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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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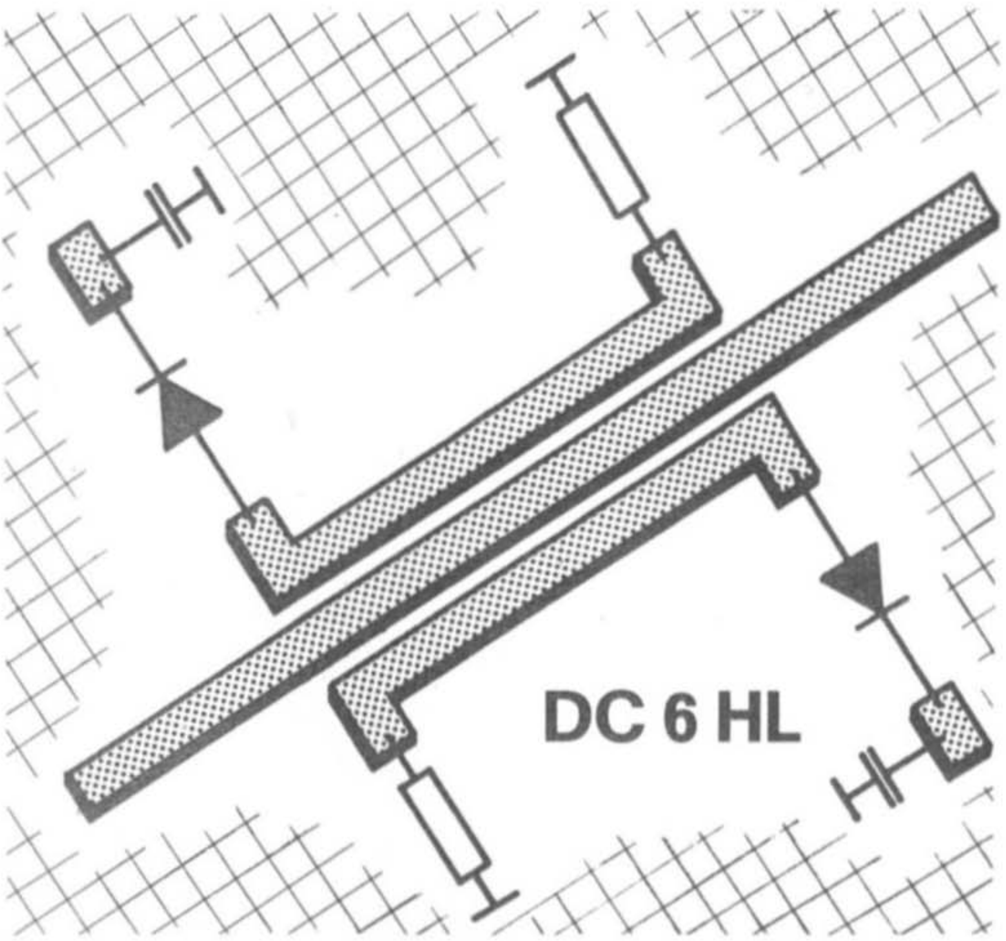

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NOTES TO OUR READERS

We are still receiving a number of money orders, transfers etc. that do not give the address of the sender or indicate what items are required. This often occurs due to the copying of the transfer in the banks. It is often necessary for us to wait until the customer complains of nondelivery in order to find out his address and the items required. Please ensure that your bank indicates your name and address as well as the required items clearly on their forms, preferably in English or German.

Address changes: Please address all changes of address to your local representative and not to the publishers. This is because the address file is held by the representative and not in Erlangen.

The French Representative will be away on holiday between September 5th and September 25th. She hopes that this will not inconvenience our French readers.

DIMENSIONING OF MICROSTRIPLINE CIRCUITS

by W. Schumacher, DJ 9 XN

The following article is to explain the fundamentals of ultra high frequency circuits. It is not only described how such circuits can be calculated and dimensioned, but information is also given regarding the theory and operation of UHF-circuits. It is intended to provide the reader with basic information from which he can carry out his own experiments. An introduction into the dimensioning of the impedance and length of striplines has already been discussed by K. Hupfer, DJ 1 EE in (4). References to this article will be made in the text.

It should be mentioned that the following theoretical considerations are equally valid for conventional circuits such as coaxial lines and for striplines. It is only necessary for the difference in line length and the other dimensions to be taken into cinsideration.

A sufficient number of diagrams are available for impedance and electrical line length (see references).

The following magnitudes and signs are used in the text:

λ	wave length	R	ohmic resistance
λ_o	wavelength in free space	X	reactance
λ_{St}	wavelength in a dielectric	Z	impedance (complex impedance $Z = R + jX$)
c	speed of light		
ω	frequency ($2\pi f$)	G	conductance (real)
β	propagation constant	B	susceptance (reactive)
ϵ_r	relative dielectric constant	Y	admittance (complex) ($Y = G + jB$)
ϵ_o	absolute dielectric constant		
Z_L	line impedance	r	reflection factor (complex)
SC	short circuit	z	space coordinate
NL	no-load (in illustrations)	l	length

1. MATERIALS

In order to obtain a high Q and low losses in a stripline circuit only low loss support materials can be used. Suitable dielectric materials are:

Teflon (PTFE) $\epsilon_r \approx 2$, $\tan \delta \approx 2 \times 10^{-4}$ at 600 MHz
Teflon (PTFE) impregnated epoxy glass board
$\epsilon_r \approx 3$, $\tan \delta \approx 20 \times 10^{-4}$
Hostalen $\epsilon_r \approx 2.7$, $\tan \delta \approx 7 \times 10^{-4}$.

In order to keep the stray field lines at a minimum, it is important for the thickness of the supporting dielectric to be as small as possible (4). However, the price of such materials is high and they are difficult to obtain, which means that such boards cannot be used by radio amateurs. For this reason, all the line circuits calculated in the article are based on a double-coated epoxy PC-board material with $\epsilon_r = 5$ (at audio frequencies), having a thickness of 1.4 mm.

2. DIAGRAMS FOR DETERMINING THE LINE DIMENSIONS

In order to complete the calculations given in (4) to include information for dimensioning the impedance and resonant length of micro-striplines, the corresponding diagrams derived from (1) are given in Figure 1 and 2. The application is described in (4), which should be used for the calculations rather than the following curves (Fig. 1 + Fig. 2).

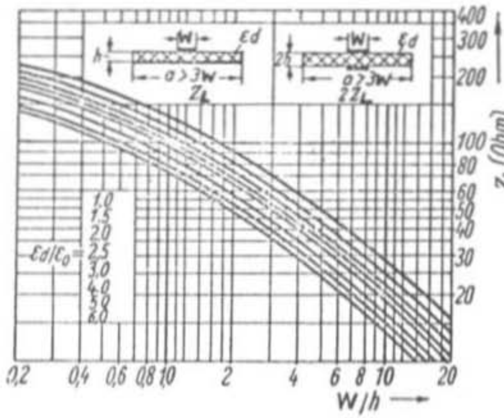


Fig. 1: Design curves (impedance) for balanced lines and microstriplines

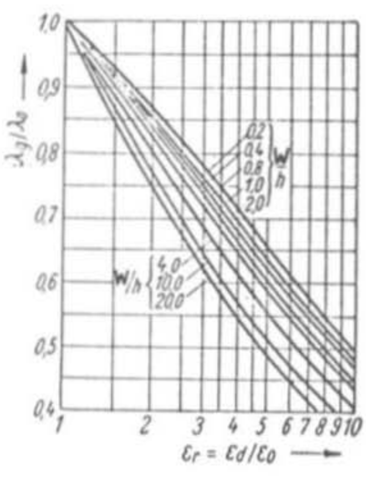


Fig. 2: Dependance of the standardized resonant length of a stripline on the relative dielectric constant.

3. CONSTRUCTION OF LINE CIRCUITS

Low loss lines can be used as transformation links. All resonant circuits constructed from lines can be explained according the special transformation processes. It is possible, for instance, for parallel resonance to be replaced by a short circuit condition between the connections in question. Any attenuation (loss) will be given as a resistor in the equivalent circuit diagram that is connected in series or parallel with the connection points.

A well known transformation link is the short circuited line (stub line). The impedance at the input of the line amounts to: $Z = jZ_L \times \tan 2\pi l/\lambda$ (1)

It will be seen from equation (1) that at $l = \lambda/4$: $Z = \infty$. At $0 < l < \lambda/4$, the stub line represents an inductance, at $\lambda/4 < l < \lambda/2$ it is a capacitance.

Figure 3 illustrates a $\lambda/4$ stripline. The connections E - E' are on the upper and lower surface of the board.

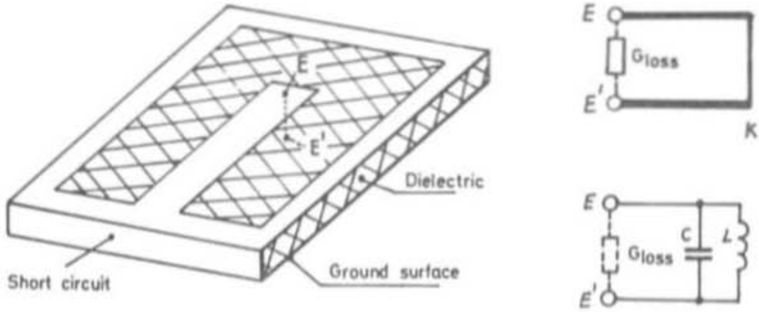
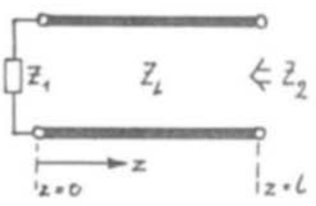


Fig. 3: $\lambda/4$ stripline circuit with equivalent diagrams

The general equation for the transformation of an impedance Z_1 into impedance Z_2 by a line of length l is (Fig. 4):



$$\frac{Z_2}{Z_L} = \frac{Z_1/Z_L + j \tan 2\pi \ z/\lambda}{1 + j \dfrac{Z_1}{Z_L} \tan 2\pi \ z/\lambda} \qquad (2)$$

Fig. 4:
Equivalent diagram for use with equation (2)

All special cases are included in equation (2), including equation (1). The different types of resonant lines (Fig. 5, 6 and 7) are derived by inserting the appropriate lengths and terminating resistances.

The impedance has an effect on the length or Q of a circuit. The relationship can be seen in equation (1). If the wavelength λ and the resonant impedance $Z = Z_E$ are kept constant, it will be seen from this relationship that the line length must be shortened on increasing the impedance. Or it can be seen that if the line length and wavelength are kept constant, the reactive component between the input connections will increase on increasing the impedance Z_L.

The reactive currents in the line will increase at high reactive impedances which means that the dielectric losses will be greater. This in turn will increase the loss, which means that it is necessary to make a compromise. However, for striplines with a relative dielectric constant of $\epsilon_r = 5$, $h = 1.4$ mm, this will not be necessary since no impedances of greater than 140 Ω can be realized. This means that the additional loss can be neglected in most cases.

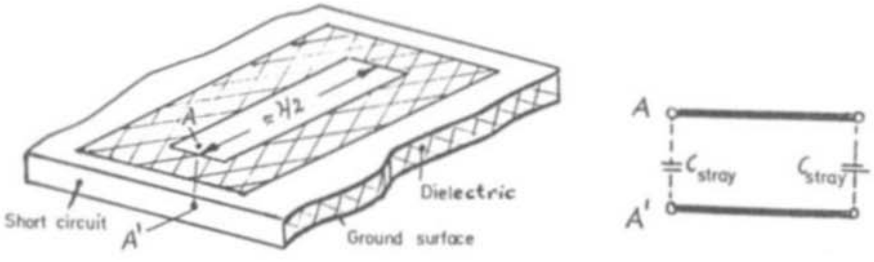


Fig. 5: λ/2 resonator (microstripline) with equivalent diagram.

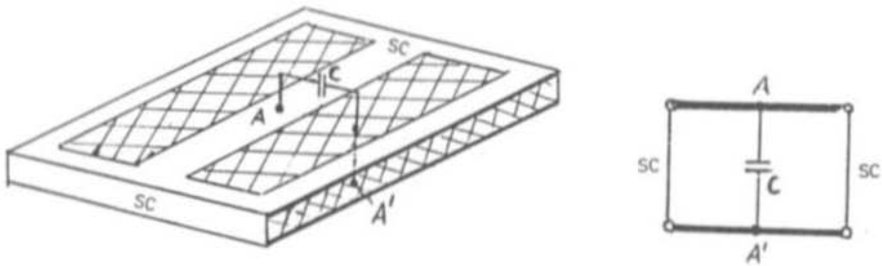


Fig. 6: Capacitively-loaded λ/2 circuit with equivalent diagram.

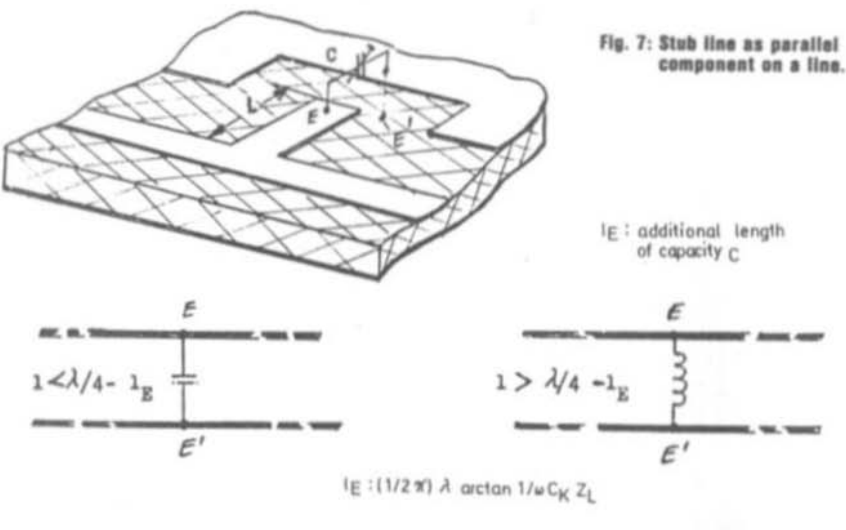


Fig. 7: Stub line as parallel component on a line.

4. DESIGN OF COUPLING LINKS

Resonant circuits are electrically or mechanically connected to other components in any circuit. The effect of the coupling on the circuit is often very difficult to calculate, which means that it is often neglected in practice. The only coupling link that can be calculated easily is that given in Figure 8. Since alignment elements are always provided, it is always possible to compensate for the detuning effect of the coupling links if a totally false dimensioning is not present.

The following types of coupling can be used:

Galvanic coupling	Capacitive coupling
Inductive coupling	Line coupling

The type of coupling is never absolutely of one type since a mixture of several types is always present. However, it can be stated that one or the other predominates.

Some of the popular coupling links are given in Figures 8 to 12.

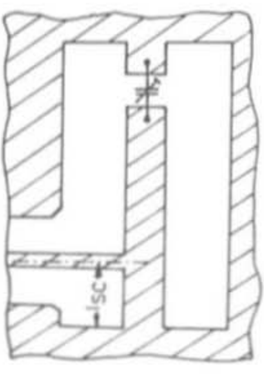


Fig. 8: Galvanic coupling.

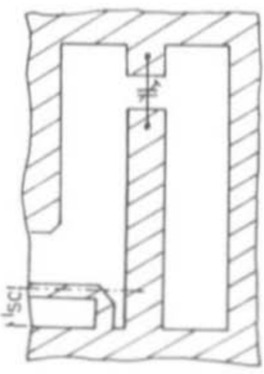


Fig. 9: Inductive coupling

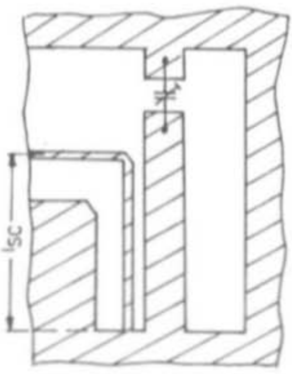


Fig. 10: Line coupling

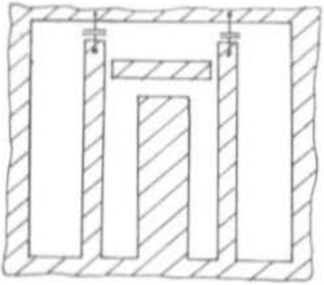


Fig. 11: Mainly capacitively-coupled filter.

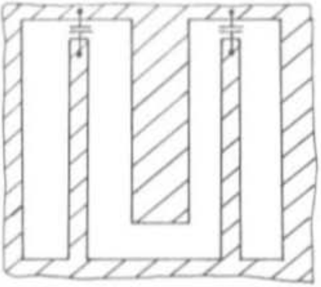


Fig. 12: Mainly inductively-coupled filter.

5. INTRODUCTION INTO THE USE OF THE LINE DIAGRAMS IN THE CALCULATION OF TRANSFORMATION CIRCUITS

5.1. GENERAL

The use of equation (2) for the dimensioning of transformation links is very complicated. However, there are methods of graphically determining the required values that are based on the fact that the reflection factor r (r = ratio between the amplitudes of forward and reflected signal) on the no-loss line is only dependent on the complex terminating impedance at position z = 0. It is only the phase of r (z) that periodically alters along the line whereas the magnitude $|r(z)|$ remains constant.

$$r(z) = |r| e^{j(\delta - 2\beta z)} \quad (3)$$

where: δ : phase angle at position z = 0
β : propagation constant $\beta = 2\pi/\lambda$

The relationship of r and Z at any point along the line is:

$$r(z) = \frac{Z(z) / Z_L - 1}{Z(z) / Z_L + 1} \quad \text{with } Z = R + jX \qquad (4)$$

The following is derived from Y = 1/Z :

$$r(z) = \frac{1 - Y \times Z_L}{1 + Y \times Z_L} \quad \text{with } Y = G + jB \qquad (5)$$

The appropriate inverted equations to (4) and (5) are:

$$Z/Z_L = (1 + r) / (1 - r) \quad (6) \qquad Y \times Z_L = (1 - r) / (1 + r) \quad (7)$$

5.2. TRANSFORMATION USING LINES

The process of transforming a terminating impedance Z (z = 0) by means of a length of line to the commencement of the line z = 1 as given in equation 2 is as follows:

The reflection factor $r(0) = |r| \exp(j\delta)$ is calculated from the known $Z = R + jX$ and Z_L.

According to equation (3), the line effects a reflection factor of $r(1) = r(0) \times e^{-2j\beta l}$ at position z = 1.

The corresponding impedance Z (1) is calculated reciprocally using equation (6) and r (1).

The described calculation with equations (4), (3) and (6) requires a considerable amount of time and is not suitable for practical use. It is better for this process to be made graphically.

The reflection factor is a dimensionless complex number which allows it to be entered in the plane of complex numbers. This is then called the reflection factor plane (r-plane; Fig. 13). Equations (4) and (5) represent reproduction rules with which any point of the Z and Y planes standardized to Z_L can be transformed to the r-plane. Since one is mainly only dealing with passive impedances (networks without current and voltage sources, impedances with positive ohmic component), only reflection factors that are less than 1 will result.

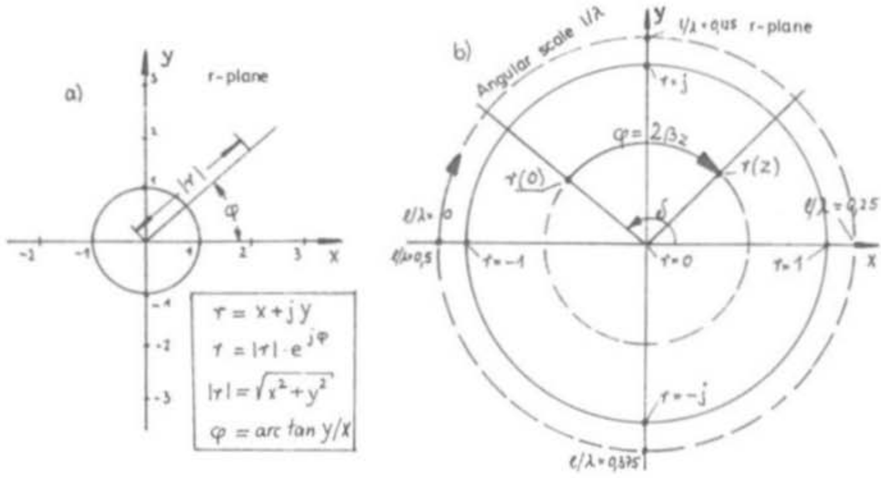


Fig. 13: a. complex numerical plane as reflection factor plane
b. transformation by a line of length z in the r-plane.

This means that the transformation can be limited to the inner part and edge of the circle of unit radius in the r-plane.

If a r-diagram was to be drawn with all R, X, B, G, Z , φ curves, the result wound be a diagram that would be very difficult to read off. According to the application, one of the Smith diagrams shown in Figures 14 to 16 is used. The previously mentioned transformation with the aid of a length of line appears as follows in the Z-type of the Smith diagram: The given impedance $Z = R + jX$ is standardized: $Z/Z_L = R/Z_L + j\,X/Z_L$.

The appropriate R and X curves are now found in the Z-type Smith diagram. The intersection of the two curves gives r (z = 0).

The line rotates r (z = 0) by angle $2\beta l$ in a clockwise direction (due to $\varphi = \delta - 2\beta z$) around the zero point of the r-plane. The zero point is designated matching point.

The numerical values are read off from the R and X curves of the new position and are resolved with Z_L.

In this manner, one will only require a Smith diagram, compass, ruler and slide rule. The procedure can also be used for other types of the Smith diagrams.

5.3. ANGULAR SCALE

All diagrams are provided with an angular scale (onto which l/λ is marked) on the edge $|r| = 1$ to facilitate reading off the angle. Most scales commence at the short circuit point. However, this is not important in practice. With the aid of this scale, it is not necessary to calculate the angle in degrees. It is only necessary to know l/λ and to place the ruler to the centre point and angular scale to find the required r-point on the straight line.

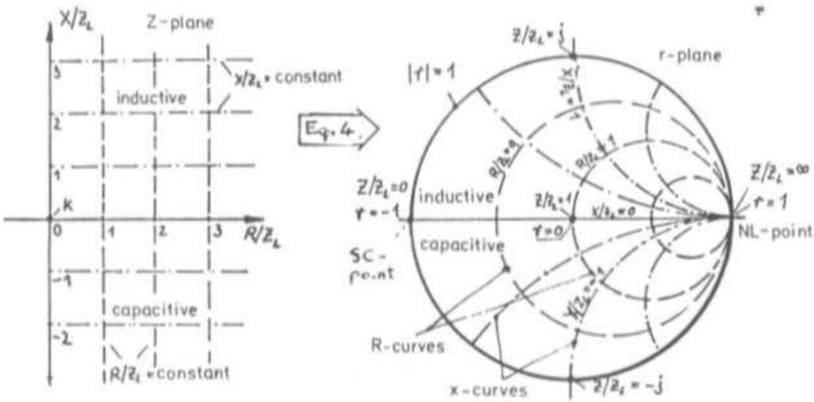


Fig. 14: Transformation of the Z-plane to the r-plane (Smith diagram, Z-type).

