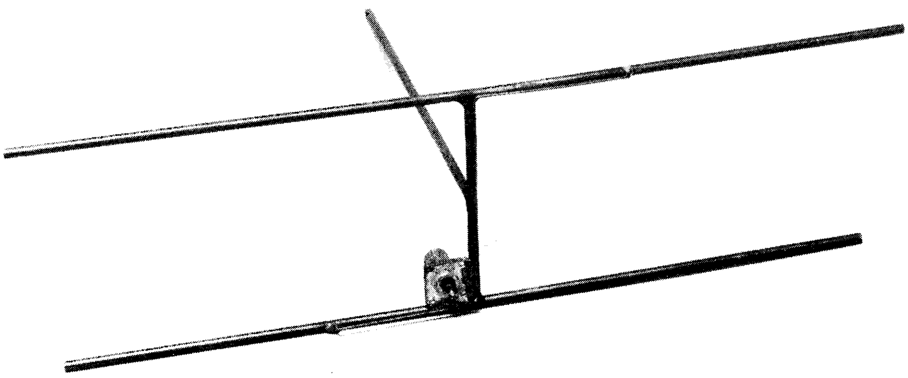
Figure 430 Drawing of HB9CV aerial with main dimensions



	2m	70cm
a	96	31
b	103	33,5
c	25	8,2
d_1	19	5,5
d_2	19	5,7
C_1	0.5-12pF	0.5-6pF

(cm)

Figure 430a Photograph of 70cm version. Spacing between feed system and aerial elements is approximately 3mm. There must not be any other contact apart from the soldered points. The aerial must stand free and is quite useless during rain, snow or - even worse - icing conditions.

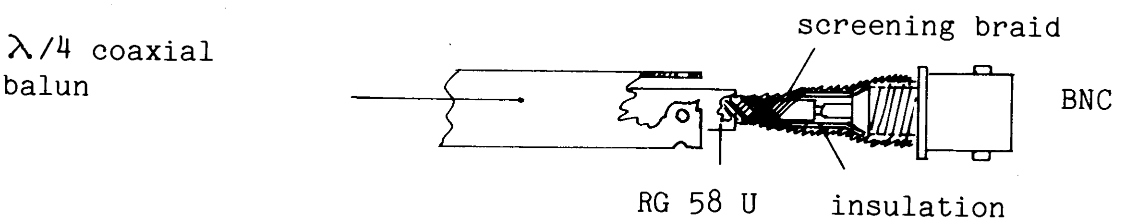


E.4.4 The corner reflector aerial

(D. Reichel DC9NL)

The high gain of corner reflector aerials claimed by ref.(1) led to the construction of such an aerial with the specified dimensions. The $\lambda/2$ dipole employed as primary feed had to be shortened by 2.4cm with respect to free-space dimensions in order to obtain optimum matching. The dipole now measures 29cm in length by 5mm diameter. The explanation is the capacitive loading by the two reflector sides.
The layout of the radiator with its coaxial balun corresponds to the reference dipole used for aerial measurements.
A piece of tubing (internal diameter 10mm) was cut to 14.5cm according to the formula given in E.3 and four holes (2.5mm) were drilled into one of the ends. These holes are required for soldering a BNC connector (single hole installation) UG 1094/U and the screening braid to the brass tube. Needless to say a square- flanged N-connector UG 58 A/U is equally suited.
The inner conductor is composed of a piece of RG 58.
After soldering the inner conductor of the cable to the connector spigot and proper insulation this assembly is pressed into the tube and soldered.
At the far end the cable must be firmly centered inside the tube by means of thermo-shrinkable tubing or PE (inner insulation of RG 214). A block of solid PVC serves for fixing the dipole to the balun.
The $\lambda/2$ radiator is composed of two pieces of 5mm diam. brass tubing and is soldered to the coaxial cable next to the balun.
All soldering points should be protected against corrosion by applying a coat of two-component resin.

Figure 431 Details regarding the mounting of the connector to the balun



The reflector sides are made from perforated aluminium sheet. Angled aluminium stock 3x5cm was riveted to the sides for reinforcement and mounting purposes. Short pieces of this angled material are riveted to the edges and allow adjustment of the angle between the reflector sides. U-channel stock is used to construct the mounting fixture between the mast and the central aluminium reinforcement.

This design exhibited a VSWR of approximately 1.5 at 430 MHz. A series of measurements was carried out (see E.6.3.1) to obtain the gain and the measured value was far from the expected value of 14.5 dB.

A value of 7 dB was achieved only after increasing the reflector angle to 90°. Even this improvement and the one that is to be expected from corrective calculation for the near-field reflections (+2dB between measurements no.8 and no.9 in E.6.3.1) does not justify the considerable constructional effort.
Having been utterly frustrated the specifications in ref. (1) were once again carefully examined and compared with the additional data and diagrams given in (2). And it soon became clear that the gain figures were based on an isotropic reference radiator (dBi) - as is common in Anglo-Saxon literature. Subtracting 2.15dB for correction results in 12dBd for 45° and 7dBd for 90° (and radiator spacing 0.6λ). The latter figure is identical to the measured value. In spite of further experimentation with opening angle and radiator spacing this value of 7dBd could not be improved upon. According to (2) gain figures of up to 18dB are obtained only if the radiator spacing is increased to 2λ which does, however, call for a reflector shank length of 4λ and the physical size of the aerial becomes excessive.

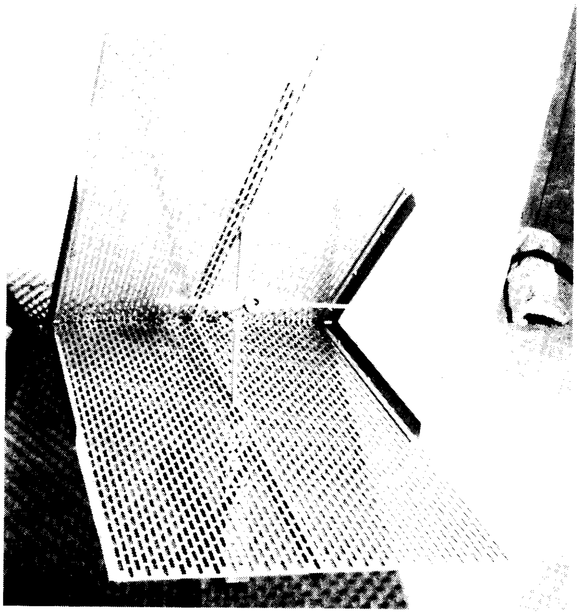


Figure 432
Experimental corner reflector aerial for the 70cm band.
The vertical element supports the radiator. The aerial may be folded after removing the feed assembly.

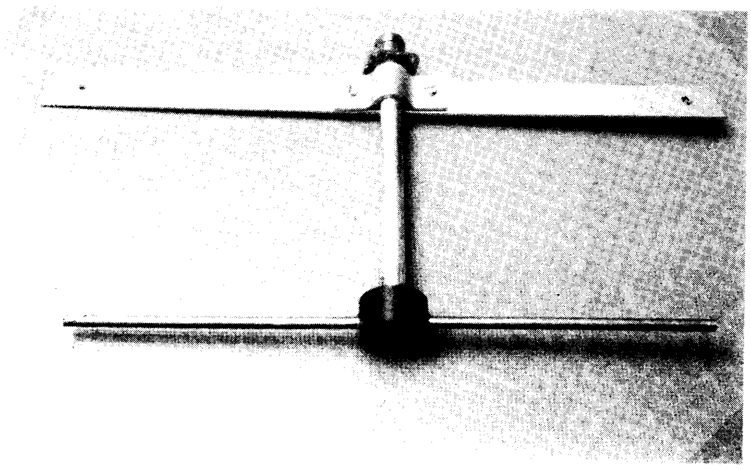


Figure 433
λ/2 dipole with λ/4 coaxial balun.
The support was turned through 90°.

References: (1) Rothammel: Antennenbuch
(2) Jasik: Antenna Engineering Handbook.

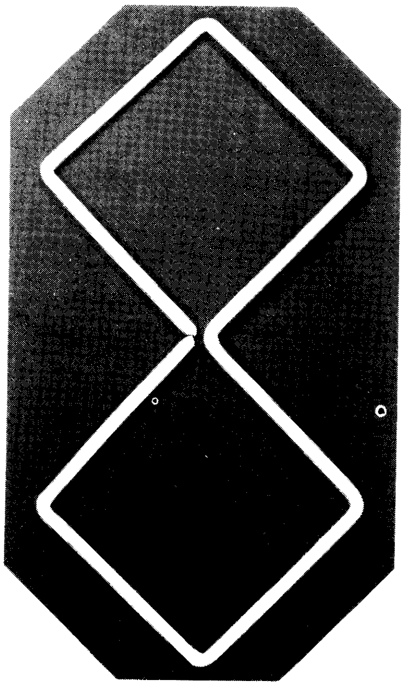
E.4.5 A twin-quad for the 2m band

The twin-quad is yet another type of proven aerial that has reasonable gain and can be constructed in minimum time and for little money. This type of aerial has been around for quite some time on 70cm and 23cm. The results are discussed in section E.4.6
OM Dieter, DL7KM, developed two versions for the 2m band and kindly provided us with the building instructions. Even though this book is basically UHF we feel that we should not withhold this from our readers. In numerous QSOs OM Rothammel, DM2ABK, remarked on the existance of this aerial which will be dealt with in the latest edition of his "Antennenbuch".

The theory of operation of this type of aerial has been adequately covered in various publications (1) and its discussion would go beyond the scope of this book. Who ever has used a quad on short wave knows of its advantages. Its nickname "Queen of DX-aerials" is well founded. Its simple construction, compact and space saving arrangement as well as its excellent relation between performance and expenditure are advantageous on the VHF/UHF bands, too.

The twin-quad now represents the variant of two stacked quads giving improved vertical beamwidth whereby the totally useless steeply upwards resp. downwards directed radiation is suppressed (see also F.9). This results in about 2.5dB of additional gain.

Quad 1



Quad 2

Figure 434
Twin quad with horizontal polarization. Feed point impedance about 60 ohms (balanced); see text for details in case of using reflectors.

During reception this narrow vertical beamwidth will reject interference (car ignition, noise from buildings etc.) originating from the noise field below the aerial - particularly in densely populated areas. These advantages can, however, be fully exploited only if horizontal polarization is chosen - as may be seen from the half-power beamwidth documented by the radiation patterns. Additional information is contained in the following figures and notes.

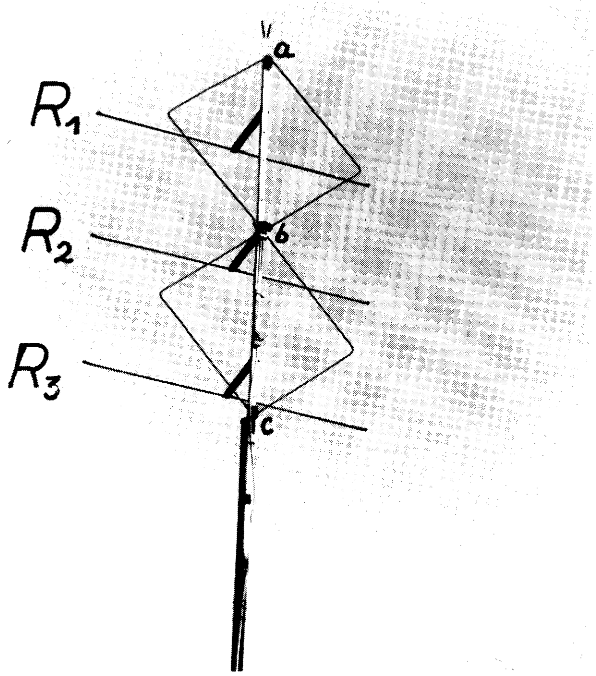


Figure 435

This photograph by DL7KM shows a simple twin-quad having a gain of ~10dB. Feed point (~60ohm) at point b.

The twin-quad element has a total circumference of 416 cm and is made from 5mm diam. copper wire. This wire may be obtained from radio and t.v. installation shops under the designation 16mm^2 insulated aerial grounding wire. Each side measures 52cm (=λ/4). The three reflectors are 1.05m long and consist of 8mm diam. aluminium tubing, wall thickness 1mm. Reflector-to-driven-element spacing is 27.5cm. The spacing between the reflectors is 51cm. The radiator is held at points a, b and c by insulated supports mounted to the boom. These supports are fabricated from plastic material and need not have high insulation values since fixing is at cold points corresponding to maximum current at minimum voltage. The hot points are at the sides.

The feed point impedance is approximately 60 ohm, balanced. It is, however, possible to connect the coaxial cable directly to the aerial without appreciable losses. All parameters remain unchanged, only the radiation pattern will be shifted slightly to the left or to the right with respect to the main axis depending on the connection of inner and outer conductor. This must be borne in mind when attempting to take bearings - an application for which the quad with its beamwidth of 67° is hardly suited.
In the interest of completeness a suggestion for balancing is included:

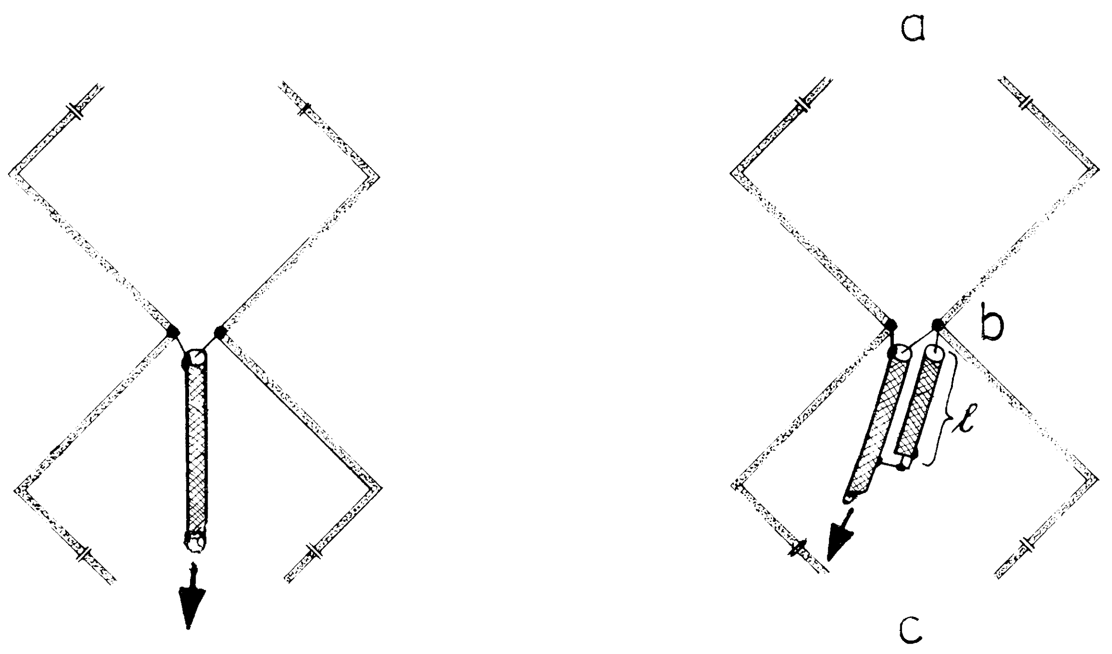


Figure 436
Unbalanced and balanced feeding of the quad aerial (horizontal polarization). The $\lambda/4$ line balancer should be cut from the same cable material that is used for the transmission line. The screening braid of this line is not soldered at point b. The spacing between the transmission line and the $\lambda/4$ section is shown for clarity only. The insulating sleeves remain on the cables. The feed assembly should be potted in resin.

$l = \lambda/4 \cdot v$
v = velocity factor of the cable material (approximately 0.66 - 0.81 depending on the dielectric - see G.1).

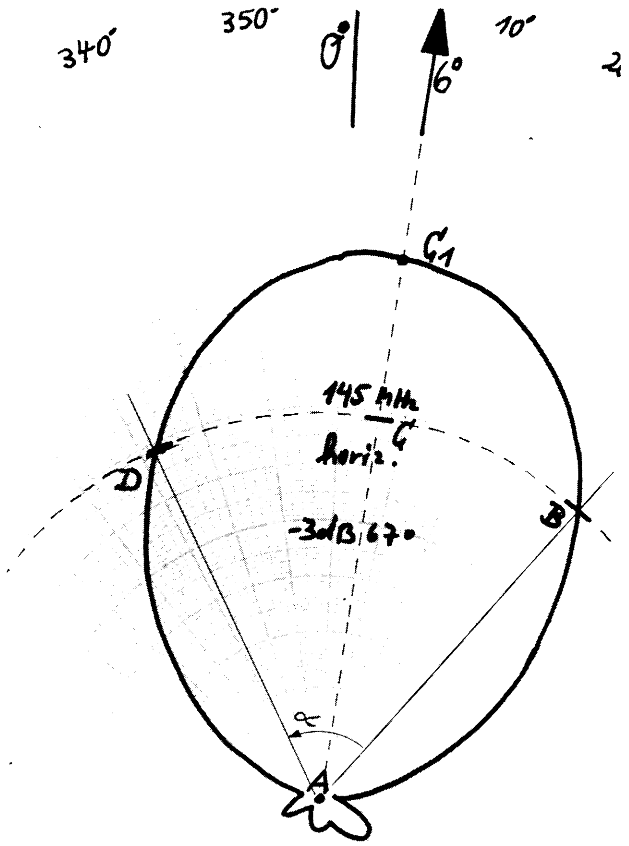


Figure 437

This diagram shows the horizontal radiation pattern of the twin-quad and the half-power beamwidth (α). This angle may be obtained by the following method: A transmitter and an aerial are positioned 500 to 800 metres away. It will give rise to a voltage at the output terminals of the aerial under test - the voltage being variable as function of the aerial direction. The voltage is measured and plotted against the aerial direction on polar co- ordinate paper (figure) to obtain the radiation pattern. The aerial is then turned so that its electrical axis points at the distant transmitter and a 3 dB attenuator is inserted into the transmission line. The measured voltage is entered into the diagram and constitutes point C. Next, draw a circle of radius AC about point A. It will cut the radiation diagram in points B and D. The angle between lines AD and AB is the half-power beamwidth (α) and its value was measured as being 67° for our specimen. The vertical half-power beamwidth was measured to be 54 degrees. The asymmetrical feed by the coaxial cable resulted in a 6° shift of the electrical axis from the geometric axis.

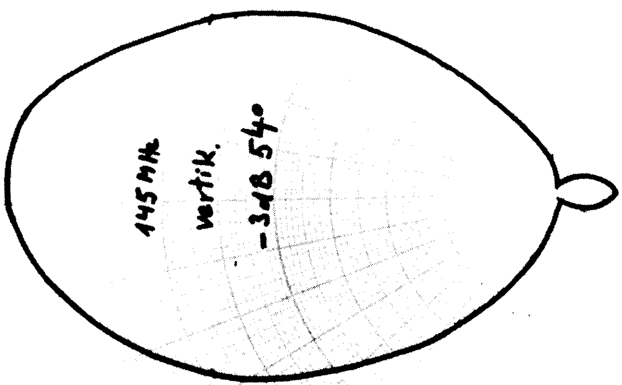


Figure 438

Vertical radiation pattern of the twin-quad

Different types of booms are used by the members of the "North- East Bavarian UHF-Group" (NOBUG). They use either a reinforced version based on mild steel tubing or the lighter version made from 20x20mm square-shaped aluminium profile with this latter version having certain advantages.

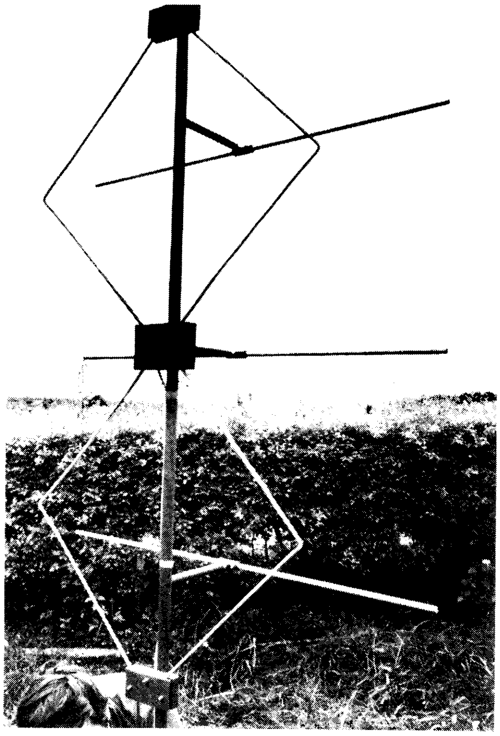


Figure 439
Reinforced version of twin-quad by OM Helmut, DK5RR
Since each quad is composed of two half-wave radiators this aerial could be termed a 7-element array.

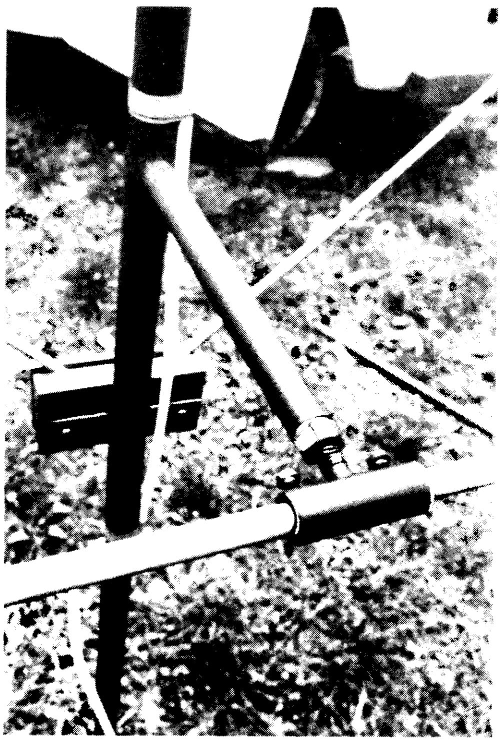


Figure 440
Reflector mount with T-element and locknut for the reinforced version. This allows the varying of the spacing. It is thus possible to match coaxial cables having characteristic impedances of between 50 and 75 ohm with optimum SWR (1.1) to the aerial.

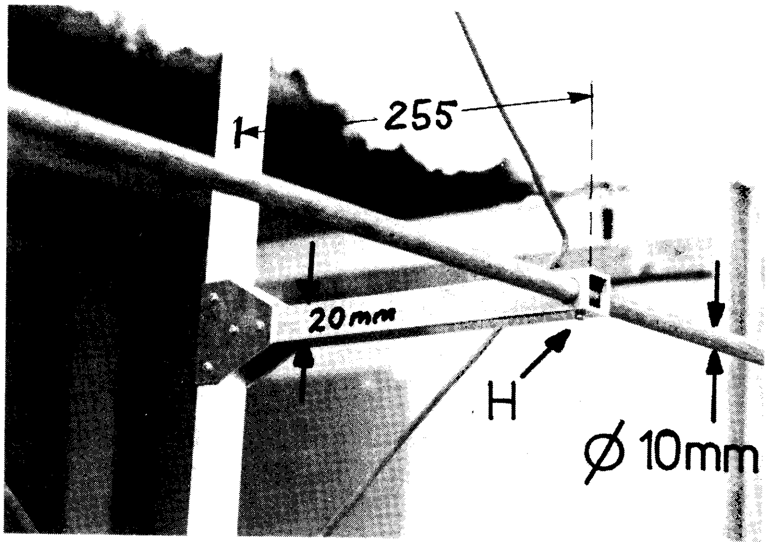


Figure 441

Light-weight version of the twin-quad built by OM Fred, DL8KH, as used by him as well as DL8KF and DJ9HO. Boom and reflector supports are made from 20 x 20 mm aluminium stock of square cross section and with 2mm wall thickness. The reflector support is fixed to the boom by means of plates cut from 2mm aluminium. Four self-tapping screws provide the required rigidity. The photograph shows one of the three reflectors. It is made from 10mm diam. aluminium rod and is secured centrally by means of a bolt in position H.

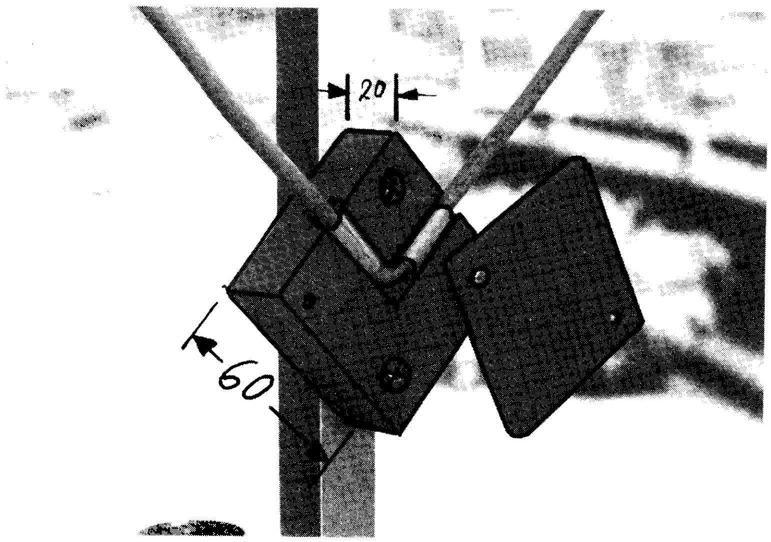


Figure 442

Mounting of the quad elements which are composed of 5mm diam. wire. The mounting block at the upper end is identical. The plastic material from which these blocks are made (shown with the cover plate removed) need not have high quality dielectric properties. Evidently the three reflectors may be replaced by a wire mesh reflector (E.4.6) having dimensions 110x180cm and 3 to 5cm width of mesh. The reflector-to-radiator spacing must be readjusted (insignificant deviation). This improves the gain by further 2dB. Wire mesh inside the house wall (hi) provides the perfect reflector for a 2m resp. 10m quad array.

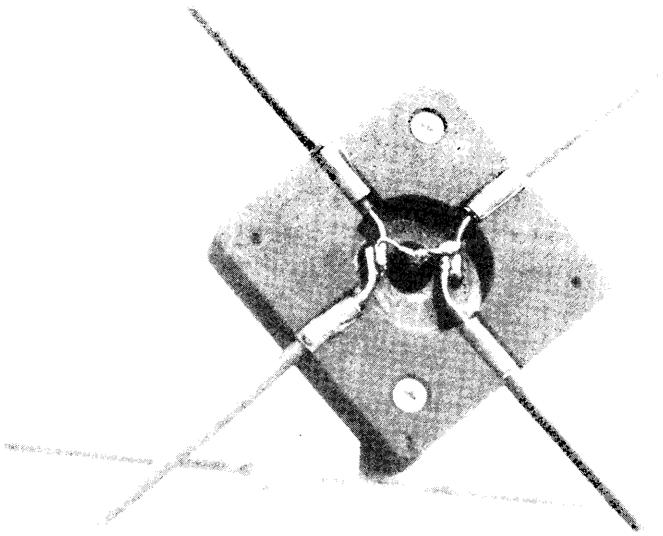


Figure 443
Feed assembly; the cover plate is removed. A round file is used to file the grooves according to the quad-wire cross section for best fit and sealing. A two-component resin is used to fill all cavities. The coaxial cable is threaded through an opening into the boom from where it is run to the station.

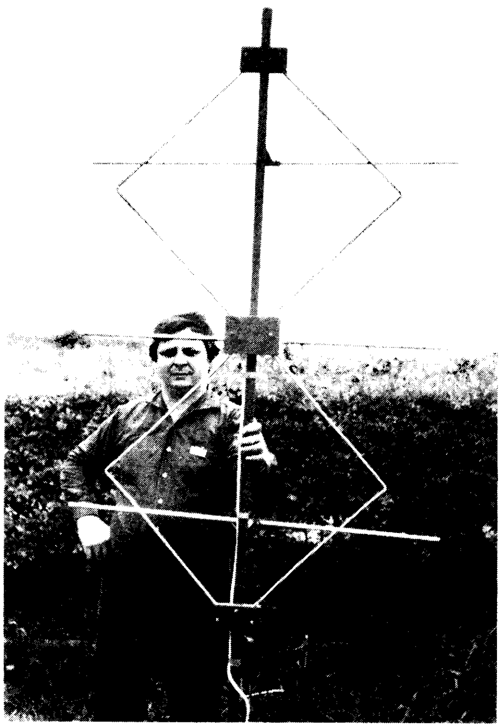


Figure 444
Comparison of dimensions: OM Helmut, DK5RR, standing behind his quad aerial (reinforced version) and looking into the main radiation direction.

It is obviously possible to combine several quads to an array. Four equal aerials (quadruple combination) result in theoretically 5 to 6dB of additional gain. Such combinations are operated in both Berlin and Dresden with a claimed gain of 16dB. One OM supplied the dimensions shown in figure 445. The vertical quad-to-quad spacing is specified as being 20cm. From theoretical considerations the best vertical spacing should be 56cm.