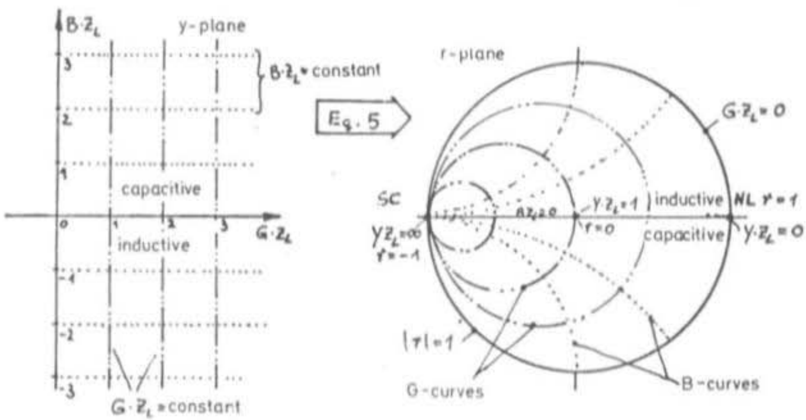


Fig. 15: Transformation of the Y-plane to the r-plane (Smith diagram, Y-type).

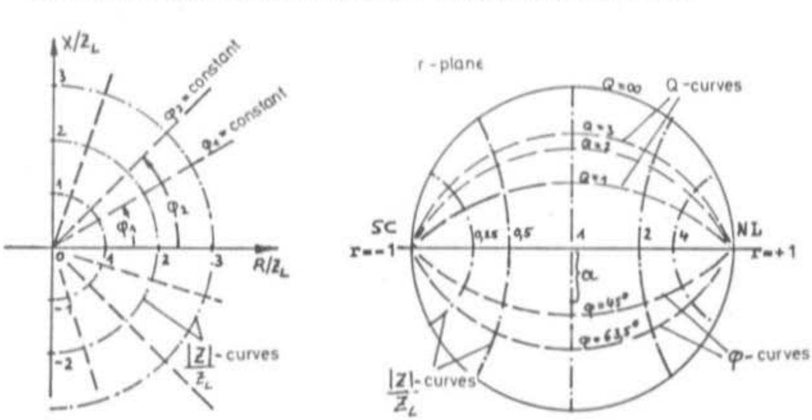


Fig. 16: Transformation of the curves /Z/ = constant, φ = constant to the r-plane (Charter diagram).

Specific angles φ	associated arc φ	and l/λ
0^{o}	0	0
90^{o}	$\pi/2$	0.125
180^{o}	π	0.25
270^{o}	$3\pi/2$	0.375
360^{o}	2π	0.5

It will be seen clearly from the list that no values of $l/\lambda > 0.5$ result. If a line is longer than $\lambda/2$, its transformation effect can always be given by a line length of:

$$z/\lambda = l/\lambda - n \times 0.5 < 0.5 \quad (n = 1, 2, 3 \ldots\ldots) \qquad (8)$$

5.4. TRANSFORMATION USING LUMPED ELEMENTS

It is also possible to perform other types of transformation processes graphically in the Smith diagram. The most important types are listed in Table 1 and the corresponding transformation paths are shown in Figure 17.

When dimensioning a transformation circuit, it should be noted that the power loss of a transformation circuit increases with the length of the transformation path in the complex impedance plane (Z-type of Smith diagram). With length one means the length of the path joining all points that must be passed. In principle, the losses therefore increase when the impedances that must be transformed differ greatly from another.

Table 1: Types of transformation in conjunction with a Smith diagram

Type of circuit		Type of Smith diagram	Transformation on ... curve	In direction	Electrical value of component
Series	R	Z	X	NL point	$R_s = R_{in} - R_{out}$
	L	Z	R	Clockwise	$X_L = X_{in} - X_{out}$
	C	Z	R	Anticlockwise to NL point	$X_c = X_{in} - X_{out}$
Parallel	G	Y	B	SC point	$G_p = G_{in} - G_{out}$
	L	Y	G	Anticlockwise	$B_L = B_{in} - B_{out}$
	C	Y	G	Clockwise	$B_c = B_{in} - B_{out}$
Ideal transformer		$\|Z\|, \varphi$	φ (Q)	$tr^2 > 1$: NL $tr^2 < 1$: SC	$Z_{in}/Z_{out} = tr^2$

Explanation of the indices:
- in: commencement value of the transformation step
- out: final value of the transformation step
- c : capacitive
- p : parallel
- s : series

The R and G values should be read off on the real axis whereas the X and B values are taken from the appropriate curves on the edge of the diagram.

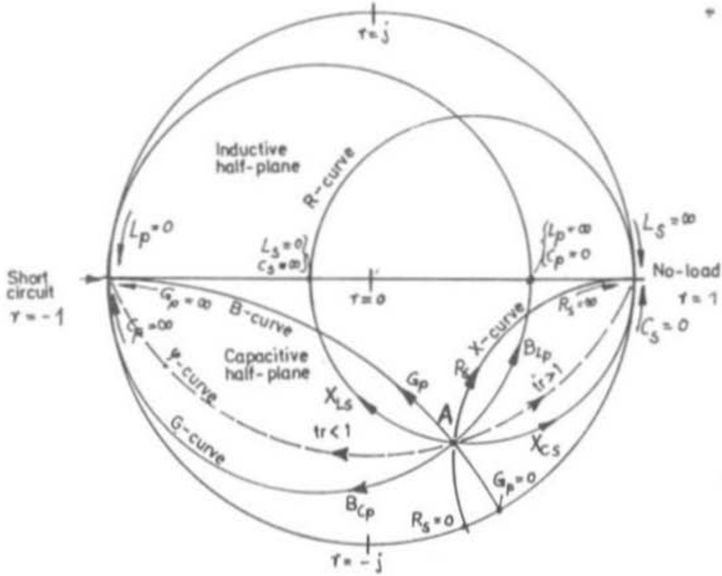
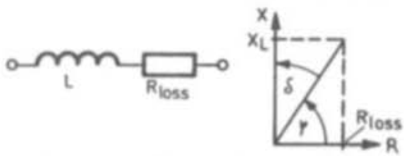


Fig. 17: Illustration of the transformation paths of the individual components in the Smith diagram.

5.5. NOTES REGARDING THE CHARTER DIAGRAM

The Q factor of a component is defined as the ratio of its reactive to real energy. The inverse value is the loss factor (d). If the energies are replaced by voltage and resistances, this will result, for instance, in the "no-load Q_o":

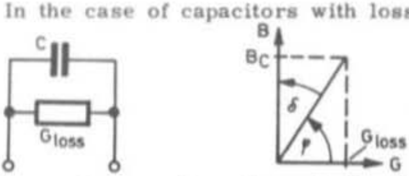
In the case of a lossy inductance:



$$Q_o = 1/d = \tan\varphi = \frac{X_L}{R_{loss}} = \frac{\omega L}{R_{loss}}$$

$$d = \tan\delta = 1/\tan\varphi$$

In the case of capacitors with lossy dielectric:



$$Q_o = \frac{B_C}{G_{loss}} = \frac{\omega_r C}{G_{loss}}$$

It can be seen from the relationship $\tan\varphi = X/R = B/G = Q$, that a Q value can be associated to any curve φ = constant in the Charter diagram (Fig. 16).

Construction of the Q-curves: The centre points are:

1. on the imaginary axis of the r-plane
2. at a spacing "a" from the zero point in an upward or downward direction.

Table 2 gives information as to the values of "a" (Fig. 16).

Table 2:

Q	φ	a/R
1	45°	1
2	63.5°	1/2
3	71.5°	1/3
4	76°	1/4
5	78.75°	1/5
etc.		

R = radius of the curve of unit radius in the r-plane in cm.
a = spacing.

The curves $\varphi(Q)$ = constant pass through the short circuit and no-load points in the Z and Y planes and must therefore pass these points in the r-plane.

An ideal transformer behaves as follows:

$$|Z_2| = tr^2 \; |Z_1|, \qquad |Z| = \sqrt{R^2 + X^2} \qquad (9)$$

This results in φ = constant. The transformation path for a transformer is therefore an arc on a φ-curve. The Q-values of a Charter diagram do not give any information as to the bandwidth of a transformation (matching) circuit.

5.6. ANALYSIS AND SYNTHESIS OF MATCHING NETWORKS USING THE SMITH DIAGRAM

A transformation problem can be solved by drawing an equivalent circuit of the network and splitting this into a number of sections (four poles) each containing only one component (inductance, capacitor, resistor or line). The interconnections between such sections should be assumed to be in one plane (designated in the following illustrations as capital letters). Starting at one end of the network, the standardized impedance is inserted in the Smith diagram. The stepwise shifting of the plane is continued and the transformation step is performed in the diagram. Before inserting the points into the diagram, it is necessary to settle the question of which standardized impedance is most favourable. If the geometric average of the real components of input and output impedance are chosen, one will have the graphical advantage that both impedances are on a r-curve in the Smith diagram.

$$Z_L = \sqrt{R_1 \times R_2} \qquad (10)$$

The reactive components of the final impedances are already part of the transformation network. Fundamentally, it is also possible to take one of the final impedances as reference Z_L. In this case, the transformation path will commence or end at the centre point of the Smith diagram. Both possibilities are given as examples in Figure 18. As long as only lumped elements and lines of the same impedance occur in one network, one will attempt to use the latter for standardizing the Smith diagram (usually 50 Ω, 60 Ω or 75 Ω). However, it often occurs that a circuit possesses junctions where two lines of different impedance meet (Fig. 19). If the reference impedance Z_L with which the junction is obtained is to be maintained, the impedance jump can be approximated by an ideal transformer with a transformation ratio $tr^2 = Z_L/Z_L$ and the Z_L line can be replaced by a line of Z_L. This procedure is only correct for ohmic characteristic impedances.

Most transformation problems are such that both input and output impedances are given (with ohmic and reactive components). In this case, the simplest, low-loss network is required that will be able to match these two impedances. The problem can be solved according to the following flow diagram (Fig. 20).

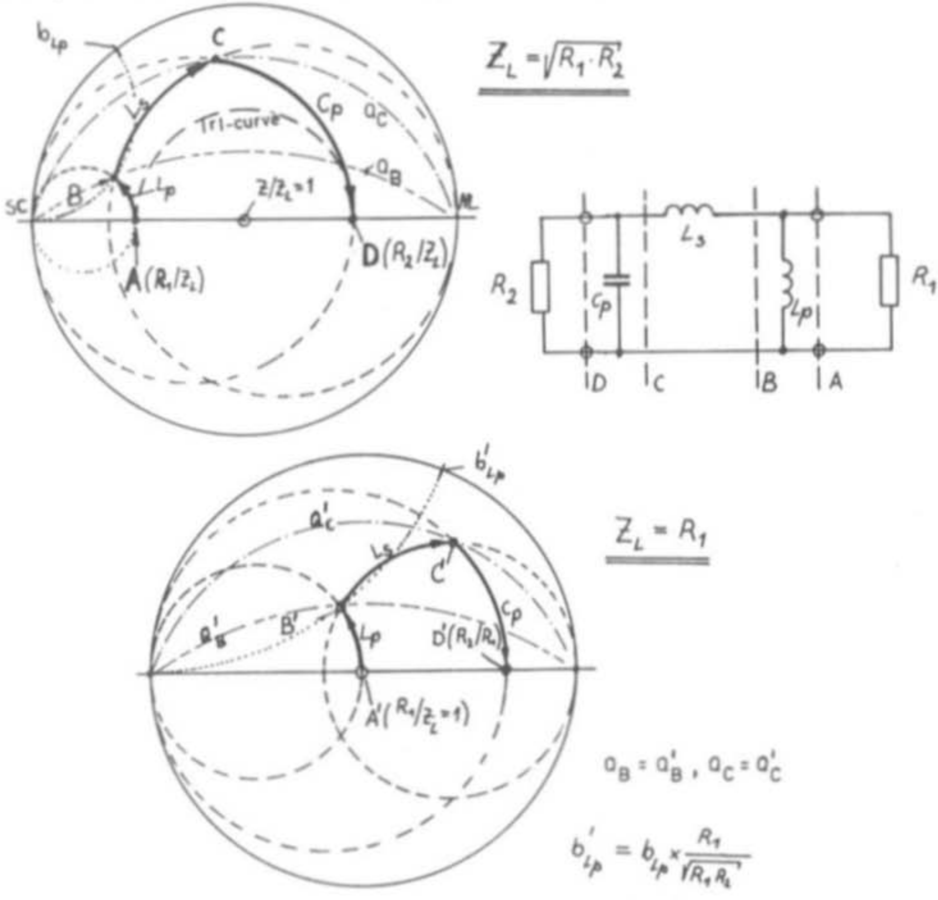


Fig. 18: Example of a matching transformation with the two reference impedances.

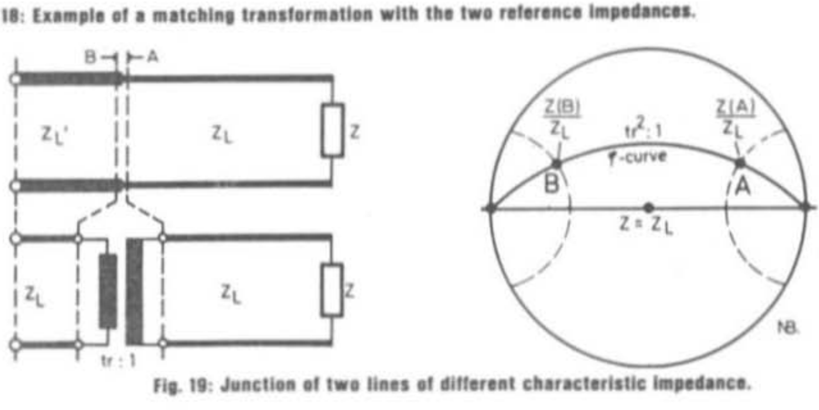


Fig. 19: Junction of two lines of different characteristic impedance.

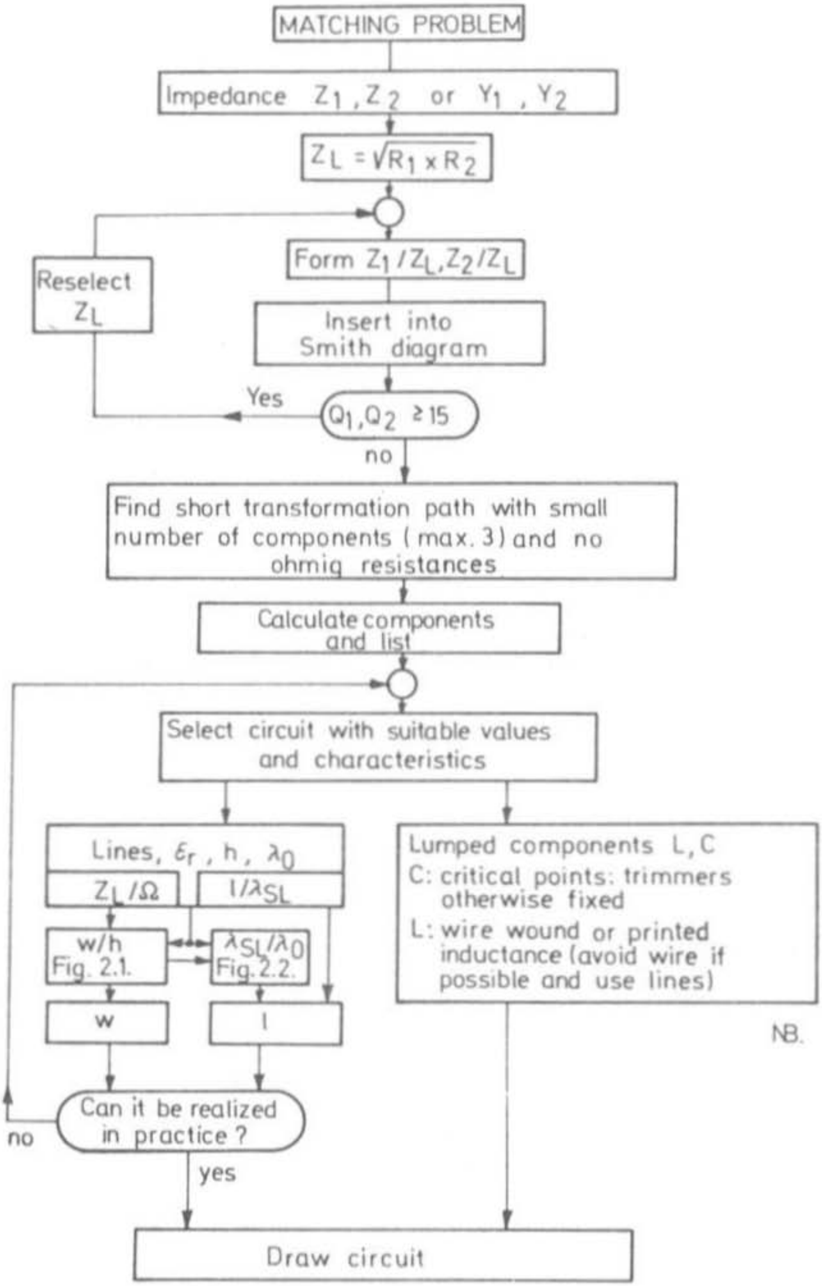


Fig. 20: Flow diagram for solving matching problems

The current and voltage sources of a network (e.g. in the equivalent diagram of a transistor) need not be taken into consideration. They do not contribute to the impedance or conductance of a circuit. If such sources are present, they are replaced before calculation:

Voltage sources by a short circuit.
Current sources by no-load.

The given method of calculating the junctions of two different impedances in a Smith diagram with the aid of a reference impedance is only suitable for calculating the line lengths of symmetrical filters. The disadvantage is that a conversion must be made before the true reflection factors in the various planes are obtained.

In the case of the method generally used for calculation of a structure with lines of different impedance, the last impedance or admittance obtained is referred to the new impedance in each transformation process. This results in the transformation ratio tr (Fig. 19).

Z_A/Z_L is valid in plane A. In plane B with Z_L' the following will result in a new diagram:

$$Z_B/Z_L' = Z_A/Z_L' = (Z_A/Z_L) = (Z_L/Z_L') = Z_A/Z_L \times tr^{2}$$

Actually, it is necessary to make a new Smith diagram after each transformation to another plane by which the impedance is altered. However, for economic reasons this is not done and no reference impedance is given for the whole diagram. It is necessary to establish which impedance is present at the location in question by consulting the appropriate circuit diagram so that the absolute Z or Y value can be read off.

The second part of this article will cover:
Examples for calculating line circuits with the Smith diagram;
Design of a capacitively shortened $\lambda/4$ resonator;
Calculation of a stripline filter for the 23 cm band;
Calculation of a single-stage preamplifier for the 70 cm band;
Improved bandpass resonator for 1296 MHz.
It will be published in one of the next editions of VHF COMMUNICATIONS.

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(1) H. Geschwinde and W. Krank: Streifenleitungen
Winter'sche Verlagshandlung, Fuessen/Germany 1960

(2) Telefunken Roehren- und Halbleitermitteilungen No. 65 09 124
Elektronisch abstimmbarer UHF-Verstärker of der Basis von doppelseitig mit leitender Folie beschichtetem Dielektrikum.

(3) Telefunken Roehren- und Halbleitermitteilungen No. 58 02 32
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(4) K. Hupfer: Striplines for VHF and UHF
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A STRIPLINE POWER AMPLIFIER FOR 70 cm USING A 2 C 39 TUBE

by A. Tautrim, DJ 2 PU

The 2 C 39 family of tubes can usually be obtained inexpensively from surplus sources since they are usually only operated for a limited period in microwave links etc. before being replaced at regular intervals by new tubes. Since these tubes are suitable for operation on the 70 cm, 24 cm and 12 cm bands, they allow a certain standardization of power supply, blowers and other fittings in the shack. The power ratings of these tubes are between that of tube types EC 8020 and 4 X 150. The marked advantages of the described power amplifier are the compact dimensions and the reasonable amount of metalwork involved in construction. Only metal plates are used and no lathing or milling is necessary. Figure 1 shows a photograph of the author's prototype.

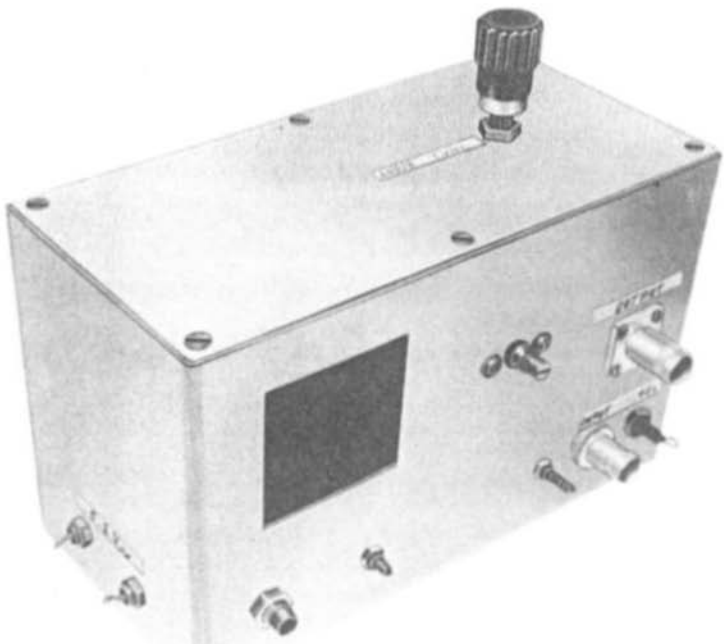


Fig. 1: Photograph of the author's prototype

1. CHARACTERISTICS

The output power of the 432 MHz power amplifier is 20 W at a plate voltage of 400 V, and 45 W at 800 V. The absolute maximum values of the plate voltage are 1000 V for unmodulated carrier operation (CW and FM) and 600 V for 100% amplitude modulation. In the SSB mode, the peak voltage appearing during plate voltage modulation of 1200 V can be classed as absolute maximum value. The data sheet of this tube states that the life is greatly dependent on the power loading of the tube, and especially on the plate voltage at higher frequencies.

It is therefore advisable to obtain the required output power at the lowest possible plate voltage. Sufficient cooling and the correct heater voltage are two important criteria with respect to the life of the tube. Further details are to be given later regarding this.

A drive power of between 1 W and 5 W is required in order to obtain the given output power levels. The author uses another, identically constructed stage as driver, which is operated at a plate voltage of only 300 V and without blower. The driver stage requires a drive power of between 50 mW and 200 mW which can be obtained easily from transistorized transverters such as DJ 6 ZZ 002 (1) or DC 6 HY 002 (2). A driver stage equipped with an EC 8020, e.g. the DC 6 HY linear amplifier (2), is equally suitable.

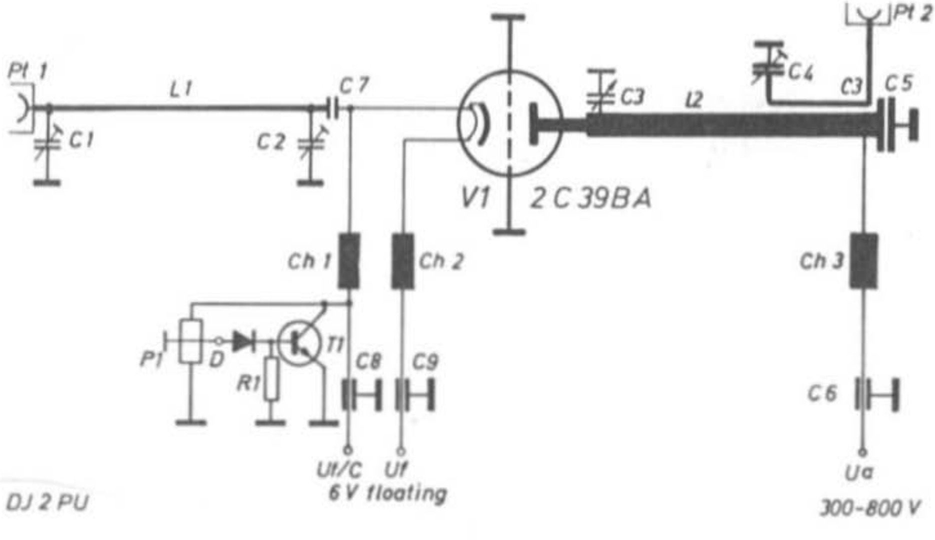


Fig. 2: Circuit diagram of the 2 C 39 linear amplifier.

2. CIRCUIT DETAILS

Figure 2 gives the circuit diagram of the grounded grid amplifier. Since the cathode is directly connected to the heater within the tube, it is not possible for the heater to be grounded since the cathode must be positively biased with respect to the grounded grid. The bias voltage is generated in a constant current two-pole circuit (2) equipped with transistor T 1. The bias can be adjusted with potentiometer P 1.

The input-impedance matching and selectivity is made in a π low-pass filter comprising C 1, L 1 and C 2. Chokes Ch 1 and 2 ensure that RF-energy is not passed onto the heater supply line.

The plate works into a $\lambda/4$ stripline circuit and the plate voltage is fed in via choke Ch 3 at the cold end of this line. A series-resonant coupling link (C 4, L 3) is used to transfer the output power to the output connector Pt 2.

When selecting components, it should be noted that capacitors C 3, C 5 and C 6 must handle the full DC plate voltage; C 3 also the peak RF-voltage. The whole RF-current of the resonant stripline circuit will flow via capacitor C 5.

3. CONSTRUCTION

Figure 3 shows a cross sectional view of the amplifier as well as the dimensions of the front panel (below). The ringed numbers designate the following:

1	Drive shaft for C 3	7	Base plate
2	Upper cover with tuning	8	Mounting screw and HT connection
3	Plate stripline	9	Disc of C 3
4	Case	10	2 C 39 tube
5	Mica plate	11	Insulation surface on disc (9)
6	Grid plate	12	Grid contact ring

The following illustrations given in Figures 4 to 11 show the individual mechanical parts and construction, as well as photographs of the author's prototype from above and below. The metalwork is made from brass plate of 1 mm, 2 mm and 3 mm thickness. The surfaces carrying RF-current are sufficiently large so that it is not really necessary to silver-plate the metal parts. However, the author silver-plated his prototype in a matt-silver solution manufactured by Degussa of Frankfurt, West Germany.

A brass tube of 19 mm outer diameter and 1 mm thick was used for the grid ring (part 12). A small tube cutter is used to prepare the inner rim. The tube cutter is placed approximately 1.5 mm from the lower edge of the tube and cuts into the tube. A rim is formed after a few revolutions of the cutter (see Fig. 3, part 12). Measure to see if the required inner diameter of 16.5 mm is obtained. If no means of accurate measurement is available, the 2 C 39 tube can be inserted until its grid ring touches the inside of the tube. The grid ring of the tube must fit into the brass tube without play. If the correct value has not been obtained cut <u>carefully</u> further until it is reached. Cut off the tube at a length of 10 mm with the tube cutter and deburr. The plate connecting ring (part 3.4.) is made in a similar manner from a brass tube of 29 mm outer diameter and 1 mm thick. The plate ring or bracket should be soldered at right angles to part 3.2. of Figure 7.

Of course, readers may have some more suitable contact material available than the home-made method described above.

The plate stripline (3.2.) forms a unit (3) together with spacer (3.3.) and the ground surface (3.1.); the stripline is fed with the plate voltage via mounting screw (8). The ground surface is insulated from the grid plate by the mica plate (5), and forms the bypass capacitor C 5. The capacitance amounts to approximately 500 pF with a mica thickness of 0.5 mm; this represents an AC resistance of less than 1 Ω.

The finest possible thread should be used for the adjustment of the plate tuning capacitor C 3. The matching nuts are slotted in order to ensure a firm fitting of the tapped bushing. After soldering the nuts to the cover, they are depressed and the screw is inserted into the tapped bushing. Check to see whether the screw is held tightly and that the adjustment is slightly stiff. After this has

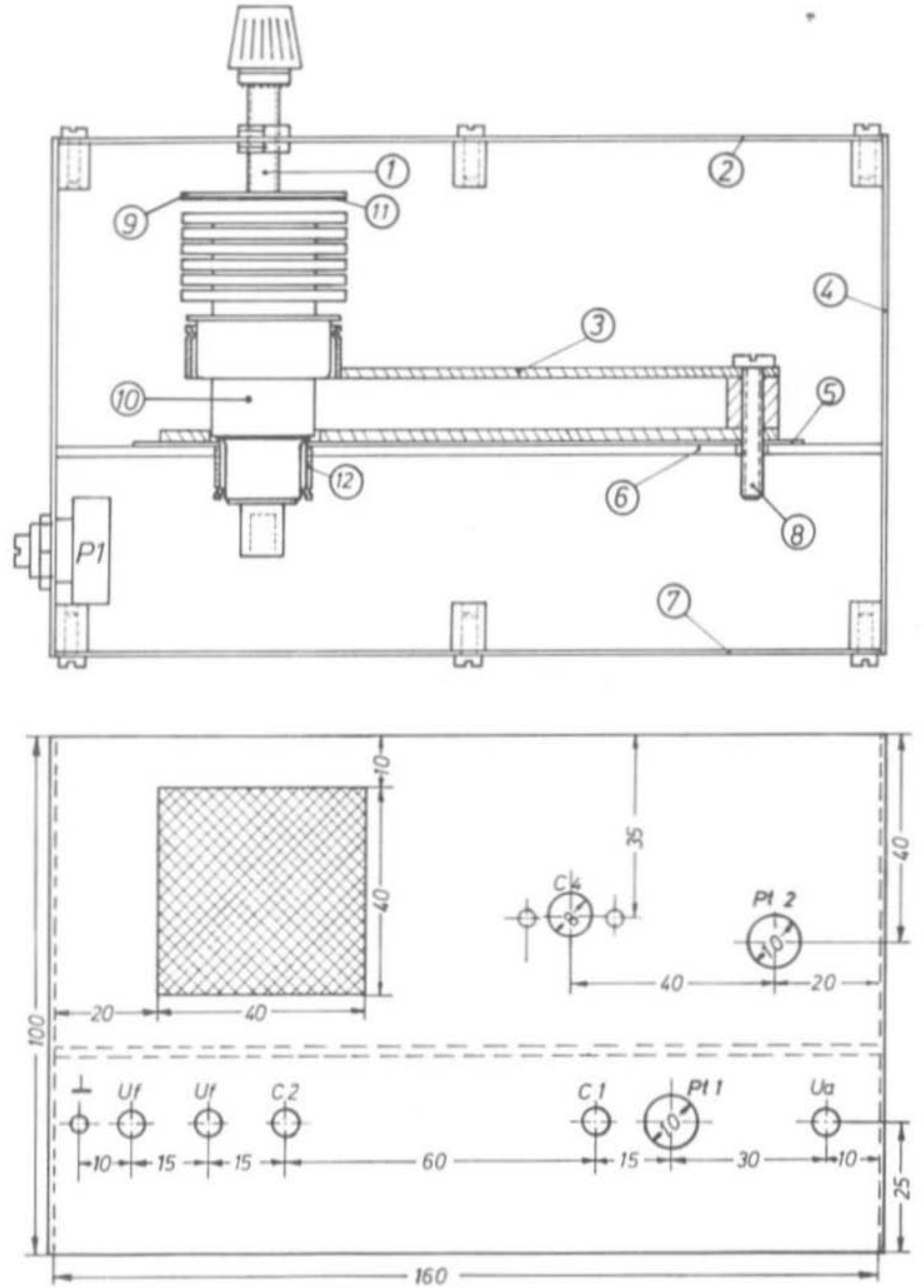


Fig. 3: Sectional diagram of the amplifier and dimensions of the front panel.

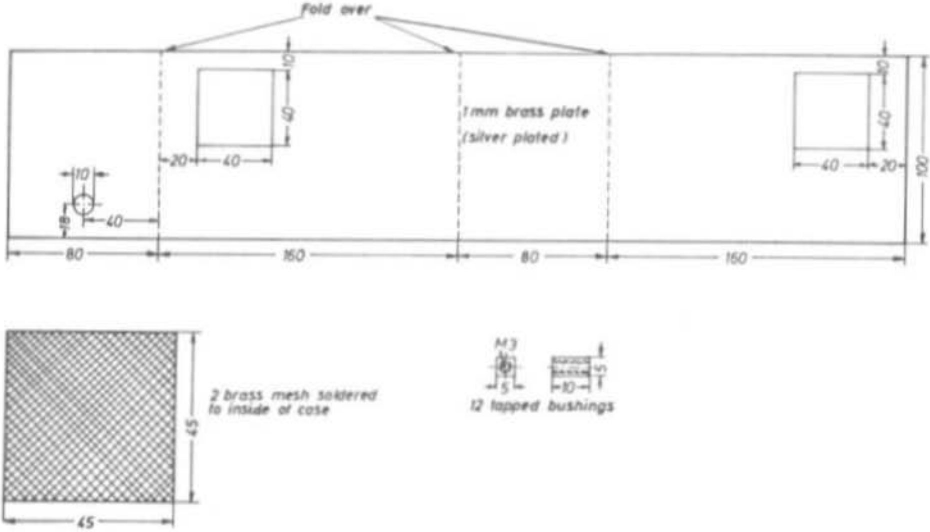


Fig. 4: Construction of the casing and air vents

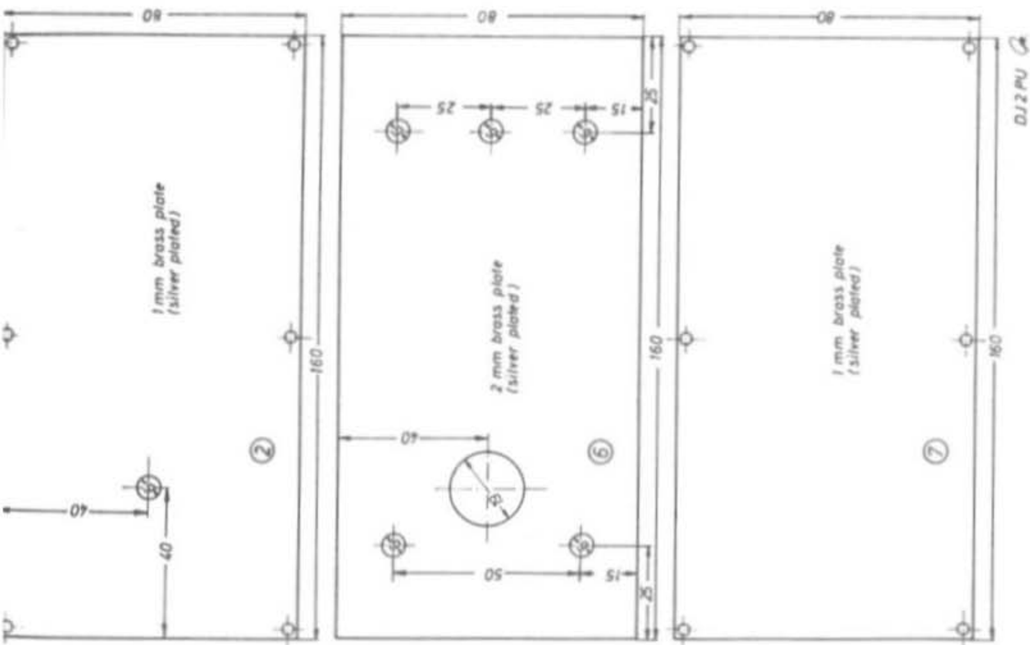


Fig. 5: Cover with tuning hole (above). Grid plate (centre). Base plate (below)

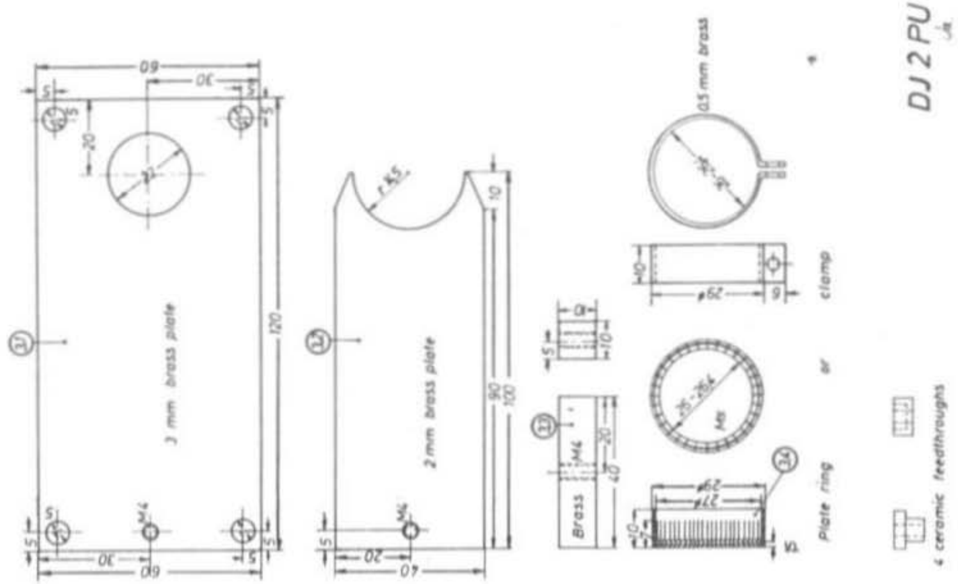


Fig. 6: Parts of the plate stripline circuit

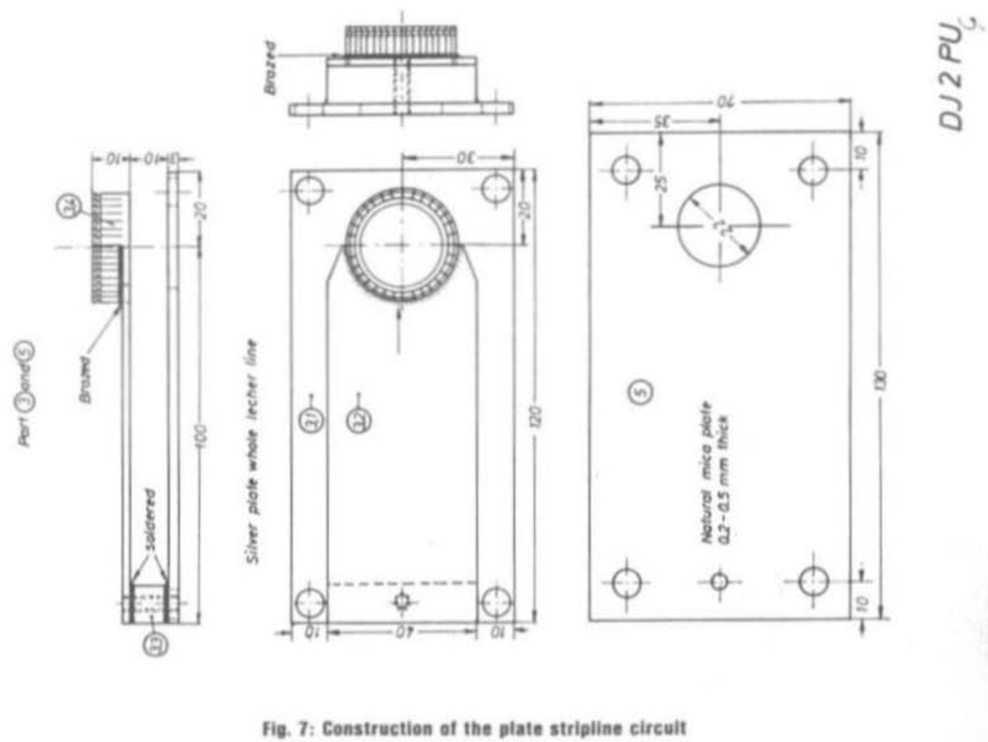


Fig. 7: Construction of the plate stripline circuit

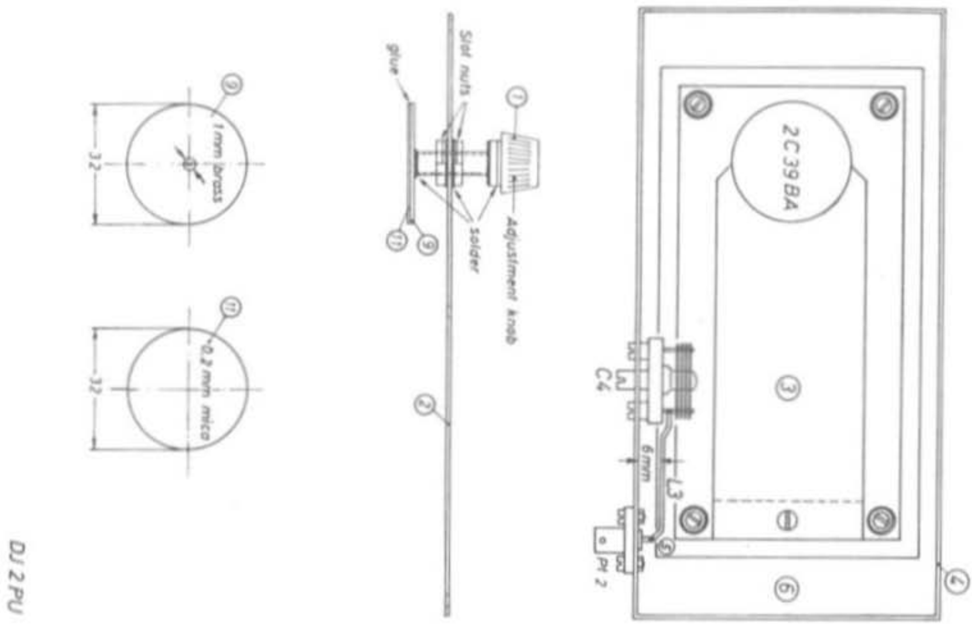


Fig. 8: Drawing of the amplifier from above and the parts for tuning capacitor C 3

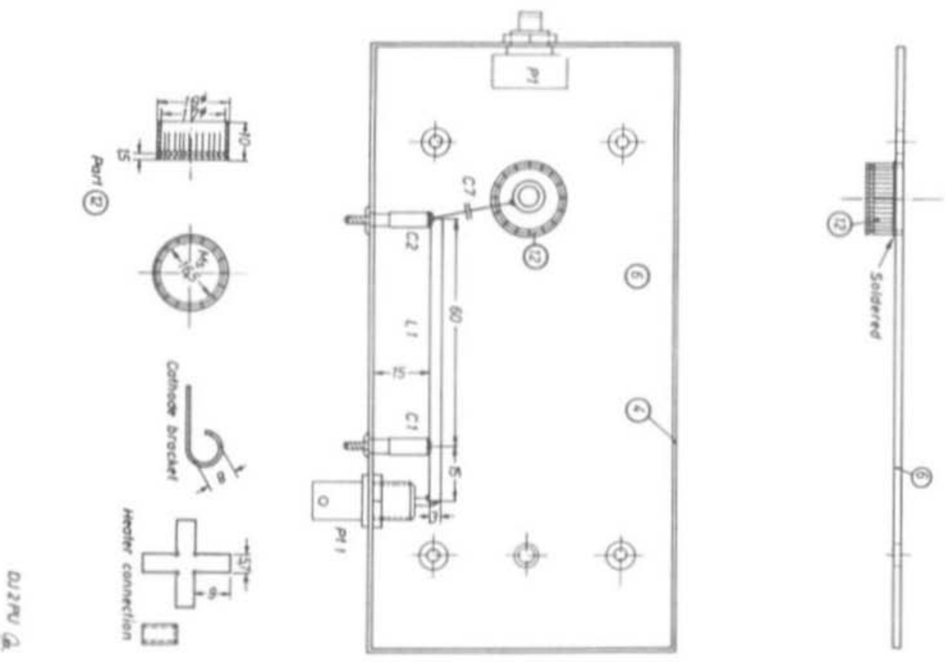
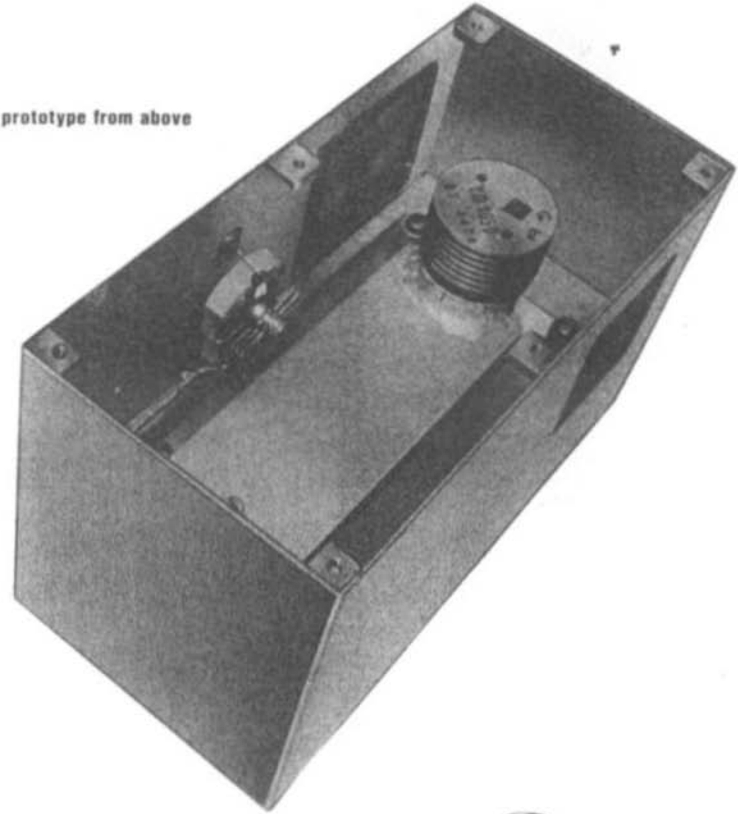


Fig. 9: Grid plate and input filter

Fig. 10: The author's prototype from above



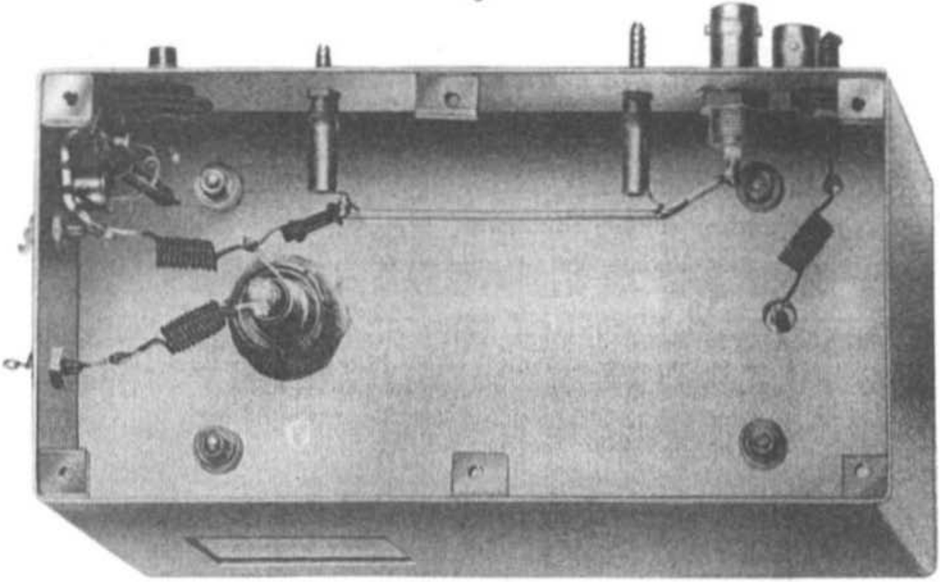


Fig. 11: The author's prototype from below

been obtained, the capacitor disc (9) can be soldered plane onto the screw. Any surplus solder should be removed, after which the mica disc (11) can be glued onto the disc (9).