Reference: (1) Rothammel, Antennentechnik

Figure 445

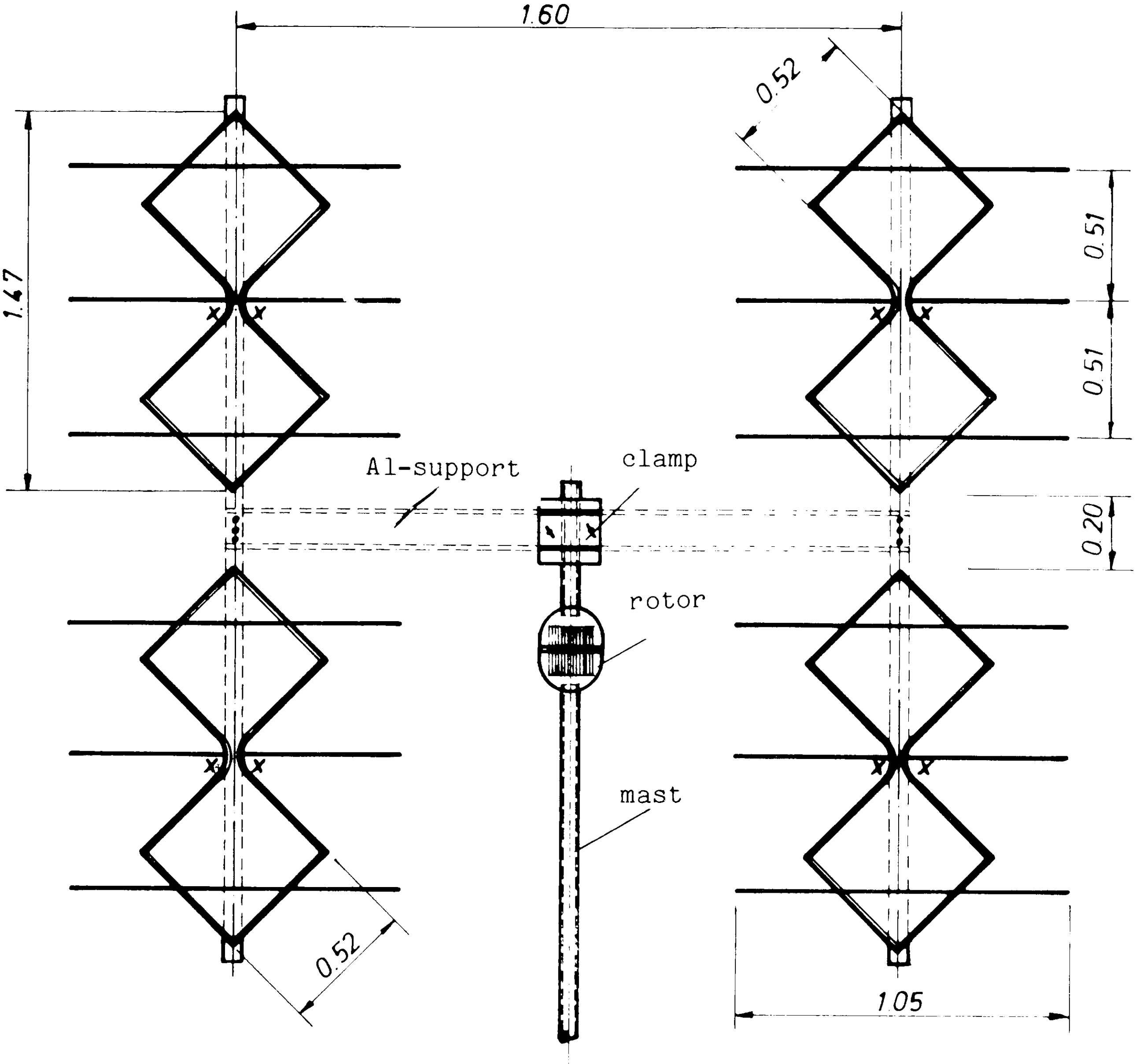
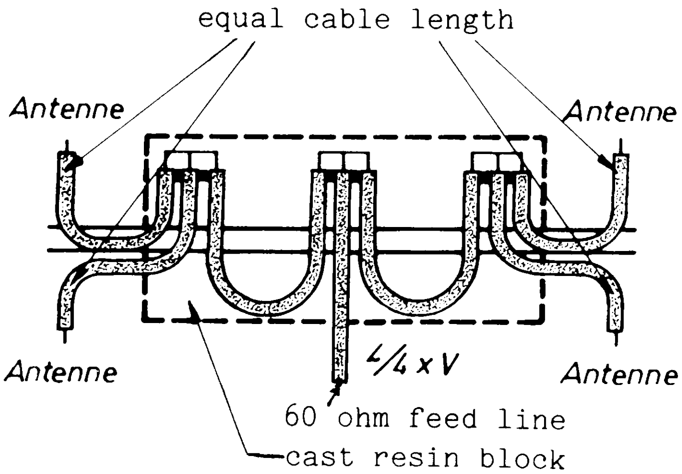


Figure 446



All dimensions in metres!

All inner conductors of cables to be connected to the same side of the radiators!

E.4.5.1 An extended twin-quad for the 2m band

Anybody who is ready for a slight additional effort to improve the gain of the twin-quad without going through the trouble of constructing the full-size beam described in E.4.5.2 should consider the extension described below.
Adding two directors per system and several reflectors will increase the gain by approximately 2 to 3db with respect to the basic version. The photograph (figure 447) shows the aerial system at DB8NP with the aerial to the right being the one in question. The aerial to the left is a twin-quad beam according to E.4.5.2 designed for the 70cm band. The dimensions of the extended twin-quad are included in the drawing prepared by DB8NP.

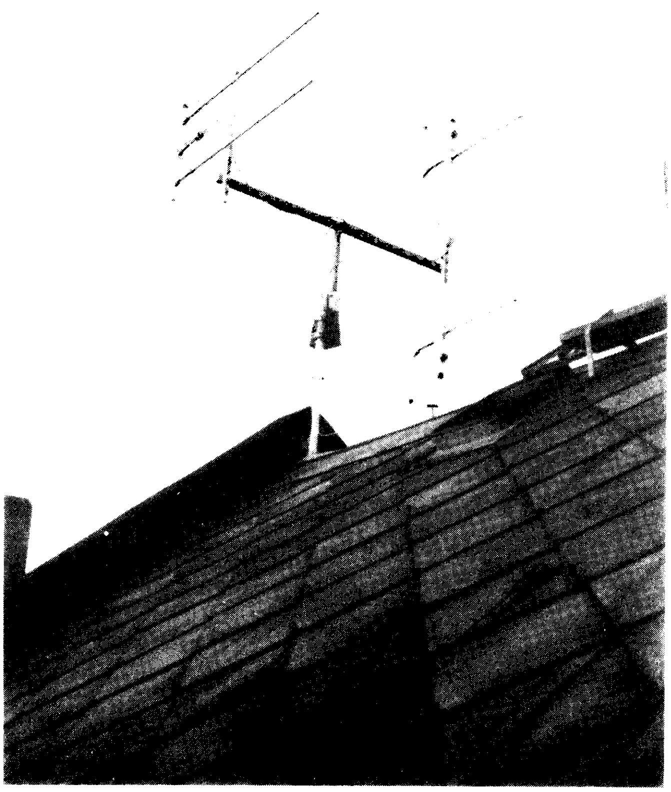
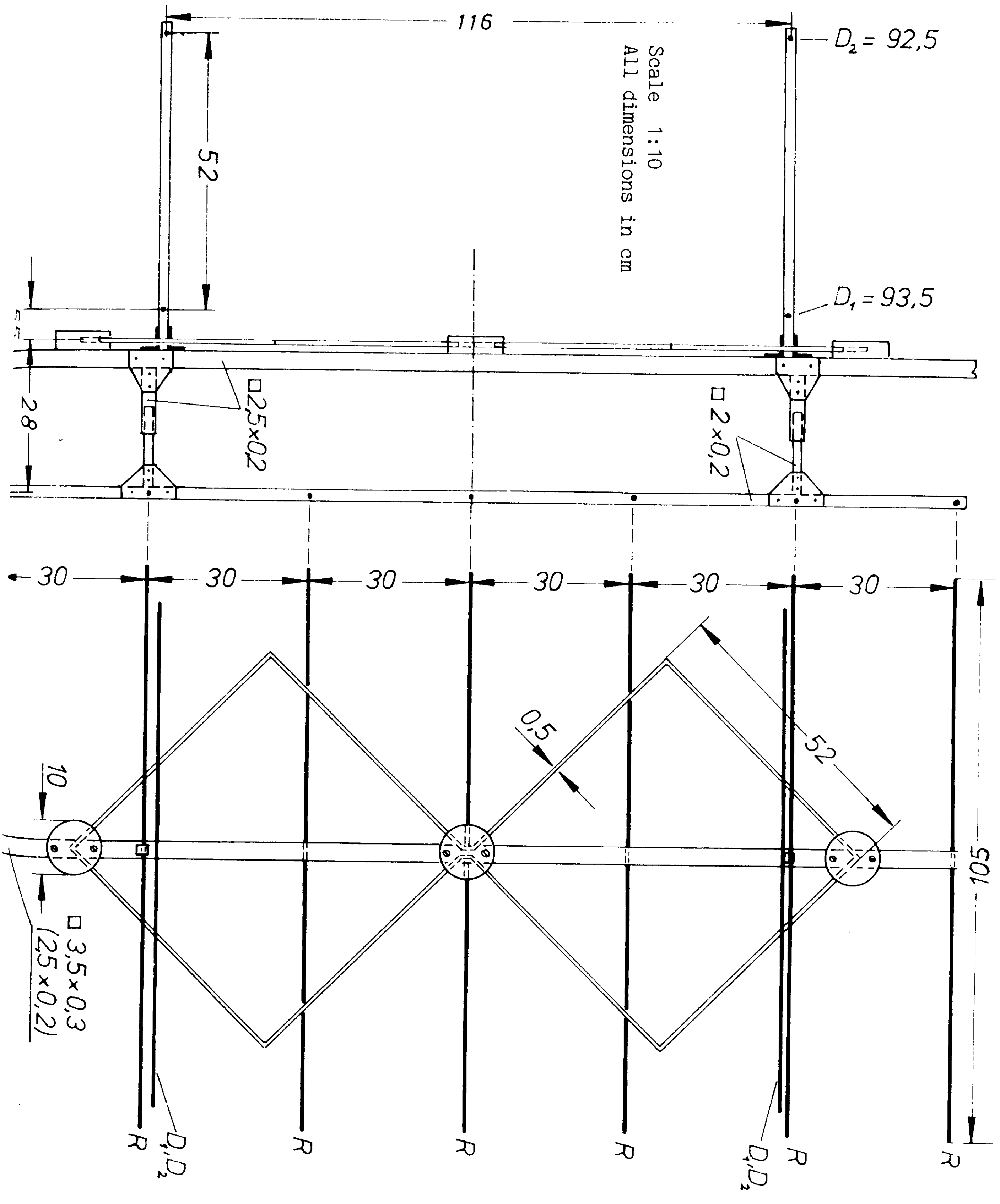


Figure 447
Aerial system at DB8NP consisting of the extended twin-quad for 2m (right) and the twin-quad beam for 70cm (left).
The 70cm version of the beam is described in E.4.6.1.

Figure 448 Dimensions of the extended twin-quad. The gain was optimized by the addition of two directors per system and a total of seven reflectors. The reflector assembly is installed by means of sliding supports which allow minimizing the SWR. The twin-quad radiator is held by blocks of plastic material. The spacing between the director groups is less than 120cm due to mechanical considerations and improved VSWR.

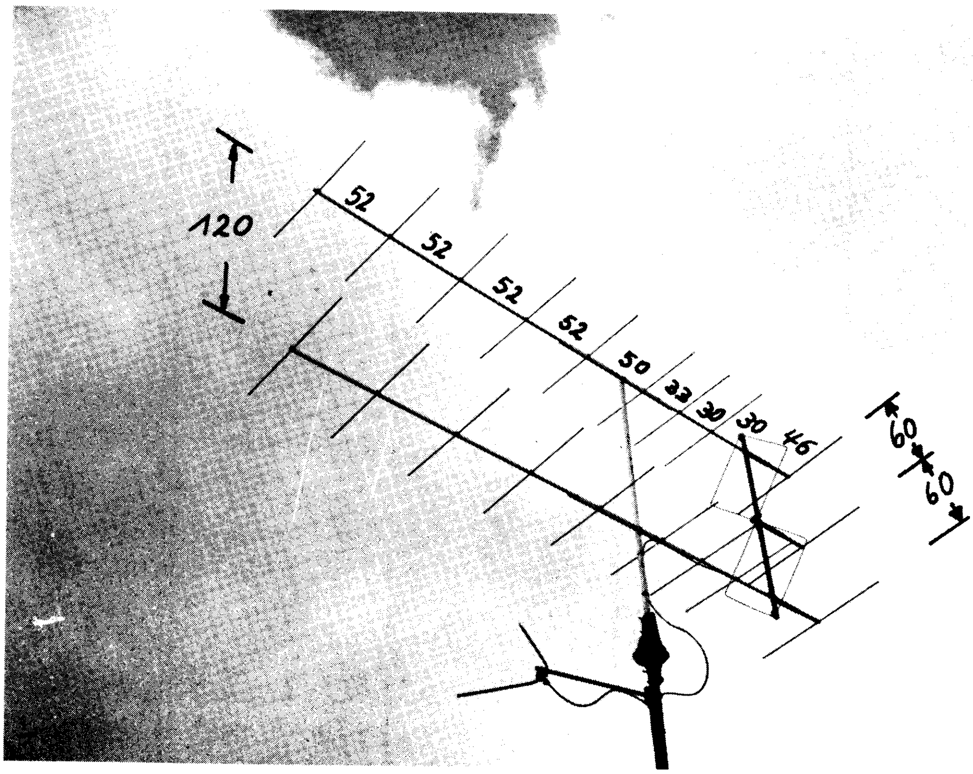


E.4.5.2 The twin-quad beam with 15dB of gain

The good results achieved with various versions of twin-quad aerials encouraged OM Dieter, DL7KM, to design a twin-quad beam which is referred to as "DL7KM-2m-Beam" in technical publications.
This version has the following distinct advantages over the quadruple combination:

1. Less material required,
2. no matching, phasing or transformation lines,
3. only one soldering/connecting point exposed to the weather,
4. the extremely narrow vertical beamwidth (~30°) reduces any energy from being radiated ineffectively upwards and downwards; it also attenuates interference (car ignition, household appliances etc.) originating from beneath the aerial and generally ensures optimum utilization of the radiated power.

Figure 449
The "DL7KM-2m-beam". All dimensions shown are in cm. OM Dieter claims a gain of 15dB. An Omni-V aerial may be seen beneath the rotor. See also modified version in figure 451 ff.



Performance figures:

Gain according to DL7KM	15 dB over dipole
Horizontal beamwidth (E-plane)	33.5 degrees
Vertical beamwidth (H-plane)	30.5 degrees
Back-to-front ratio	25 dB (approximately)

Physical data:
All directors and reflectors are cut from 8mm diam. tubular aluminium stock, wall thickness 1mm. A stronger construction is presented below.

Required material:

3 elements of length 1050 mm R1, R2, R3
2 elements of length 935 mm D1
2 elements of length 930 mm D2
2 elements of length 925 mm D3
2 elements of length 920 mm D4
2 elements of length 915 mm D5
2 elements of length 910 mm D6
2 elements of length 905 mm D7
2 elements of length 890 mm D8

The two director assemblies are identical. DL7KM mounted all directors and reflectors to the top of the boom as is common practice in t.v.aerials. The quad element has the above mentioned dimensions U = 4.16 m, s = 52 cm, diam.= 5 mm. It is recommended to start with a length of 420cm and to provide 4cm of overlap at the feed point. A coat of protective varnish should be applied to the aerial if non-insulated copper wire is used. The reflector-to-reflector spacing of 60cm deviates from the simple quad as well as the reflector-to-radiator spacing.
The booms are 410cm long pieces of 15mm square-section aluminium stock having a wall thickness of 2mm. The booms are reinforced around the centre of gravity by 120cm long sections of the same material. The vertical support (mounted to the rotor) is again square-section aluminium stock 35 x 35mm with a wall thickness of 3mm.
The vertical member to which the quad element is fixed has a square cross-section of 10x10mm and a wall thickness of 2mm. Holes of 10mm diameter are drilled 46cm off the rear end of the booms and filed to a square shape. The vertical member is then pushed into these holes and secured by bolts.

It could be advantageous - particularly for the use in a rough climate - to construct a stronger version of this aerial; since the spacing between the quad-cum-reflector assembly and the director arrangement should be adjustable for optimum tuning and SWR the authors recommend the construction of a stronger version and in four steps according to the following scheme:

1st step:

Twin-quad and reflectors with support up to point "A" in figure 450. (Attention: Reflector spacing here 46cm).

2nd step:

Main vertical support (~2.50m) with the recesses at the top end, the cover (B) and the square-shaped passage (15x15mm) which is again obtained by drilling and filing.

3rd step:

Installation and fixing of the booms to the main vertical support and installation of the diagonal braces (A1, see detail C).

4th step:

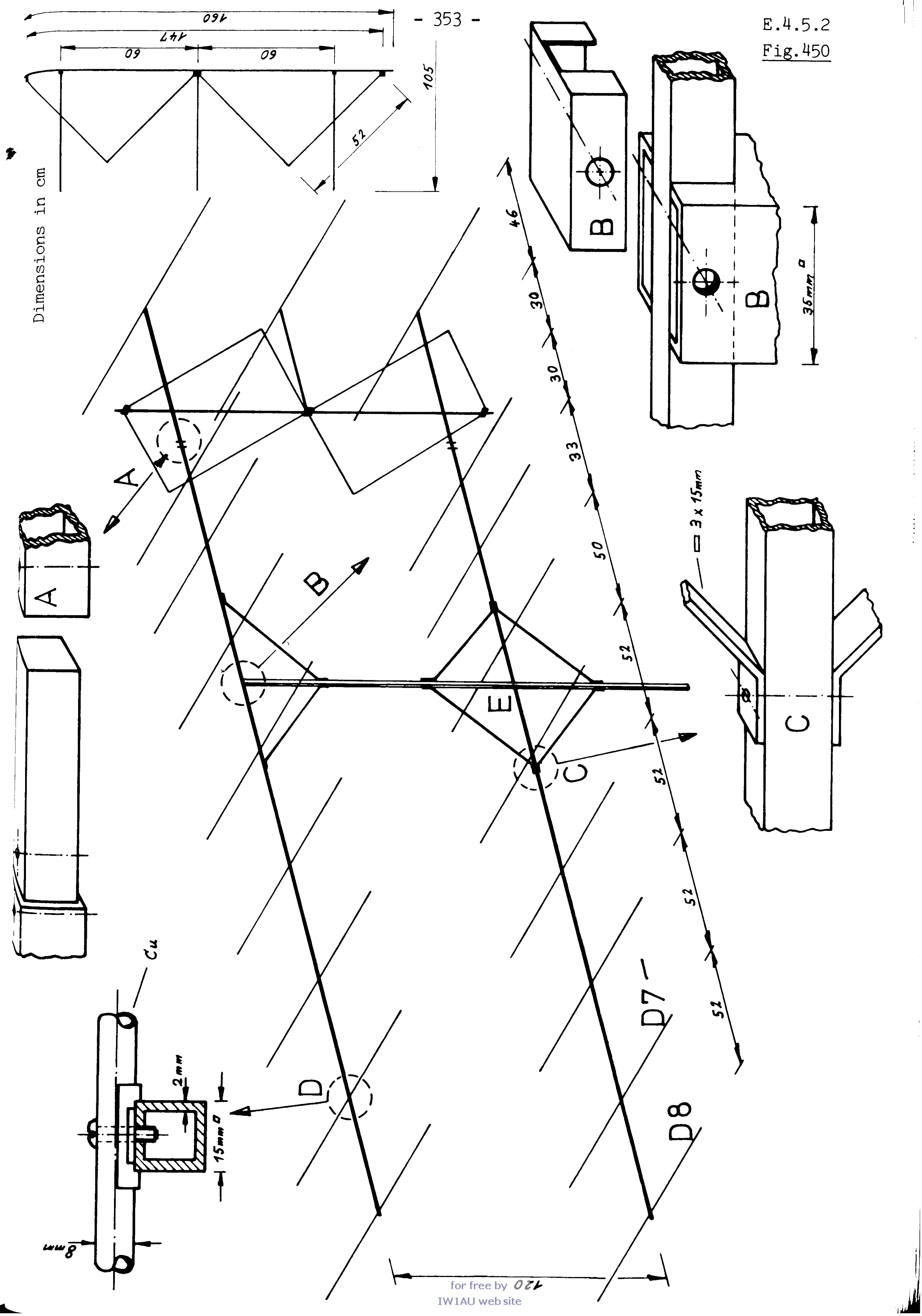
Installation and securing of the directors (see detail D).

The coaxial transmission line is then connected as described at the beginning of this section and the spacing between the first director and the quad-cum-reflector assembly is adjusted for best SWR. For this the aerial should be 2.5 x λ off the ground. Should this turn out to be impossible it will suffice to point the aerial 45^o upwards - making sure that there are no obstacles in front of the aerial. Obstacles in the near- field affect the radiation properties and thus the standing wave ratio.

Apart from that all details are enclosed in figure 450. Furthermore there are no limits to even stronger construction. It is merely essential to maintain the element dimensions and spacings.

E.4.5.2

Fig.450



Dimensions in cm
160
147
60
60
105
52
A
B
C
D
E
D7
D8
46
30
30
33
50
52
51
52
51
120
36mm□
□ 3 x 15mm
Cu
2mm
15mm□
8mm

The description is rounded off by mention of a modified version which already includes the stronger construction. Seven reflector elements were installed for increased gain. Different spacings were thus to be expected.
The following pictures show the modified version for the 2m band. It is possible to reach each aerial from the roof top since they are mounted onto a transverse support.
It is possible to add a dish aerial for the 23cm band at the intersection of the vertical and the transverse supports.

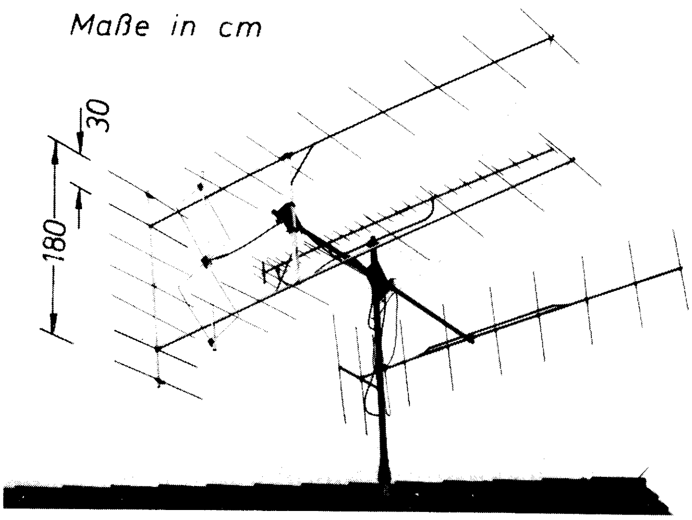


Figure 451
Arrangement and dimensions of the reinforced version of the twin-quad beam with seven reflectors. The gain was compared with that of the eleven-element Yagi visible in the background which was at the time arranged for horizontal polarization and was found to be 6dB over that of the Yagi.

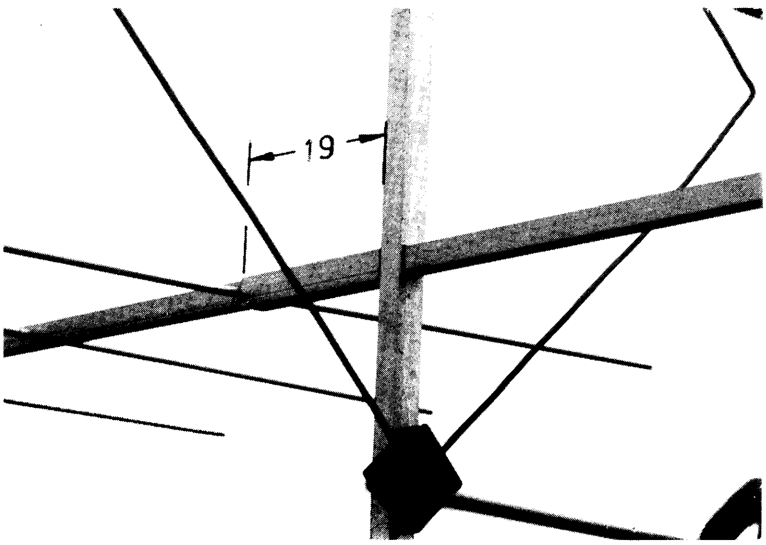


Figure 452
Enlarged portion of figure 451 showing the different spacing compared against that of the DL7KM beam. The vertical quad- element support carries the boom. The square-shaped passage for the boom was obtained by drilling and filing.

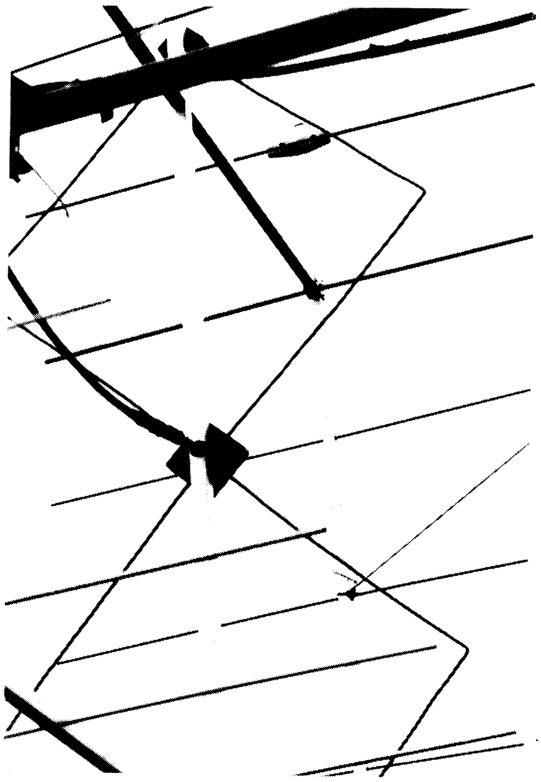


Figure 453

The photograph shows the feed of the twin-quad. The coaxial cable is insulated and runs through the main vertical support. At no point must it come into contact with metal parts. All parasitic elements - except the top and bottom reflectors - are pushed through the respective supports and secured by means of self-tapping screws. All director and reflector dimensions remain unchanged from the basic version described. It is advisable to seal all screw-connections with spray-on varnish or automobile underbody sealant.

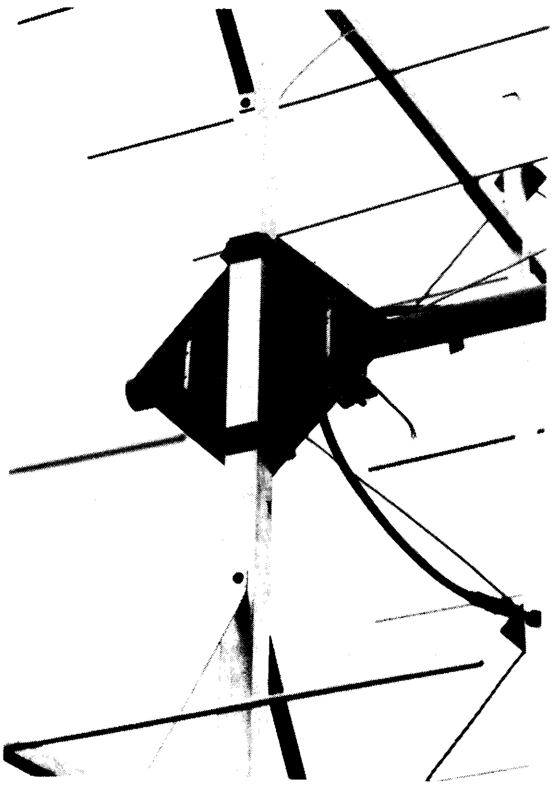


Figure 454

Mounting assembly at the transverse support. It consists of a coated square-shaped steel plate (25x25cm, thickness 4mm). The clamps holding the vertical aluminium support are also formed from 4mm mild steel sheet. Two threaded (M8) studs are welded to the clamps, pushed through corresponding holes in the steel plate and secured by nuts from the rear. The mounting assembly between the transverse and main vertical support follows the same design. The four diagonal braces from aluminium strip are fixed by bolts going right through one brace, the support and the opposite brace.