As has been mentioned previously, an identical amplifier stage can be constructed for use as driver. Since a plate voltage of only 300 V is required for the driver, no cooling will be required. However, the extra work involved for making the cooling cutouts when using the arrangement shown in Figure 12 makes cooling of both driver and power amplifiers from one single blower a simple matter. It is only necessary for the connectors, trimmers and feedthrough capacitors of the driver to be placed on the other side. The construction of the cone for fitting the blower is shown in Figure 13. It is soldered to the blower. Of course, a different cone will be necessary to suit other types of blowers.

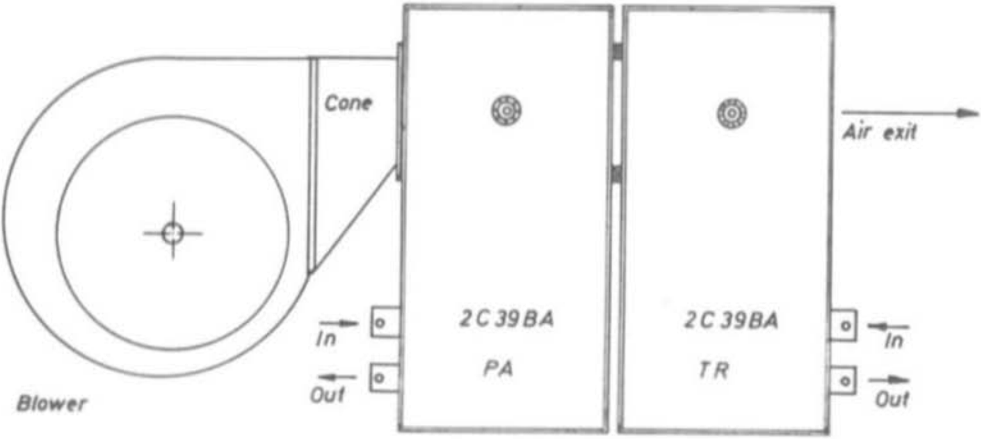


Fig. 12: Driver and power amplifier located next to each other with a common blower.

3.1. COMPONENT DETAILS

Pt 1, Pt 2: BNC or similar connectors

C 1, C 2: 1-12 pF ceramic tubular trimmers (Philips 2222 802 20004 or 20015)
C 3: see Fig. 8
C 4: 3.5-25 pF airspaced trimmer (Philips 2222 804 01004 or 01009)
C 5: see Fig. 7
C 6: 1 nF feedthrough capacitor, screw-fitting, for at least 500 V
C 7: 270 to 1000 pF ceramic disc or tubular capacitor
C 8, C 9: 1 nF or more, feedthrough capacitor, screw fitting (Siemens B 37020 - B 5102 - S 000)

R 1: 10 kΩ
P 1: 1 kΩ trimmer potentiometer
T 1: 2 N 1613, 2 N 2219
D 1: 1 N 914, 1 N 4148 or any silicon diode

L 1: see Fig. 11; L 2: see Fig. 7; L 3: see Fig. 8

Ch 1 - Ch 3: 11 turns of 0.5 mm dia. (24 AWG) enamelled copper wire wound on a 5 mm former, close wound, self-supporting.

4. POWER SUPPLY

Figure 14 gives the circuit diagram of a power supply suitable for the described power amplifier. A power transformer with centre-tapped secondary is used so that two voltages with a ratio of 1 : 2 are obtained. The plate voltage of the power amplifier is 650 V under load. This transformer will not be suitable if the maximum tube ratings are to be utilized (1200 V) or if the EC 8020 is to be used as driver (200 V).

The 2 C 39 requires that the heater voltage is adjusted exactly to the values given in Section 5 due to return emission (electrons that hit the cathode and not the plate). This is not necessary with the EC 8020.

5. PREPARATIONS

5.1. HEATER VOLTAGE AND HEATER CURRENT

Attention must be paid that all heater windings for the 2 C 39 are floating (not grounded. The cathode is directly connected to the heater within the tube. The heater current should be adjusted to 1 A with the aid of resistors P 2 and P 3 and an ammeter connected in series with the heater line. The plate voltage should not be applied during this adjustment.

5.2. PLATE VOLTAGE AND PLATE CURRENT

Heat the 2 C 39 tube for approximately 1 minute so that the cathode is able to reach its operating temperature. Switch on the blower.

Switch on the plate voltage (300 V for the driver or 600 V for the power amplifier). The quiescent plate current should be adjusted to approximately 20 mA with resistor P 1.

5.3. TUNING OF THE POWER AMPLIFIER

Connect the output of the power amplifier to an artifical antenna and output meter, or if not available via a calibrated reflectometer (power meter) to an antenna. Heat the 2 C 39 tube for one minute before switching on the plate voltage. Switch on the blower of the power amplifier. Drive the unit with a 432 MHz signal (approx. 50 - 20 mW for the driver; 1 to 5 W for the PA).

Align the power amplifier to resonance by adjusting the plate tuning capacitor C 3. This should result in a reading on the reflectometer. Align capacitors C 1, C 2, C 3, C 4 and the position of the output link for maximum output reading. Repeat this alignment until no further improvement can be obtained. The final plate current will be approximately 30 mA for the driver and approximately 100 mA for the power amplifier.

The output power of the driver stage is in the order of 1 to 4 W and 20 to 50 W for the power amplifier, which is dependent on the drive power and the plate voltage.

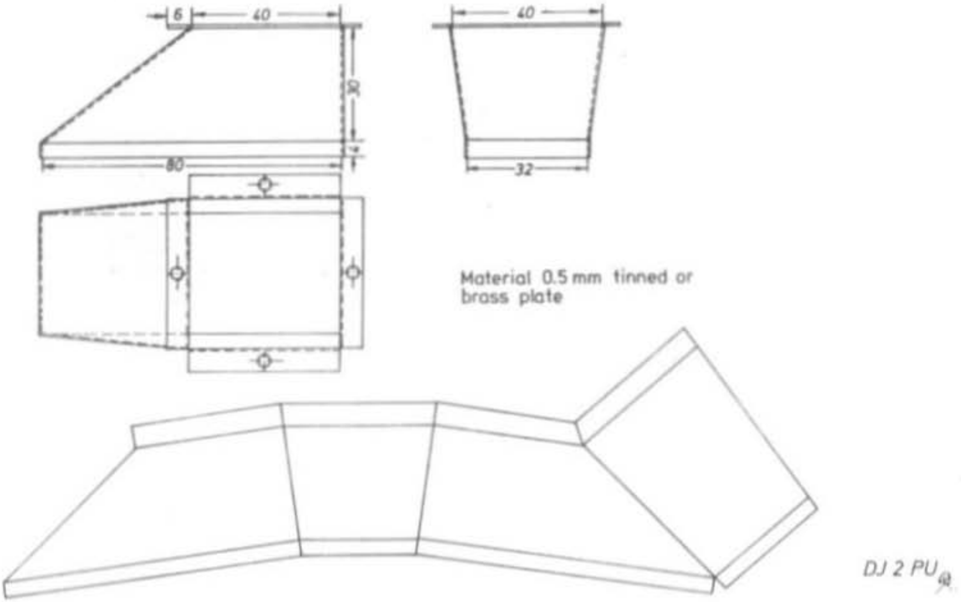


Fig. 13: Blower cone.

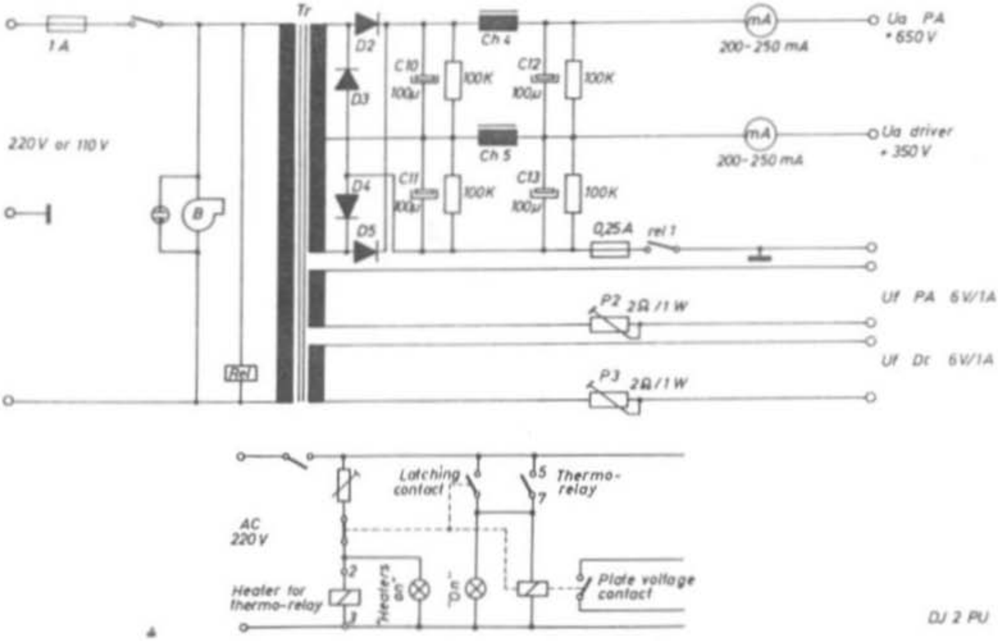


Fig. 14: Power supply suitable for PA and driver stage

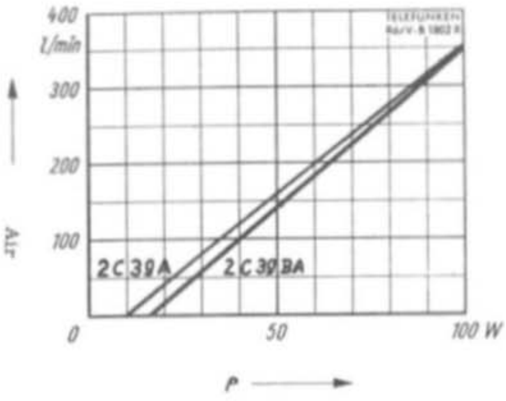
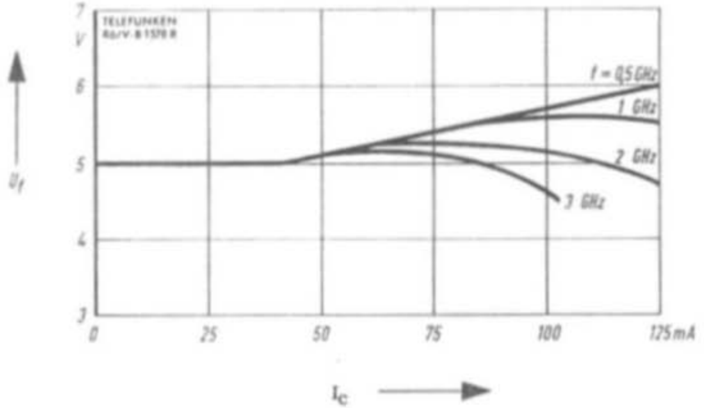
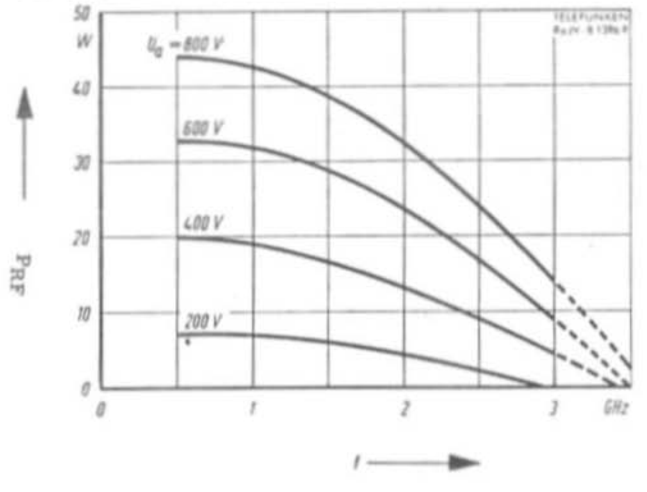


Fig. 15: Volume of air for cooling



$U_f = f\ (I_c)$
f = Parameter

Fig. 16: Reduction of the heater voltage as a function of frequency and cathode current.



2 C 39 BA

P_{RF} = f (f)
U_a = Parameter
I_a = 100 mA

Fig. 17: Output power as a function of frequency at a plate current of 100 mA

TABLE 1

AIR-COOLED POWER TRIODES FOR MICROWAVES UP TO 3 GHz

Type	Uf[1] V	If A	Prelim. heating	Limit values Ua[2] unmod.	Ua modul.	P (W)	Limit values Ugp+ V	Ugp- V	Pg W	Ig mA	Ic mA	t max. 3)	Sl. mA/V	Ug V	Cg/p pF	Cg/c pF	Cp/c pF
2 C 39 A	6.3 V	1 A	1 min.	1000 V	600 V	100	+25 V	-400 V	2 W	50	125	175°C	25	-150	2 pF	6 pF	0.035 pF
2 C 39 BA	6 V	1 A	1 min.	1000 V	600 V	100	+30 V	-400 V	2 W	50	125	200°C	25	-150	2 pF	6 pF	0.035 pF
3 CX 100 A5	6 V	1 A	1 min.	1000 V	600 V	100	+30 V	-400 V	2 W	50	125	250°C	25	-150	2 pF	6 pF	0.035 pF
YD 1050	6 V	1 A	1 min.	800 V	600 V	100	+25 V	-400 V	2 W	50	125	200°C	27	-150	2 pF	6 pF	0.045 pF
YD 1051	6 V	1 A	1 min.	1000 V		100	+30 V	-400 V	2 W	50	190	250°C	30	-150			
YD 1053	6 V	1 A	1 min.	800 V		100	+25 V	-400 V	2 W	50	125	250°C	27	-150			
7211	6.3 V	1 A	1 min.	1000 V		100	+30 V	-400 V	2 W	45	190	250°C	30	-150			
7289	6 V	1 A	1 min.	1000 V	600 V	100	+30 V	-400 V	2 W	50	125	300°C	25	-150	2 pF	6 pF	0.035 pF

1) It may be necessary to reduce the heater voltage Uf at frequencies over 400 MHz (see manufacturers instructions). Normally, a heater voltage fluctuation of ± 10% is permissible. However, a limitation to ± 5% will increase tube life.

2) Tube life is greatly dependent on the loading of the tube, especially on the plate voltage at higher frequencies. It is therefore better to obtain the required output power at the lowest possible plate voltage.

3) The permissible maximum temperature (t_{max}) may not be exceeded at any part of the tubes surface. It is important to always remain below this temperature in order to obtain a long tube life. Approximate air volumes for sufficient tube cooling should be taken from the instructions of the manufacturers.

5.4. NOTES

The author was not able to test the power amplifier at plate voltages higher than 600 V since the feedthrough capacitor C 6 had a maximum rating of 600 V. Also, a drive level of only approximately 100 mW was available. However, it is assumed that still higher output powers should be obtained at higher drive and plate voltage levels.

The described power amplifier has proved itself in operation over balloon-carried repeaters and UHF-DX (F, HB, LX, OE, PA). The author drives the driver-power amplifier combination with the 28-432 MHz transmit converter DJ 6 ZZ 002 (1) and the output power of 35 W is fed into a 25 element long yagi antenna.

6. SPECIFICATIONS OF THE 2 C 39 FAMILY

Since it is often difficult for radio amateurs to obtain extensive information on microwave tubes, the most important details of these tubes are to be given.

Table 1 lists various specifications which do not differ greatly between the various tube types. The data given in Figures 15, 16 and 17 are virtually valid for all types.

The RF-output power amounts to 45 W at 500 MHz, 800 V and 100 mA plate current. This represents an efficiency of 56% at the DC input power of 80 W. The plate dissipation will only be 35 W which is far below the permissible maximum limit of 100 W. The full plate dissipation can only be utilized at higher plate voltages. However, one will soon leave the guarantee range of the tube which will also have an adverse effect on tube life. The maximum cathode current ratings should also not be exceeded. Table 1 also lists two types having a higher cathode current rating (YD 1051 and 7211).

7. AVAILABLE PARTS

See material list.

8. REFERENCES

(1) F. Weingaertner: A 28 - 432 Transmit Converter with FET Mixer
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 99-106

(2) K. Eichel: Stripline Transverter for 70 cm
VHF COMMUNICATIONS 2 (1970), Edition 4, Pages 224-239

PORTABLE SSB TRANSCEIVER FOR 144 - 146 MHz
WITH FM ATTACHMENT

PART III: FM ATTACHMENT

by G. Otto, DC 6 HL

7. FM-ATTACHMENT

The SSB transceiver is provided with an independent IF module for the FM mode that is connected just before the SSB crystal filter. Of course, this module is equally suitable for modifying other SSB transceivers for FM reception that have an intermediate frequency of 9 MHz. A small frequency-modulated 9 MHz oscillator is injected after the SSB crystal filter in the transmit mode and replaces the 9 MHz SSB signal. The 9 MHz FM signal is then processed in the transceiver in the conventional manner.

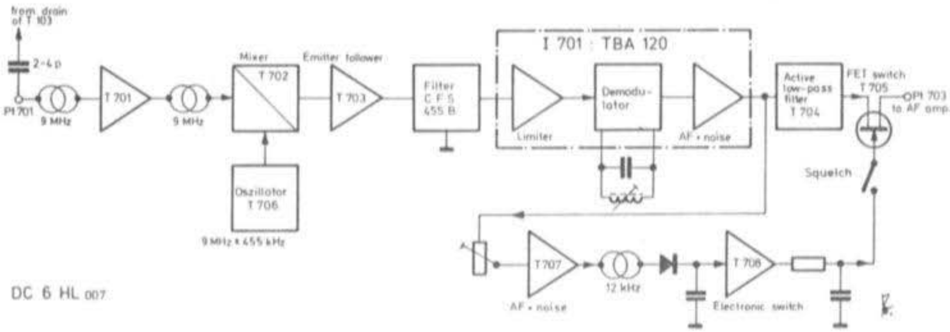


Fig. 28: Block diagram of the FM IF module

7.1. FM IF MODULE

Figure 28 shows the block diagram of the FM IF module. The wideband signal converted down to 9 MHz in the receiver is fed via the tuned amplifier T 701 to the mixer T 702, where it is converted to the second IF of 455 kHz by beating with the local oscillator signal from T 706. The selectivity is gained at 455 kHz using a ceramic filter having a 3 dB bandwidth of 20 kHz (50 kHz at 70 dB). The ceramic filter is followed by a limiter, coincidence demodulator and AF amplifier, which are combined in one integrated circuit (I 701). An active audio filter (T 704) is used to suppress the high-frequency noise components. The module also possesses a squelch circuit which operates as noise amplifier and actuates an electronic switch when no carrier signal is present.

The FM IF module represents a single-conversion superhet tuned to the intermediate frequency of the receiver module (9 MHz). For simplex operation in the transceive mode. It is necessary for the local oscillator frequency to be 9 MHz $\pm$ 455 kHz in order to obtain the required IF of 455 kHz. However, due to the increasing use of repeater stations, it may be required to have a constant transmit-receive frequency spacing of say 600 kHz, which is the most popular spacing at present. This can easily be obtained by shifting the local oscillator frequency from 9 MHz $\pm$ 455 kHz by 600 kHz or the required offset frequency. If a frequency of 9.600 MHz $\pm$ 455 kHz is selected, for instance, this will result in a receive frequency that is 600 kHz higher than the transmit frequency. If the local oscillator is made variable, it is possible for simplex or any frequency offset to be obtained according to the frequency variation range of the local oscillator. It would, of course, also be possible for a crystal oscillator equipped with two crystals to be used so that one is able to switch from simplex to duplex operation with the required frequency spacing. This FM-IF module only requires an input voltage of 1 μV which makes it extremely versatile.

The circuit diagram of the FM IF module is given in Figure 29. The input (Pt 701) is loosely coupled via a capacitor of 2 to 4 pF to the drain of mixer transistor T 103 on the main board. Since this coupling is at high impedance, it is not possible for coaxial cable to be used, and the interconnection must be as short as possible. Three ready-wound IF filters as used on the main board DC 6 HL 001 are used in the input stages of the IF module (L 701 - L 703). Ready-wound IF filters are also used in the three 455 kHz stages after the mixer, and for the coincidence demodulator. These miniature IF filters with built-in capacitors save a considerable amount of work during construction. It is only necessary for the inductances for the local oscillator (9.000 $\pm$ 455 kHz or required offset frequency, L 707), and noise amplifier (squelch) (L 708, 12 kHz) to be wound by hand. Since it is important that the resonant frequency of the noise amplifier is exactly 12 kHz, the circuit has been provided with a ceramic capacitor of 10 nF.

Figure 30 gives the PC-board DC 6 HL 007 and shows the component locations; Figure 31 shows a photograph of the authors prototype. The oscillator should be screened; this is not shown in the photograph.