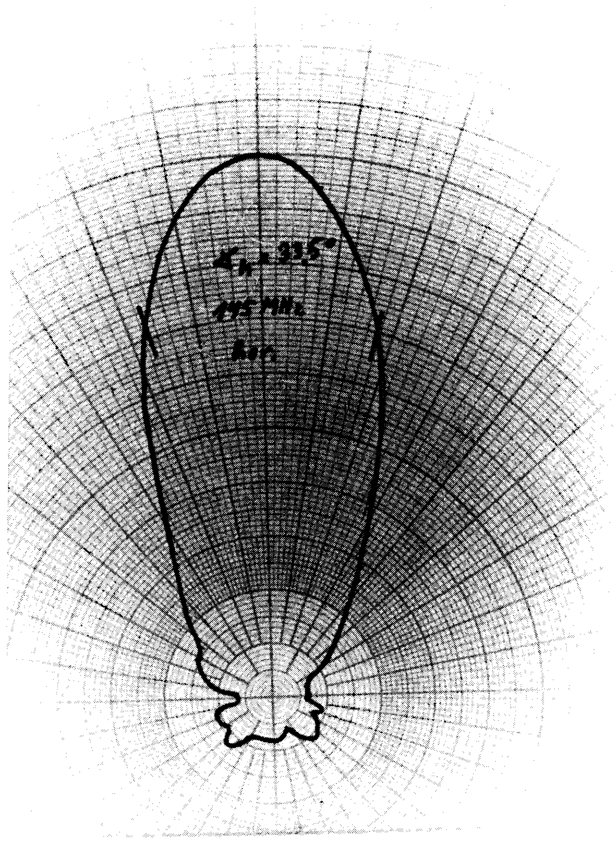
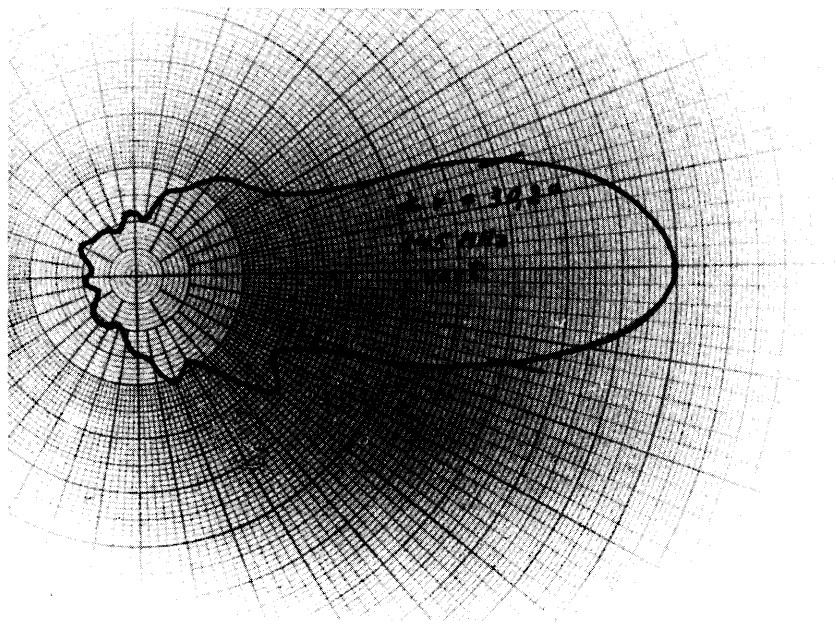


Figure 455
The photograph was provided by DL7KM and presents the horizontal radiation pattern of the twin-quad.
The measured data of both the horizontal and vertical patterns indicate a sharply focused main beam with very low sidelobes.

Figure 456 shows the vertical radiation pattern with a half- power beamwidth of approximately 30°. (See E.1 for definition of half-power beamwidth and the diagrams in E.4.5).



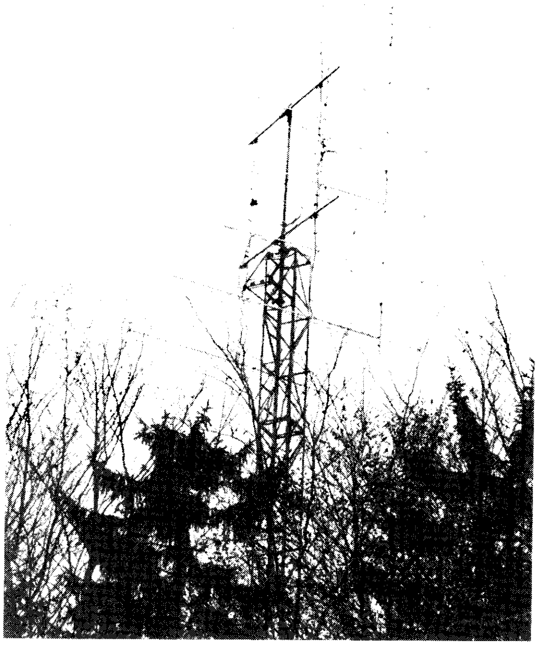


Figure 457
Stacked array of six twin-quad beams. The two photos by DB5NQ show the aerial system of DC9NI and DL4NA. The system has the following specification:

Length: 4.0 m
Height: 6.0 m
Width: 3.15 m
Weight: 50 Kg
SWR: 1.2

The horizontal half-power beamwidth is approximately 16 degrees, the corresponding value for the vertical beamwidth is approximately 10 degrees. The resultant gain is 20 dB which corresponds to a power amplification factor of 100. The vertical spacings are 120cm. The vertical braces are also made from aluminium. This array is composed of reinforced versions of the twin-quad.

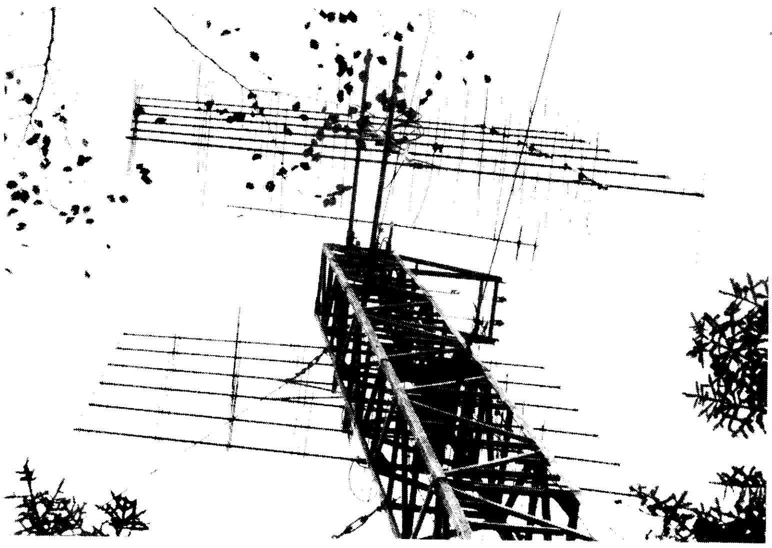


Figure 458
The aerial system seen from below. The 70cm Yagi installed in the centre looks rather meagre.

E.4.6 Twin-quads for 70cm and 23cm

When operating from exposed locations it is essential to own simple light-weight aerials of reasonable gain. The OM is frequently prevented from participating in contests or activity periods by weight or dimensions of his aerials.

Two versions by DJ9HO are presented which fulfil the requirements stated above. The radiation pattern of the system was measured during the 1975 aerial measuring campaign of the "UHF-Group Munich". The reflector is made from copper-clad pertinax board. This provides good conductivity (see skin- effect A.1.3) at low weight and a gain increase of 2 to 3dB. The quad-to-reflector spacing may be varied by means of just one screw. It is thus possible to achieve optimum matching of coaxial cables of 50 to 75 ohm.

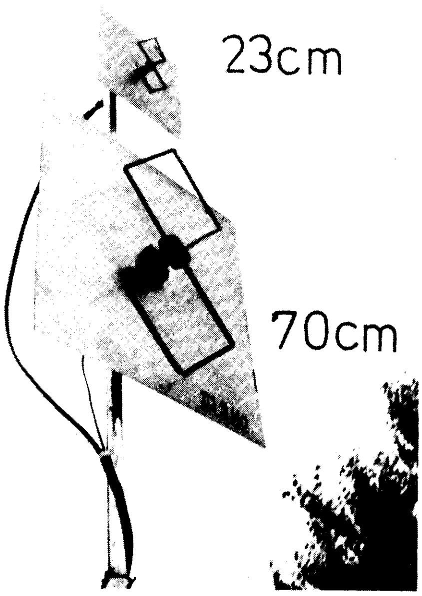


Figure 459
Aerial system for two UHF-bands.
Gain = 11 dB
The twin-quad is comparable to a 10-element Yagi. Its broad horizontal radiation pattern and the well-known DX-properties may even be regarded as advantageous over the Yagi (see also section F.9).

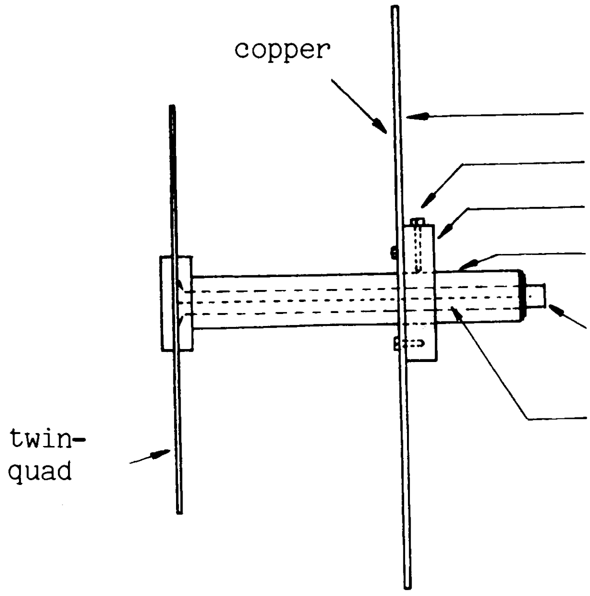


Figure 460

Identification of components

reflector (copper-clad board)

setscrew

plastic block

quad support tube (can slide in plastic block)

BNC-connector installed inside the support, soldered to coax.

Coaxial cable inside the plastic support tubing (connects BNC-connector with twin- quad).

Description

The square-shaped reflector is made from copper-clad pertinax (or fine wire mesh). The copper foil is coated with protective varnish. The size of the plastic blocks depends on the diameter of the quad support tube which in turn depends on one's possibilities for obtaining such material. The tube which also houses the coaxial cable may have about 15 to 25mm diameter. In our specimen it was turned from plastic material (23cm aerial) resp. its square cross-section was retained. The following photographs illustrate the various possibilities.

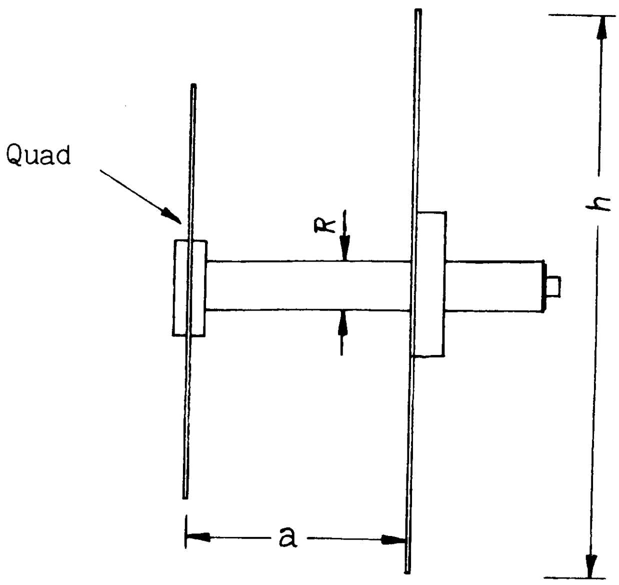
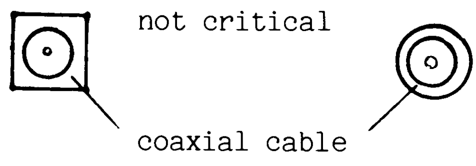


Figure 461 Dimensions

	70cm band		23cm band
h	55 cm		25 cm
a	10 cm		3 cm
R	2 cm	to	1.8 cm

shape of tube not critical



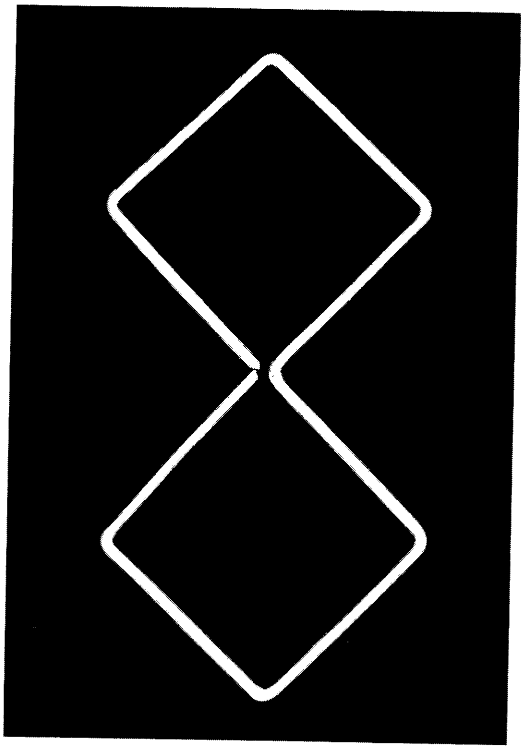


Figure 462
Dimensions of the quad element

Band:	70 cm	23 cm
width of side (inside)	17-17.5 cm	6 cm

Material: Copper, 4mm diam.

The quad radiator is bent from a single piece of wire as shown by the photograph. The transmission line is connected at the centre with the inner conductor being connected to the left and the outer conductor connected to the right (or vice-versa).

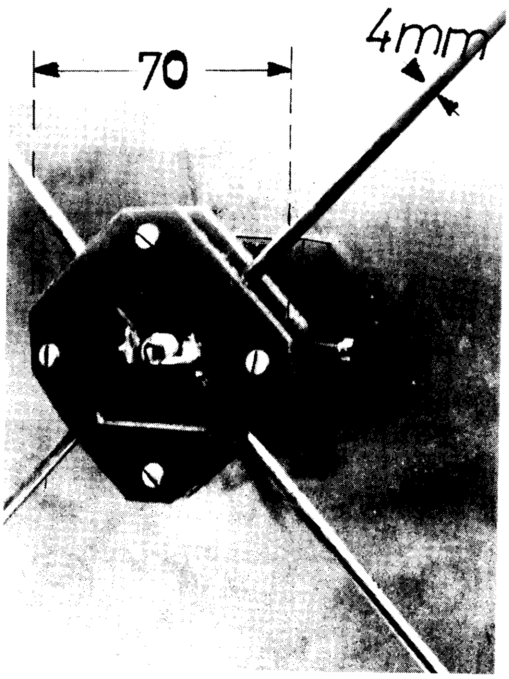
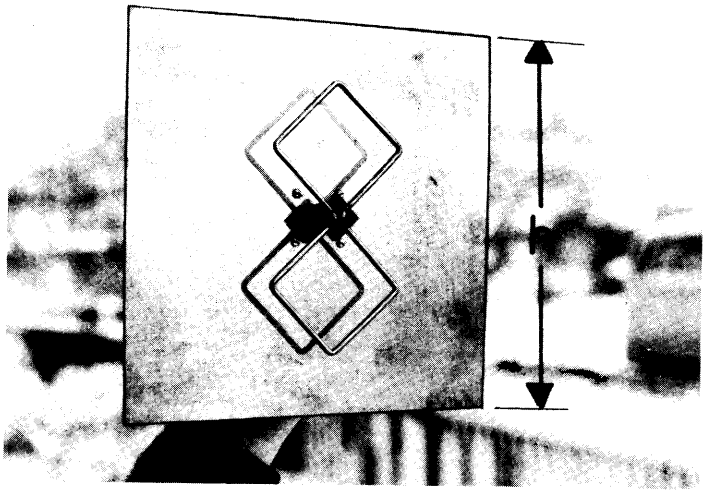
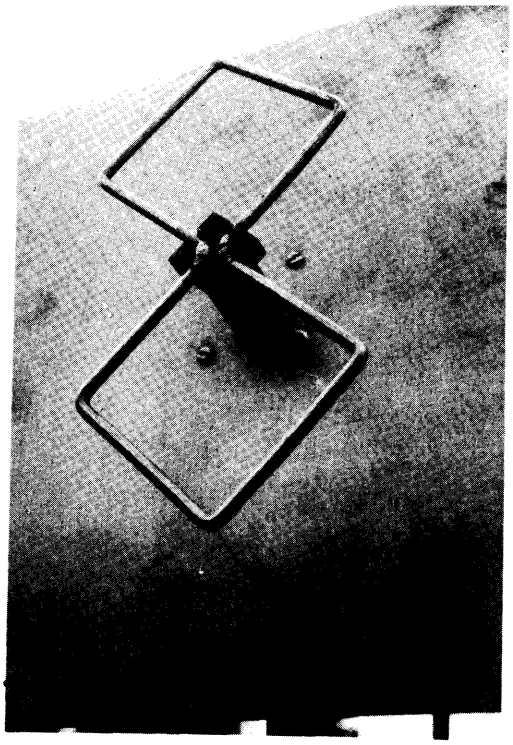


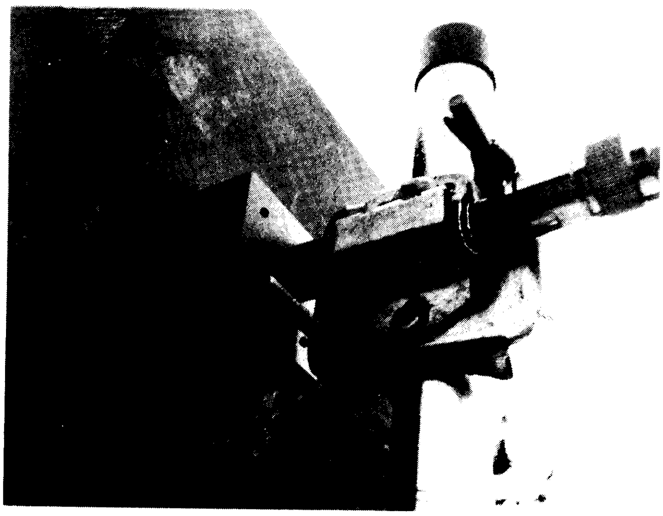
Figure 463
Quad support of the 70cm version. The driven element is clamped between two PVC plates in this particular case. Other synthetic materials may be employed since voltages are minimal at this "cold" point. The feed assembly should be sealed in the case of permanently installed aerials.



<u>Figure 464</u>
Overall view of the 23cm version. See figure 461 for dimension h.

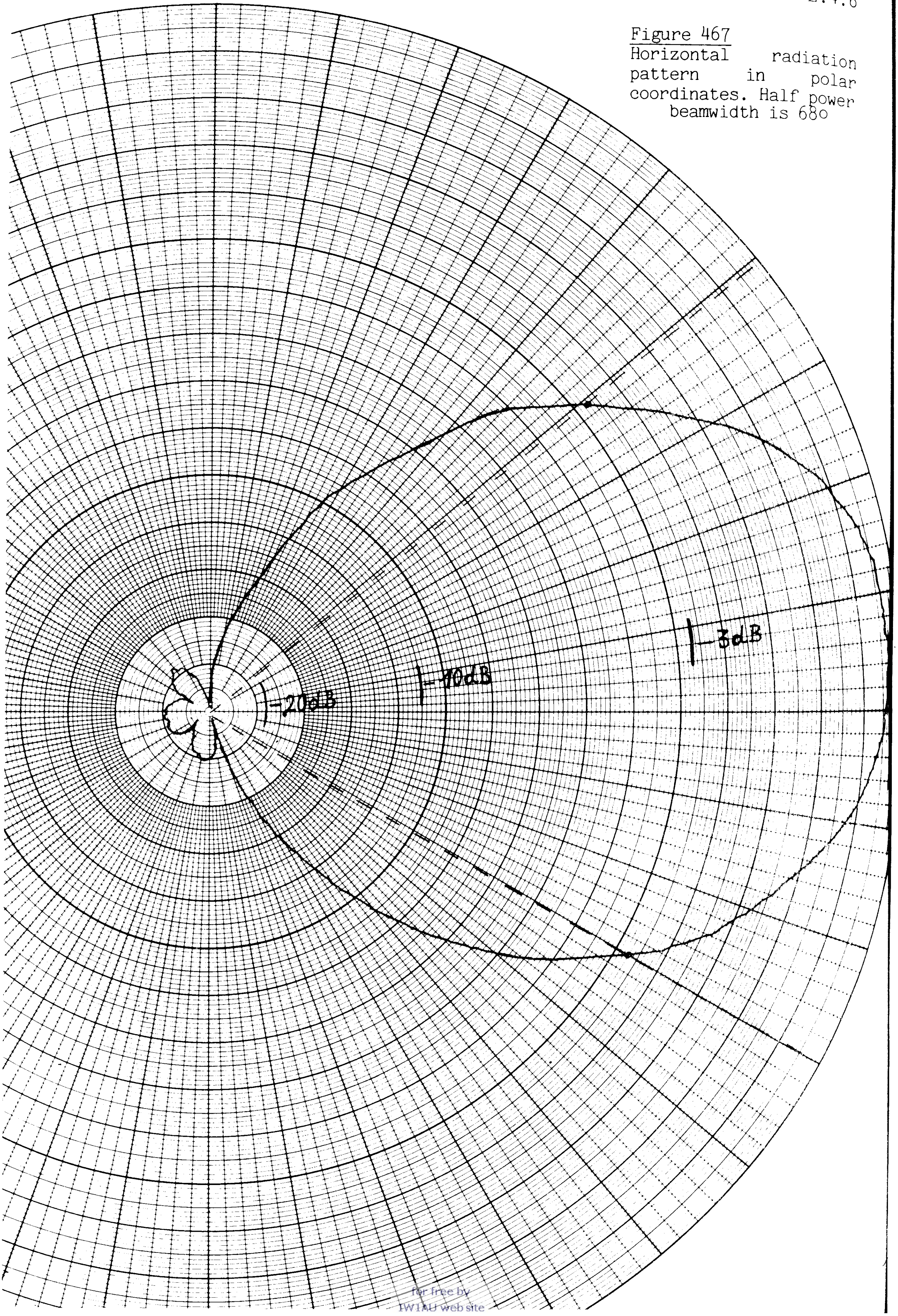


<u>Figure 465</u>
The coaxial cable (such as RG-8, RG-9, RG-213 or equiv.) is routed through the plastic tubing and soldered to the twin-quad (inner conductor to the right, screening to the left). The feed point of a permanently installed aerial should be sealed against water intrusion.



<u>Figure 466</u>
Rear view. A normal holder as customary for t.v.aerials is used to clamp the aerial to the aluminium mast. An N-connector (G.4) terminates the Flexwell cable used in our sample at the rear.

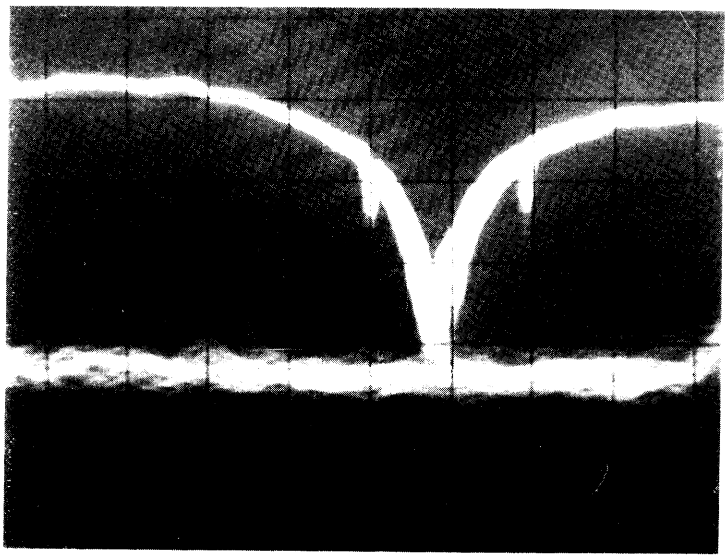
Figure 467
Horizontal radiation pattern in polar coordinates. Half power beamwidth is 68o



The UHF version with reflector plate designed by DJ9HO was additionally optimized by means of a commercial (HP) return-loss instrument. The shown trace was painted onto the scope of the instrument. The horizontal axis represents the frequency range 1.0 to 1.5 GHz. The vertical axis represents the logarithm of the power reflected by the aerial. It is clear to see that the return loss is approximately 30 dB. This means that the power reflected by the aerial is 30 dB below the incident power. The voltage across the "reverse detector" of a VSWR meter at an incident power of 10 watts into a correctly tuned aerial is so small that there is practically no movement of the needle, i.e. practically all the power is absorbed by the aerial (as in a dummy load), amplified according to the gain and emitted into space as effective radiated power.

An example based on this 23cm aerial and an assumed incident power should clarify this and aquaint the OM with the terminus "return loss".

TX power (incident)	return loss	at approx. freq. (MHz)	power refl. by aerial	power emitted by aerial
10 watt	3 dB	1200/1500	5 watt	5 watt
10 watt	6 dB	1240/1400	2.5 watt	7.5 watt
10 watt	10 dB	1250/1350	1.0 watt	9.0 watt
10 watt	20 dB	1270/1310	0.1 watt	9.9 watt
10 watt	30 dB	1290/1300	0.01 watt	9.99 watt



1 GHz 1,3 GHz 1,5 GHz

Figure 468

Scale:

horizontal: 50 MHz/div.

vertical: 10 dB/div. return loss

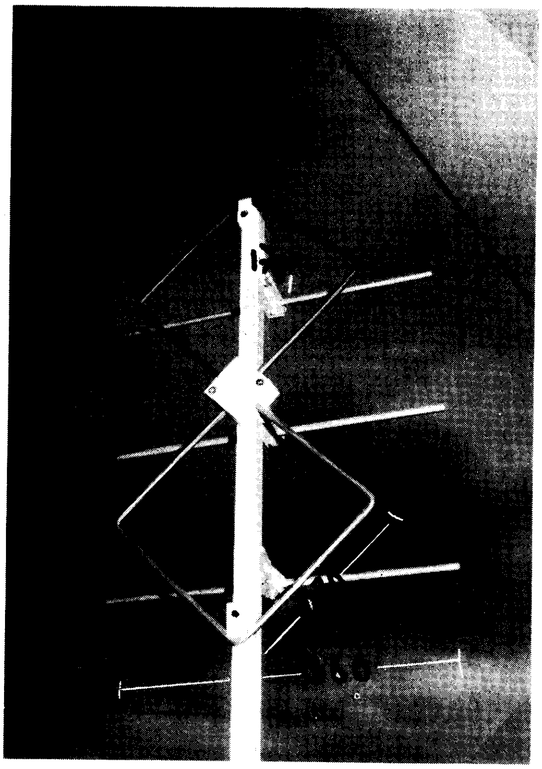
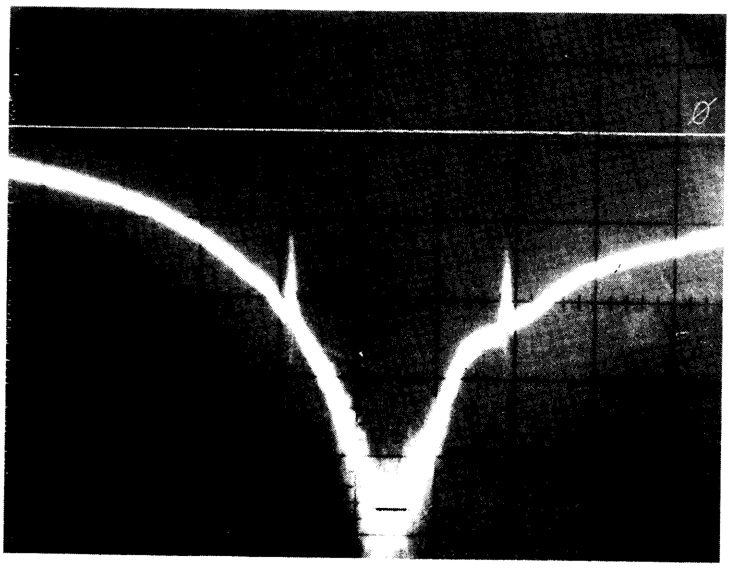


Figure 469 shows construction and dimensions of a horizontally polarized twin-quad for the 70cm band. It is simple to construct, light weight and small in dimension. It is quite inconspicuous when mounted to the mast of the t.v.aerial. Its considerable horizontal beamwidth (60°) covers a large sector without having to rotate it. The support is made of 15x15mm aluminium. The radiator is plastic coated copper wire of about 4mm diameter. The plastic mounting blocks are screwed to the support as described for the 2m version.

The following picture shows the frequency response of return loss of a 13cm twin-quad. Its construction is according to figure 459ff. The reflector is made from copper-clad board (width 70mm, height 95mm). Support of the aerial which is rather small because of the wavelength is by a semi-rigid cable (2mm outside diam.). Radiator-to-reflector spacing is approximately 16mm. The cable screening is simply soldered to the copper foil. The little twin-quad is made from silver plated copper wire of 1mm diam and is connected according to figure 462.



2100 MHz 2500 MHz

Figure 470
Frequency response of return loss of a twin-quad for the 13cm band.

h: ~ 50 MHz/div.
v: 5 dB/div.

markers: 2250 MHz and 2400 MHz

Return loss: ~ 24 dB

E.4.6.1 A twin-quad beam for the 70cm band

This description suggests how an interested 70cm amateur may construct an aerial which has small physical dimensions and guarantees 16dB of gain (amplification factor 40) at moderate cost. Having an overall length of 133 cm, a height of 40 cm resp. 62 cm (see diagram) it may be installed inconspicuously anywhere.

The basic concept is that of the reinforced 2m version. Seven rather than three reflectors are employed. Once again the spacing between the reflector array and the radiator is somewhat unusual. The twin-quad is mounted to the director support and the spacing may be varied. The directors are held by two booms (one per bay) of square cross-section 15x15mm. They are secured by self tapping screws and sealed against water intrusion. The slightly thicker material 20x20mm of the reflector support slides over the booms. The spacing between the directors and the reflector/radiator assembly may thus be adjusted for optimum SWR. The 70cm twin-quad beam, too, has a good SWR of 1.2.

Figure 471

The photograph shows the 70cm twin-quad beam as part of the aerial system at DB8NP.

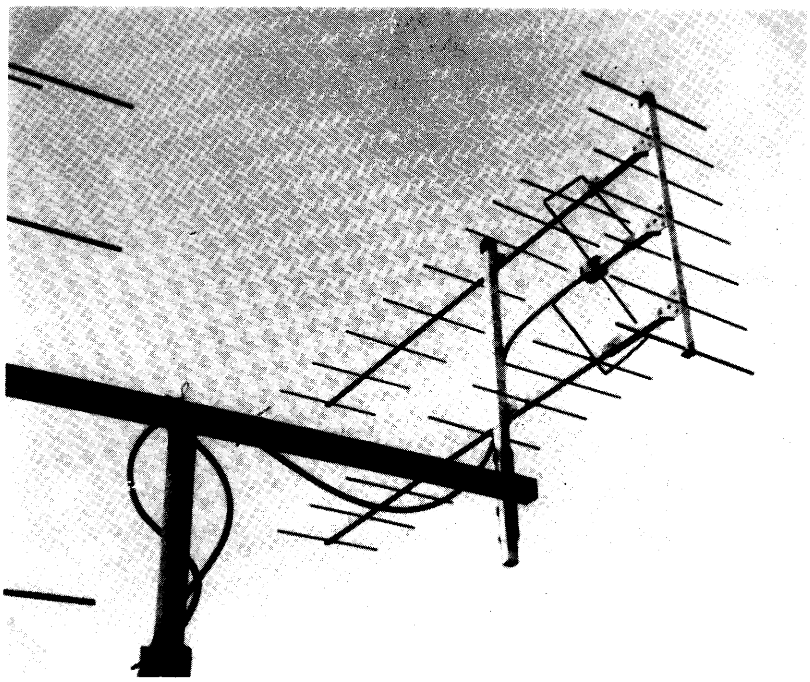


Figure 472

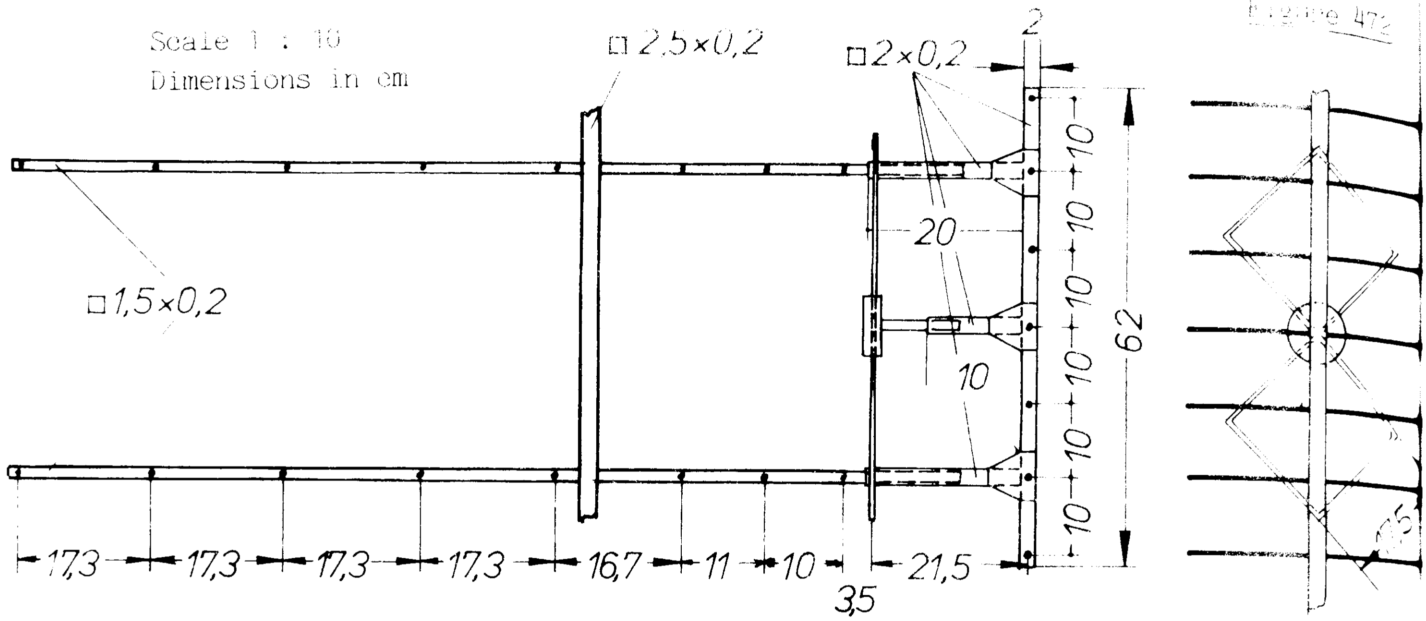
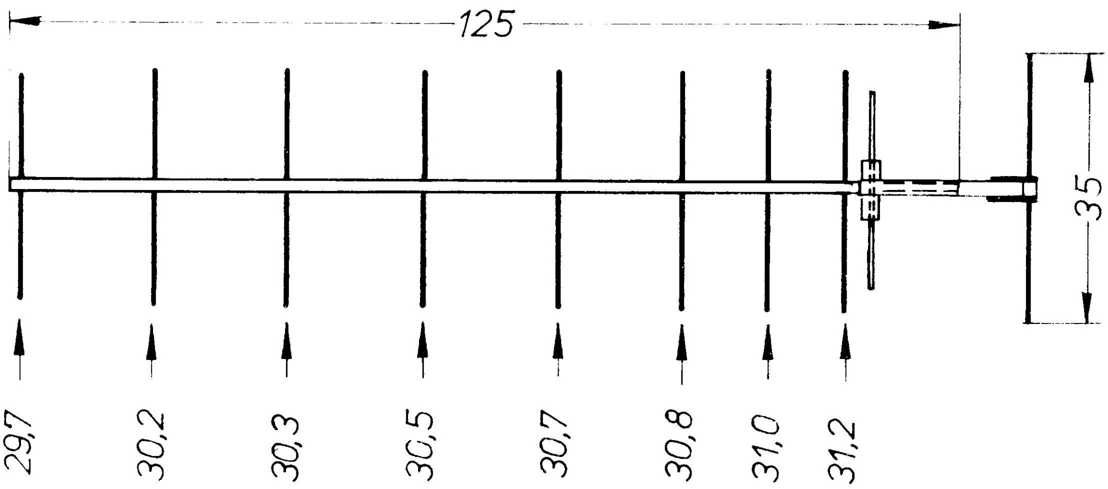
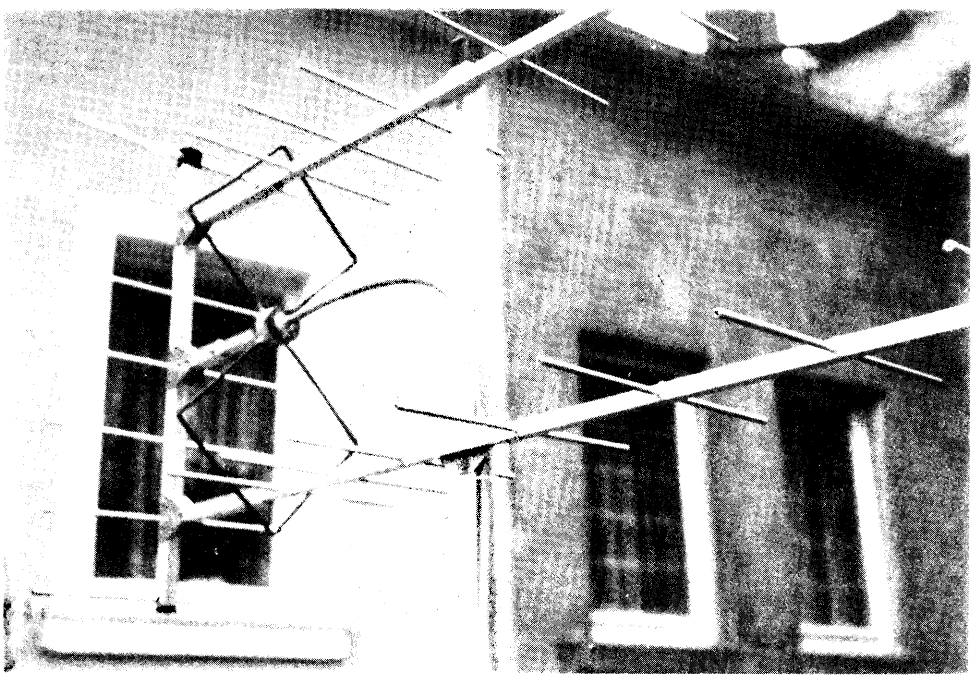


Figure 473



70 cm Doppelquad - Beam

Figure 474



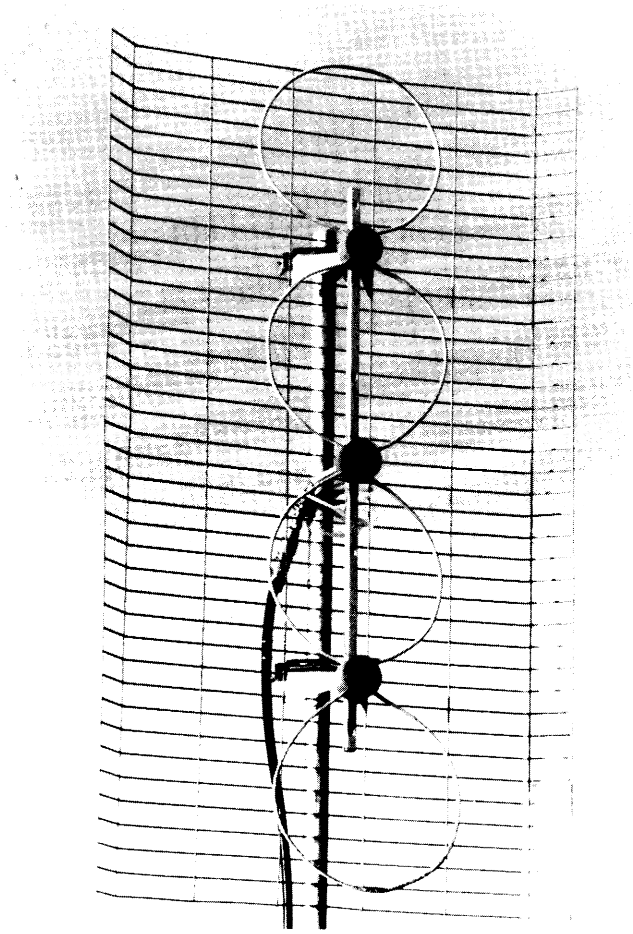
70cm Tw[illegible]-quad be

E.4.7 Quadruple-quad for 432 and 1296 MHz (DJ9HO twin-eight)

After achieving results with the twin-quad that are quite good, considering the small effort involved, stacking was taken into consideration. One of the goals was to get away with just one feed point for the aerial system since the matching and distribution network would cancel some part of the additional gain.
Days of empirical testing were spent to find an arrangement that would turn the calculated element dimensions into an array with additional gain. The stated requirements (horizontal polarization and one feed point only) implied stacking in the vertical direction.
Deviating from the customary shape, a circular form was selected with one each full-wave section added to both top and bottom. The grid reflector of a t.v.aerial was used as reflector which has the dimensions 100 x 60 cm. The active element is made from insulated 2mm copper wire. It is held by split plastic blocks at the points corresponding to the current antinodes. Each one of the four loops has a circumference of 68cm. Measurements indicated that this shape is not perfectly suited in this case. The obtained gain was only around 9dB. A simple loop shaped twin-quad was quickly built and yielded a 9.5dBd gain after optimizing the aerial. It was then obvious that the ideal shape had not yet been found. The following photograph shows the twin-eight design and the wiring.

Figure 475/476

Experimental aerial of stacked twin-quads. The advantage is in the single feed point and the simplicity of the construction. The mechanics should be arranged so as to allow one to vary the radiator-to-reflector spacing.



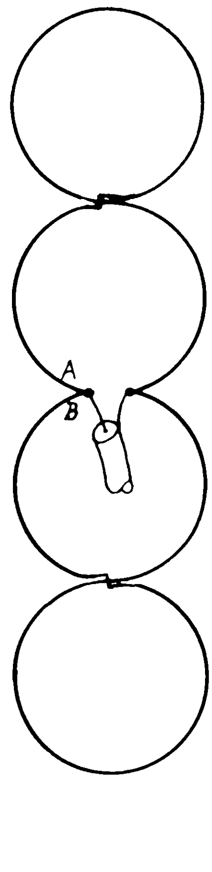
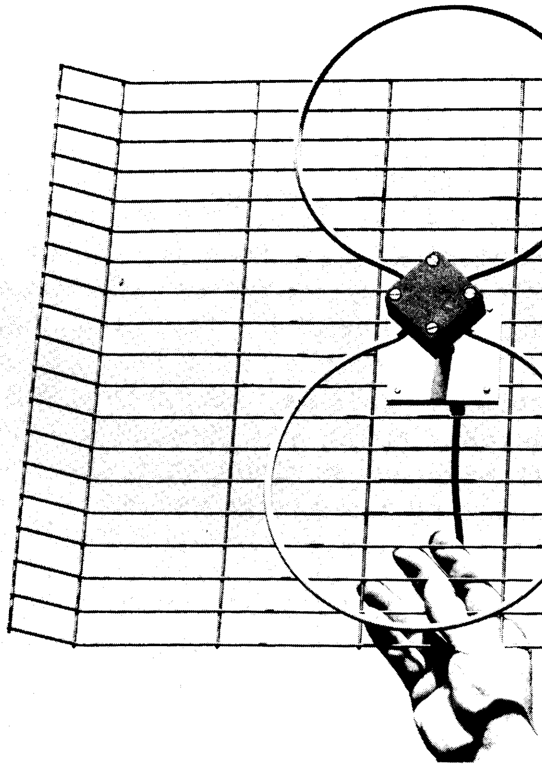
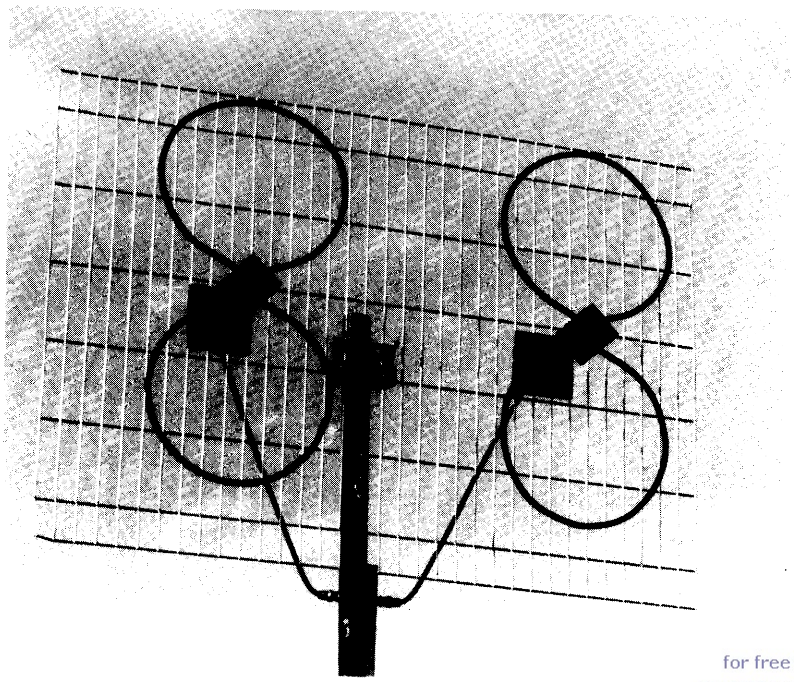


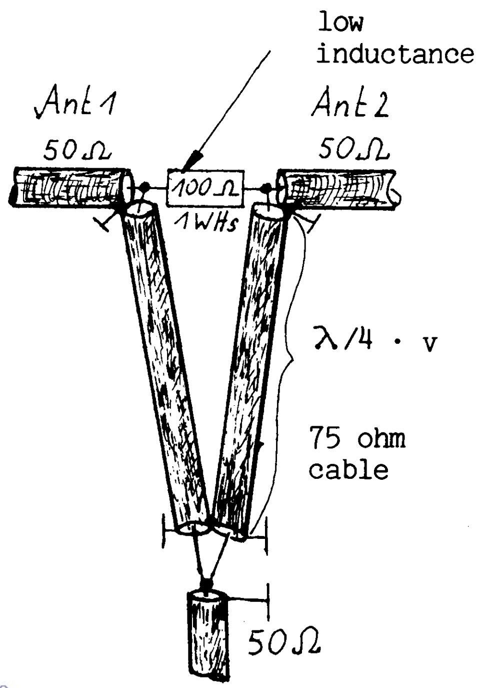
Figure 477 shows the simple twin-loop version in front of a grid reflector. Each loop has a circumference of 68cm. Feed arrangement as shown in figure 436. The reflector dimensions are 50x60cm. The active element is held by an aluminium tube of 12mm diameter through which the feed cable (RG58/U) is routed. The radiator of this sample is made from insulated grounding wire (copper, 4mm diam.). Gain: 9.5dB. The half-power beamwidth may be assumed to be as shown in figures 437/438.



Horizontal stacking in addition to vertical stacking was tried. The reflector was once again the grid 60x100cm. It was confirmed (see measurements in E.6.3) that the grid rectangles should be parallel to the polarization plane (figure 477). When using a plate or correct grid-type reflector a gain of 11 to 12dB may be expected.

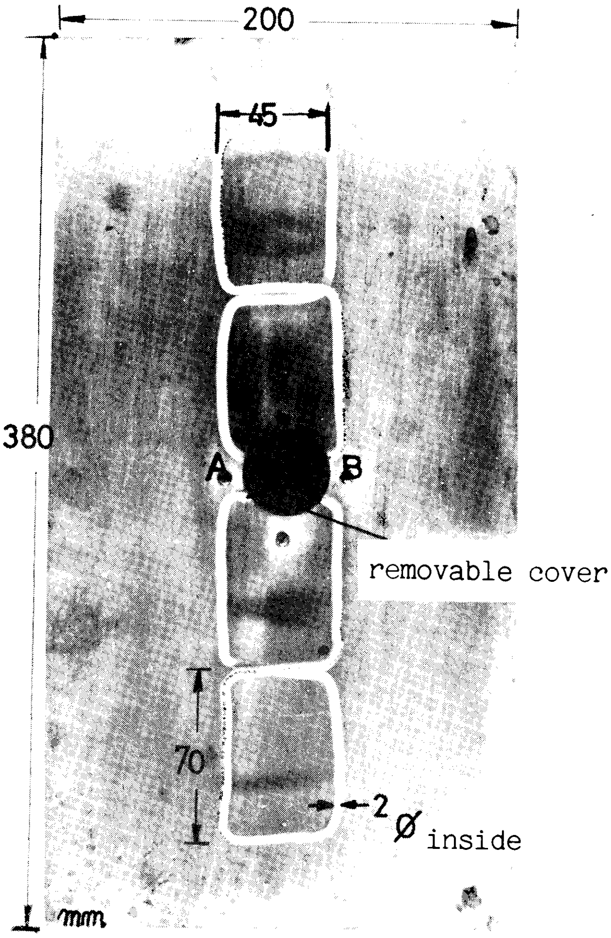
Figure 478/479
Horizontal stacking of twin-loop aerials. The interconnecting cables running to the feed points should be of equal length for both bays. The distribution network for 50 ohm cables is ilustrated by the drawing.

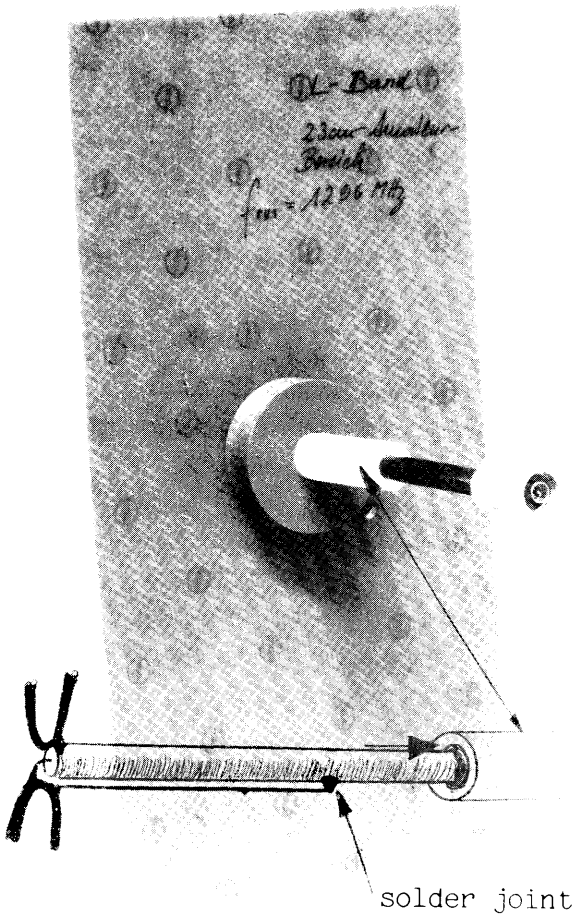




Further optimization of the "twin-loop" was carried out on a 23cm version because of the lower mechanical effort in case of modifications. The starting point was once again the circular loop version which eventually had to make room for the rectangular version which exhibited better gain and improved SWR.

Figure 480/481 Front and rear view of the stacked twin-eight array for the 23cm band. The gain is 14.5 dBd, the half-power beamwidth a) horizontal: 32° and b) vertical: 16°.





DJ9HO would like to take this opportunity of thanking Mr. Wiesbeck (Telefunken) for executing the measurements in a reflection-free room for aerial measurements in comparison with a reference horn. DJ9HO repeated the measurements according to the method described in section E.6 which confirmed the excellent results. The spacing between the driven elements and the copper clad reflector surface was 30mm. It is, however, necessary to make this distance variable. The 12mm diam. RG214/U cable is run through the plastic tubing and soldered according to figure 436. Balancing is achieved through a 6cm long piece of stranded copper wire (2mm diam.). One of its ends is connected to the inner conductor of the coaxial cable, is then run insulated and parallel to the latter and finally soldered to the outer sleeve of the coaxial cable.

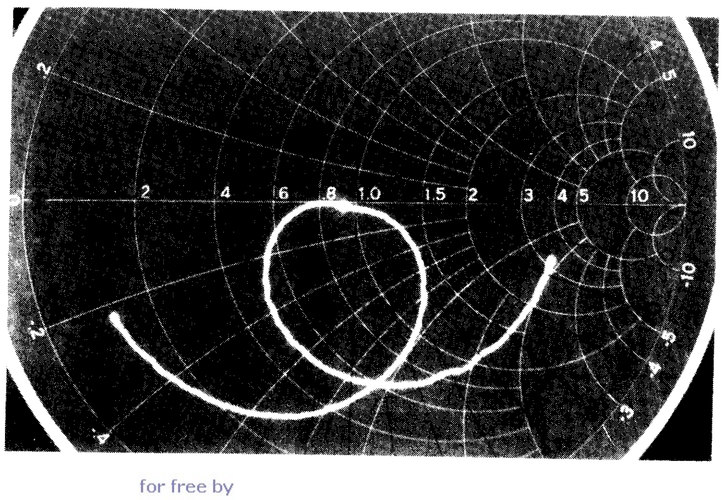


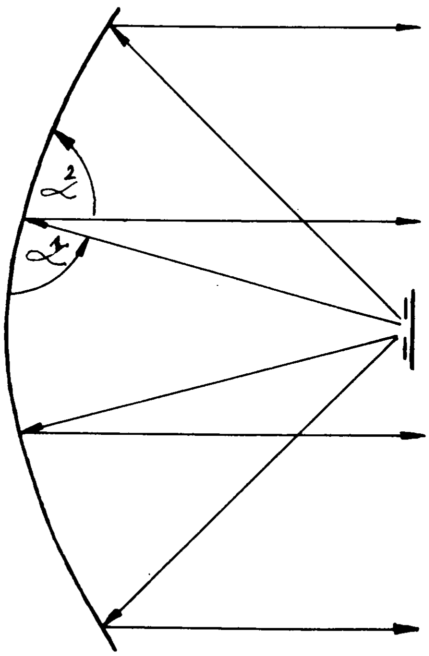
Figure 482 Aerial matching as represented by the Smith chart. Frequency range 1200 to 1400MHz. The illuminated spot shows the properties at 1296MHz.

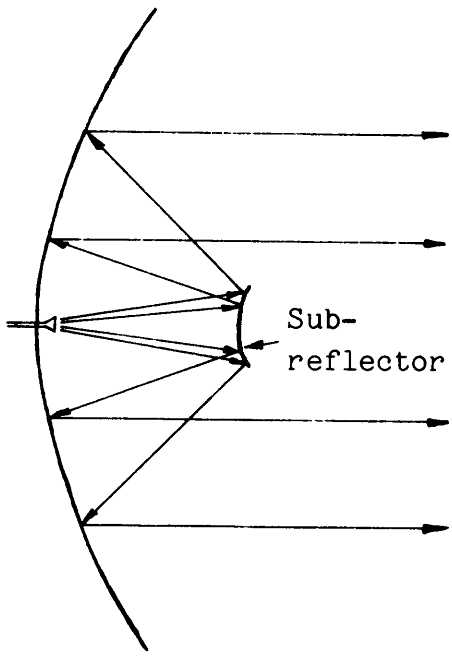
E.4.8 Parabolic aerials in theory and practice

The parabolic aerial is one of the highest gain systems and is very popular with UHF amateurs. Its principle of operation may be described like this:

The electromagnetic radiation which originates from a point source (feed in the focal point of the mirror) is transformed into a coherent electromagnetic field and focused in a dominant direction. The energy which is emitted by the feed and directed into the paraboloidal reflector is thus reflected according to optical laws (angle of incidence equals the angle of reflection; $\alpha 1 = \alpha 2$) and is re-radiated as plane wave front in the desired direction. The OM should be quite familiar with this principle from electric torches and car headlights.

Figure 482/483 shows simple and double reflecting systems (Cassegrain). In a Cassegrain system the waves are radiated into the paaraboloidal reflector by means of a hyperboloidal sub-reflector.



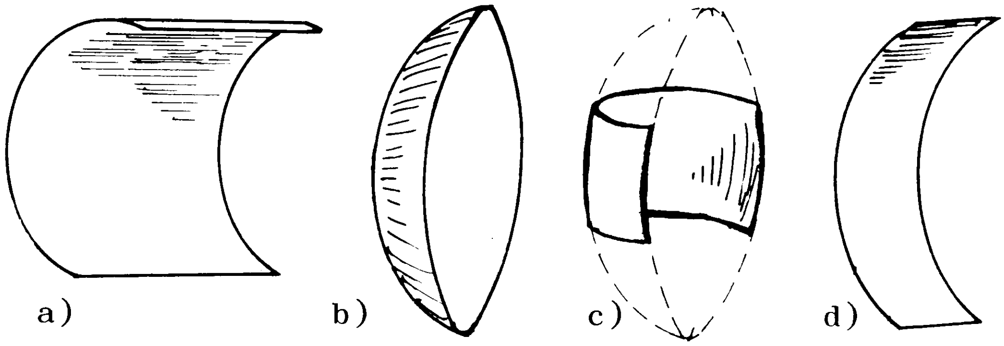


The shape of the reflector defines both the horizontal and the vertical beamwidth (-3dB points) of the aerial. This and the types of feed cause the differences between the reflector aerials.

Figure 484
Some of the commonly used reflector shapes and their focusing properties are presented.

Beamwidth:

	a)	b)	c)	d)
horizontal:	minimum	maximum	maximum	minimum
vertical:	maximum	maximum	minimum	maximum



Cylindric paraboloid   Paraboloid of rotation resp. segments thereof

Each type of reflector requires the design of a suitable feed to obtain the maximum possible gain. A dipole with reflector is thus suited for types a) and d) in case of horizontal polarization and for type c) in case of vertical polarization. Apart from that a horn feed (coffee-tin or beer-can feed) may be employed for type b) and the twin-quad for type c). The feed systems are held either by central or lateral supports and the feed cables run along any of the braces.

Illumination of the reflector should be arranged so (1) that the -10db points of the radiation pattern of the primary radiator are not outside the main reflector. Further details are discussed in (1) and (2).

The gain in dB is given by (1)

$$G\ (dBi) = 10 \log_{10} \frac{4 \pi A}{\lambda^2}$$

Apart from that this value may be read off the diagram in G.2. For various reasons the area efficiency may be assumed to be 55%. An area efficiency of 70 to nearly 80% may be achieved with special types of feed horns and closely controlled apperture illumination.