Fig. 31: A photograph of the completed module

7.1.1. SPECIAL COMPONENTS FOR THE FM IF MODULE DC 6 HL 007

I 701: TBA 120 (Siemens)

T 701, T 705: BF 256, BF 245 (TI), W 245 (Siliconix)
T 702 : 40604, 40673 (RCA) or similar dual-gate MOSFET
T 703 : BF 115, BF 224 or similar silicon NPN RF transistor
T 704, T 706, T 707: BC 108 B, 2 N 2926 or similar silicon NPN AF transistor

D 701, D 702: 1 N 4148, 1 N 914 or similar silicon diode

L 701-L 703: 10.7 MHz IF transformers (type FM-FB, with built-in capacitors)
L 704-L 706: 455 kHz IF transformers (with built-in capacitor)
L 707: 14 turns of 0.1 to 0.2 mm dia. (32-38 AWG) enamelled copper wire on a 4 mm dia. coil former with SW core (vary according to offset frequency)
L 708: 18 mH; 268 turns of 0.1 mm dia. (38 AWG) enamelled copper wire in a ferrite potted core. Material N 28, AL-value: 250. Size 14 x 8 mm (Siemens: B 65 541 - K 0250 - A028)

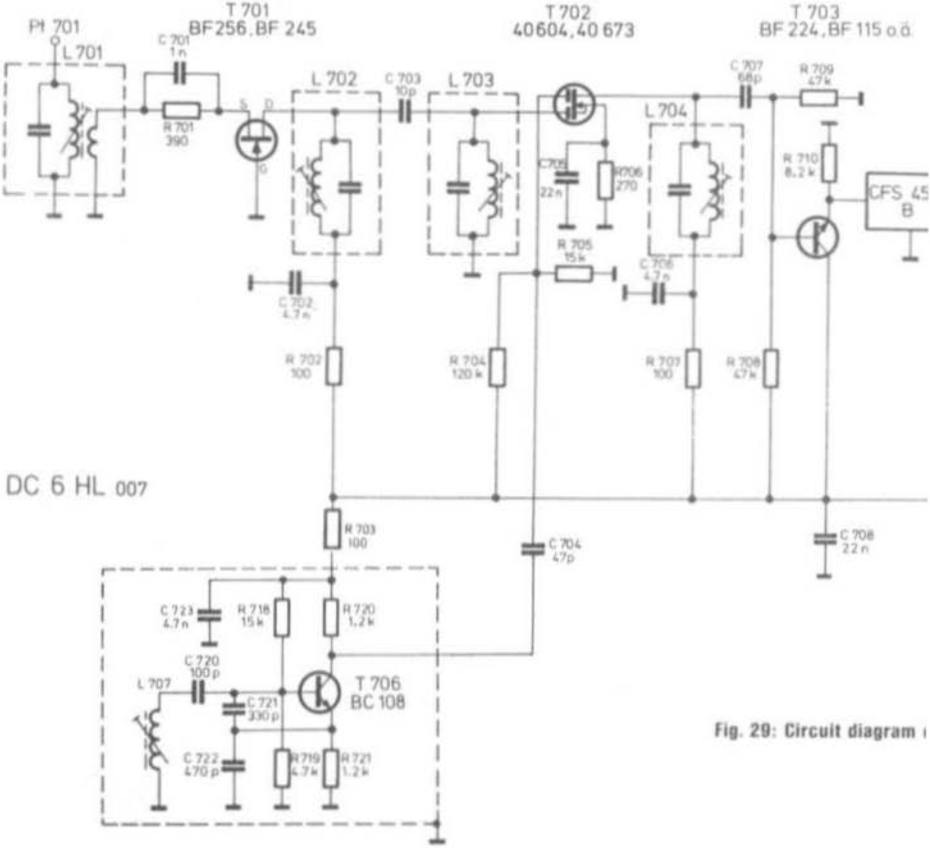


Fig. 29: Circuit diagram

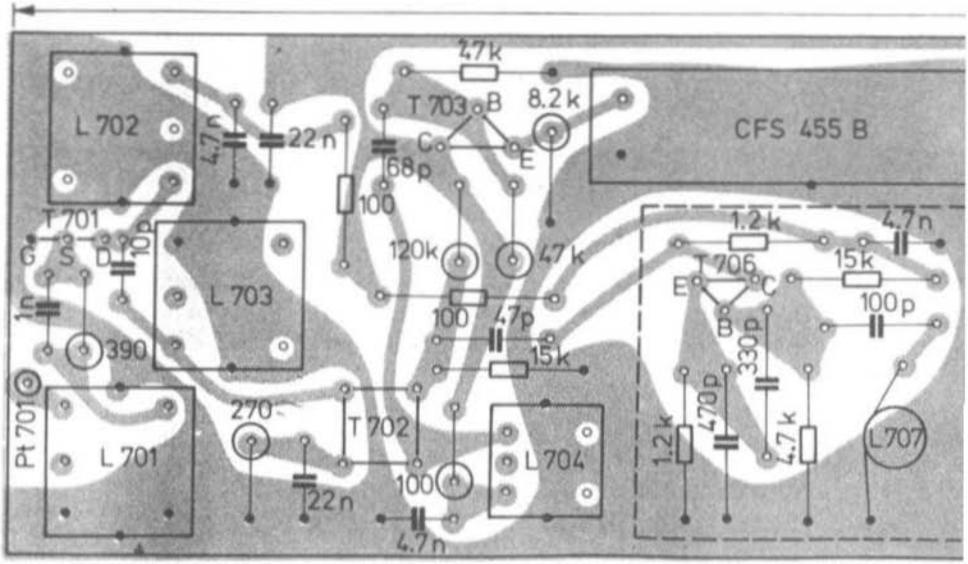
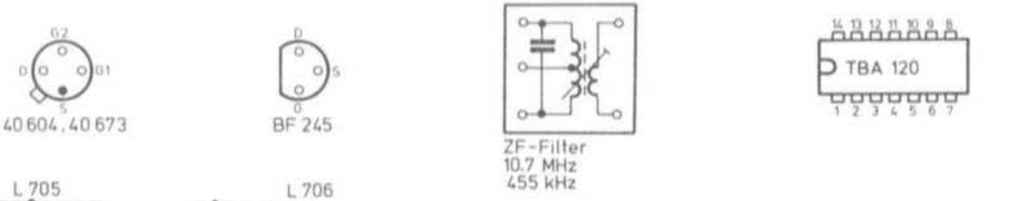
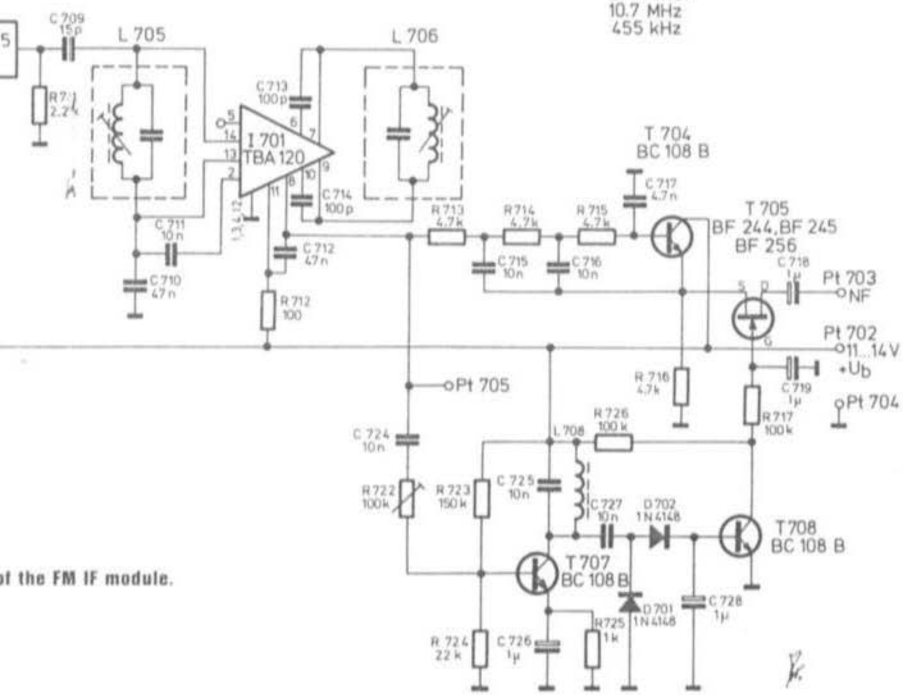
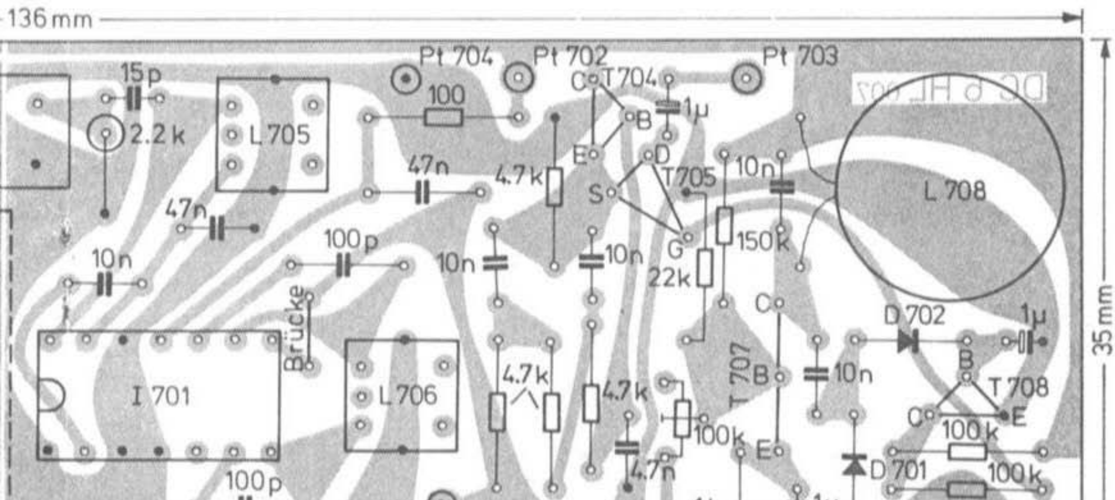


Fig. 30: PC-boa





of the FM IF module.



rd DC 6 HL 007 with component plan.

Styroflex capacitors: 3 x 100 pF; 1 x 330 pF; 1 x 470 pF
1 miniature trimmer potentiometer 100 kΩ for vertical mounting.
Spacing 2.5/5 mm.

All electrolytics are tantalium types whose value is not critical.

Bypass capacitors: 1 nF, 10 nF, 22 nF, 47 nF with 5 mm spacings.

7.1.2. ALIGNMENT

Only one special point must be noted with the alignment. The resonant circuit comprising L 706, which effects the phase shift for the coincidence demodulator, should be aligned for best reproduction of a strong FM signal, e.g. for minimum distortion. The signal will be louder and distorted at both sides of the correct alignment. All other stages should be aligned for maximum with a low level input signal which does not cause the IF-module to limit. The centre frequency of strong signals is indicated when the DC voltage at test point Pt 705 falls to half the operating voltage.

The threshold voltage of the squelch circuit can be adjusted with trimmer potentiometer R 722.

7.2. ADDITIONAL FM OSCILLATOR

The output signal of the FM oscillator module DC 6 HL 008 replaces the 9 MHz SSB signal in the FM transmit mode. The circuit diagram of this oscillator is given in Figure 32. The oscillator operates at 9 MHz and is frequency modulated with the aid of a varactor diode. The actual oscillator (T 801) is followed by an emitter-follower, buffer stage (T 802). The output of this module is fed via a capacitance of 10 pF (a trimmer may be used) to the base of transistor T 113 on the main board (DC 6 HL 001). This interconnection is at high impedance which means that it should be as short as possible. However, it can remain connected at all times, which means that it is only necessary to switch the operating voltage of the appropriate module in order to switch from the SSB to FM mode etc.

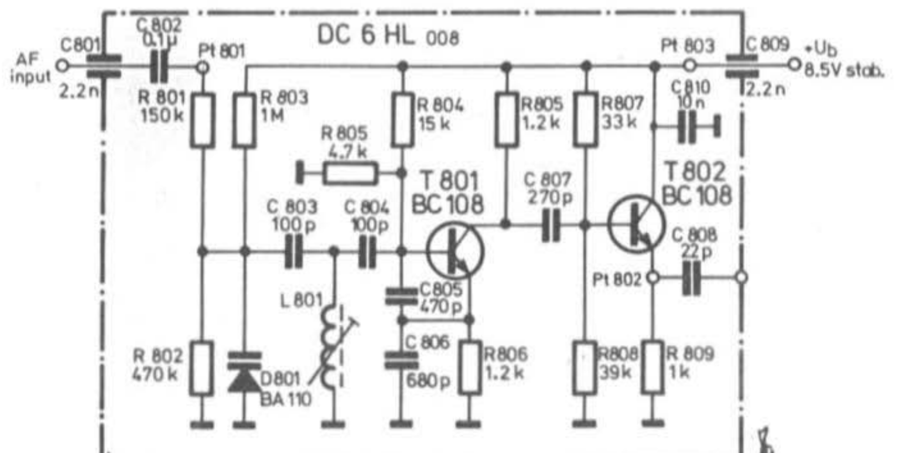


Fig. 32: Circuit diagram of the transmit oscillator

The AF input signal is taken from the emitter of transistor T 111 on the main board DC 6 HL 001 and is fed via a potentiometer for adjustment of the frequency deviation and a isolating capacitor to Pt 801 of the FM oscillator board. The value of this potentiometer is not critical and can be in the range of 1 kΩ to 25 kΩ.

The alignment is extremely simple. It is only necessary for the frequency to be adjusted by aligning the core of inductance L 801 to 9 MHz. The value of the external coupling capacitor (trimmer) to the main board should be reduced to a point just before the output voltage of the transmitter starts to fall.

Figure 33 illustrates PC-board DC 6 HL 008 and shows the component locations. This board is double-coated and the components are directly soldered on to the conductor lanes. No drilling is required. The same principle was also used for modules DC 6 HL 002 and 003. The screening should be fitted with the required two feedthrough capacitors and a low-capacitive feedthrough.

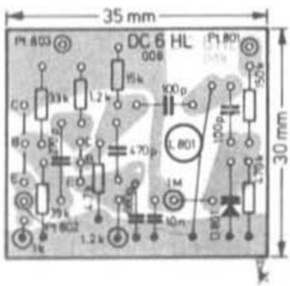


Fig. 33: PC-board DC 6 HL 008 and component plan for the FM transmit oscillator

7.2.1. COMPONENTS FOR DC 6 HL 008

T 801, T 802: BC 108, 2 N 2926 or similar silicon AF transistor

D 801: BA 110 (ITT-Intermetall), BA 121, BA 149 (Tfk) or similar (approx. 10 pF) varactor diode

L 801: 33 turns of 0.35 mm dia. (27 AWG) enamelled copper wire wound on a 5 mm dia. coilformer with SW core.

Styroflex capacitors: 2 x 100 pF; 1 x 270 pF; 1 x 470 pF; 1 x 680 pF.

1 plastic foil capacitor 0.1 μF.

8. AVAILABLE PARTS

Please see price list.

9. REFERENCES

G. Otto: A Portable SSB Transceiver for 144 MHz - 146 MHz with FM Attachment
Part 1: VHF COMMUNICATIONS 4 (1972), Edition 1, Pages 2-15
Part 2: VHF COMMUNICATIONS 4 (1972), Edition 2, Pages 66-79

HOME MADE REFLECTOMETER FOR 100 - 1400 MHz

by R. Griek, DK 2 VF

A reflectometer is to be described that allows the matching and output power to be measured in a frequency range of approximately 100 to 1400 MHz. Fig. 1 shows a photograph of the reflectometer. It is provided with built-in demodulating inserts and only requires two inexpensive BNC-connectors for the through-line. This through-line consists of the inner conductor and dielectric of a coaxial cable (e.g. RG-8/U). This line is then embedded in an aluminum block that is provided with two cavities for the demodulator inserts. These cavities are deep enough to allow a spacing of approximately 2 mm between the auxiliary lines and the inner conductor.

The author was not able to measure the coupling attenuation and directivity as a function of frequency. However, the coupling attenuation seems to be very high since a power output of approximately 4 W results in a reading of only 100 μA at 144 MHz. At 1296 MHz, the losses in the coupling loop and demodulator circuit will have increased so that the reading is even lower than at 144 MHz. The directivity can be estimated if a good terminating resistor is available. It has been found that the return power reading deviates only slightly from zero, even at 1296 MHz.

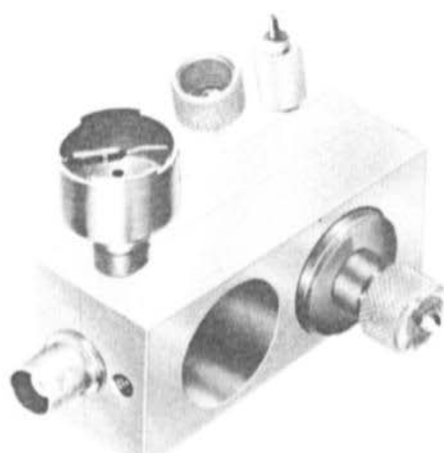


Fig. 1: Home-made wideband reflectometer

1. CONSTRUCTION

Figure 2 shows an exploded diagram of the reflectometer. The reflectometer is constructed in the following order:

An aluminum block of 75 mm in length (the other dimensions can deviate from those given in Figure 2) is sawn through in a longitudinal direction (lengthwise) after which the surface is milled. The thickness of the two halves should then be approximately 10 mm, and 15 mm. Tapped holes should be provided at the corners so that the two halves can be screwed together. The accuracy of the reflectometer will be improved if two pins are provided in the centre.

This is followed by drilling a hole corresponding to the diameter of the cable dielectric in a longitudinal direction through the block. Both ends of this block should be tapped to provide the required threading for the BNC-connectors (type UG - 1094/U for single-hole mounting). After this, the 25 mm diameter cavities are milled into the aluminum. The depth of these cavities is such that

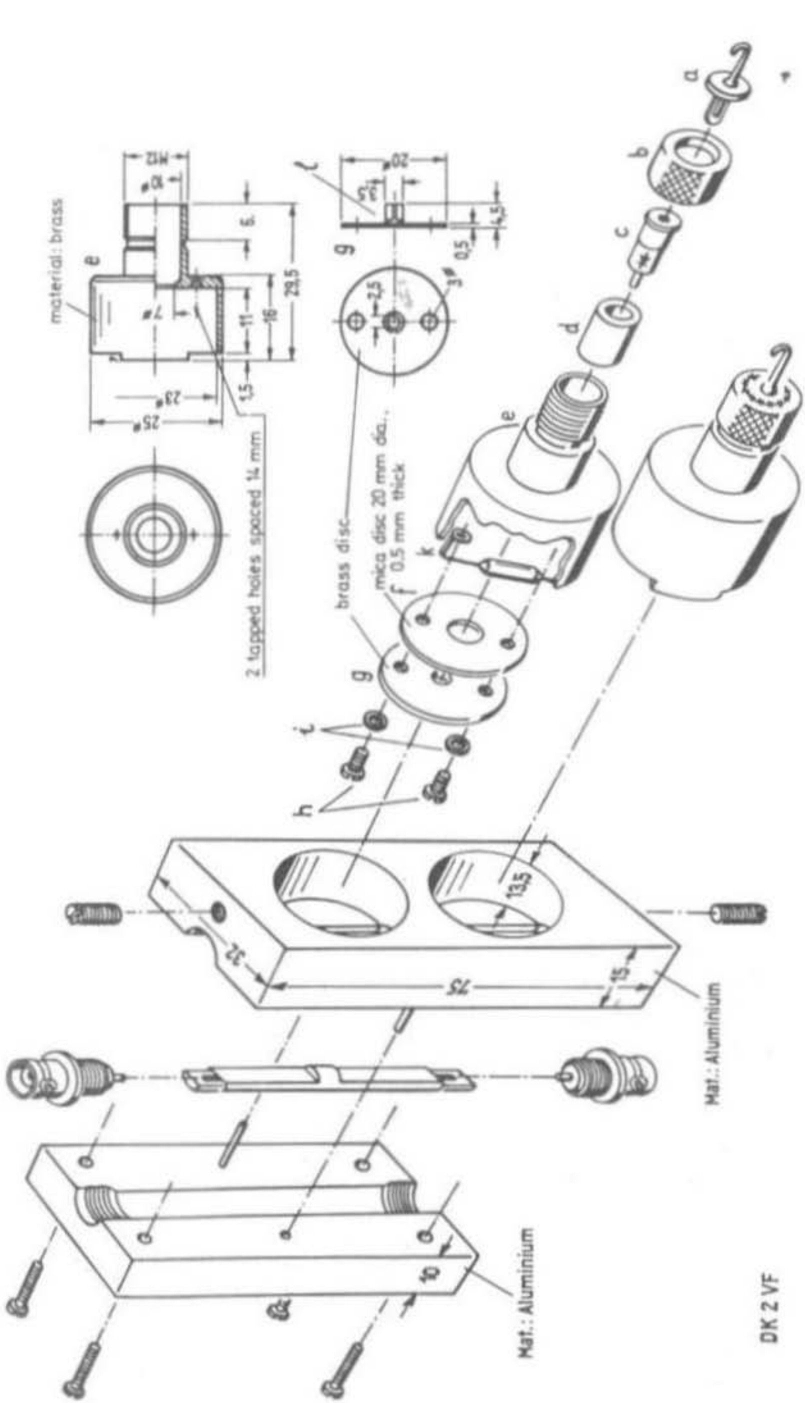


Fig. 2: Constructional details of the reflectometer

only a thin layer of the dielectric of the cable (approx. 0.5 - 1 mm) remains. This is obtained most favourably by inserting the inner conductor with the full dielectric into the aluminum block and for the dielectric to be removed in the same milling process as the aluminum. Finally, the two tapped holes for the grub screws holding the demodulator inserts should be made, the BNC-connectors should be inserted and soldered into place. After this, the main body of the reflectometer will be complete.

The two demodulator inserts are identical. They are both lathed from 25 mm diameter brass bars. Further details are given in Figure 2. A mica feedthrough capacitor (a) is soldered as cover to tube (b), which has been tapped with a thread matching part (e). A teflon tube (d) is placed over the diode (c) so that it fits into the 10 mm hole in part (e) and insulates the diode.

The connection pin of the diode protrudes through the brass disc (g), which is insulated by the mica disc (f) from the main body (e). Solder tag (k) is placed under an insulating ring which is in contact with disc (g) but not with screw (h). A 150 Ω carbon resistor is soldered to this tag with one relatively long connection so that the other lead can be soldered to ground directly adjacent to the body of the resistor, i.e. to part (e). The longer lead of the approximately 12 mm long resistor represents the auxiliary line. It must be near and parallel to the inner conductor of the cable and be bent so that it runs near to the edge of part (e) and brought to disc (g). This disc is provided with a slotted collar (1), which provides a good connection for the pin of the diode. The wider contact of the diode touches the shortened and bent solder tag of the mica capacitor (a). The ball-shaped edge of part (e) is shortened by 1.5 mm over most of its circumference; only two tongues remain for maintaining the spacing between coupling loop and inner conductor.

The mechanical design is especially suitable for diodes of the 1 N 21 series or similar types. The polarity results in the negative, rectified voltage at the connections of the demodulator inserts. If the transmitter is connected to the left hand coaxial socket, the left hand demodulator insert will indicate the forward power. The right hand demodulator insert (located on the load side) will indicate the reflected power. The meters should be sensitive enough so that full scale deflection can be attained at relatively low power levels.

The reflectometer is aligned by indicating forward power and rotating the demodulator insert for maximum reading. This will be the case when the coupling loop is directly parallel to the inner conductor and points towards the connectors. The reflectometer is now connected in the opposite direction and the second demodulator insert rotated for maximum reading in the same manner. By suitable selection of the diodes (assuming that the resistors are of the same value and that the coupling links are equally large) it is possible for the reading to be brought to the same value in both cases. If this should not be possible, this will mean that the spacings between coupling loop and inner conductor are different. In this case, the more sensitive demodulator insert should be pulled out slightly.