The -3dB (half-power) beamwidth may be calculated from (1)

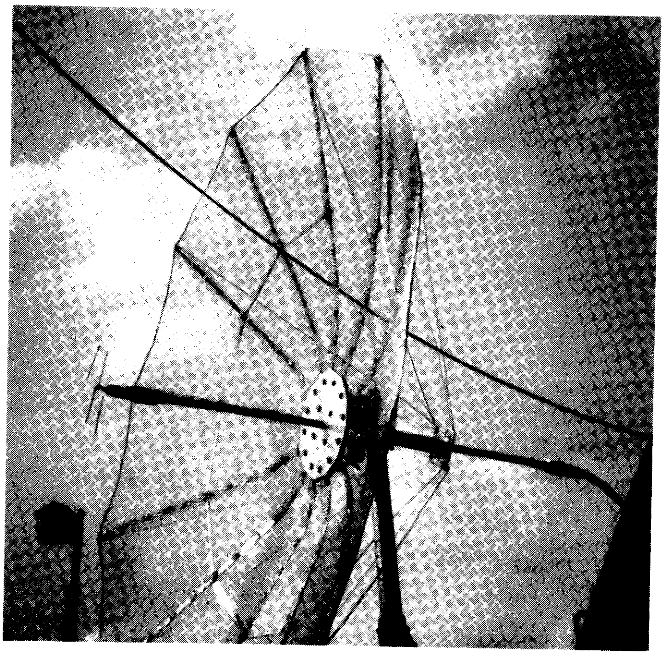
$$\Theta = 70 \frac{\lambda}{D} \qquad D = \text{reflector diameter}$$

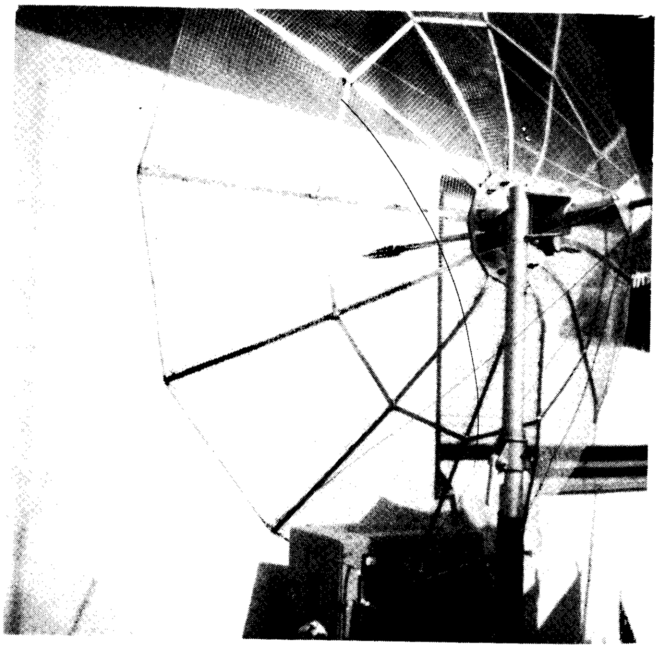
Practical construction: A relatively simple version of a parabolic aerial was described in (2). This is supplemented by the description of a reflector (1.2m diameter) made from galvanized wire mesh. As in (2) the reflector is composed of 12 trapezoidal segments (b1 = 6cm; b2 = 34cm; height = 55cm).

They are joined with 1cm overlap and soldered to struts of galvanized mild steel (dimensions 550x15x3mm). The whole assembly is finally held at the centre between two aluminium plates (twelve- cornered; 200x4mm) by means of twelve galvanized steel screws M6. Two hollow square-shaped table legs are welded together and form the support. The rigid feed cable (which also carries the feed) rests inside. Galvanized steel wire (2mm diam.) is connected at several points along the reflector rim and braced to the rear end of the support to ensure adequate strength of the structure. The vertical mounting clamp is a 90° fitting from the range of t.v. installation accessories. The gain reduction due to the non-perfect paraboloidal shape should not be over emphasized. The whole aerial is coated with thick grease upon completion.
An exact copy of the design described in (2) and with a twin-quad as feed provided a gain of 23.6dB with respect to a reference horn. The gain of this version is approximately 21dBd.

Figure 485/486

The two photographs give a good idea of the construction. The feed is a twin-quad without reflector. This resulted in a fairly wide low power search beam and a steep gain increase in the main radiation axis. The focal point of the reflector (spacing between feed and centre-boss) is 50cm.





References: (1) Fa. Andrew Bulletin 1076
(2) VHF Communications 1/1979 pages 8-12
(3) G.v.Trentini Übersicht der...Mikrowellenantennen "Frequenz" volume 6/1975

E.4.9 Feed systems for parabolic and corner reflector aerials

Half-wave dipoles according to E.4.4 and twin-quads are reasonably well suited for corner systems and reflector segments (E.4.8). The section marked "l" is made from stiff "semi-rigid" cable. The inner conductor is connected to the right-hand side of the radiator, the outer conductor to the left-hand side (see photograph). The component marked "r" is a quarter-wavelength element from solid material. It runs in parallel (3mm spacing) with the feed cable and is soldered to its sleeve at point "a" $\lambda/4$ (60mm at 1296MHz) to the rear.

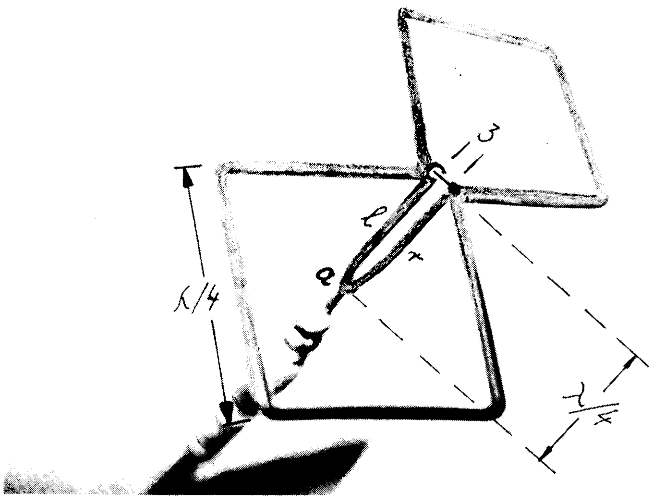


Figure 487
Twin-quad as feed for the reflector segments of E.4.8

So-called "beer can" or "coffe tin" feed systems are suited for complete paraboloidal reflectors. In principle they provide the transition from coaxial cable to circular waveguide with a beamwidth of 130° at -10dB. The VSWR is better than 1.5 at 50 ohms. OM Michael (DB6NT) provided the following data obtained by him and based on (1).

Frequency:	h	a	d	e	
1296 MHz	45 mm	57 mm	155 mm	231 mm	beer-can 4 Liters
2304 MHz	29 mm	55 mm	83 mm	107 mm	coffee-tin 200 g *)
3456 MHz	19 mm	21 mm	58 mm	87 mm	} 1 mm brass sheet
5760 MHz	10 mm	13 mm	37 mm	57 mm	

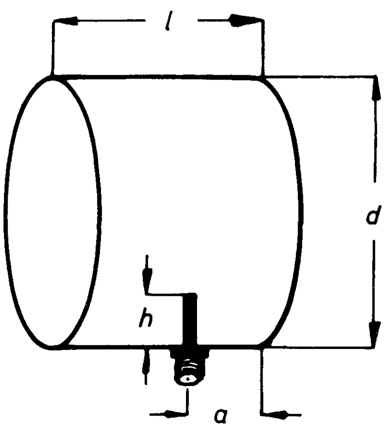


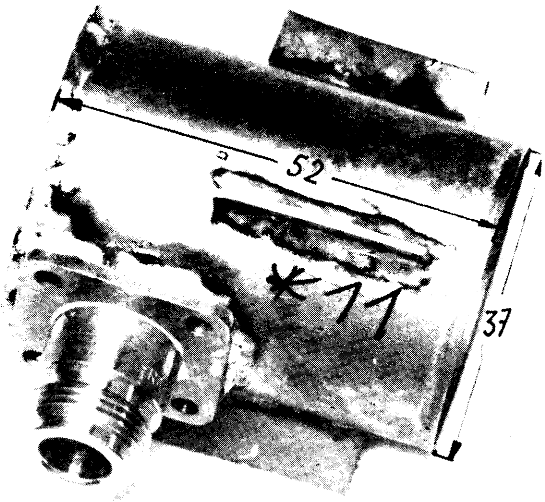
Figure 488
The radiator for the 23cm band has a diameter of 6mm; apart from that 2mm diam CuAg is used.

*) This feed for 2304MHz is not yet optimized (measurements indicated a VSWR of >2). The VSWR of all others is less than 1.5.

Reference:

(1) H.J.Griem: Tubular feeds.. VHF Communications volume 1/76

Figure 489
Primary feed for 5760MHz with mounting flanges attached.
Attention!
The radiator shown in figure 488 must be horizontal for horizontal polarization resp. the feed cable must be entered from the side.



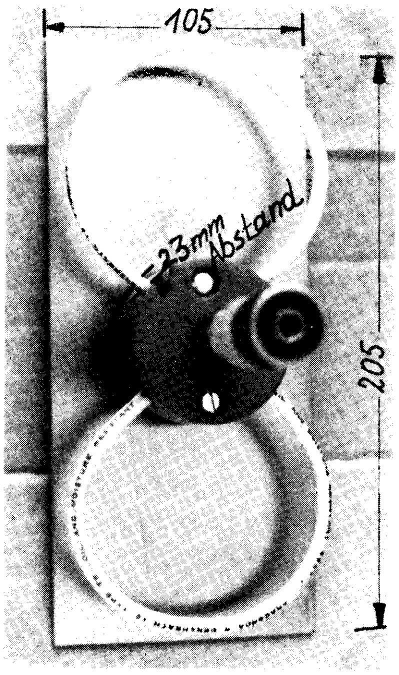


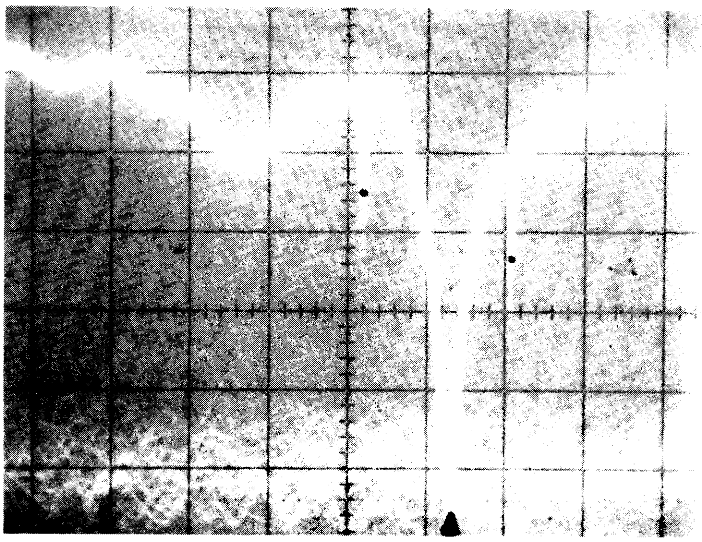
Figure 490
Twin-quad used as primary feed. A loop shape instead of the rectangular shape according to figure 487 was chosen. The shown balancing assembly was also deleted. Inner and outer conductor were simply connected to the right - resp. left - sides as discussed in E.4.5.
The shown reflector plate turned out to be too small. The dimensions are, however, correct. The required total length of wire (2mm diam. copper with insulation) is 2xλ. The plastic mounting block is first drilled crosswise and then split by means of a saw cut. Radiator and feed cable are then clamped to the rear part (with reflector) through two screws.

Polarization: horizontal

Figure 491
Response of the twin-quad feed. Frequency range 1 to 1.5GHz. The return loss is displayed. Its value at 1300MHz is 25dB. The VSWR is thus less than 1.25

h: approx 50 MHz/div.
v: 5 dB/div.

markers: 1250 MHz
1350 MHz



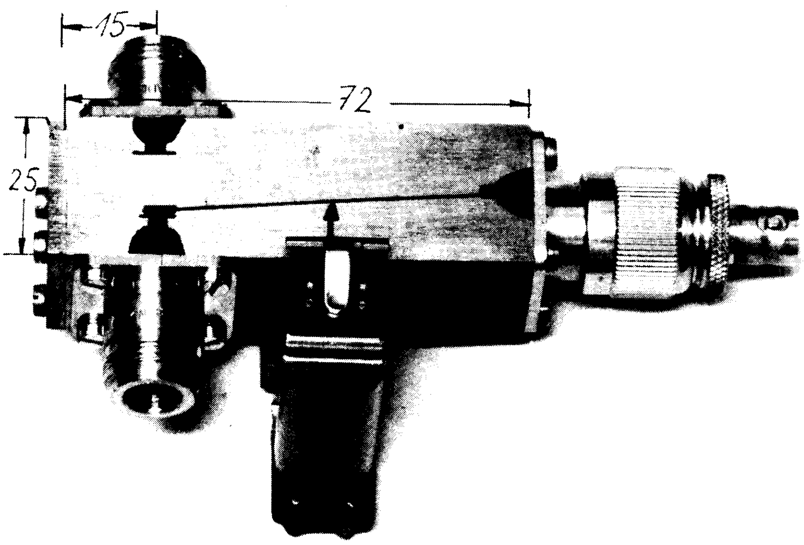
E.5 Building your own coaxial relay

A simple station concept according to D.7 requires just one coaxial relay. It will switch the aerial between the receiving and transmitting chains. Problems encountered with relays are: Power handling properties, good contact and low cross-talk. The cross-talk attenuation rules the general plan of power levels. A value of merely 30dB and a transmitter power of 100 watts gives rise to 100 milliwatts at the receiver input which could result in the destruction of an expensive input transistor. The first receiver stage connected to the coaxial relay must under all circumstances be disconnected from the power supply when transmitting - even with a cross-talk attenuation of 40dB. The power level at the receiver input would still be 10 milliwatts under these circumstances - sufficient to drive most preamplifier transistors into saturation (see compression point in A.4.3) and possible subsequent destruction (depending on the transistor properties). With the supply voltage removed there is little risk from an input power of 10 milliwatts.

The coaxial relay described below has a cross-talk attenuation of 43 dB and will handle more than 150 watts of r.f.power. The upper limit was not investigated. Change over is performed by a plastic pin which transfers the actuating force to the switching arm.

Figure 492

The relay in the excellent workmanship of DL8KH. All terminals should be N-connectors for this sort of power. Actuation is performed by a standard 12 volt relay as used in telephone installations.



The brass block measures 72 x 25 x 25 mm and has a longitudinal bore hole of 8mm diameter. The N-connectors for the two outputs are installed approximately 15 mm off the end. The two holes leading to the centre cavity have diameters of 9 mm. The dimensions for other characteristic impedances may be obtained from section A.2.3. They govern the dimensions of the bore holes and the berillium switching arm (55 x 6.5 x 0.3 mm). Gold or silver plated contacts must be soldered to both the output N-connector spigots and the arm. They may be salvaged from old relays unless one happens to be on good terms with a relay manufacturer. The spacing between the contacts of the two output connectors is 4 mm. This then defines the distance through which the switching arm must travel. The arm is now soldered to the input N-connector, pre-stressed and centered (no contact with the sides whilst making or breaking). The rear end is finally covered by means of a brass plate to preclude the intrusion of dust etc.

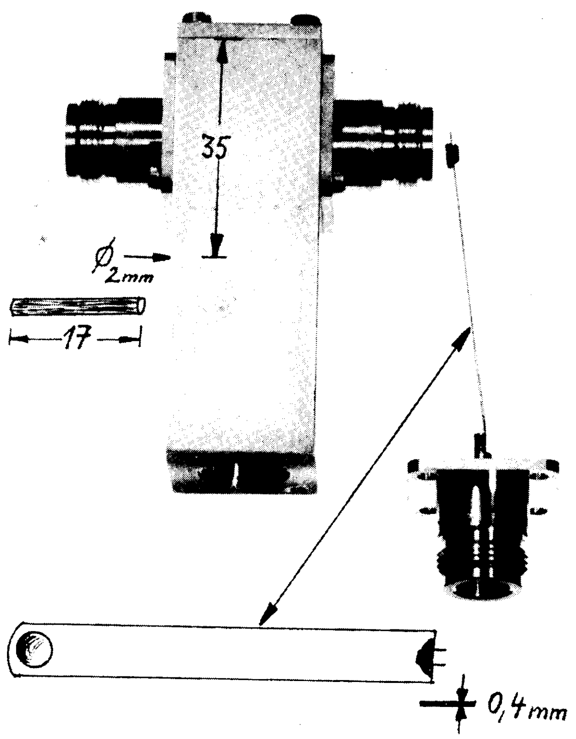


Figure 493
Brass block and the vertically installed switching arm.

The cross-talk attenuation of the relay shown in figures 492 and 493 is reduced to 33 dB when operating in the 23 cm band. A design with two switching arms was therefore tried and gave 42 dB of cross-talk attenuation. In this design one of the arms will make while the other one breaks. The problem is once again to obtain good contact. The aerial is joined to the centre connector. The spigot of the BNC- or N-connector constitutes the stop. Two Teflon blocks prevent the switching arms from turning since the spigots of the connectors to which they are soldered are not locked.

The drawing in figure 494 shows only one screw (ES); a second screw for the other arm is placed correspondingly. These screws are adjusted so that the arm not in contact with the middle spigot will be shorted to ground further away from the centre. This measure increases the cross-talk attenuation by further 10 dB.

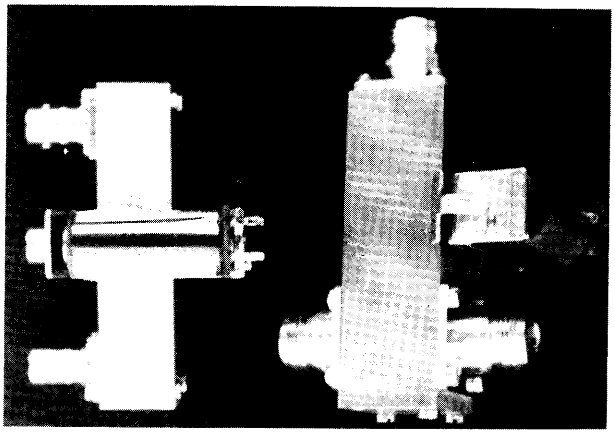
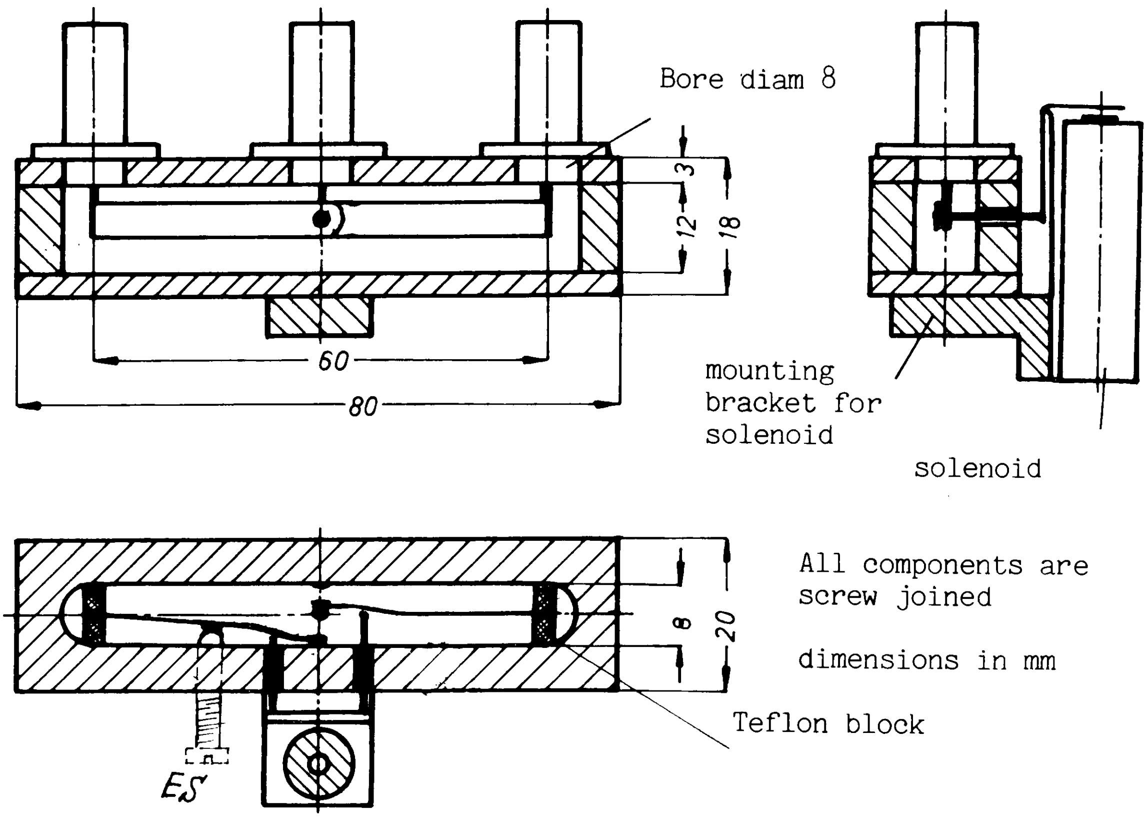


Figure 494

Both relays in comparison. The drawing contains all relevant dimensions. Operating frequency up to 3.4GHz.

Maximum possible power not evaluated.



E.6 Aerial measurements
(D. Reichel, DC9NL)

E.6.1 General remarks

When comparing aerials the amateur has to rely primarily on the makers gain specifications and he should start with them for further considerations.
The gain is, however, not always stated with respect to the half- wave dipole but frequently to the isotropic radiator (which exists in theory only) respectively the Hertzian radiator. The isotropic radiator is used almost entirely in English technical literature. It is therefore necessary to apply a correction to allow direct comparison between aerials.

$$G_{Iso} = 1.64\ G_D = 1.5\ G_{Hz} \qquad G_{Iso} - 2.15\ \text{dB} = G_D$$

$$G_{Hz} - 1.77\ \text{dB} = G_D$$

If attempting to obtain the gain figure of a home made-aerial the amateur will quickly run into difficulties since he can usually just give a subjective evaluation which is obviously prone to errors.

Literature for r.f.technicians (1) describes four methods which are difficult to duplicate by the amateur since he normally lacks a room with low reflectivity.

E.6.2 Measuring methods

E.6.2.1 The most elaborate method is to measure phase and amplitude distribution and to compute the gain by integrating the measured data. Here we are not going to treat this method in any detail.

E.6.2.2 The absolute method
This method calls for two identical aerials. The spacing (d) between them depends on the physical dimensions of the aerials.

$$d = k\ \frac{a^2}{\lambda}$$

a = maximum dimension
k = 2...4 depending on aerial type

This setup is used to measure transmitted power (P_T) and received power (P_R) by means of a bolometer power meter (which converts the r.f. into heat energy) to achieve the required accuracy at the generally low power levels. The losses of the aerial feeder cables must obviously be taken into account in all methods.
The gain is then calculated according to:

$$G_{Iso} = \frac{4\pi d}{\lambda} \sqrt{\frac{P_R}{P_T}}$$

E.6.2.3 Reflected power method

In this method only one aerial is used. The energy emitted by this aerial is reflected by a conducting surface and directed back into the aerial where it is measured by means of a probe inserted between the transmitter and the aerial. The received power increases as the aerial gain is increased. This then results in a deteriorating SWR. Prior to taking measurements the aerial must be matched perfectly under free space conditions (aerial pointing vertically upwards). The gain is calculated according to:

$$G_{Iso} = \frac{8\pi r x}{\lambda}$$

X = aerial-to-reflector spacing
r = reflection factor U_{incid}/U_{refl}

Several measurements with varying distances (x) should be executed because of possible multiple reflections. The resultant sinusoidal interference curve may then be graphically integrated. Only now may the aerial gain be assessed fairly accurately. The reflecting plane should not be too small in order to ensure a credible result (see E.6.3.2).

E.6.2.4 Method of comparison with a reference aerial

The comparison with an aerial of known gain might pose the least problems to the amateur. This method requires, however, also extensive test equipment and the application of correction factors to compensate for the influences of ground properties and various reflections until a useful result is reached. The latter could be disregarded if a non-reflecting room was available.

When considering a practical setup one realizes that the transmitting and receiving aerials are not situated in an ideal free-space environment but at a finite height above ground. This gives rise to twin-path propagation (either directly or via the more or less reflecting ground surface between the two aerials). Assuming perfect reflection without any loss the phase difference along the two paths at the receiving aerial will lead to doubling of the field strength or complete cancellation of any value in between.

$$E = E_{dir} + E_{refl} \quad [E \text{ in uV/m}] \quad \text{(vector sum)}$$

Since the ground has generally imperfect reflection properties it is possible to calculate the free-space field strength from the minimum and maximum measured values.

$$E_F = E_{max} - E_r = E_{min} + E_r$$

The reader is once again reminded of the fact that field strength E refers to voltages. This then means that 6dB voltage increase corresponds to 3dB power increase.

$$E = \frac{E_{max} + E_{min}}{2} = E_{max}\left[\frac{1}{2} + \frac{E_{min}}{2\,E_{max}}\right] = E_{max}\left[\frac{1 + E_{min}/E_{max}}{2}\right]$$

Since $F_{(dB)} = 20 \log E$ and $X = 10^{\log x}$ we obtain from:

$$\frac{E_{min}}{E_{max}} = 10^{\log E_{min}/E_{max}} = 10^{\frac{20 \log E_{min}/E_{max}}{20}} = 10^{\frac{F_{min} - F_{max}}{20}}$$

$$F_F = F_{max} + 20 \log \left[\frac{1 + 10^{\frac{-\Delta F}{20}}}{2}\right]$$

This implies: The free-space field strength F_F - expressed as logarithm of voltage ratios - equals the sum of maximum measured value and a negative correction factor. This factor results from the difference between the lowest and the highest measured value divided by 20 as exponent to base 10. Since this value is negative the logarithmic expression will be negative in spite of the positive sign.
Doubling the field strength expressed as logarithm means +6dB. The correction factor will thus vary between 0 and 6dB.

This equation yields the following table of correction factors:

F_{dB}	1-3	4-6	7-9	10-14	15-23	24...
negat. correct. factor	1	2	3	4	5	6

(see example in measured diagrams)

This table lists the rounded correction factor which has to be subtracted from the maximum value of the field strength.

The discussion of this method shows quite clearly that it is an illusion to compare aerials (that are installed on one mast but at different heights) and to expect the result to be correct.

E.6.3 Practical experiences with aerial measurements

E.6.3.1 Measurements with reference aerials

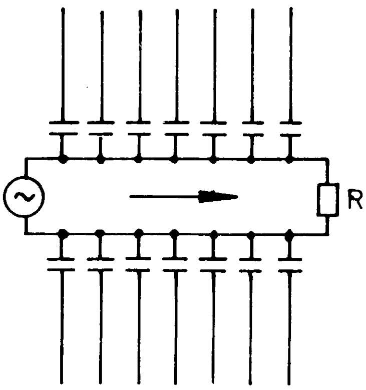
Initial attempts at measuring home-made aerials (70cm) were rather amateurish. It was noted that a $\lambda/2$ dipole (see E.4.4) showed only occasionally significant losses in comparison with a reference aerial. This experience was made at various locations and it was decided to document several series of measurements since little material on this subject has appeared in amateur radio publications.

A variable attenuator was placed between an SSB-receiver and the aerial which was directed at a weak but constant signal (beacon). The attenuation was then increased until the signal disappeared below the noise. The value set at the attenuator is a measure of gain in comparison with other aerials. An additional error is introduced by the rather subjective judgement as to whether the signal is still audible. A rough estimate accurate to within ±2dB is quite realistic after a little experience.
The same arrangement with a strong signal and the S-meter as indicator will yield accuracies comparable with that of the variable attenuator. Most S-meters are, unfortunately, merely guesswork and the indicated S-increments are unsuited for direct calibration.

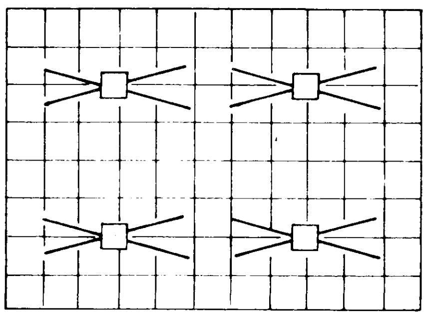
The most efficient method is based on displaying the r.f.signal by means of a panoramic receiver or on one of the hard-to-come- by spectrum analyzers.