3. REFERENCES

R. Griek: Simple Stripline Reflectometers for 144 and 432 MHz
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 89-92

RECOMMENDED STANDARDS FOR FM REPEATERS AND FIXED CHANNEL FM STATIONS IN THE 2 m BAND

by T. Bittan, G 3 JVQ/DJ ∅ BQ

New regulations were adopted at the May 1972 Conference of IARU Region 1 in Scheveningen/Holland. It was decided not to adopt the West German system for repeater operation due to the interference which this is stated to cause to sub-bands in other countries. The following frequency plan was recommended for FM transmissions. There has been many comments regarding the 70 cm repeater recommendations which have a frequency spacing of 7.6 MHz. It is assumed that the reason why the 0.6 MHz used at 144 MHz could not also be adopted at 70 cm is because a frequency spacing of 0.6 MHz only amounts to just over 0.1% at 432 MHz and would probably require separate location of the repeaters transmitter and receiver.

1. FM REPEATER STATIONS

1.1. REPEATER FREQUENCIES

Output frequency 600 kHz above input frequency

Channels:	R 1	145.025	-	145.625
	R 2	145.050	-	145.650
	R 3	145.075	-	145.675
	R 4	145.100	-	145.700
	R 5	145.125	-	145.725
	R 6	145.150	-	145.750
	R 7	145.175	-	145.775
	R 8	145.200	-	145.800
	R 9	145.225	-	145.825

1.2. POLARISATION: vertical

1.3. REPEATER OPERATION:

Without a new selective call the operating time for a repeater is limited 3-10 minutes. The frequency of the selective call is 1750 ± 50 Hz. When the signal to be relayed has disappeared or the operating time has come to the end, the repeater station will send it's own call for 15 seconds after which the transmission is interrupted. The automatic identification cannot be interrupted by a selective call. F 2 is used as identification modulation.

1.4. POWER TO BE USED IN THE REPEATER SERVICE

Recommended effective radiated power of the repeater transmitter shall not exceed 15 watts. When working through a repeater the lowest usable power consistent with good communication is recommended.

1.5. MODE OF TRAFFIC

Simplex on one channel, demodulate/remodulate systems.

1.6. DEVIATION OF REPEATER TRANSMITTER

± 3 kHz, 12 F 3

1.7. L.F. RESPONSE

300-3000 Hz. Outside this it is attenuated by 12 dB/octave.

1.8. PRE-EMPHASIS: + 6 dB/octave in the transmitter.

1.9. DE-EMPHASIS: - 6 dB/octave in the receiver.

1.10. RESPONSIBILITY FOR REPEATER OPERATION

The repeater shall be under the control of the national IARU member society or its agent. The member society shall be responsible for the allocation of the adopted frequency channels.

2. FIXED CHANNEL AMATEUR STATIONS

2.1. SIMPLEX CHANNELS

S 0	145.000	Mobile calling frequency
S 21	145.525	
S 22	145.550	
S 23	145.575	
S 24	145.600	
S 25	145.625 +	
S 26	145.650 +	
S 27	145.675 +	
S 28	145.700 +	+ Also repeater output
S 29	145.725 +	frequencies
S 30	145.750 +	
S 31	145.775 +	
S 32	145.800 +	
S 33	145.825 +	

2.2. MODE OF TRAFFIC: Simplex on one channel

2.3. DEVIATION: $\pm$ 3 kHz, 12 F 3

2.4. PRE-EMPHASIS: + 6 dB/octave in the transmitter

2.5. DE-EMPHASIS: - 6 dB/octave in the receiver

2.6. L.F. RESPONSE: 300 - 3000 Hz. Outside this attenuated by 12 dB/octave.

3. UHF REPEATER CHANNELS

Channel	Input (MHz)	Output (MHz)
70	431.05	438.65
72	431.10	438.70
74	431.15	438.75
76	431.20	438.80
78	431.25	438.85
80	431.30	438.90
82	431.35	438.95

Simplex 435.00 MHz.

MODIFYING THE DL 6 HA 001/28 DUAL-GATE MOSFET CONVERTER FOR RECEPTION OF WEATHER SATELLITES AND OTHER SPACE VEHICLES

by T. D. Bittan, G 3 JVQ/DJ Ø BQ

INTRODUCTION

The DL 6 HA 001/28 MOSFET converter can be easily modified for reception of weather satellites and other space vehicles. With the described modification, it is only necessary to alter the inductance values and two capacitors. The same crystal of 38.667 MHz can be used as in the original description. This results in an intermediate frequency range of 20 to 22 MHz, which can be covered by virtually any general coverage receiver.

The original circuit diagram is given in Fig. 1. All component values except for the inductances and capacitors C 12 and C 15 are still valid. The input frequency range is 136 to 138 MHz and the IF output is 20 to 22 MHz.

It is advisable for circular-polarized antennas to be used for reception. The same was found for the OSCAR V satellite. This is because the satellites tend to rotate in space or for them to be between vertical and horizontal polarization. This can be achieved with a cross yagi (3).

COIL DATA

L 1: 2 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound onto L 2.
L 2: 6 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, coil length 10 mm, centre tap, self-supporting.
L 3 and L 5: 6 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm coil former with core, coil length 15 mm.
L 4 and L 6: As L 2 but without centre tap.
L 7 and L 8: 17 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound on a 3.5 mm dia. coilformer with core.
L 9 and L 10: 7 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, coil length 10 mm, self-supporting.

CAPACITORS

C 12: 20 pF
C 15: 85 pF

REFERENCES

(1) G. Laufs: The 144 MHz Converter with Dual-Gate MOSFET Mixer
VHF COMMUNICATIONS 2 (1970), Edition 1, Pages 1-11

(2) Drs. F. M. Schimmel and Drs. W. Janssen: Weersatellieten Waarnemen
Radio Electronica 16 April 1971, Pages 304-306

(3) E. Reitz: A Tiltable Antenna with Selectable Polarity
VHF COMMUNICATIONS 2 (1970), Edition 1, Pages 12-20.

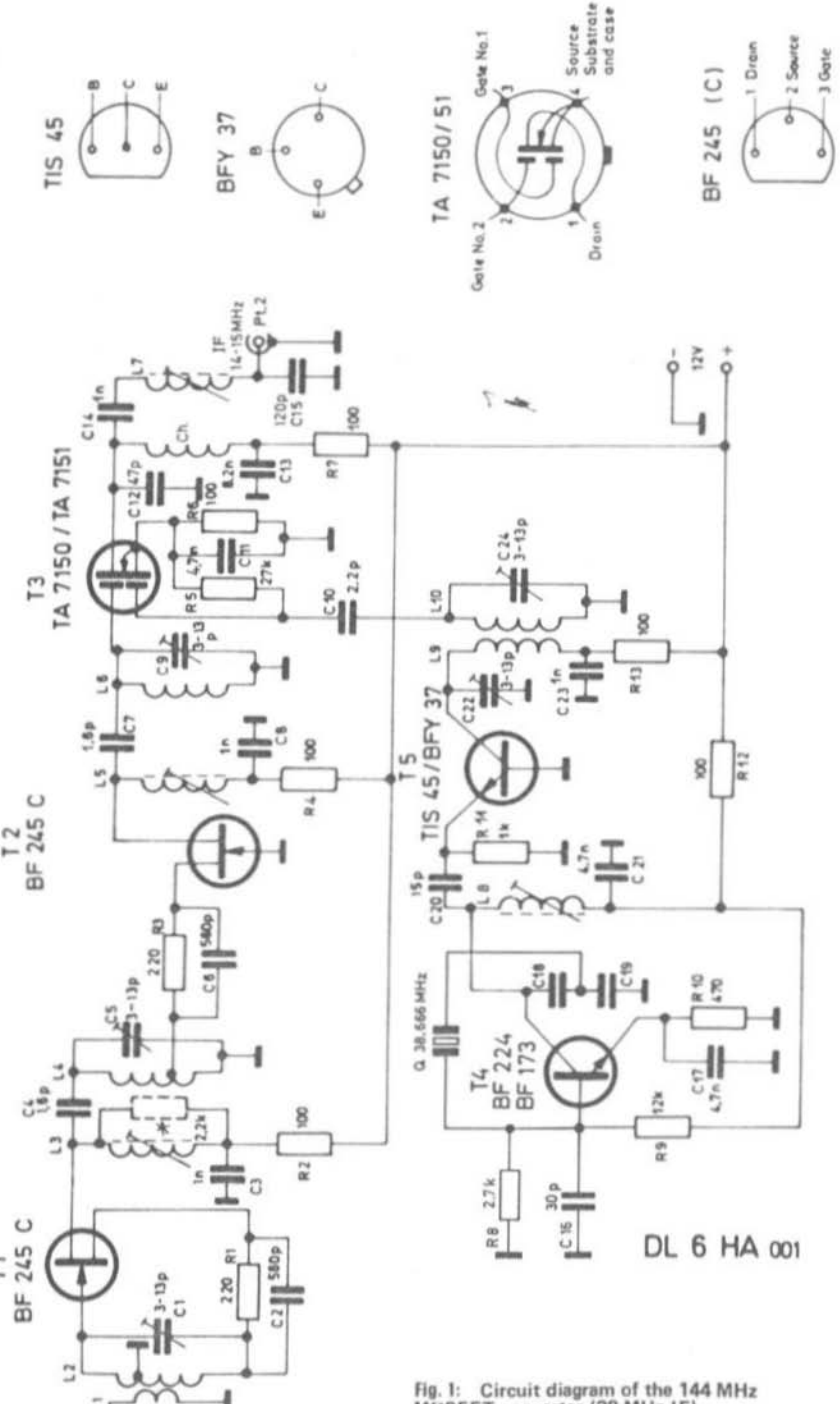


Fig. 1: Circuit diagram of the 144 MHz MOSFET converter (28 MHz IF)

DIPLEXER AMPLIFIER FOR 28 - 30 MHz

by G. Rühr, OH 2 KT (DL 7 IM)

It is very advantageous, especially during contests or satellite translator operation for several shortwave receivers to be connected to one VHF converter. Required is a special distributor amplifier at the intermediate frequency (28 MHz) which is provided with one input connected to the converter and several outputs which can be fed to a number of shortwave receivers tuned to the IF-range.

1. CIRCUIT DETAILS

Fig. 1 shows the circuit diagram of such a distributor amplifier with three outputs. The input of the amplifier is aperiodic (untuned) since VHF converters usually possess at least one tuned circuit at the output. Inductance L 1 represents a wideband toroid transformer which matches the integrated differential amplifier CA 3028 A to the low-impedance converter output. The transformation ratio amounts to 1 : 5 so that an impedance transformation of 1 : 25 takes place.

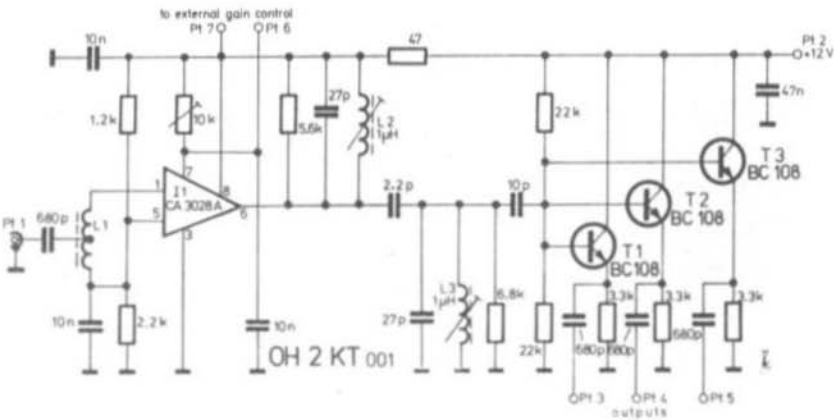


Fig. 1: Circuit diagram of the IF distributor amplifier

A capacitively coupled two-section bandpass filter is provided at the output of the integrated circuit. It is damped with resistors which ensures a flat passband characteristic (+ 1 dB) in the frequency range of 28 MHz to 30 MHz. Three emitter follower stages are connected to the secondary of this bandpass filter. They are each provided with a decoupled output with a low source impedance (less than 60 Ω).

The gain of the amplifier is variable between 0 and 20 dB with the aid of the 10 kΩ trimmer resistor at connection 7 of the CA 3028 A (I 1). If the gain is not to be preset to a certain value but should be adjustable from the front panel, it is possible for an external 10 kΩ potentiometer to be connected to connections Pt 6 and Pt 7. The trimmer resistor can then be deleted or adjusted to its maximum value. The noise figure of the IF-amplifier was 7 dB.

2. COMPONENTS

I 1: CA 3028 A (RCA)

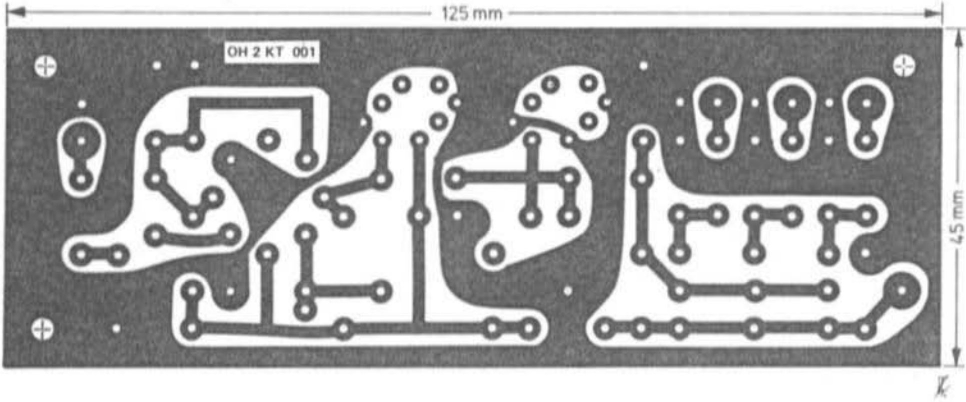
T 1 - T 3: BC 108, 2 N 3904 or other silicon NPN AF/VHF types.

L 1: Ferrite toroid, size and material is not critical. For instance, 9 mm outer diameter, 6 mm inner diameter, 3 mm in height.
Material: ferrocarit Fi 1e7 or Fi 05f7.

Five silk-covered and enamelled copper wires of approx. 0.2 mm dia. (32 AWG) should be wound together and pulled through the toroid core three times. Connect all five windings in series by appropriately soldering the starts and ends. The first soldered connection from the cold end is the input (Pt 1).

L 2, L 3: 1 µH, approx. 10 to 14 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound on a 5 mm diameter coilformer with shortened shortwave core and screening can.

All capacitors are ceramic tubular or disc types.



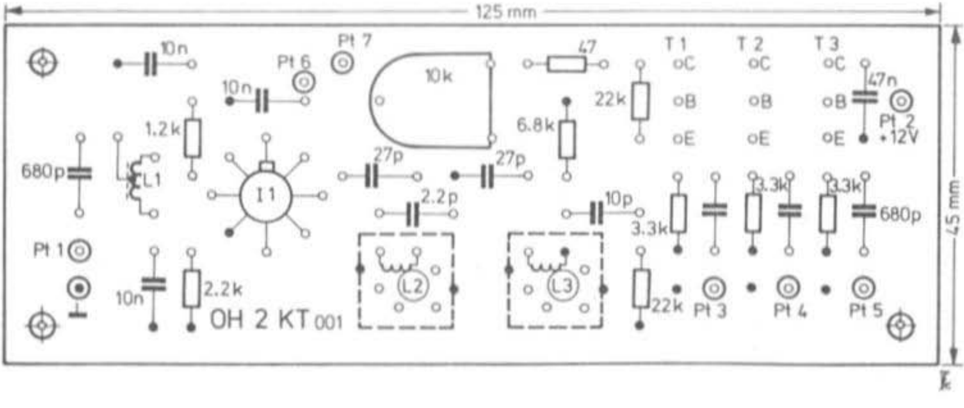


Fig. 2: Printed circuit board OH 2 KT 001 and component locations

3. CONSTRUCTION

The IF-amplifier can be built up on a small PC-board that has been designated OH 2 KT 001. Fig. 2 shows this printed circuit board together with component location plan. The construction is not critical. The alignment only consists of tuning the two resonant circuits comprising inductances L 2 and L 3. The damping resistor of the primary circuit is not accommodated on the PC-board but connected to the connections of the 28 pF capacitor. This allows the most favourable value for obtaining a flat passband characteristic to be found.

Fig. 3: Photograph of the author's prototype

4. REFERENCES

RCA ICAN-5337: Applications of the RCA CA 3028 Integrated Circuit RF Amplifier in the HF and VHF Ranges.

LIST OF THE TEKO MODULES ALREADY DESCRIBED AND FUTURE ADDITIONS

by D. E. Schmitzer, DJ 4 BG

Several of the modules for the modular system introduced in (1) and (2) have already been described. Since they have been described as individual modules without indicating the combination of these with future modules to form a complete system, it was thought that a general description should be brought listing all modules. Another author, DK 1 PN (3), has also developed TEKO modules which complement the present programme and allow alternate approaches.

Figures 1 and 2 give the block diagrams of a crystal-controlled FM receiver and SW/VHF receiver that can be easily constructed from the TEKO-modules:

RF:	Universal RF amplifier/mixer DJ 4 BG 011 (in preparation)
FM IF:	FM IF strip with crystal discriminator DJ 4 BG 008 in edition 1/72
CO:	Triple overtone crystal oscillator DJ 4 BG 014 (in preparation)
AF:	AF amplifier and voltage stabilizer DJ 4 BG 007 in edition 1/72
VFO:	Universal VFO with varactor tuning DJ 4 BG 012 (in preparation)
SSB/CW IF:	IF strip for SSB and CW DJ 4 BG 013 (in preparation) or IF strip DK 1 PN 003 (caution different voltage levels)
BFO:	Triple crystal oscillator DJ 4 BG 009 in this edition

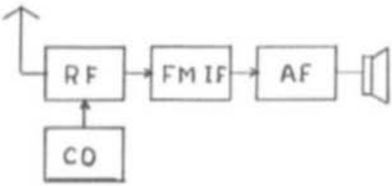


Fig. 1: Simple crystal-controlled FM receiver

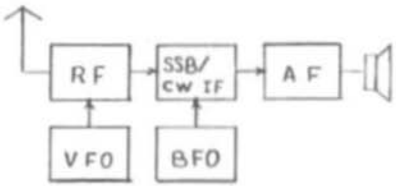


Fig. 2: SSB/CW receiver

A CRYSTAL OSCILLATOR MODULE WITH THREE INDEPENDENT OSCILLATORS

by D. E. Schmitzer, DJ 4 BG

The module to be described contains three, separate crystal oscillators that can operate with fundamental wave crystals in HC 25/U holders at frequencies between 3 and 25 MHz. The circuit can be used for a multiple of applications since each crystal is provided with its own oscillator circuit. Each of the three oscillators possesses a trimmer capacitor for aligning the frequency so that each crystal can be pulled onto the nominal frequency independent of another. All three oscillators drive a common buffer stage which provides a low-harmonic output voltage of 200 to 300 mV (peak-to-peak) at a source and load impedance of approximately 60 Ω. The switching of the individual oscillators is made with the aid of DC-lines with which the base bias of the required oscillator is fed.

1. OSCILLATOR CIRCUIT

The circuit is based on the well-known parallel-resonance circuit as shown in Figure 1. This circuit was modified by addition of a series trimmer capacitor for aligning the crystal frequency which allows the voltage divider connected to the transistor to exhibit higher capacitances than the conventional load capacitance of 30 pF. In order to ensure that the oscillating characteristics are not adversely affected, it is necessary for the collector of such circuits to be terminated at a relatively low impedance for alternating voltages. This is obtained here by connection of a common base circuit which serves as buffer amplifier. It provides a RF voltage at the collector and a high degree of isolation (Fig. 2). This circuit is similar to a cascode circuit, which is also known for its good isolation qualities.

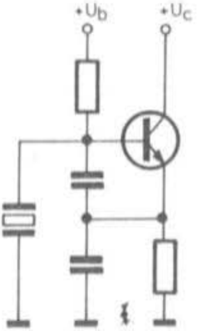


Fig. 1: A crystal oscillator circuit for parallel-resonance

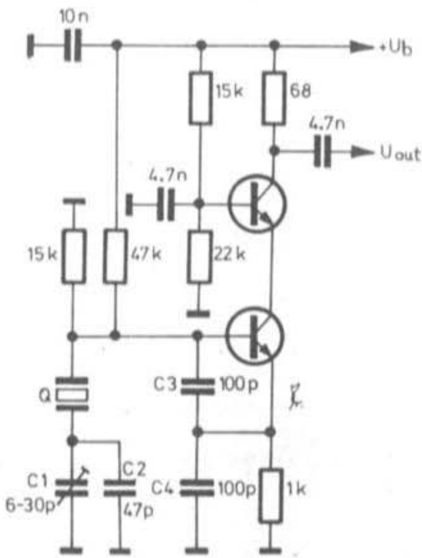


Fig. 2: Modified circuit with alignment capacitor and buffer stage

The complete circuit diagram of the crystal oscillator module is given in Figure 3. It will be seen that it comprises three identical crystal oscillator circuits that feed into a common buffer stage. The required oscillator is switched on by connecting the base voltage line to the operating voltage of + 12 V. A Pi-filter is connected to the output of the buffer stage in order to suppress any harmonics present on the output voltage, which is somewhat distorted on the collector resistor R 6.

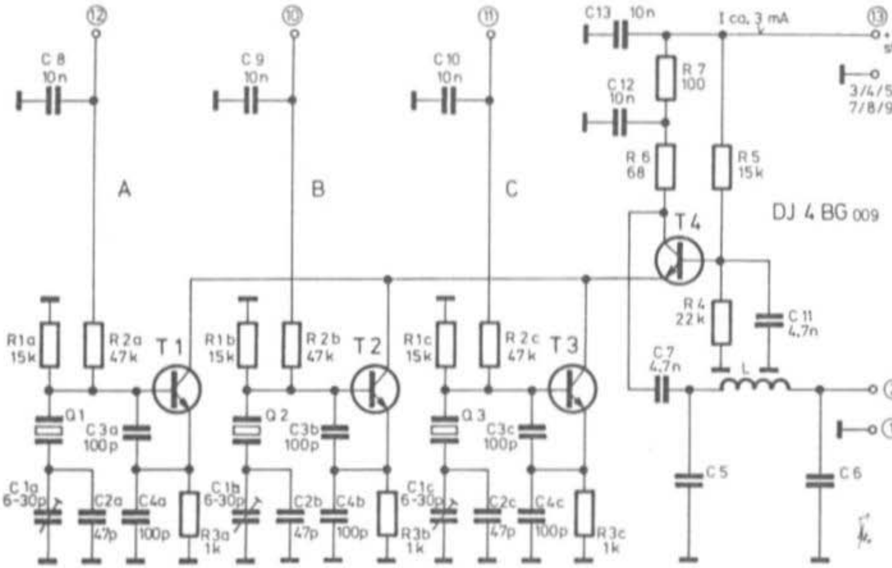


Fig. 3: Circuit diagram of the multiple crystal oscillator

2. SPECIAL FEATURES OF THE CIRCUIT

The adjustment range of the trimmer capacitors C 1a, C 1b and C 1c has been kept relatively low so that the nominal frequency can be adjusted easily. If the adjustment range is not sufficient to align the required frequency due to variance of the crystals or of other components, it is possible for C 2a, C 2b or C 2c to be reduced or increased (e.g. to 33 pF or 68 pF).