The following aerials were compared:		claimed gain: (433 MHz)
1. Home-made aerial:	$\lambda/2$ dipole (E.4.4)	0 dB
2. Reference aerial:	WISI FT 01	7.5 dB
3. UHF, 350-850 MHz	WISI FT 04	2.5 dB
4. Twin-quads	(section E.4.7)	10.0 dB

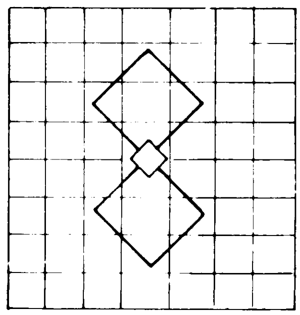
Figure 495
Measured aerials



Fishbone aerial
WISI FT 04



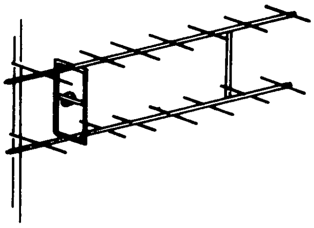
Array aerial
WISI FT 01



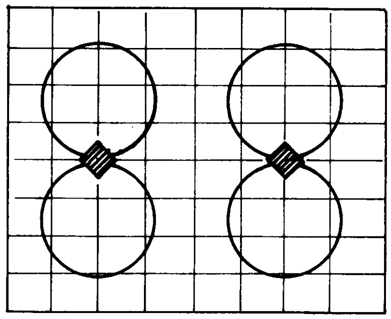
Twin-quad

5. Stacked quad array (E.4.7)
6. Yagi, 11 elements, home-made
7. Twin-quad, loop-shaped, 2x horizontal, fed through $\lambda/4$ balun
8. Twin-quad loop-shaped 2x horizontal, fed through $6/4\lambda$ separatn.filter
9. Corner reflector aerial (see E.4.4)
10. Hirschmann UHF-t.v.aerial FESA 418 UN 46 with twin-loop feed
11. Hirschmann FESA 717 UN 37 unmodified
12. Skeleton slot aerial D 8/70cm (~ 5 years old)

Figure 496 Measured aerials

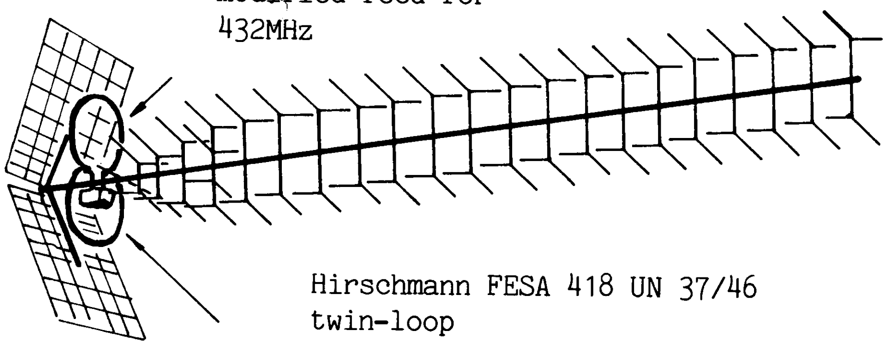


J-beam
D 8/70cm

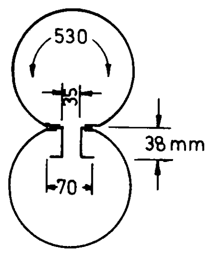


twin quad (loop), stacked horizontally

modified feed for 432MHz



Hirschmann FESA 418 UN 37/46
twin-loop



The aerial output from a beacon (19 kilometres away) was measured in 2.5 resp. 4.5m height at a location with open surroundings and an undisturbed line of sight. None of the aerials gave constant results at all points and it may be concluded that near-field reflection was present. Under these circumstances even the arithmetic mean values were not particularly trustworthy.
The same series of measurements was repeated at a different location approximately 25m above ground on top of a wooded hill (9km from beacon DB0KI). The tendency of the results was the same and the measuring error again 2dB (dipole with respect to FT04). A repetition with significantly higher power level between the two outdoor-locations and the receiving aerials (at a height of 6m above ground level) merely confirmed the existence of highly elevation-dependent signal levels.

Aerial type	location: FK69b					FK69a	FK69 (a → b)		
Test number	1.	2.	3.	4.	5.	6.	7.	8.	9.
λ/2 dipole	0dB	0dB	0dB	0dB	0dB	0dB	0dB	0dB	0dB
WISI FT 04	-1	-1	1	0	-0.5	0		0	2
WISI FT 01							7.5	6	8
Yagi	7	6	8	5	6.5	8	7.5	7	9
skeleton slot	9	8	12	11	9.5	10	10	10.5	12.5
stckd. quad arr.	9	6	9	9	7.5	9	8	7.5	9.5
2xD.quad (hor. λ/4 balun)		3	5	6	5.5	9		5.5	7.5
as above but w. transform.(D.5.9	5	4	5	7		8.5		6	8
corner 45° (90°)	2.5	2	4	5	3	2.5		4(7)	
418 UN 46 (8)						10.5	13.5	10.5	12.5
elevation above ground in metres	2	2	2.5	4	4	25	2.5	4	6

Close examination of this table reveals that a reference aerial with little or no directivity is hardly suited - at least if placed low above the ground. The dipole (in horizontal polarization) has, for instance, identical field strength along the desired radiation axis as well as along the vertical and the rear axis. This means that there is no vertical directivity. It is thus strongly influenced by reflections from the near field both in front and in the rear of the aerial. This explains why the WISI FT 04 has no gain with respect to the reference aerial even though this aerial is claimed to have a gain of 2.5dB above the dipole.
It is also worth mentioning that the skeleton slot aerial reached its specified gain only in test number 9 which also provided correct figures for the dipoles FT 04 and FT 01. It is thus reasonable to assume that the measuring errors are small for the other aerials as well; this test should thus be representative. It is unfortunate that such a large number of measurements is required to obtain satisfactory results.

With the intention of performing quick and reliable aerial measurements a 10m telescopic mast was obtained and instrumented for near-field measurements in the open.

A skeleton slot aerial was fixed to a mast, positioned five metres above ground level and connected to a transmitter. The transmitter was equipped with automatic level control to ensure constant output power. The telescopic mast was erected at a distance d of 25 metres. A spectrum analyzer served as receiver. One after the other the various test objects were installed and measured. Readings of the output levels as function of the elevation were taken and the results documented.

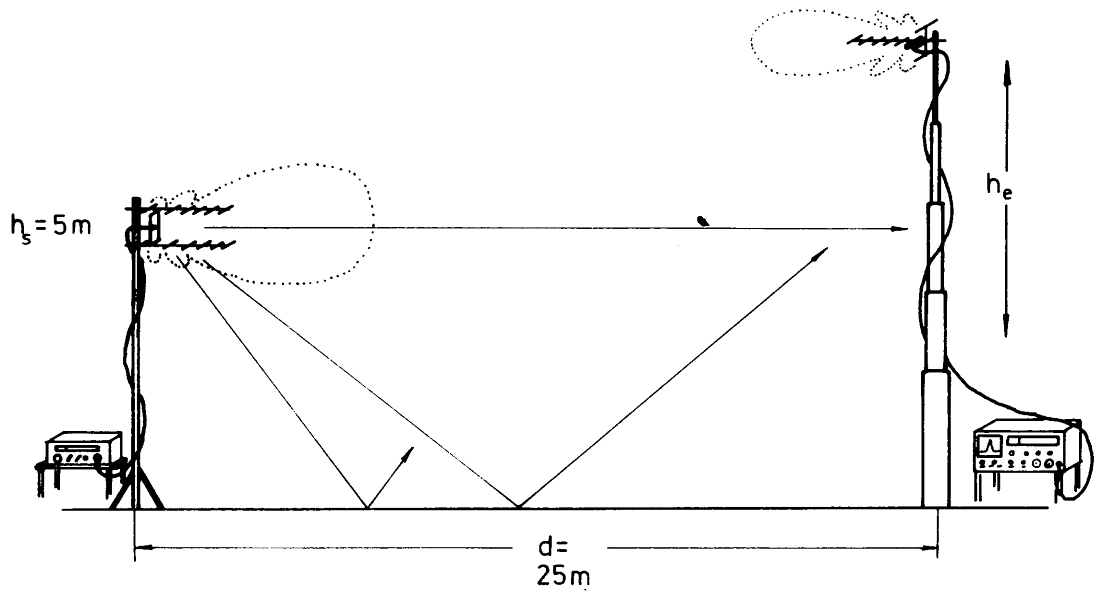


Figure 497
Test set-up for nearfield measurements (with radiation patterns).

Looking at the set up it is evident why total cancellation even under most unfavourable conditions cannot be expected: The aerials employed possess vertical characteristics (E-plane) with directivity. The energy radiated from the aerial towards the ground and reflected according to the factor of reflectivity (and possibly routed to the receiving aerial) is already attenuated by several dB due to the radiation pattern. The reflected energy is furthermore attenuated by the vertical radiation pattern of the receiving aerial with respect to the direct path. Even if the phase conditions for cancellation or enhancement are met complete cancellation or energy doubling can thus not be expected (see E.6.2.4).

The results of the measurements are plotted to give the following information:

1. Point-by-point plot of received field strength as function of elevation for each individual aerial type.

2. Overlay of all results of all aerials for gain computation. That particular diagram contains the results of the various aerials drawn to the same scale.

Both diagrams indicate that the minima are rather more pronounced than the maxima.

Figure 498

Measured aerial voltages of the reference aerials. By means of these curves the reference point 0 dBd was established since the gain figures of these aerials were known.
Clearly noticeable are the significant differences between the maxima and minima of the array aerial (horizontally stacked) and the respective values of the dipole. These differences remain essentially unchanged with increasing height. This behaviour points towards the large vertical beamwidth of the aerial (360°).

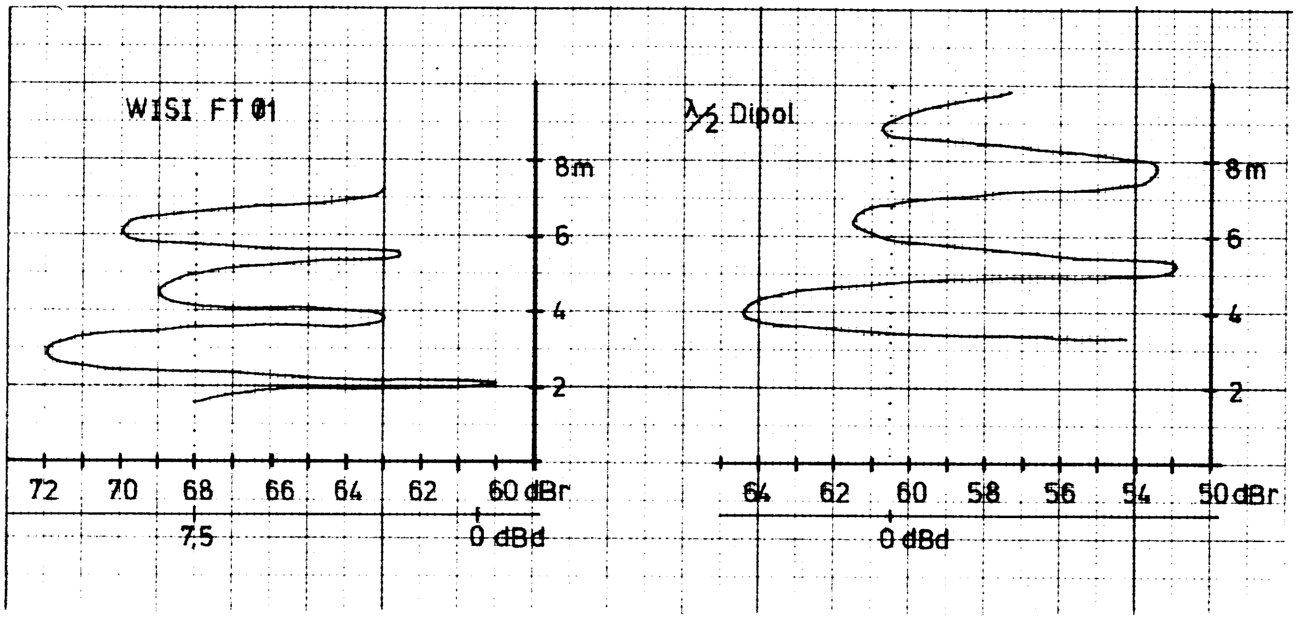


Figure 499

These aerials are types with increased vertical directivity. The measured gain of the skeleton slot aerial was about 2dB below the specified figure; this might be due to corrosion of the balun, of the aerial itself (5 years in the open) and the feed cable fixed to the aerial (same type and length as the cables used in conjunction with the other aerials).
It was therefore decided not to use this aerial for reference purposes.

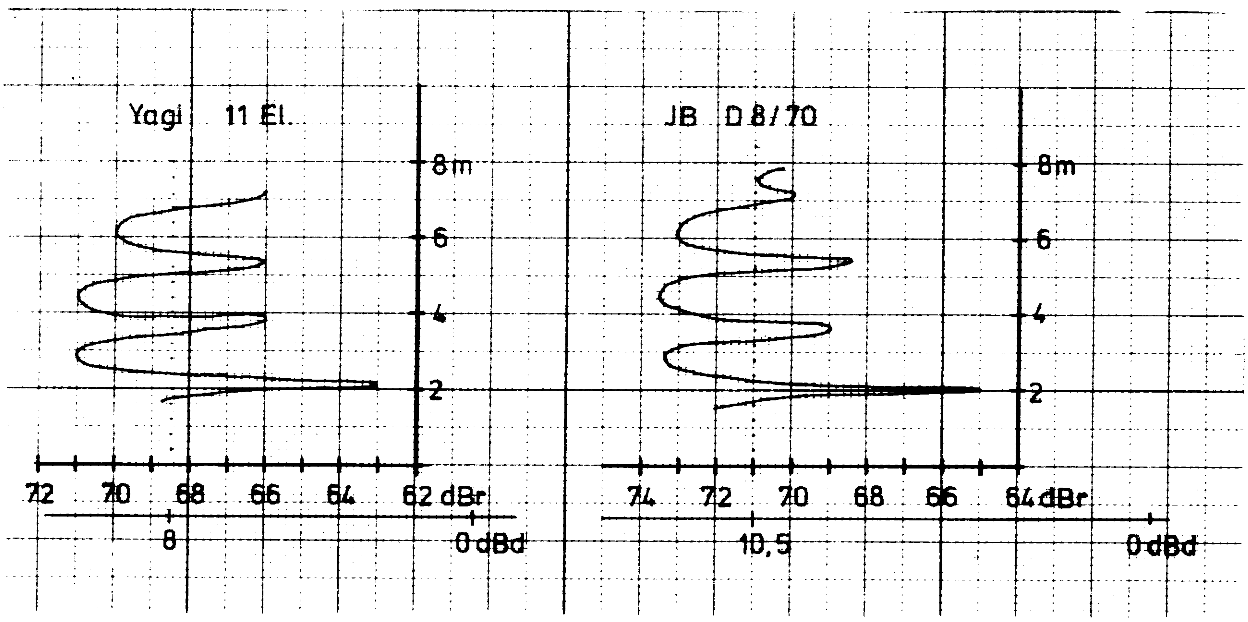


Figure 500

Properties at 432 MHz of a t.v.aerial produced by Hirschmann. The diagram to the right represents the unmodified version whereas the diagram to the left is that of the same aerial but equipped with a modified feed and connecting box. The feed is a twin-loop made from aluminium strip (1 x 5mm) and balancing is by a $\lambda/4$ detuning sleeve. This adds up to 8dB of additional gain.

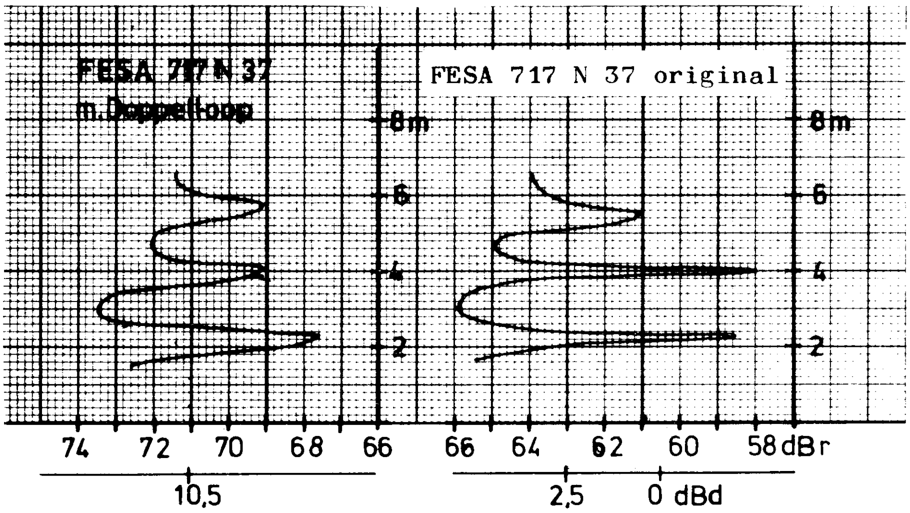


Figure 501

Another t.v.aerial of different design was equipped with the twin- loop feed and its performance measured. The obtained curve does, however, point to significant sidelobes in the vertical radiation pattern since the aerial voltage is clearly reduced by ground reflections - particularly at a height corresponding to that of the transmitting aerial.

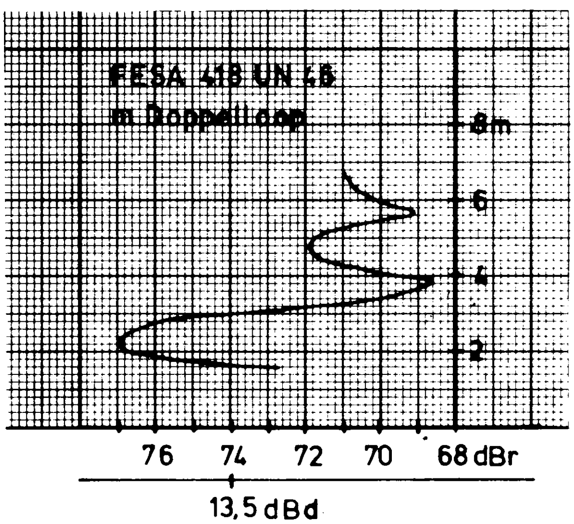


Figure 502
Height-dependent aerial voltage of a twin-loop aerial still under development. The dimensions are not yet optimized and yield a gain of 3.7dB. Upon installation of an additional feed of identical design in front of an enlarged reflector and combining the two feeds through a $\lambda/4$ transformer the expected gain increase of nearly 3dB was obtained.

twin-loop 2 x twin-loop, stacked horizontally

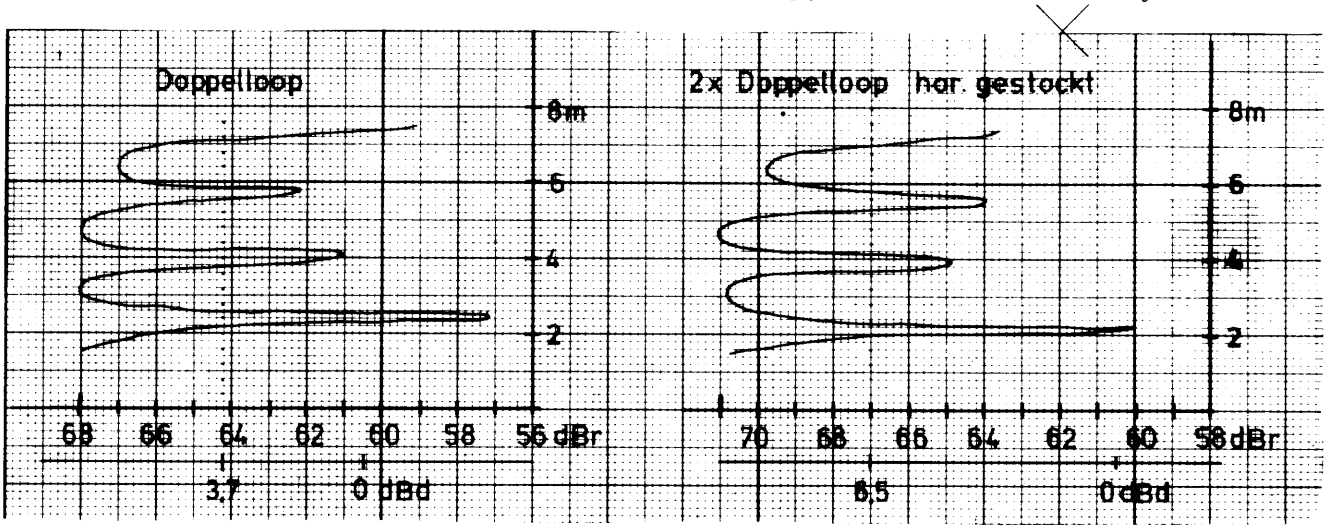


Figure 503
Stacked quad-array (DJ9HO) in front of the reflector grid of a t.v.aerial. The 70cm aerial is not yet optimized. A comparable arrangement for the 23cm band provided a gain of 14.5dBd and was measured in the anechoic chamber of Telefunken. This value was eventually confirmed, albeit with the improved radiator concept.

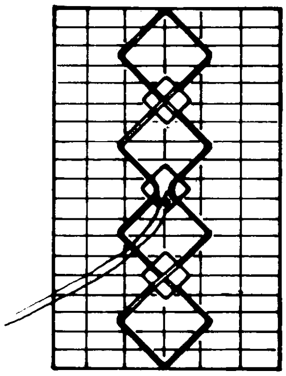


Figure 504
The vertically stacked quad array possesses distinct vertical directivity (reduced difference between maxima and minima). The gain figure of 7.7dB is not yet representative since the aerial was still in its design phase at the time of the tests. The right-hand diagram represents a simple twin-quad; after further optimization a gain of 9.5dB was eventually obtained.

stacked quad array twin-quad

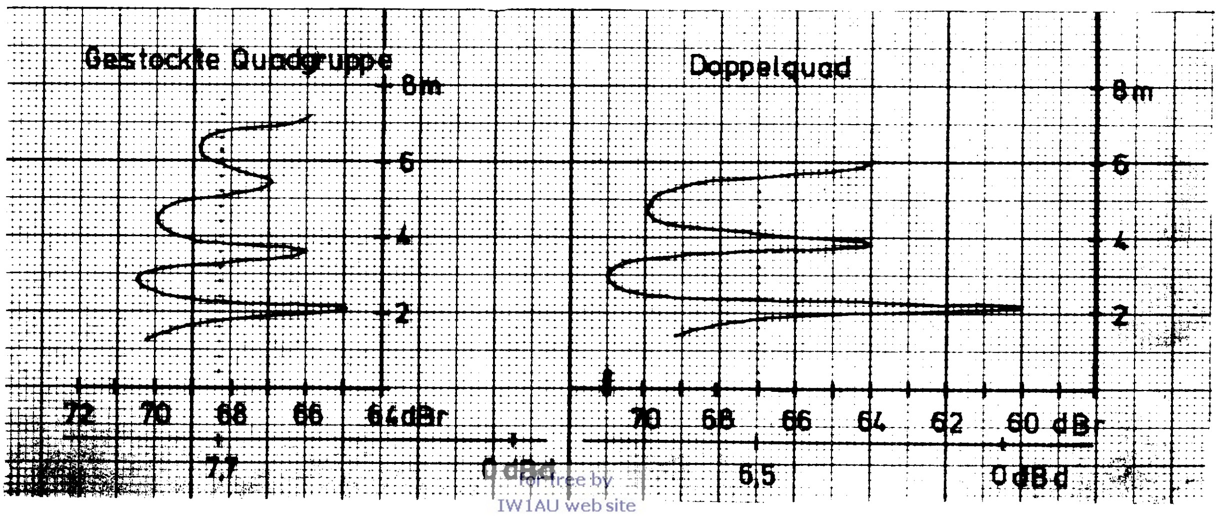
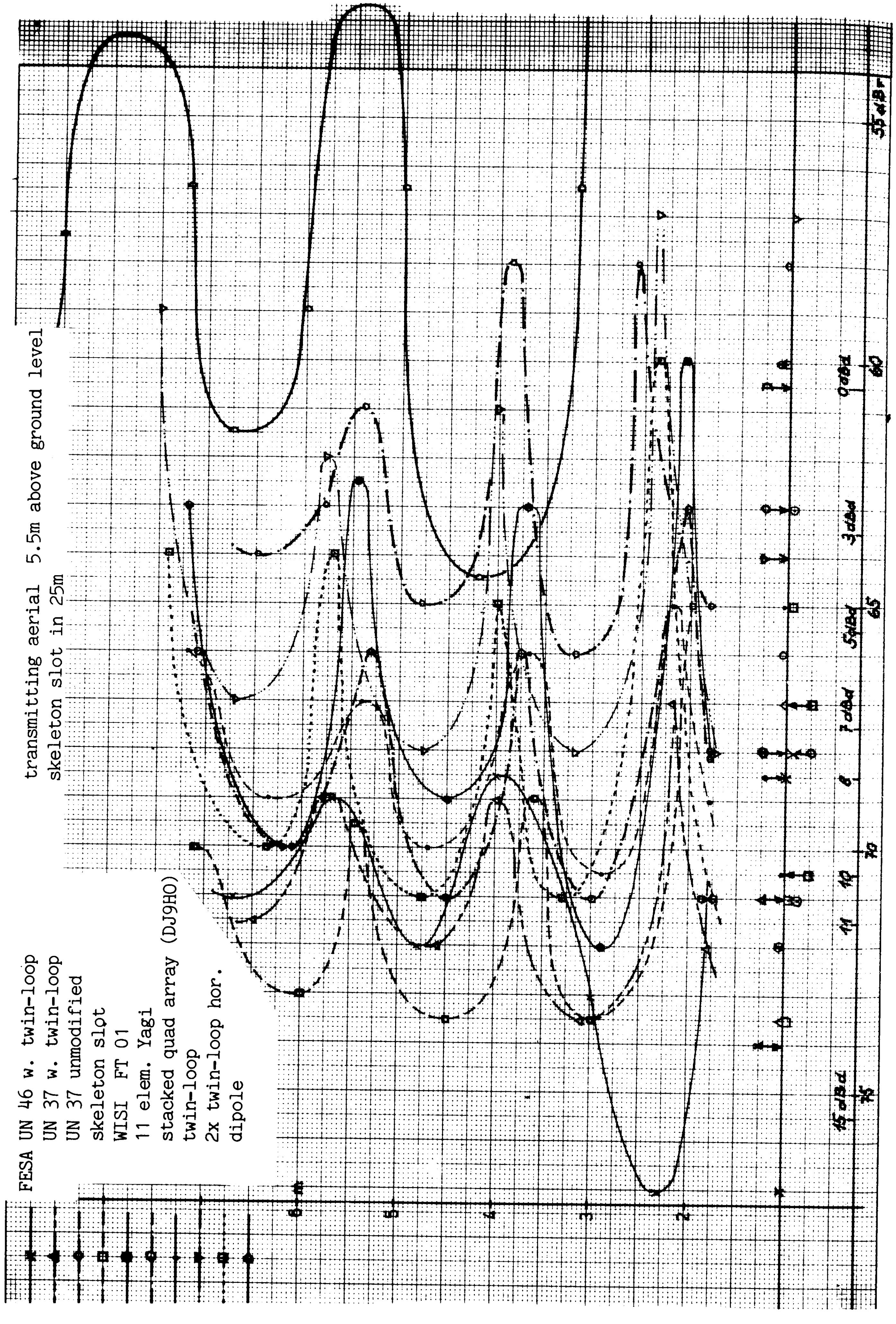


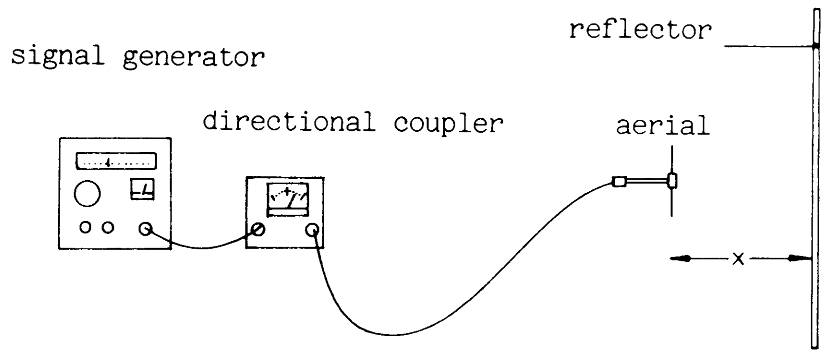
Figure 505 Interlaced presentation of all results



E.6.3.2 Gain computation from reflection measurements

A major advantage of this method is the possibility of performing the measurements inside the house or the garage - at least on aerials of reasonable physical dimensions (higher frequencies).
On 435 MHz various objects were tried as reflectors in this set-up and the results were compared. It was observed that (steel- reinforced) concrete (floor), fire-resistant steel doors, a metal plate or household aluminium foil (spread out on the floor) yielded comparable results.
In order to avoid lateral influences (reflections from walls, floor or other objects) the aerial was tied to the ceiling - the main radiation axis pointing downwards. By means of a piece of string it was now possible to adjust the aerial-to-reflector spacing to any desired value within the height of the room.

Figure 506 Test set-up for reflection measurements



Any reflecting object in the close vicinity of the aerial and protruding into the radiation lobe affects the matching and thus the VSWR. Gain computation by this method requires obtaining the reflection factor. To measure the reflection factor "r" we need a directional coupler having an instrument with a scale of 100 increments (such as a 100uA meter). At the beginning of the test series the signal generator is switched on, the aerial is disconnected (open circuit) and the instrument is adjusted to read 100% reflected power.

After connecting the aerial to the coupler a series of measurements may be performed ("r" as function of spacing "x"). The interesting and for the gain calculation useful range up to a distance of $\lambda/2$ from the reflector yields r-values of 25 to 100%.