2.1. PI-FILTER

The Pi-filter has been dimensioned for a Q of approximately 2. This means that capacitors C 5 and C 6 are approximately 470 pF at 10.7 MHz and approximately 220 pF at 18 MHz. At other frequencies, the values of C 5 and C 6 should be modified correspondingly. The effective capacitance of the circuit is equal to half the value of C 5 or C 6 so that the inductivity can be calculated according to the resonance formula:

$$L = \frac{1}{C\,(2\pi f)^2}$$

For alignment, C 7 should be disconnected and no load present. The circuit formed by C 5, C 6 and L 1 can then be aligned with the aid of a dipmeter to the nominal frequency.

2.2. FREQUENCY RANGE

Basically, the module can be used over the whole frequency range for which fundamental wave crystals in HC 25/U holders are available. This is usually in the range of approximately 3 to 25 MHz. Cheaper crystals for amateur applications are offered in the range of 8 to 20 MHz. It is important that not only the frequency but also the following data is mentioned on ordering: Fundamental wave, parallel resonance, loading 30 pF.

3. CONSTRUCTION

The multiple crystal oscillator circuit is built up on a printed circuit board of 65 mm x 90 mm which is suitable for use in the modular receiver system described in (1). Figure 4 shows the printed circuit board which has been designated DJ 4 BG 009.

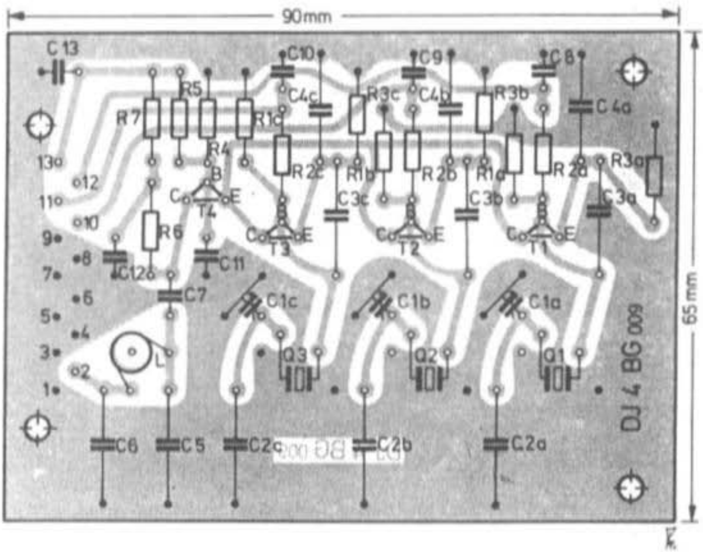


Fig. 4: Printed circuit board DJ 4 BG 009 showing the component positions

3.1. COMPONENT DETAILS

With the exception of the crystal holders and the trimmers, the construction does not possess any special features. If the module is to be used in the modular receiver system, the board can be provided with the 13-pole connectors.

Styroflex capacitors are recommended for C 2, C 3 and C 4 of the resonant circuit. Transistor type BF 224 can be used for transistor T 1 - T 4. However, since a large number of transistor types can be used equally well, the printed circuit board has been designed for the more popular base connection order of E-B-C. This must be taken into consideration when using transistors with different connections. At lower frequencies, for instance at 9 MHz, even AF-type transistors such as the BC 108 can be used. Such types often exhibit transit frequencies of up to 150 MHz. However, when the buffer is also to be used for frequency multiplication, it is very advisable for RF transistors such as BF 115, 2 N 706, 2 N 708, 2 N 914 or 2 N 918 to be used.

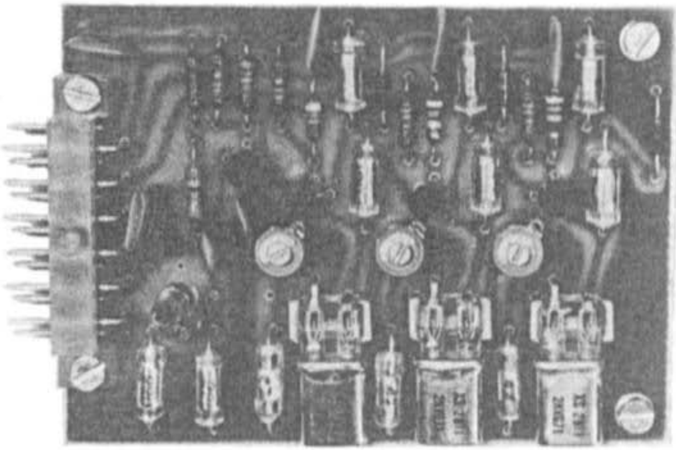


Fig. 5: Photograph of the author's prototype

4. APPLICATIONS

The module DJ 4 BG 009 was mainly designed as the local oscillator in the modular receiver system described in (2). In this concept, crystals in the order of 17 MHz are used(Example: channel 145.150 MHz; receiver IF: 9 MHz; crystal frequency: 17.01875 MHz). Of course, this module is equally suitable as exciter for a transmitter if a suitable phase modulator is used. In order to obtain a sufficient frequency deviation at good linearity, crystals for 8 MHz (x 18) or 9 MHz (x 16) should be used. Of course, the module is also suitable for other applications as control or auxiliary oscillator for 144 MHz or 432 MHz. The output can be modified so that the Pi-filter is replaced by a resonant circuit (the tap should be made so that maximum voltage is obtained at an output impedance of 60 Ω). The resonant circuit filters out the required harmonic which means that the signal is directly multiplied. A recommendation for this is given in Figure 5.

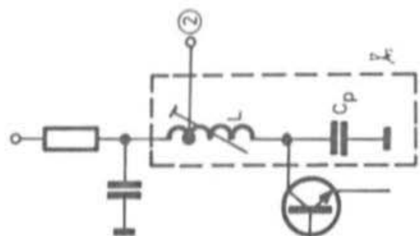


Fig. 6: Modification of the buffer stage for frequency multiplication

The module DJ 4 BG 009 can also be used in a receiver for the SSB and CW modes by equipping it with the beat frequency crystals for CW, upper and lower sidebands.

Furthermore, it can be used as an exciter in a miniature CW transmitter on the shortwave bands. This does not only provide the beginner with a very easy method of getting on the air, but also provides expedition stations with a very interesting possibility. Normally, such expeditions operate on a previously published frequency which means that it is sufficient for the crystal oscillator module to be equipped with the nominal frequency and with two other crystals with frequencies deviating by 1 kHz to 2 kHz so that it is possible to shift frequency slightly in the case of interference. It is possible to directly key the oscillators in the base lines. Only an extremely low frequency shift (chirp) exists. Of course, if a subsequent stage of the transmitter is keyed, a better CW tone should result. In order to ensure that the signal is not directly coupled through the transmitter during the off-periods (BK-keying), it is possible for the oscillator to be directly keyed and appropriate RC-link be provided in a subsequent stage so that there is a very slight delay in keying. When using this module as a crystal-controlled exciter for 15 m or 10 m, the previously mentioned possibility of frequency multiplying can be used for instance for tripling from 7 MHz to 21 MHz or doubling from 14 MHz to 28 MHz.

5. AVAILABLE PARTS

See advertising page.

6. REFERENCES

(1) D. E. Schmitzer: Plug-In Modular Equipment
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 107-109

(2) D. E. Schmitzer: A Modular Receiver System
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 110-114

A SIMPLE FET-TESTER

by H. Matuschek, DJ 3 MY

The favourable characteristics of field-effect transistors are well known. Their advantages can be utilized to the full if the most favourable transistor for a particular circuit can be selected from a number of transistors. It is also quicker to test all components before installation than to find the fault after the circuit has been completed. A FET-tester would therefore be of great assistance. A simple unit is to be described which can usually be built up from parts available in the shack. Even the more sophisticated tester, which allows measurement of the I_D/U_{GS} (drain current/gate source voltage) characteristic, can be constructed easily.

1. THEORY OF OPERATION

The polarity of the voltages in the FET-tester are provided for the popular N-channel field effect transistors. For testing of the less popular P-channel FETs, it is necessary for the polarity of the voltage sources and the meter connections to be reversed.

The operation and polarity of N-channel field effect transistors can be compared to those of triode and tetrode tubes. In this case the following electrodes have similar functions: source = cathode; gate 1 = grid 1 (control grid); gate 2 = grid 2 (screen grid); drain = anode. It is possible that this comparison may simplify the understanding of field effect transistors for some readers.

1.1. A SIMPLE TESTER

Figure 1 shows the circuit diagram of a simple FET tester. Two flat batteries are used for provision of the two, separate operating voltages. The voltage between drain and source (U_{DS}) is positive with respect to the source; the gate voltage U_{G1S} is negative with respect to the source. The gate voltage U_{G2S} can be varied from negative values via zero and up to approximately 1.5 V. The voltage U_{DS} remains constant at 4.5 V. A meter shunted for a full scale deflection of 5 mA indicates the drain current I_D. The two gate voltages are adjustable with the aid of 1 MΩ resistors.

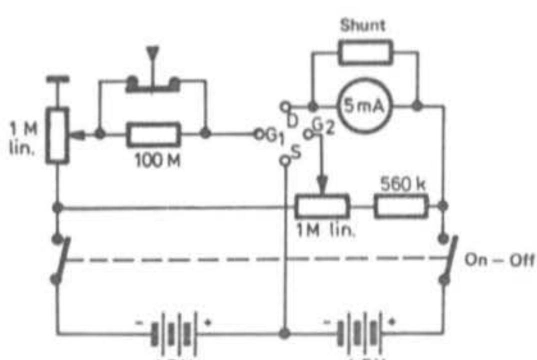


Fig. 1: Simple FET tester

The test procedure is very simple with single-gate field effect transistors: the drain current is greatest at $U_{GS} = 0$ V and will be reduced if U_{GS} becomes more negative. This means that the basic function can be tested and the transistor sorted roughly into the I_D/U_{GS} group. In addition to this, one obtains an impression as to the slope characteristics.

In order to examine the static input impedance, e.g. at DC, a 100 MΩ resistor is connected in the gate 1 connection and is normally bridged. On depressing the push button (break contact), the 100 MΩ resistor is placed in series with the gate voltage, which means that the gate voltage source is provided with an impedance of 100 MΩ. With good field effect transistors, the previously adjusted and indicated drain current will not be altered on increasing the impedance of the gate-voltage source from a few 100 kΩ to 100 MΩ. If, however, the drain current is reduced, e.g. by 10%, this would indicate that the input impedance of the field effect transistor is only ten times this value, e.g. 1000 MΩ (1 GΩ) and thus well under the typical values.

1.2. EXTENDED TESTER

The characteristics of field effect transistors vary greatly from transistor to transistor. It is possible, for instance, for the drain current of the popular transistor type BF 244/BF 245 (2 N 5345) to vary between 2 and 25 mA (at $U_{DS} = 15$ V and $U_{GS} = 0$ V). The gate voltage required for blocking the transistor ($U_{DS} = 15$ V and $I_D = 10$ nA) can be between -0.5 V and -8 V. Figure 2 shows the tolerance field for the I_D/U_{GS} characteristic of this type of transistor.

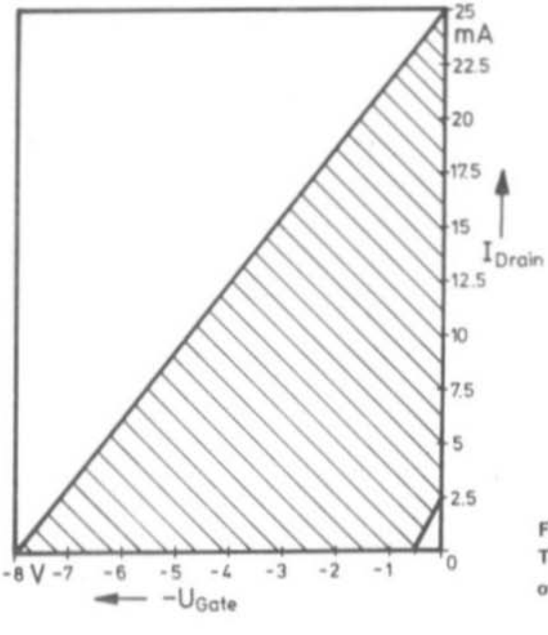


Fig. 2:
Tolerance field for the I_D/U_{GS} characteristic of the field effect transistors BF 244/245

For the transistor types BF 244/245 (well known from the DL 6 SW FET-Converter), three groups are valid:

Group	I_D at U_{DS} = 15 V, U_{GS} = 0 V	$-U_{GS}$ for I_D = 200 μA at U_{DS} = 15 V
A	2.0 - 6.5 mA	0.4 - 2.2 V
B	6.0 - 15.0 mA	1.6 - 3.8 V
C	12.0 - 25.0 mA	3.2 - 7.5 V

However, many other transistor types are not divided into such groups.

In order to obtain the most favourable characteristics for the application, the field effect transistor should be selected according to its characteristics. For stages where a high cross or intermodulation rejection is required, a field effect transistor with a long, flat characteristic is more favourable, whereas a steep characteristic is preferable for high gain.

Figure 3 shows a more elaborate tester for determination of the characteristic curve. The drain voltage U_{DS} is increased to 15 V and the gate voltage U_{GS} to -8 V with the aid of a small power supply that provides the two voltages. The drain voltage is stabilized with the aid of a pass transistor and a zener diode. The potentiometers for adjustment of the two gate voltages are calibrated with the aid of a valve voltmeter (VTVM) in steps of 0.5 V. The drain current meter is provided with an additional shunt which can be switched into circuit to provide a 20 mA range.

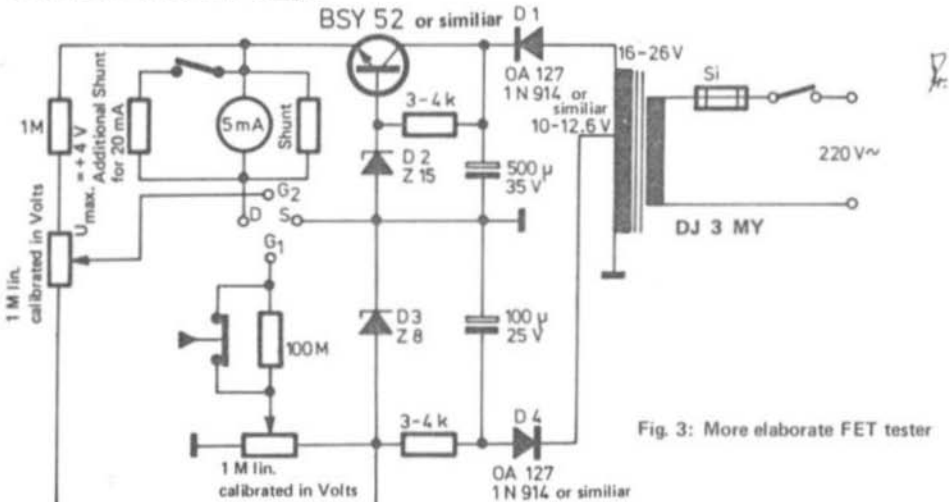


Fig. 3: More elaborate FET tester

With these measuring ranges, it is possible for the previously mentioned characteristics of field effect transistors to be measured. However, the power dissipation limits must be taken into consideration at high drain currents. In the case of the BF 245, this is 300 mW. This means that the drain current should not exceed a value of 20 mA at a test voltage of 15 V.

The static (DC) input impedance, which need only be measured for one gate, is carried out in the same manner as with the simple tester, by placing a 100 MΩ resistor in series with gate 1.

2. DUAL GATE FIELD EFFECT TRANSISTORS

Whereas gate 2 of a dual gate junction FET can be driven between -7 V and 0 V in the same manner as gate 1, this is not valid for the dual gate MOSFET. Gate 1 of a MOSFET can be driven from several minus volts to approximately +1 V, gate 2, on the other hand, up to approximately 30% of U_{DS} (e.g. up to approx. +4 V when U_{DS} = 15 V). The value of 30% - 40% of U_{DS} should not be exceeded under any circumstances. For this reason, both test circuits provide dropper resistors in the voltage source for gate 2.

Figure 4 shows the I_D/U_{G1S} characteristics of the dual gate MOSFET 3 N 140. The parameter is the gate 2 voltage U_{G2S}. The dependence of the drain current on the two gate voltages can be seen clearly.

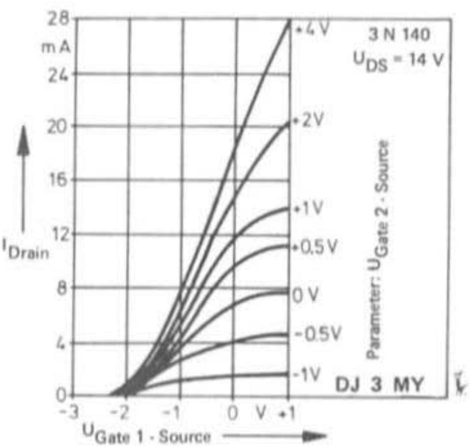


Fig. 4: Characteristic curves of a dual gate MOSFET

When testing the dual gate junction field effect transistors manufactured up till now, it is important that gate 2 is not accidentally driven into the positive voltage range. When testing dual gate MOSFETS, it is only necessary for the maximum permissible voltages and currents, and the resulting power dissipation to be taken into consideration. Of course, the described test circuits are also suitable for FETs with integrated diodes for gate protection (e.g. 40673).

3. CONSTRUCTION

Since only DC voltages and currents are present, the construction of the two FET testers is not critical. All methods from breadboard to printed circuit construction are possible. It is only important to have a clear meter reading and carefully calibrated potentiometers. It is, of course, advisable for a socket to be provided for the transistor under test.

The transformer provided in the second, more elaborate tester only has to provide a power of 1 W. This means that the smallest transistor type is sufficient. The most simple manner is probably to provide two identical windings for 12 V/25 mA (approx. 0.2 mm dia. (32 AWG) enamelled copper wire). The pass transistor BSY 52 can be replaced by type 2 N 1613 (BSY 53) or, in the given circuit by types in a TO-18-casing such as BC 108, 2 N 708, etc. The zener diodes are low power types for 8 V or 15 V. Due to the low voltage and current values, all universal silicon diodes can be used as rectifiers.

AMATEUR TELEVISION

by T. Bittan, G 3 JVQ/DJ ∅ BQ

1. INTRODUCTION

Unfortunately, there are a very large number of different television standards is use throughout the world. This has meant that radio amateurs have also tended to use the TV-standard prevailing in their country.

The different television standards differ in many ways: The most obvious distinctions are the number of lines, which vary from 405 to 819, and the frame frequency which is either 50 Hz or 60 Hz (according to the power line frequency). However, even the video (picture) modulation varies between positive and negative amplitude modulation (AM). The same is valid for the sound signal which can be either AM or FM and be spaced 4.5 MHz, 5.5 MHz, 6 MHz or 6.5 MHz above or below the video carrier. Even the widespread 625 line system has a large number of variants.

When the UHF television commenced in Europe it was decided that the 625 line system was to be used in all European countries. The result could have been a common television standard for UHF. Unfortunately, even this has not been possible and there are now five different 625 line systems operative in Europe.

The most widespread television standard is the CCIR 625 line system used in most European countries.

This system is virtually the European version of the American 525 line system. The difference in the number of lines is mainly due to the difference between the power-line frequency of 60 Hz in the USA and 50 Hz in Europe. The increase in the number of lines entailed an increase in bandwidth which is no doubt the reason, why the channel bandwidth of the CCIR-standard was increased from 6 MHz to 7 MHz and the video-sound carrier spacing from 4.5 MHz to 5.5 MHz.

Since all UHF television transmissions with the exception of America and Japan are already using a 625 line system it is recommended that this standard be adopted as television standard for amateur radio transmissions.

The difference in the video-sound carrier spacings is not too important since TV-amateurs often use a different transmitter, or even a different band, for the sound transmission.

The difference to the 525 line system is also not too important since this is mainly dependent on the power-line frequency and it is not to be expected that amateur television communication will take place between countries using the 525 and 625 line systems.

The CCIR-standard uses negative amplitude modulation for the video signal, which has several advantages over positive AM. The main advantage is the improved synchronization under low-signal conditions.

2. THE TELEVISION SIGNAL

2.1. FORMING THE IMAGE

In order to convert a visual image into an electrical signal it is necessary for the image to be divided into individual points which are scanned in the television camera. In the case of the CCIR-standard described here, the image is scanned in 625 lines.

A frame frequency of 15 Hz would be sufficient for transmission of moving images. However, in order to avoid the unpleasant flicker-effect, 25 complete frames are provided per second (50 Hz). Each frame is divided into two interlaced fields of 312.5 lines each which are scanned one after another. A total of 50 individual fields are scanned each second. The horizontal deflection frequency is 50 x 312.5 = 15.625 Hz.

After the first field of 312.5 lines has been scanned with double line spacing, the second field is scanned between the lines of the first field. Due to the lag of the human eye and the persistence of the picture tube, the two individual, interlaced fields appear as a complete image of 625 lines.

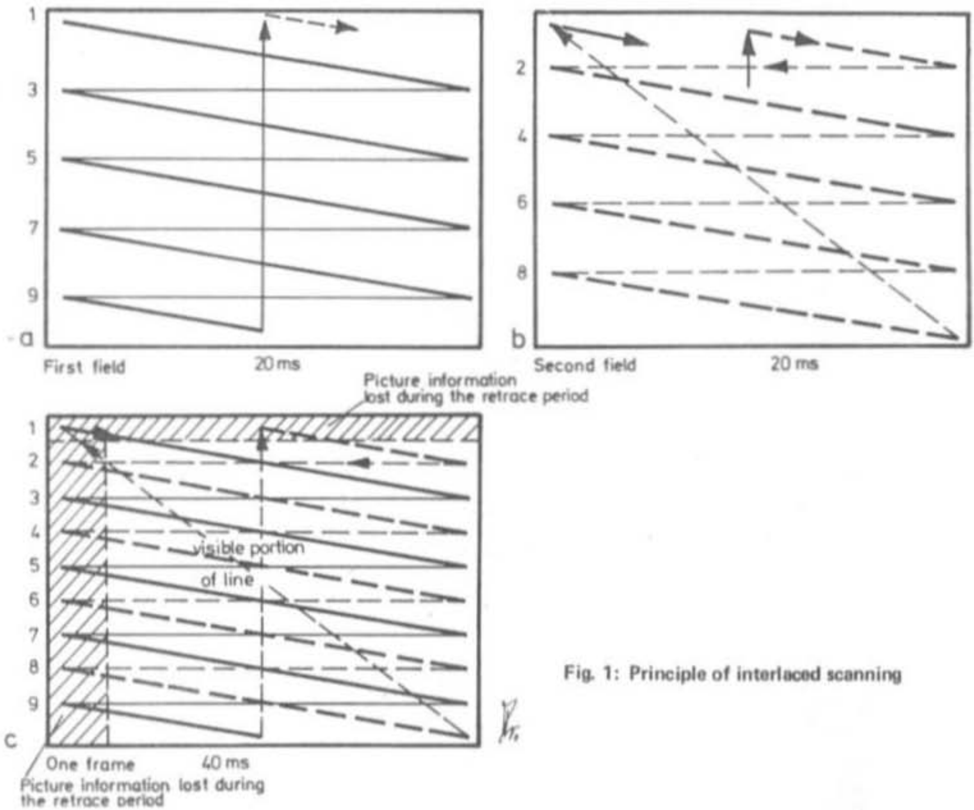


Fig. 1: Principle of interlaced scanning

After the electron beam of the television camera or picture tube has completed the first line, a horizontal retrace pulse is fed from the synchronizing circuits to the deflection coils which deflects the electron beam back to the lefthand side of the image. The horizontal retrace period $t_{retr.}$ amounts to approximately 18% to 22% of that of the normal scan $t_{for.}$. With a total scanning period $t_{hor} = t_{for.} + t_{retr.} = 64\,\mu s$, the horizontal retrace period will amount to 11.5 to 12 μs.