Spacing	$\lambda/2$-dipole		$\lambda/2$-dipole with reflect.		twin-quad	
x [cm]	r [%]	G_i	r [%]	G_i	r [%]	G_i
5	75	1.36	95	1.73		
10	30	1.1	90	3.27	85	3.05
15	12	0.66	80	4.37	70	3.77
20	15	1.1	70	5.1	62	4.45
25	30	2.73	50	4.55	56	5.03
30	28	3.06	25	2.73	50	5.39
35	26	3.31	10	1.27	48	6.03
40	25	3.64	10	1.46	46	6.89
45	18	2.95	30	4.91	42	6.79
50	7	1.27	45	8.2	38	6.82
55	5	1.0	48	9.61	38	7.5
60	17	3.71	35	7.65	35	7.54

The reflection factor "r" is converted according to the equation (see E.6.2.3)

$$G_i = \frac{\pi \cdot r \cdot x}{\lambda}$$

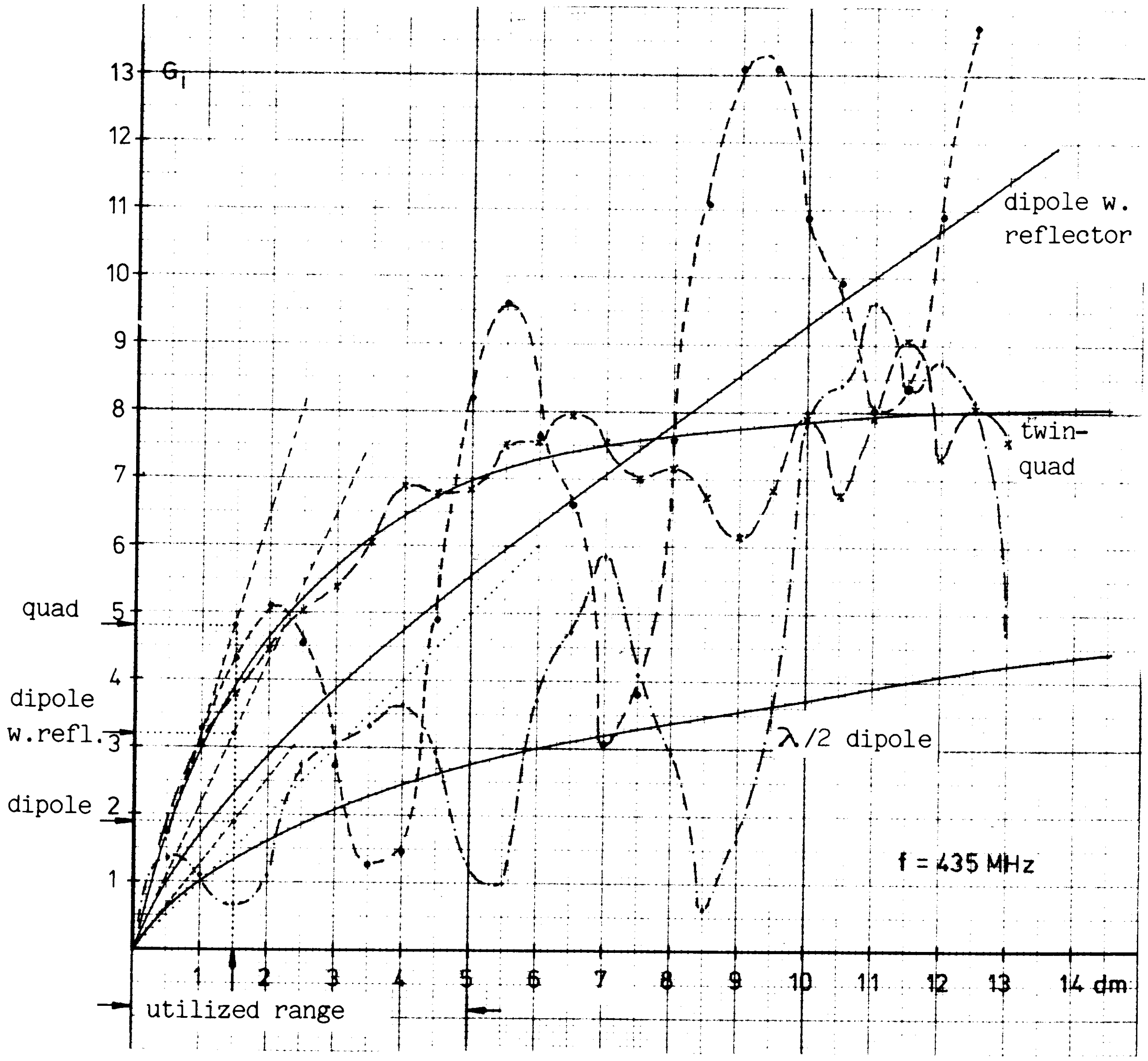
and displayed graphically. The result is an <u>initially</u> steeply ascending curve which is superimposed by a number of sinusoidal waves of various phase relations. <u>Each</u> curve is then subdued to graphical integration in order to obtain smooth curves. The gain may be calculated from the slope of the curves at the origin (may be obtained by drawing a straight line through the origin such that it is tangent to the respective curve). Assuming a scale corresponding to that of the diagram in figure 507 (1cm along the abscissa = 0.1m and 1 cm along the ordinate = gain factor 1) a line of slope 45° represents a gain of 0dBi.

$G_{i[dB]}$ = 20 log {1[cm] : 1[cm]} = 0 dBi

dBi = gain with respect to isotropic radiator
dBd = gain with respect to $\lambda/2$-dipole

Examples of gain computations from the following diagram:
Distance x = 15cm = 1.5dm was chosen as the starting point for calculating the slopes of the various curves. One should go vertically upwards from that starting point until the respective line is reached and read off the corresponding value from the side. This is at 1.9 in the case of the λ/2 dipole. From this we obtain:
$G_i[dB] = 20 \log (1.9 : 1.5) = 20 \log 1.27 = 2.05$ dBi ~ 0 dBd.
The intersection is at 4.8 in the case of the twin-quad and we obtain:
$G_i[dB] = 20 \log (4.8 : 1.5) = 20 \log 3.2 = 10.1$ dBi ~ 8 dBd

Figure 507 Graphical presentation of gain factor according to E.6.2.3



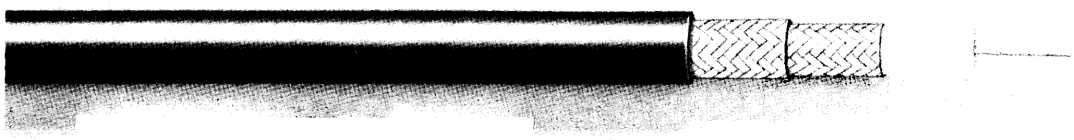
Recommended reading:
(1) Meinke Grundlach, Taschenbuch der Hochfrequenztechnik
(2) Rothammel, Antennenbuch
(3) Jasik, Antenna Engineering Handbook

G.1 Transmission lines (materials, attenuation and power specifications)

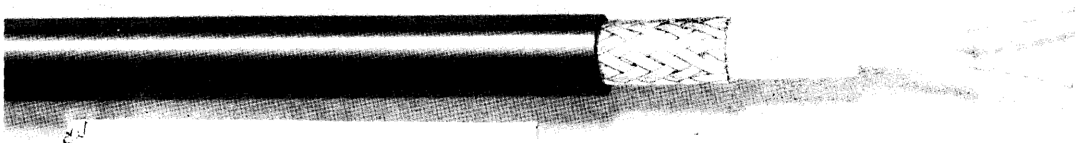
This section is supposed to provide all necessary information. Your local club may obtain a cable catalogue of the company Suhner through their representative at: 8000 München 90, Pfälzer-Wald-Str.68. The products of the company Hirschmann may be obtained from t.v.service dealers. A few of the commonly used types of the RG-series as well as the "Semi-rigid- cables" are presented albeit in short form.

Figure 508

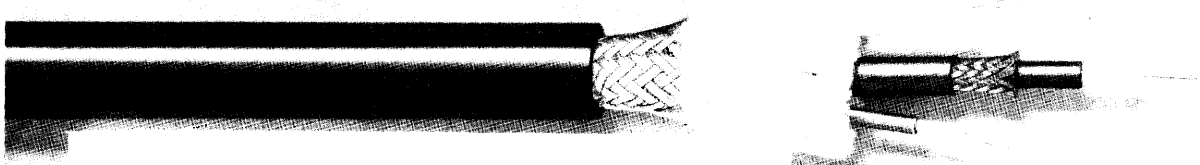
= single screening



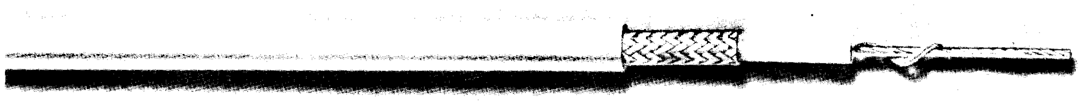
= double screening



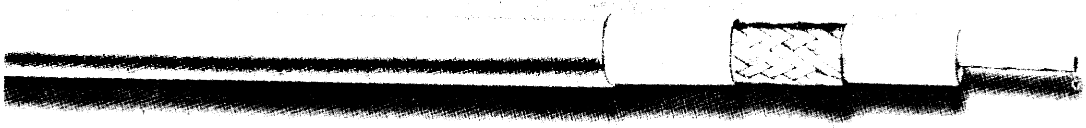
= symmetrical cable



= coaxial cable with control lines



= heat resistant cable

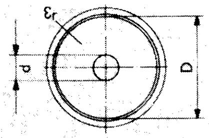


RF characteristics, cut-off frequency f_G

The construction and insulating materials used in radio frequency transmission lines makes them suitable for transmitting low-loss, broad-band signals. The applicable frequency range lies between direct current and the upper cut-off frequency where the propagation mode of the transmitted electromagnetic waves change. Flexible RF cables can be economically used up to frequencies of approx. 5 GHz.

Grenzfrequenz von Koaxialkabeln	*Cut-off frequency of coaxial cables*

$$f_G = \frac{2 \cdot c}{\pi \sqrt{\varepsilon_r}} \; \frac{1}{D+d} \; [\mathrm{Hz}]$$



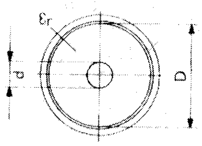
ε_r = Relative dielectric constant c = Velocity of light

Characteristic impedance Z_O

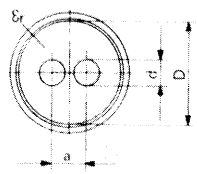
For reflection free RF transmission, Z_O at the input, along the cable and at the cable output must be uniform. Depending on the value of Z_O, an RF cable can be optimally dimensioned with respect to signal loss, dielectric strength or power transmission. Depending on the cable design and the insulating material used, different values of Z_O are preferred.
The characteristic impedance of RF cables between frequencies of 1 MHz and f_G is practically constant.
The following empiric equations apply:

Koaxial *Coaxial*	Symmetrisch abgeschirmt *Balanced, screened*

$$Z_O = \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \frac{D}{d} \; [\Omega]$$



$$Z_O = \frac{120}{\sqrt{\varepsilon_r}} \cdot \ln \left[\frac{2a}{d} \cdot \frac{D^2 - a^2}{D^2 + a^2} \right] \; [\Omega]$$



Typical characteristic impedances:
Coaxial: 50, 60, 75, 95, 120 ohm
Balanced: 95, 125, 150, 240, 300 ohm

Power rating

The power of a signal transmitted in an RF cable is limited by the resulting heat and the softening temperature of the insulation. The heat loss in the cable grows as frequencies, the dielectric constant ε_r and the dissipation factor tgδ increase. The permissible power transmission depends on the diameter ratio D/d as well as on the diameter D alone.
In addition, ambient temperature and voltage reflections also influence the power rating.

Typical diameter ratios D/d

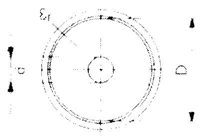
Isolationsmaterial *Insulation material*	ε_r	koaxial/*coaxial* 50 Ω D/d	60 Ω D/d	75 Ω D/d	symmetrisch/*balanced* 95 Ω a/d	95 Ω D/a
Voll-PE/ *Solid PE*	2,28	3,6	4,7	6,7	1,9	3,2

Capacitance C

The capacitance increases proportionally with cable length (pF/m). As Z_o, C depends on ε_r, the diameter and ratio D/d. Above 1 MHz it is independent of the frequency.

Kapazität von Koaxialkabeln *Capacitance of coaxial cables*

$$C = \frac{\varepsilon_r}{18 \cdot 10^{-3} \ln \frac{D}{d}}$$



Typical vaues of C Coaxial and Balanced

Isolationsmaterial *Insulation material*	ε_r	50 Ω pF/m	60 Ω pF/m	75 Ω pF/m	95 Ω pF/m
Voll-PE Solid PE	2,28	101	84	67	54
PTFE/FEP	2,10	96		64	

Attenuation

Signal loss in RF cables is caused by conductor losses (skin effect) increasing proportionally to $\sqrt{f}$, dielectric losses which occur above 10 MHz only and increase proportionally to f.
The signal loss depends on the diameter ratio D/d as well as on the diameter D. In addition, temperature and cable aging influence the attenuation.
α is calculated as the logarithmic ratio of the signal voltage at the cable input (U_1) and the cable output (U_2).
The signal loss calculated with natural logarithms is expressed in Nepers (N).
The loss based on decadic logarithms is expressed in decibles (dB).
α [N/m] equals 8.686 α [dB/m] α [dB/m] equals 0.115 α [N/m]

Velocity of signal propagation v_r

Electromagnetic waves ideally move at the speed of light (c).
In RF cables:

$$v_r = \frac{c}{\sqrt{\varepsilon_r}}$$

A typical feature of RF cables is that signals at all frequencies of the applicable band width (DC...F_G) propagate at the same velocity. This means that broad band signals are practically free of phase distortion.

Typical values of v_r

Isolationsmaterial	ε_r	v_r [% c]	Signal Laufzeit [n sec/m]
Voll-PE	2,28	66,2	5,03
PTFE/FEP	2,1	69,0	4,83
Schaum-PE	1,5	81,6	4,08
Luft, Vakuum	1,0	100	$3,3\dot{3}$

$$\alpha\,[N/m] = \ln \frac{U_1}{U_2} \qquad \alpha\,[dB/m] = 20 \log_{10} \frac{U_1}{U_2}$$

Availability of a cable catalogue and suppliers have been mentioned before. Some information on popular cable types is, however, included for convenience. The diagram provides quick and easy estimation of attenuation versus frequency. The attenuation is governed by the diameter of the insulating material; since there are no significant differences between 50 ohm and 75 ohm cables the diagram for 50 ohm is sufficient.

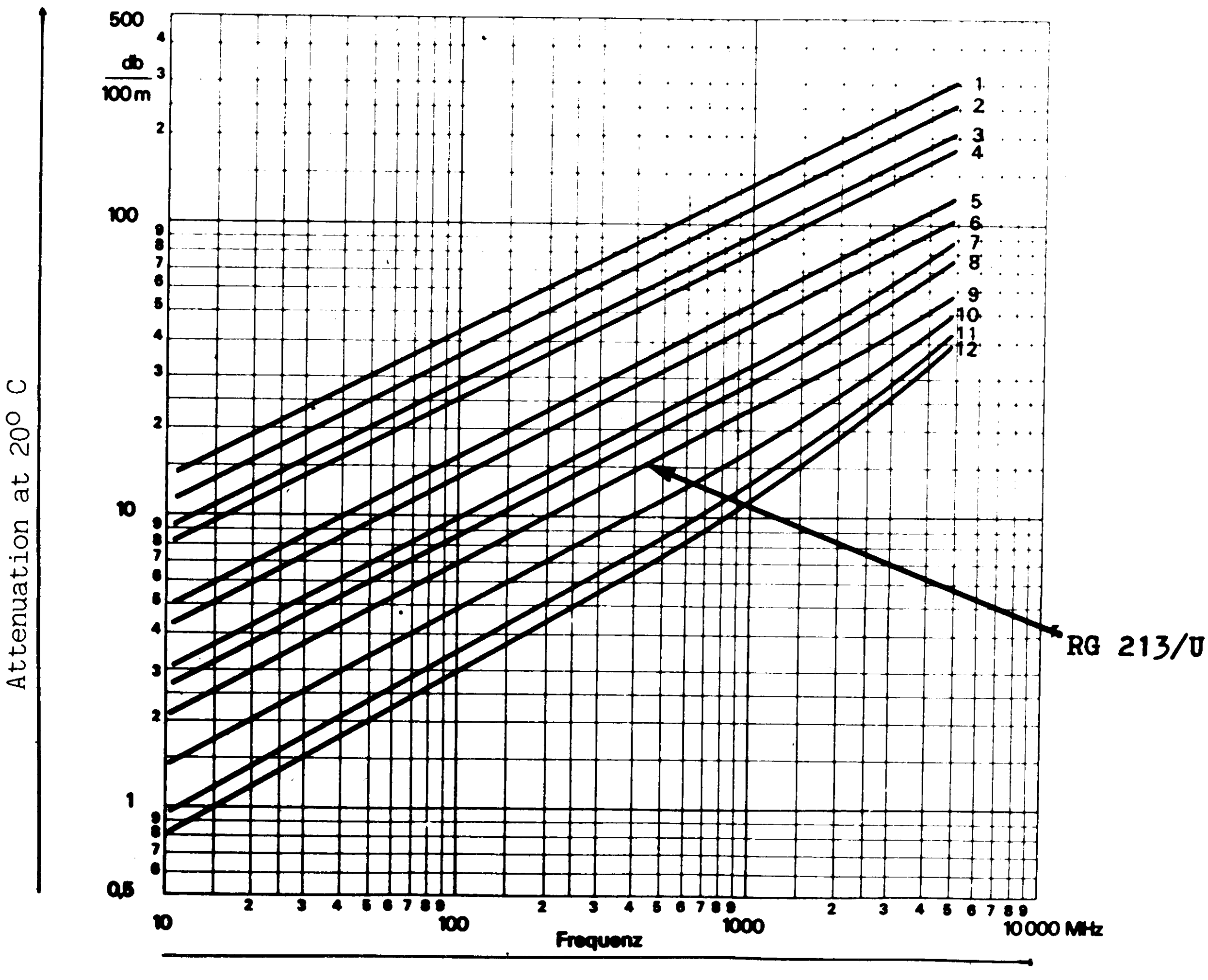
cable type	imped. [ohm]	capacit. [pF/m]	U_{max} [kV]	inner cond. mater.diam. [mm]	insulat. mat.diam. [mm]	screeng. mat.diam [mm]
RG-8A/U	52 ± 2	97	5	Cu 7x0.72	PE 7.25	Cu 8.1
RG-9A/U	51 ± 2	98.5	4	Ag 7x0.72	PE 7.1	AgAg 8.6
RG58C/U	50 ± 2	101	1.9	Sn 19x0.18	PE 2.95	Sn 3.6
RG59B/U	75 ± 2	67	2.3	CuSt solid	PR 3.70	Cu 4.5
RG213/U	50 ± 2	101	5	Cu 7x0.75	PE 7.25	Cu 8.1
RG214/U	50 ± 2	101	5	Ag 7x0.75	PE 7.25	AgAg 8.7

PE = polyethylene
Ag = copper wire, silver plated
Sn = copper wire, tinned
Cu = copper wire, bare
CuSt = steel wire, copper plated
Velocity factor of these types is 0.662 (solid PE)

Attenuation and power
Coaxial cables 50 ohm, polyethylene insulation.
Temperature coefficient of attenuation = 0.002 / degree C

cable specifications:

curve no.	inner conduct.	insul.[mm]
1	solid wire	0.87
2	solid wire	1.05
3	stranded wire	1.5
4	solid wire	1.5
5	stranded wire	2.95
6	solid wire	2.95
7	stranded wire	4.8
8	solid wire	4.8
9	stranded wire	7.25
10	stranded wire	11.5
11	solid wire	17.3
12	solid wire	23.1



Semi-rigid-Kabel — Typenübersicht

	SR 04	SR 06	SR 1	SR 2	SR 2-75	SR 3	SR 3-1	SR 3-75	SR 5	SR 7
Electrical data				RG-405/U		RG-402/U			RG-401/U	
Impedance (Ω)	50 ±2	50 ±2	50 ±2	50±1,5	75±1,5	50 ±1	50 ±0,5	75 ±1	50 ±0,5	50 ±0,5
Capacity (pF/m)	96	96	96	96	64	96	96	64	96	101
Signal delay (ns/m)	4,80	4,80	4,80	4,80	4,80	4,80	4,80	4,80	4,80	5,03
Velocity factor (%)	69,5	69,5	69,5	69,5	69,5	69,5	69,5	69,5	69,5	66,2
Operating voltage V eff. 50 Hz.	200	400	800	1500	1200	2.500	2.500	2.000	3.000	4.500
Mechanical data										
Temp.range (°C)	-70/+85	-70/+100	-70/+150	-70/+150	-70/+150	-70/+150	-70/+150	-70/+150	-70/+150	-40/+70
Inner cond. Ø (mm)	0,11	0,20	0,29	0,51	0,29	0,91	0,91	0,51	1,63	2,1
Inner conductor	CuStAg	CuStAg	CuStAg	CuStAg	CuStAg	CuStAg	CuStAg	CuStAg	Ag	Ag
Dielectric Ø (mm)	0,38	0,64	0,89	1,68	1,68	3,00	3,00	3,00	5,33	7,25
Dielectric	PTFE	PTFE	PTFE	PTFE	PTFE	PTFE	PTFE	PTFE	PTFE	PE
Outer diam. (mm)	0,51	0,86	1,19	2,20	2,20	3,58	3,58	3,58	6,35	8,25
Bending radius (mm)	≧1,6	≧3,2	≧3,2	≧5,0	≧5,0	≧6,5	≧6,5	≧6,5	≧95,0	≧50,0
Weight (kg/100m)	0,15	0,2	0,8	2,1	2,0	4,8	4,8	4,6	15,0	17,0
Available length nominal (m) (Deviat. reserved)	ca. 1	ca. 3	ca. 5	≧6	≧6	≧6	≧6	≧6	≧6	≧6
Suitable connectors			MCX	BNC, C, N, SMA, SMB, SMC, SMS, TNC, MCX	BNC, C, N, TNC	BNC, C, N, SMA, TNC PC 7	BNC, C, N, SMA, TNC PC 7	BNC, C, N, TNC	C, N	C, N

Waranty: All specified technical data are guaranteed for a mechanically undamaged cable that was bent only once to the minimum bending radius.

Explanations: CuStAg = silvered copper-plated steel wire, Ag = silver-plated copper wire, PE = polyethylene, PTFE polytetrafluorethylene

Special cables: On request (surcharge). Impedances and dimensions that differ from standard specifications on request.

Surface plating: silver "Ag", gold "Au", tinn "Sn", SUCOPLATE "SP". Add symbol when ordering; example: SR3/Ag

Fitting of connectors: All listed types may be delivered finished and equipped with suitable SUHNER connectors according to customers specifications. Delivery: In rings.

suhner
elektronik
gmbh HF-STECKER HF-KABEL

Postfach 90 06 60, 8000 München 90
Telefon 089/68 10 48-49, Telex 05 29 767 sem-d
[089/68 10 11]

HF-Kabel

Koaxiale HF-Kabel

Hirschmann

						
Typ	**Koka 14**	**Koka 711**	**Koka 712**	**Koka 792**	**Koka 741**	**Koka 751**
Bestell-Nr.	**198 061-000**	**198 711-000**	**198 712-000**	**198 792-001**	**198 741-000**	**198 751-000**
Wellenwiderstand	60 Ohm	75 Ohm				
Anwendung	Aufputz, Imputz, in Rohren, Installationen			Empfänger-Anschluß-Kabel (Bausätze)	Aufputz Imputz Erdkabel	Luftkabel
Besonderheit	im Stammnetz Anschluß von Antennensteckdosen		in (Haupt-) Verstärkernähe und Verteiler; doppelt geschirmt	hochflexibel	starker Schutzmantel	Tragseil 4 mm ø Belastung 560 kp
Lieferlängen	Ringe zu 50 und 100 m				auf Anfrage	
Abmessungen: Innenleiter ø mm	1	0,7	0,75	0,6	1,1	
Abmessungen: Abschirmung ø mm (innen)	3,9	4,3	4,8	3,3	7,3	
Abmessungen: Außenhülle ø mm (außen)	5,8	6,3	6,8	5,0	10,6	
Material: Innenleiter	Kupfer versilbert	Kupfer blank		Kupferlitze blank 7 x 0,2 mm ø	Kupfer blank	
Material: Isolierung	Vollpolyäthylen					
Material: Abschirmung	Kupfergeflecht versilbert	Kupfergeflecht verzinnt	Kupferfolie + Kupfergeflecht blank	Kupfergeflecht blank	glattes Kupferband, blank längslaufend, überlappend	
Material: Mantel	Kunststoff, weiß				Polyäthylen, schwarz	
Minimaler Biegeradius: für einmaliges Biegen	1 cm	1 cm	3,5 cm	1 cm	10,5 cm	
Minimaler Biegeradius: für mehrmaliges Biegen	10 cm	10 cm	30 cm	10 cm	57,5 cm	
Gleichstromwiderstand*	3,75 Ω	5,55 Ω	4,6 Ω	10,02 Ω	2,31 Ω	
Verkürzungsfaktor	0,67	0,66	0,66	0,66	0,66	
Kopplungs-Widerstand	< 200 m Ω	< 200 m Ω				
max. Betriebspegel	105 dBμV	110 dBμV	130 dBμV	105 dBμV	130 dBμV	
Schirmungsmaß	50 dB	55 dB	75 dB	50 dB	75 dB	

* Gleichstromwiderstand der Schleife aus Innenleiter und Abschirmung bei 100 m Kabel und + 20° C

Koaxiale HF-Kabel

Typ	Dämpfung (dB) pro 100 m bei + 20° C und bei 1 MHz	50 MHz	100 MHz	200 MHz	500 MHz	600 MHz	700 MHz	800 MHz
Koka 14	1,0	7,0	10,0	15,0	25,0	27,5	30,5	33,5
Koka 711	0,9	7,0	10,5	14,5	24,0	26,3	28,6	31,0
Koka 712	0,8	5,6	8,4	11,8	19,3	21,0	22,9	25,2
Koka 792	1,1	8,0	12,0	18,0	30,0	34,0	37,0	41,0
Koka 741 **Koka 751**	0,5	3,5	5,3	7,7	12,9	15,0	15,7	17,2

Frequenzen der Fernsehkanäle und des UKW-Bereichs

Bereich I

VHF
47-68 MHz 6,4–4,4 m

47-54 MHz Kanal 2
54-61 MHz Kanal 3
61-68 MHz Kanal 4

Bereich II

UKW-Rundfunk
87,5–104 MHz 3,4–2,9 m

Bereich III

VHF
174–230 MHz 1,7–1,3 m

174–181 MHz Kanal 5
181–188 MHz Kanal 6
188–195 MHz Kanal 7
195–202 MHz Kanal 8
202–209 MHz Kanal 9
209–216 MHz Kanal 10
216–223 MHz Kanal 11
223–230 MHz Kanal 12

Bereiche IV/V

UHF
470-790 MHz 64-38 cm

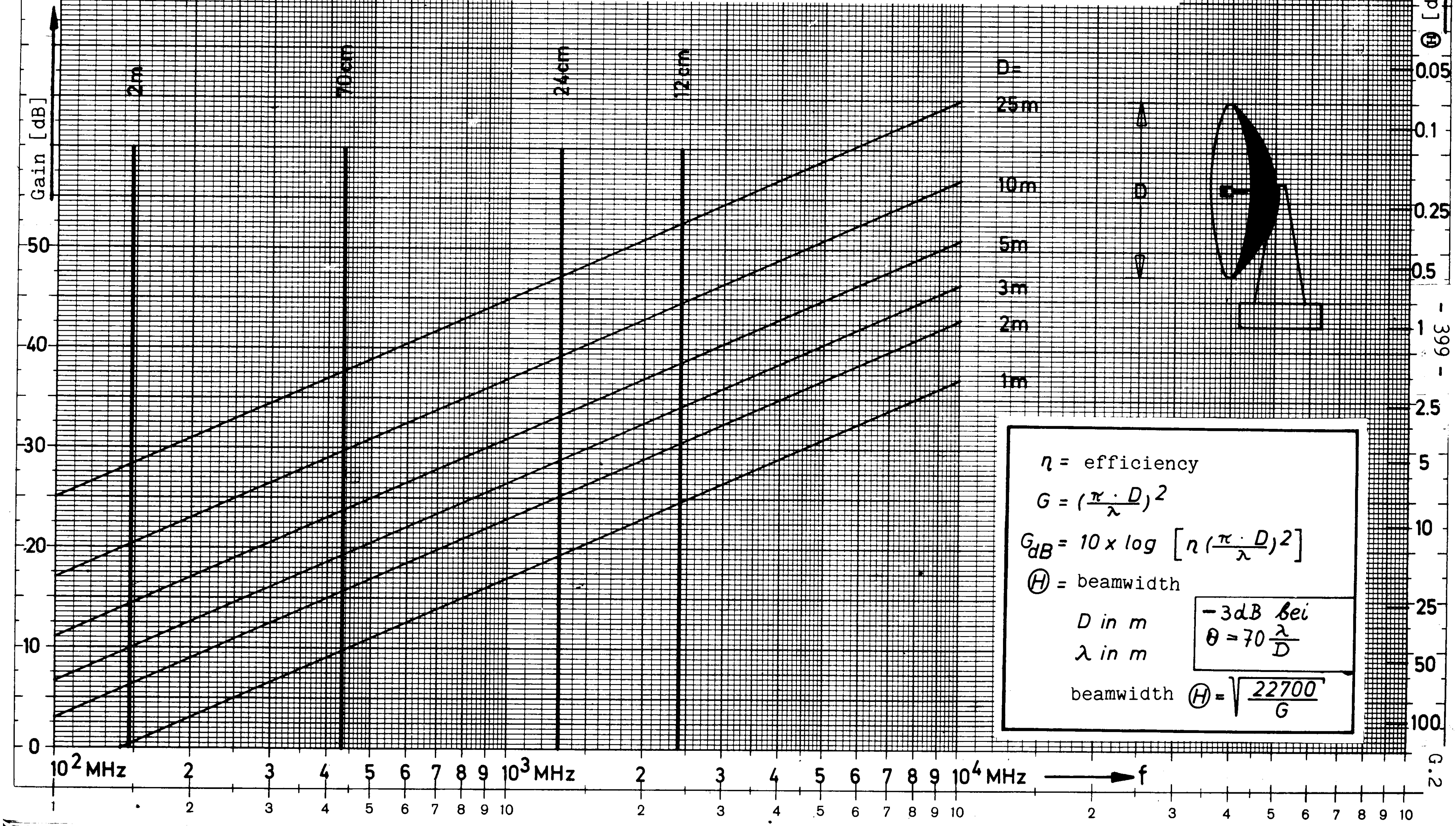
Bereich IV

470–478 MHz Kanal 21
478–486 MHz Kanal 22
486–494 MHz Kanal 23
494–502 MHz Kanal 24
502–510 MHz Kanal 25
510–518 MHz Kanal 26
518–526 MHz Kanal 27
526–534 MHz Kanal 28
534–542 MHz Kanal 29
542–550 MHz Kanal 30
550–558 MHz Kanal 31
558–566 MHz Kanal 32
566–574 MHz Kanal 33
574–582 MHz Kanal 34
582–590 MHz Kanal 35
590–598 MHz Kanal 36
598–606 MHz Kanal 37
606–614 MHz Kanal 38
614–622 MHz Kanal 39

Bereich V

622–630 MHz Kanal 40
630–638 MHz Kanal 41
638–646 MHz Kanal 42
646–654 MHz Kanal 43
654–662 MHz Kanal 44
662–670 MHz Kanal 45
670–678 MHz Kanal 46
678–686 MHz Kanal 47
686–694 MHz Kanal 48
694–702 MHz Kanal 49
702–710 MHz Kanal 50
710–718 MHz Kanal 51
718–726 MHz Kanal 52
726–734 MHz Kanal 53
734–742 MHz Kanal 54
742–750 MHz Kanal 55
750–758 MHz Kanal 56
758–766 MHz Kanal 57
766–774 MHz Kanal 58
774–782 MHz Kanal 59
782–790 MHz Kanal 60

G.2 Size, gain and beamwidth of parabolic aerials, efficiency 55%



G.3.1 dBm/Watt conversion table

Ringmixers are being increasingly used in amateur circles. Data - in particular the required injection signal power level - are specified in dBm by many manufacturers (7 dBm, 17 dBm etc.). This value "dBm" refers to power with respect to one milliwatt.

```
+----------------+
| 0 dBm = 1 mW  |
+----------------+
```

The following list should help the OM to become aquainted with the dBm:

dBm	Milliwatt	dBm	Milliwatt	dBm	Milliwatt	dBm	Milliwatt	dBm	Milliwatt	dBm	Milliwatt	dBm	Milliwatt
-18.0	.0158	-11.1	.0776	-4.2	.380	2.7	1.86	9.6	9.12	16.5	44.7	23.4	219
-17.9	.0162	-11.0	.0794	-4.1	.389	2.8	1.91	9.7	9.33	16.6	45.7	23.5	224
-17.8	.0166	-10.9	.0813	-4.0	.398	2.9	1.95	9.8	9.55	16.7	46.8	23.6	229
-17.7	.0170	-10.8	.0832	-3.9	.407	3.0	2.00	9.9	9.77	16.8	47.9	23.7	234
-17.6	.0174	-10.7	.0851	-3.8	.417	3.1	2.04	10.0	10.0	16.9	49.0	23.8	240
-17.5	.0178	-10.6	.0871	-3.7	.427	3.2	2.09	10.1	10.2	17.0	50.1	23.9	245
-17.4	.0182	-10.5	.0891	-3.6	.437	3.3	2.14	10.2	10.5	17.1	51.3	24.0	251
-17.3	.0186	-10.4	.0912	-3.5	.447	3.4	3.19	10.3	10.7	17.2	52.5	24.1	257
-17.2	.0191	-10.3	.0933	-3.4	.457	3.5	2.24	10.4	11.0	17.3	53.7	24.2	263
-17.1	.0195	-10.2	.0955	-3.3	.468	3.6	2.29	10.5	11.2	17.4	55.0	24.3	269
-17.0	.0200	-10.1	.0977	-3.2	.479	3.7	2.34	10.6	11.5	17.5	56.2	24.4	275
-16.9	.0204	-10.0	.100	-3.1	.490	3.8	2.40	10.7	11.7	17.6	57.5	24.5	282
-16.8	.0209	-9.9	.102	-3.0	.501	3.9	2.45	10.8	12.0	17.7	58.9	24.6	288
-16.7	.0214	-9.8	.105	-2.9	.513	4.0	2.51	10.9	12.3	17.8	60.3	24.7	295
-16.6	.0219	-9.7	.107	-2.8	.525	4.1	2.57	11.0	12.6	17.9	61.7	24.8	302
-16.5	.0224	-9.6	.110	-2.7	.537	4.2	2.63	11.1	12.9	18.0	63.1	24.9	309
-16.4	.0229	-9.5	.112	-2.6	.550	4.3	2.69	11.2	13.2	18.1	64.6	25.0	316
-16.3	.0234	-9.4	.115	-2.5	.562	4.4	2.75	11.3	13.5	18.2	66.1	25.1	324
-16.2	.0240	-9.3	.117	-2.4	.575	4.5	2.82	11.4	13.8	18.3	67.6	25.2	331
-16.1	.0245	-9.2	.120	-2.3	.589	4.6	2.88	11.5	14.1	18.4	69.2	25.3	339
-16.0	.0251	-9.1	.123	-2.2	.603	4.7	2.95	11.6	14.5	18.5	70.8	25.4	347
-15.9	.0257	-9.0	.126	-2.1	.617	4.8	3.02	11.7	14.8	18.6	72.4	25.5	355
-15.8	.0263	-8.9	.129	-2.0	.631	4.9	3.09	11.8	15.1	18.7	74.1	25.6	363
-15.7	.0269	-8.8	.132	-1.9	.646	5.0	3.16	11.9	15.5	18.8	75.9	25.7	372
-15.6	.0275	-8.7	.135	-1.8	.661	5.1	3.24	12.0	15.8	18.9	77.6	25.8	380
-15.5	.0282	-8.6	.138	-1.7	.676	5.2	3.31	12.1	16.2	19.0	79.4	25.9	389
-15.4	.0288	-8.5	.141	-1.6	.692	5.3	3.39	12.2	16.6	19.1	81.3	26.0	398
-15.3	.0295	-8.4	.145	-1.5	.708	5.4	3.47	12.3	17.0	19.2	83.2	26.1	407
-15.2	.0302	-8.3	.148	-1.4	.724	5.5	3.55	12.4	17.4	19.3	85.1	26.2	417
-15.1	.0309	-8.2	.151	-1.3	.741	5.6	3.63	12.5	17.8	19.4	87.1	26.3	427
-15.0	.0316	-8.1	.155	-1.2	.759	5.7	3.72	12.6	18.2	19.5	89.1	26.4	437
-14.9	.0324	-8.0	.158	-1.1	.776	5.8	3.80	12.7	18.6	19.6	91.2	26.5	447
-14.8	.0331	-7.9	.162	-1.0	.794	5.9	3.89	12.8	19.1	19.7	93.3	26.6	457
-14.7	.0339	-7.8	.166	-0.9	.813	6.0	3.98	12.9	19.5	19.8	95.5	26.7	468
-14.6	.0347	-7.7	.170	-0.8	.832	6.1	4.07	13.0	20.0	19.9	97.7	26.8	479
-14.5	.0355	-7.6	.174	-0.7	.851	6.2	4.17	13.1	20.4	20.0	100	26.9	490
-14.4	.0363	-7.5	.178	-0.6	.871	6.3	4.27	13.2	20.9	20.1	102	27.0	501
-14.3	.0372	-7.4	.182	-0.5	.891	6.4	4.37	13.3	21.4	20.2	105	27.1	513
-14.2	.0380	-7.3	.186	-0.4	.912	6.5	4.47	13.4	21.9	20.3	207	27.2	525
-14.1	.0389	-7.2	.191	-0.3	.933	6.6	4.57	13.5	22.4	20.4	110	27.3	537
14.0	.0398	-7.1	.195	-0.2	.955	6.7	4.68	13.6	22.9	20.5	112	27.4	550
13.9	.0407	-7.0	.200	-0.1	.977	6.8	4.79	13.7	23.4	20.6	115	27.5	562
13.8	.0417	-6.9	.204	0.0	1.00	6.9	4.90	13.8	24.0	20.7	117	27.6	575
13.7	.0427	-6.8	.209	0.1	1.02	7.0	5.01	13.9	24.5	20.8	120	27.7	589
13.6	.0437	-6.7	.214	0.2	1.05	7.1	5.13	14.0	25.1	20.9	123	27.8	603
13.5	.0447	-6.6	.219	0.3	1.07	7.2	5.25	14.1	25.7	21.0	126	27.9	617
13.4	.0457	-6.5	.224	0.4	1.10	7.3	5.37	14.2	26.3	21.1	129	28.0	631
13.3	.0468	-6.4	.229	0.5	1.12	7.4	5.50	14.3	26.9	21.2	132	28.1	646
13.2	.0479	-6.3	.234	0.6	1.15	7.5	5.62	14.4	27.5	21.3	135	28.2	661
13.1	.0490	-6.2	.240	0.7	1.17	7.6	5.75	14.5	28.2	21.4	138	28.3	676
13.0	.0501	-6.1	.245	0.8	1.20	7.7	5.89	14.6	28.8	21.5	141	28.4	692
12.9	.0513	-6.0	.251	0.9	1.23	7.8	6.03	14.7	29.5	21.6	145	28.5	708
12.8	.0525	-5.9	.257	1.0	1.26	7.9	6.17	14.8	30.2	21.7	148	28.6	724
12.7	.0537	-5.8	.263	1.1	1.29	8.0	6.31	14.9	30.9	21.8	151	28.7	741
12.6	.0550	-5.7	.269	1.2	1.32	8.1	6.46	15.0	31.6	21.9	155	28.8	759
12.5	.0562	-5.6	.275	1.3	1.35	8.2	6.61	15.1	32.4	22.0	158	28.9	776
12.4	.0575	-5.5	.282	1.4	1.38	8.3	6.76	15.2	33.1	22.1	162	29.0	794
12.3	.0589	-5.4	.288	1.5	1.41	8.4	6.92	15.3	33.9	22.2	166	29.1	813
12.2	.0603	-5.3	.295	1.6	1.45	8.5	7.08	15.4	34.7	22.3	170	29.2	832
12.1	.0617	-5.2	.302	1.7	1.48	8.6	7.24	15.5	35.5	22.4	174	29.3	852
12.0	.0631	-5.1	.309	1.8	1.51	8.7	7.41	15.6	36.3	22.5	178	29.4	871
11.9	.0646	-5.0	.316	1.9	1.55	8.8	7.59	15.7	37.2	22.6	182	29.5	891
11.8	.0661	-4.9	.324	2.0	1.58	8.9	7.76	15.8	38.0	22.7	186	29.6	912
11.7	.0676	-4.8	.331	2.1	1.62	9.0	7.94	15.9	38.9	22.8	191	29.7	933
11.6	.0692	-4.7	.339	2.2	1.66	9.1	8.13	16.0	39.8	22.9	195	29.8	955
11.5	.0708	-4.6	.347	2.3	1.70	9.2	8.32	16.1	40.7	23.0	200	29.9	977
11.4	.0724	-4.5	.355	2.4	1.74	9.3	8.51	16.2	41.7	23.1	204	30.0	1000
11.3	.0741	-4.4	.363	2.5	1.78	9.4	8.71	16.3	42.7	23.2	209		
11.2	.0759	-4.3	.372	2.6	1.82	9.5	8.91	16.4	43.7	23.3	214		

dBm	Watts	dBm	Watts	dBm	Watts	dBm	Watts	dBm	Watts	dBm	Watts
30.1	1.02	36.8	4.79	43.5	22.40	50.2	105.00	56.9	490.00	63.6	2290.00
30.2	1.05	36.9	4.90	43.6	22.90	50.3	107.00	57.0	501.00	63.7	2340.00
30.3	1.07	37.0	5.01	43.7	23.40	50.4	110.00	57.1	513.00	63.8	2400.00
30.4	1.10	37.1	5.13	43.8	24.00	50.5	112.00	57.2	525.00	63.9	2450.00
30.5	1.12	37.2	5.25	43.9	24.50	50.6	115.00	57.3	537.00	64.0	2510.00
30.6	1.15	37.3	5.37	44.0	25.10	50.7	117.00	57.4	550.00	64.1	2570.00
30.7	1.17	37.4	5.50	44.1	25.7	50.8	120.00	57.5	562.00	64.2	2630.00
30.8	1.20	37.5	5.62	44.2	26.30	50.9	123.00	57.6	575.00	64.3	2690.00
30.9	1.23	37.6	5.75	44.3	26.90	51.0	126.00	57.7	589.00	64.4	2750.00
31.0	1.26	37.7	5.89	44.4	27.50	51.1	129.00	57.8	603.00	64.5	2820.00
31.1	1.29	37.8	6.03	44.5	28.20	51.2	132.00	57.9	617.00	64.6	2880.00
31.2	1.32	37.9	6.17	44.6	28.80	51.3	135.00	58.0	631.00	64.7	2950.00
31.3	1.35	38.0	6.31	44.7	29.50	51.4	138.00	58.1	646.00	64.8	3020.00
31.4	1.38	38.1	6.46	44.8	30.20	51.5	141.00	58.2	661.00	64.9	3090.00
31.5	1.41	38.2	6.61	44.9	30.90	51.6	145.00	58.3	676.00	65.0	3160.00
31.6	1.45	38.3	6.76	45.0	31.60	51.7	148.00	58.4	692.00	65.1	3240.00
31.7	1.48	38.4	6.92	45.1	32.40	51.8	151.00	58.5	708.00	65.2	3310.00
31.8	1.51	38.5	7.08	45.2	33.10	51.9	155.00	58.6	724.00	65.3	3390.00
31.9	1.55	38.6	7.24	45.3	33.90	52.0	158.00	58.7	741.00	65.4	3470.00
32.0	1.58	38.7	7.41	45.4	34.70	52.1	162.00	58.8	759.00	65.5	3550.00
32.1	1.62	38.8	7.59	45.5	35.50	52.2	166.00	58.9	776.00	65.6	3630.00
32.2	1.66	38.9	7.76	45.6	36.30	52.3	170.00	59.0	794.00	65.7	3720.00
32.3	1.70	39.0	7.94	45.7	37.20	52.4	174.00	59.1	813.00	65.8	3800.00
32.4	1.74	39.1	8.13	45.8	38.00	52.5	178.00	59.2	832.00	65.9	3890.00
32.5	1.78	39.2	8.32	45.9	38.90	52.6	182.00	59.3	851.00	66.0	3980.00
32.6	1.82	39.3	8.51	46.0	39.80	52.7	186.00	59.4	871.00	66.1	4070.00
32.7	1.86	39.4	8.71	46.1	40.70	52.8	191.00	59.5	891.00	66.2	4170.00
32.8	1.91	39.5	8.91	46.2	41.70	52.9	195.00	59.6	912.00	66.3	4270.00
32.9	1.95	39.6	9.12	46.3	42.70	53.0	200.00	59.7	933.00	66.4	4370.00
33.0	2.00	39.7	9.33	46.4	43.70	53.1	204.00	59.8	955.00	66.5	4470.00
33.1	2.04	39.8	9.55	46.5	44.70	53.2	209.00	59.9	977.00	66.6	4570.00
33.2	2.09	39.9	9.77	46.6	45.70	53.3	214.00	60.0	1000.00	66.7	4680.00
33.3	2.14	40.0	10.00	46.7	46.80	53.4	219.00	60.1	1020.00	66.8	4790.00
33.4	2.19	40.1	10.20	46.8	47.90	53.5	224.00	60.2	1050.00	66.9	4900.00
33.5	2.24	40.2	10.50	46.9	49.00	53.6	229.00	60.3	1070.00	67.0	5010.00
33.6	2.29	40.3	10.70	47.0	51.10	53.7	234.00	60.4	1100.00	67.1	5130.00
33.7	2.34	40.4	11.00	47.1	51.30	53.8	240.00	60.5	1120.00	67.2	5250.00
33.8	2.40	40.5	11.20	47.2	52.50	53.9	245.00	60.6	1150.00	67.3	5370.00
33.9	2.45	40.6	11.50	47.3	53.70	54.0	251.00	60.7	1170.00	67.4	5500.00
34.0	2.51	40.7	11.70	47.4	55.00	54.1	257.00	60.8	1200.00	67.5	5620.00
34.1	2.57	40.8	12.00	47.5	56.20	54.2	263.00	60.9	1230.00	67.6	5750.00
34.2	2.63	40.9	12.30	47.6	57.50	54.3	269.00	61.0	1260.00	67.7	5890.00
34.3	2.69	41.0	12.60	47.7	58.90	54.4	275.00	61.1	1290.00	67.8	6030.00
34.4	2.75	41.1	12.90	47.8	60.30	54.5	282.00	61.2	1320.00	67.9	6170.00
34.5	2.82	41.2	13.20	47.9	61.70	54.6	288.00	61.3	1350.00	68.0	6310.00
34.6	2.88	41.3	13.50	48.0	63.10	54.7	295.00	61.4	1380.00	68.1	6460.00
34.7	2.95	41.4	13.80	48.1	64.60	54.8	302.00	61.5	1410.00	68.2	6610.00
34.8	3.02	41.5	14.10	48.2	66.10	54.9	309.00	61.6	1450.00	68.3	6760.00
34.9	3.09	41.6	14.50	48.3	67.60	55.0	316.00	61.7	1480.00	68.4	6920.00
35.0	3.16	41.7	14.80	48.4	69.20	55.1	324.00	61.8	1510.00	68.5	7080.00
35.1	3.24	41.8	15.10	48.5	70.80	55.2	331.00	61.9	1550.00	68.6	7240.00
35.2	3.31	41.9	15.50	48.6	72.40	55.3	339.00	62.0	1580.00	68.7	7410.00
35.3	3.39	42.0	15.80	48.7	74.10	55.4	347.00	62.1	1620.00	68.8	7590.00
35.4	3.47	42.1	16.20	48.8	75.90	55.5	355.00	62.2	1660.00	68.9	7760.00
35.5	3.55	42.2	16.60	48.9	77.60	55.6	363.00	62.3	1700.00	69.0	7940.00
35.6	3.63	42.3	17.00	49.0	79.40	55.7	372.00	62.4	1740.00	69.1	8130.00
35.7	3.72	42.4	17.40	49.1	81.30	55.8	380.00	62.5	1780.00	69.2	8320.00
35.8	3.80	42.5	17.80	49.2	83.20	55.9	389.00	62.6	1820.00	69.3	8510.00
35.9	3.89	42.6	18.20	49.3	85.10	56.0	398.00	62.7	1860.00	69.4	8710.00
36.0	3.98	42.7	18.60	49.4	87.10	56.1	407.00	62.8	1910.00	69.5	8910.00
36.1	4.07	42.8	19.10	49.5	89.10	56.2	417.00	62.9	1950.00	69.6	9120.00
36.2	4.17	42.9	19.50	49.6	91.20	56.3	427.00	63.0	2000.00	69.7	9330.00
36.3	4.27	43.0	20.00	49.7	93.30	56.4	437.00	63.1	2040.00	69.8	9550.00
36.4	4.37	43.1	20.40	49.8	95.50	56.5	447.00	63.2	2090.00	69.9	9770.00
36.5	4.47	43.2	20.90	49.9	97.70	56.6	457.00	63.3	2140.00	70.0	10000.00
36.6	4.57	43.3	21.40	50.0	100.00	56.7	468.00	63.4	2190.00		
36.7	4.68	43.4	21.90	50.1	102.00	56.8	479.00	63.5	2240.00		

G.3.2 Diagrams on cavity resonators and transmission line circuits

The design of directional couplers, VSWR-meters etc. necessitates wiring arrangements with physical dimensions D/d that avoid impedance discontinuities throughout the entire cable run into which the device is inserted (50, 60 or 75 ohm). The realization of this requirement is simplified by using a diagram (1) in place of calculations (see also A.2.3).

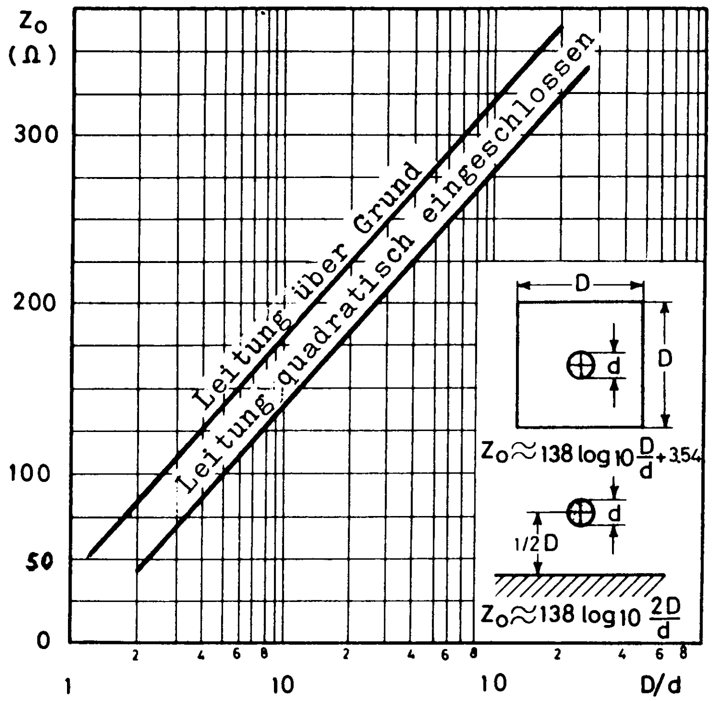


Figure 509
Diagram for calculation of line impedance.
- Line above a conducting surface.
- Line enclosed by a square-shaped cavity.

UHF cavity resonators are designed for higher impedances. The following diagram allows to read-off length of line and capacity (C) for a cavity resonator impedance (Z_O) of 130 ohm (D/d = 9).

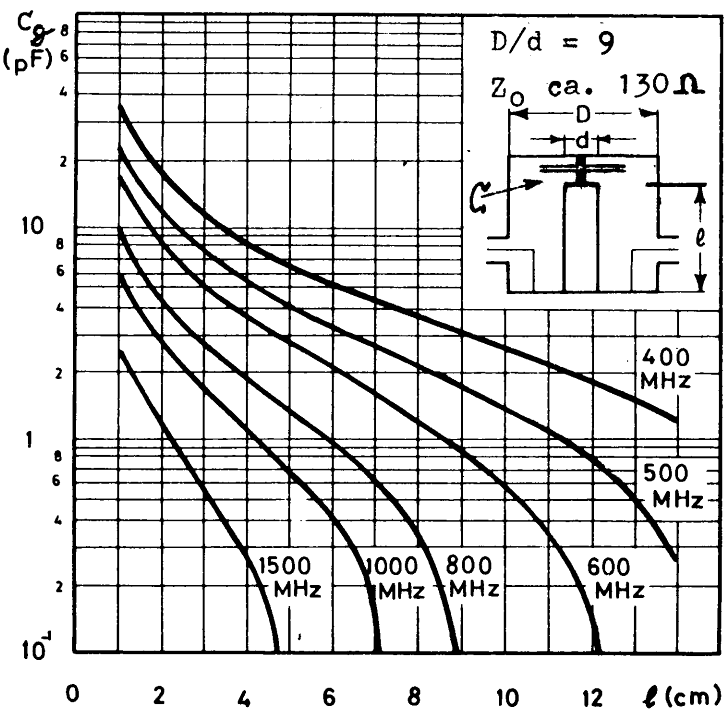


Figure 510
Diagram for quick estimation of length of line (l), circuit capacitance (C) and f_{res} of the resonator.
C must take all capacitances of influence on the resonator into account.

Directional couplers, mixers, power splitters and frequency sensitive circuits for microwaves are predominantly constructed in stripline technique. The impedance of such lines is defined by the ratio of linewidth (W) and distance (h) from the conducting surface as well as the dielectric between the two conductors.
Ceramic material and Teflon - apart from others - are used as substrate. Epoxy based p.c.boards should no longer be used in input circuits of 70cm converters. The losses are quite high.

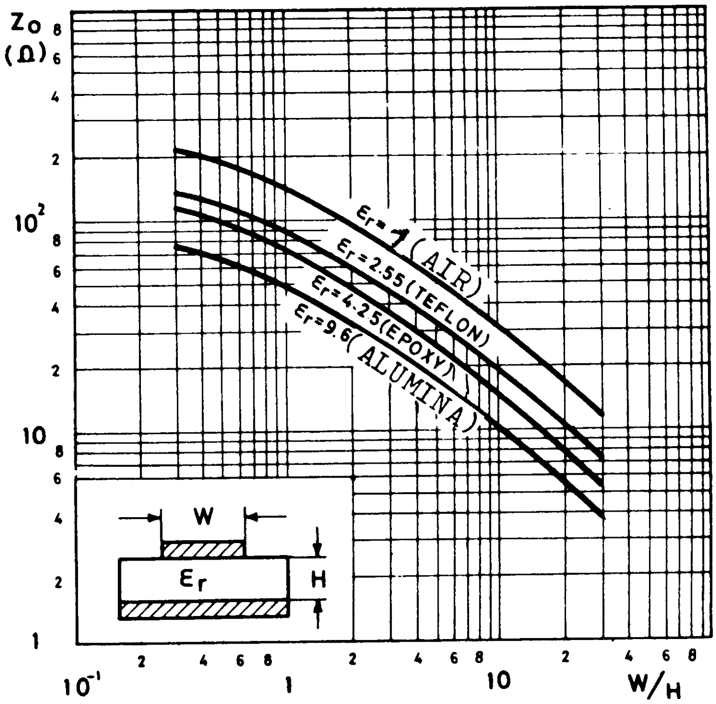


Figure 511
Diagram for the estimation of stripline impedance for various strip widths and different types of dielectric between the line and the conducting surface.

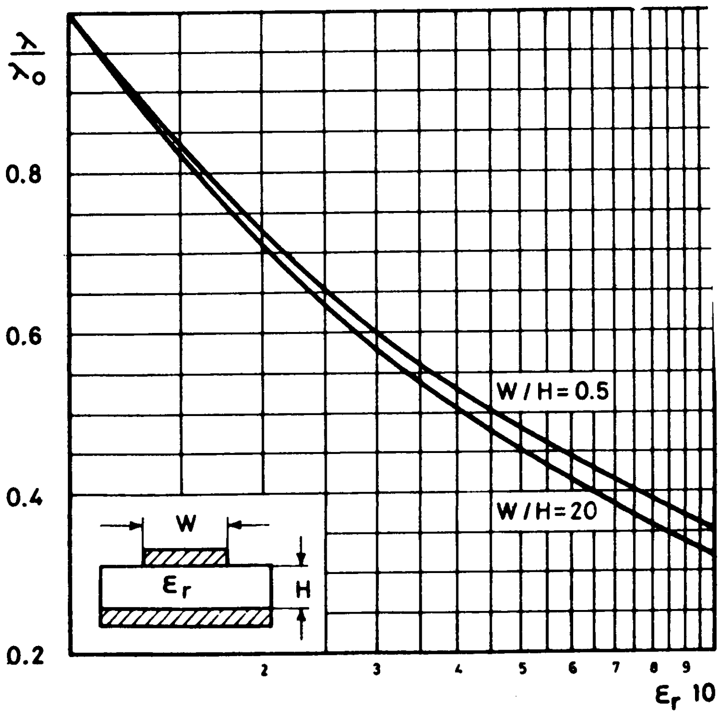


Figure 512
Diagram for the estimation of the velocity factor of a stripline. It is evident that the velocity factor is predominantly defined by the ε_r of the substrate with the ratio W/h being of little significance.
(See also A.2.4).

Reference:

(1) SGS/ATES SILICON