The electron beam is blanked during the retrace period so that it will not be seen on the cathode-ray tube of the TV-receiver. After reaching the lefthand of the image, the electron beam commences the third line, and after the subsequent horizontal retrace, the fifth line, and so on.

The last line of the first field in only deflected up to the centre of the line where the vertical retrace pulse returns the electron beam to the top of the image. The electron beam is also blanked during the vertical retrace period, which amounts to approximately 6% of the vertical deflection period. At a vertical deflection frequency of 50 Hz, the field deflection period is 20 ms which means that the vertical retrace period will amount to approximately 1.2 ms.

The horizontal deflection commences once again as soon as the electron beam reaches the centre of the top edge of the image. As can be seen in Fig. 1c, the lines of the second field are interlaced between the lines of the first field. The time difference between both fields occurs automatically since the vertical retrace pulse appears after 20 ms $\div$ 0.064 ms = 312.5 lines.

In order to maintain the exact position of both fields, it is necessary to have a phase-locked relationship between the horizontal and vertical synchronization. This relationship is obtained using a master timing generator from which both horizontal and vertical synchronizing pulses for the deflection circuits of the TV-camera are derived. These synchronizing pulses are superimposed on the video signal for synchronization of the TV-monitor or TV-receiver.

2.2. BANDWIDTH OF THE TELEVISION SIGNAL

The video bandwidth required for a television signal is dependent on the number of image points that are to be transmitted per second. In order to determine the maximum video frequency occuring during the scanning process, it should be assumed that each line consists of a sequence of white and black square image points.

At an aspect ratio (image width to height) of 4 : 3, 25 frames per second and 625 lines, the maximum video frequency will be 7.5 MHz.

However, the line structure causes a reduction of the vertical resolution. Since the subjective impression of the definition is always determined by the direction of lowest resolution, it is possible to reduce the horizontal resolution without deteriorating the overall impression of the image.

With identical resolution in both the horizontal and vertical directions, it is necessary for a correction factor to be included in the equation for determining the maximum video frequency. This factor is called the Kell factor and amounts to 0.67 with the CCIR 625 line standard. The maximum video frequency to be transmitted is therefore 5 MHz. Since the DC-components of the TV-image must also be transmitted, the total bandwidth is from DC to 5 MHz.

Of course, such high resolution is not necessary for amateur television transmissions and bandwidths of 3.5 or 4 MHz will provide sufficient resolution for most amateur applications. Even bandwidths down to approximately 2.5 MHz can be used.

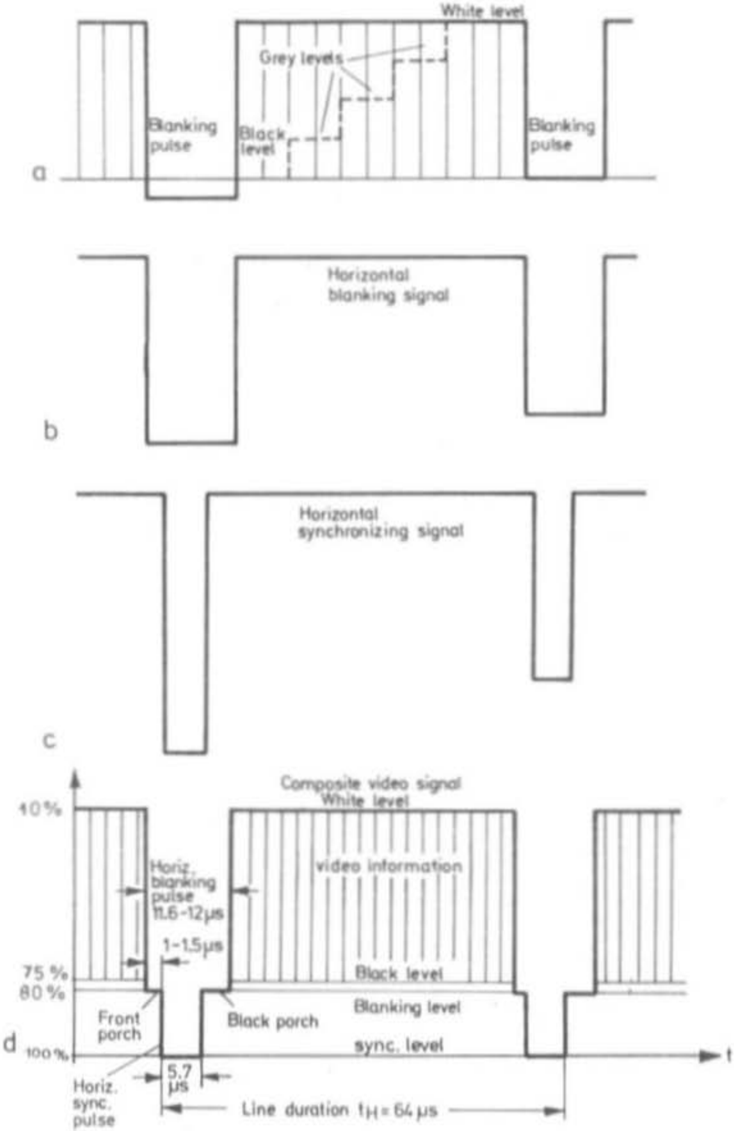


Fig. 2: Oscillogram of the composite video signal over one line showing video information with horizontal blanking and synchronizing pulses

2.3. THE COMPOSITE VIDEO SIGNAL

It has already been explained that the original image was divided into individual image points in the TV-camera. This means that it is necessary to provide some means of synchronizing the scanning of the television receiver to that of the TV-camera. This achieved by superimposing synchronizing pulses on to the video signal. A horizontal retrace pulse is given at the end of each line and a vertical retrace pulse at the end of each field; these are used to synchronize the horizontal and vertical deflection of the TV-receiver. Both pulses are generated in the master timing generator of the TV-camera.

The horizontal and vertical blanking pulses are also superimposed on the video signal. The level of these pulses is such that the electron beam of the TV-camera and cathode-ray tube are blanked during the horizontal and vertical retrace periods.

The actual video information is contained between the blanking pulses (Fig. 2a). The white level corresponds to 10% and the black level to 75%, the total amplitude of the blanking and synchronizing pulses is 100%.

2.3.1. HORIZONTAL BLANKING AND SYNCHRONIZING SIGNAL

These pulses are shown in Fig. 2b and 2c together with the video signal. The duration of the horizontal blanking pulse is 11.5 to 12 μs and the level is 80% of the total amplitude of blanking pulse and synchronizing pulse (100%).
The synchronizing pulse commences 1 μs to 1.5 μs after the blanking pulse and reduces the amplitude of the blanking pulse to between 80% and 100%. The duration of the horizontal synchronizing pulse is 5.7 μs.

The blanking period before and after the synchronizing pulse is called the front and back porch. The black level of the TV-signal corresponds to 75%.

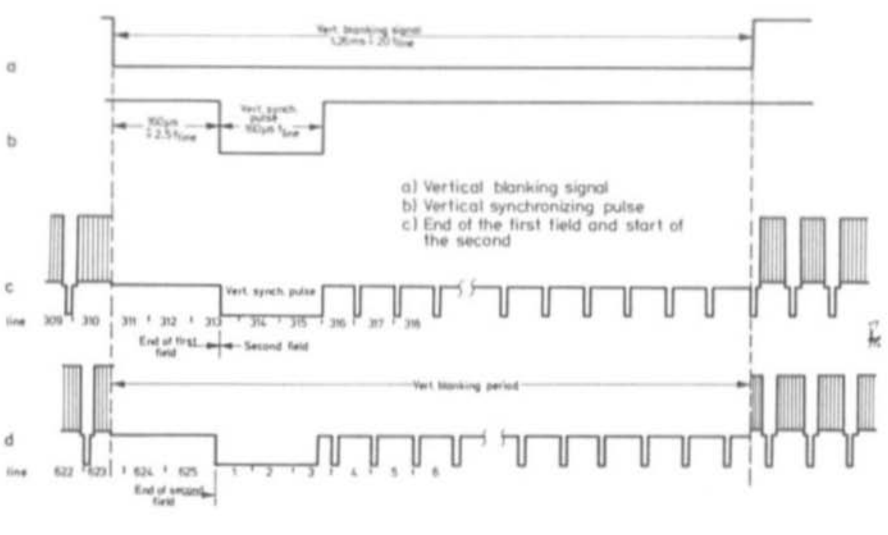


Fig. 3: Oscillogram of the composite video signal over one field showing the vertical blanking and synchronizing pulses

2.3.2. VERTICAL BLANKING AND SYNCHRONIZING SIGNAL

A simplified vertical synchronizing signal such as used in closed-circuit television networks is shown in Fig. 3. The complete vertical synchronizing signal comprises the vertical blanking pulse with a period of 18 to 22 times the line duration (approx. 1.25 ms), and the vertical synchronizing pulse with a duration of 160 μs. The horizontal synchronizing pulses are blanked previous to the commencement of the vertical blanking pulse until after completion of the vertical synchronizing pulse. The video signal is blanked during the whole period of the blanking signal.

The CCIR television standard uses equalizing pulses before and after the synchronizing pulse and also divides the synchronizing pulse into five individual pulses of half the line duration for maintaining the horizontal synchronization during the vertical retrace period. These additional pulses are often deleted in closed-circuit systems.

2.3.3. VIDEO SIGNAL

The video information is transmitted between the blanking pulses (Fig. 2a and 2d). The amplitude can vary between the black level at 75% and the white level at 10% of the 100% level during the synchronizing pulses. The various grey levels of the monochrome image represent intermediate values between 75% and 10%.

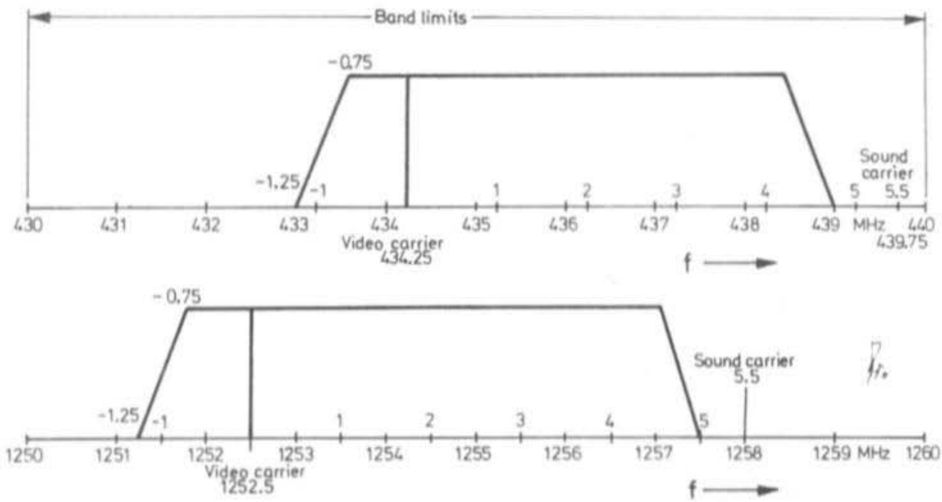


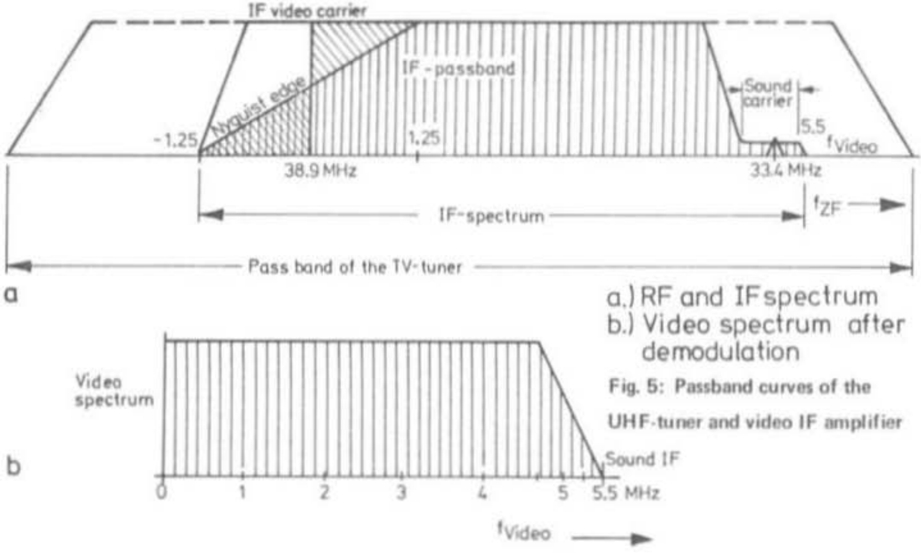
Fig. 4: Spectra of an amateur television signal and sound carrier on the 70 cm and 24 cm bands

3. TRANSMISSION CHANNELS

The channels for amateur television transmissions on the 70 cm and 24 cm bands are given in Fig. 4. On the 70 cm band, the video carrier is at a frequency of 434.25 MHz and the sound carrier at 439.75 MHz; whereas the video carrier is at 1252.5 MHz and the sound carrier is at 1258 MHz on the 24 cm band.

3.1. VIDEO TRANSMISSION RANGE

In order to accommodate the greatest number of television stations in a given frequency range, television transmitters operate according to the vestigial sideband technique, which is basically the same principle as that of single sideband transmissions. The lower sideband cannot be suppressed completely due to the relatively wide bandwidth required, but is suppressed further in the intermediate frequency amplifier of the receiver. The television transmitter therefore radiates the upper sideband and video carrier at full amplitude, whereas the lower sideband is transmitted down to 0.75 MHz below the video carrier after which it is suppressed.



a.) RF and IF spectrum
b.) Video spectrum after demodulation

Fig. 5: Passband curves of the UHF-tuner and video IF amplifier

In order to do this, the response of the IF-passband curve of the receiver is aligned as shown in Fig. 5. Assuming an intermediate frequency with the video carrier at 38.9 MHz, the video carrier is located half way up the skirt of the passband curve so that the addition of the lower frequency components of the lower sideband during the detection process is compensated for due to the reduction of the low-frequency components of the upper sideband. At modulation frequencies greater than 1.25 MHz, only the upper sideband will be operative. This means that a constant amplitude is obtained over the video band (Fig. 3b).

3.2. SOUND CHANNEL

It is favourable for a sound channel to be radiated in addition to the video signal. In order to be able to use a conventional television receiver in conjunction with a frequency converter for reception of amateur television transmissions, the sound carrier should be spaced 5.5 MHz above the video carrier.

Of course, the sound channel can be transmitted using a separate transmitter. However, it can also be modulated onto the video carrier using the intercarrier method. In order to ensure that no interference is heard on the sound channel, the amplitude of the video carrier must not be allowed to exceed the 10% white level. The amplitude of the sound carrier is reduced to approximately 5% the video carrier in the IF-amplifier (Fig. 5b).

- to be continued -

MATERIAL PRICE LIST OF EQUIPMENT

described in Edition 3/72 of VHF COMMUNICATIONS

70-cm-POWER AMPLIFIER FOR 2 C 39			Ed. 3/72
Silver-plated, ready-to-operate but without tube, power supply and fan			DM 398.--

DC 6 HL 007	FM-ATTACHMENT		Ed. 3/72
PC-board	DC 6 HL 007	(with printed plan)	DM 10.--
Semiconductors	DC 6 HL 007	(8 transistors, 1 IC, 2 diodes)	DM 27.10
Minikit 1	DC 6 HL 007	(6 IF filters, 1 coilformer with core, 1 potted core set, 1 pot.)	DM 22.40
Minikit 2	DC 6 HL 007	(5 styroflex, 13 ceramic, 4 tantalium caps.)	DM 19.--
Ceramic filter	CFS-455 B		DM 75.50
Kit	DC 6 HL 007	with above parts	DM 152.--

DC 6 HL 008	9 MHz FM OSCILLATOR		Ed. 3/72
PC-board	DC 6 HL 008	(double-coated)	DM 7.--
Minikit	DC 6 HL 008	(2 transistors, 1 coilformer with core, 5 styroflex, 2 ceramic and 2 feedthrough capacitors)	DM 15.70
Kit	DC 6 HL 008	with above parts	DM 22.--

OK 2 KT 001	28-30 MHz DIPLEXER AMPLIFIER		Ed. 3/72
PC-board	OH 2 KT 001	(with printed plan)	DM 10.--
Minikit	OH 2 KT 001	(2 coilformers with core and screening cans, 1 ferrite toroid)	DM 2.15
Semiconductors	OH 2 KT 001	(3 transistors, 1 IC)	DM 14.80
Kit	OH 2 KT 001	with above parts	DM 26.80

DL 6 HA 001/28	SATELLITE CONVERTER		Ed. 3/72
PC-board	DL 6 HA 001	(with printed plan)	DM 6.--
Minikit	DL 6 HA 001	(4 coreformers with core, 5 trimmers)	DM 5.20
Semiconductors	DL 6 HA 001	(5 transistors)	DM 20.50
Crystal	38.6667 MHz	(HC-6/U)	DM 13.70
Kit	DL 6 HA 001	with above parts	DM 45.40

DJ 4 BG 009	TRIPLE CRYSTAL OSCILLATOR MODULE		Ed. 3/72
PC-board	DJ 4 BG 009	(with printed plan)	DM 9.--
Minikit	DJ 4 BG 009	(1 coilformer with core, 3 trimmers, 3 crystal holders, 13-pole connectors and TEKO case 3 A)	DM 25.80
Semiconductors	DJ 4 BG 009	(4 transistors)	DM 23.20
Kit	DJ 4 BG 009	with above parts	DM 56.00

SPECIAL OFFER

We have a number of 2 metre tube transmitters AT 201 manufactured by STE in stock. These units provide a carrier output power of 12 W. Crystals required are 8 or 12 MHz types which makes the transmitter suitable for FM transmissions if a suitable phase-modulator is added. A matching amplitude modulator AA 12 is also available for AM transmissions. These are not shop soiled items but completely new units.

Prices:			
	AT 201 Transmitter	(without tubes)	DM 55.--
	AA 12 AM Modulator	(without tubes)	DM 40.--
	Modulating transformer		DM 19.--

Please add an extra DM 5.-- for post and packing. Full details and circuit available on request.

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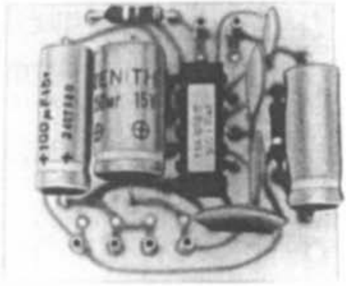
Unfortunately, there have been two increases in the postal rates this year and we are afraid that we must also be forced to raise our charges for post and packing. The new rates are DM 3.00 per parcel up to a value of DM 80.00 and DM 4.00 when registered post is used for higher valued shipments. If express shipment is required, this can be provided for an extra DM 2.00 on the normal charge.

AD 4



AA 1

FM Limiter and Discriminator AD 4: DM 30.35

FM discriminator for the AR 10 and other receivers with an intermediate frequency of 455 kHz. Advantage over FM demodulation using the IF-slope: the receiver need not be tuned away from the signal, and ignition interference is suppressed by the limiter (AM-suppression: 40 dB, limiter threshold: 100 µV).

Audio Amplifier AA 1: DM 29.70

Miniature integrated AF-amplifier with an output power of 1.5 W at 12 V. Ideal for many applications.

Transistor transmitter AT 210: DM 152,90

4-stage crystal-controlled transmitter with modulating transformer and antenna relay. Especially suitable for portable and mobile operation from 12 V. Crystals in 72 MHz range (HC-25/U). Connection available for a 24 MHz VFO.

Specifications:

Frequency range:	144—146 MHz	Operating voltage:	12 V/400 mA (max. 15 W)
Output power (unmodulated carrier):	2.2 W at 12 V	Dimensions:	150 mm x 46 mm x 32 mm

Modulator and AF-amplifier AA3: DM 82,50

4-stage modulator and audio amplifier matching the transmitter AT 210 or receiver AR 10. Built-in relay switches the modulator so that it can be used as audio output for the receiver. Output power more than sufficient for mobile operation.

Specifications:

Output power:	2.8 W at 12 W
Output impedance:	3 Ohm
Sensitivity:	2 mV for 2.8 W
Frequency response:	300—3000 Hz (—3 dB)
Distortion:	< 2 % at 2.8 W/1000 Hz
Dimensions:	150 mm x 46 mm x 32 mm

The above mentioned modules can be combined to form a complete two metre transceiver. They are available ex stock in Erlangen. Each module is accompanied by a description including circuit diagram, list of components and connection details. These modules can also be purchased from the representatives of VHF COMMUNICATIONS in the following countries: Denmark & Sweden, Norway, Spain & Portugal, South Africa. Other countries please inquire directly to

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CRYSTAL FILTERS - FILTER CRYSTALS - OSCILLATOR CRYSTALS

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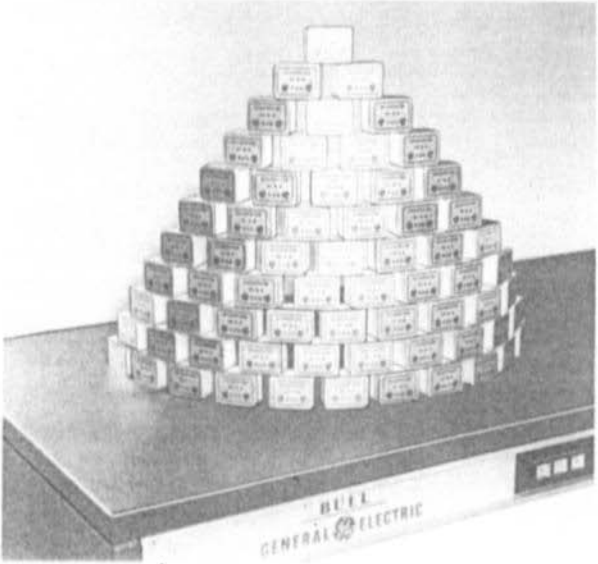
PIEZO-ELECTRIC PRESSURE TRANSDUCERS

Listed is our well-known series of

9 MHz crystal filters for SSB, AM, FM and CW applications.

In order to simplify matching, the input and output of the filters comprise tuned differential transformers with galvanic connection to the casing.



Filter Type		XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9M
Application		SSB-Transmit.	SSB	AM	AM	FM	CW
Number of Filter Crystals		5	8	8	8	8	4
Bandwidth (6dB down)		2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz
Passband Ripple		< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 1 dB
Insertion Loss		< 3 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 3 dB	< 5 dB
Input-Output	Z_t	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω	500 Ω
Termination	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
Shape Factor		(6:50 dB) 1.7	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6: 0 dB) 2.5
			(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:60 dB) 4.4
Ultimate Attenuation		> 45 dB	> 100 dB	> 100 dB	> 100 dB	>90 dB	> 90 dB

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