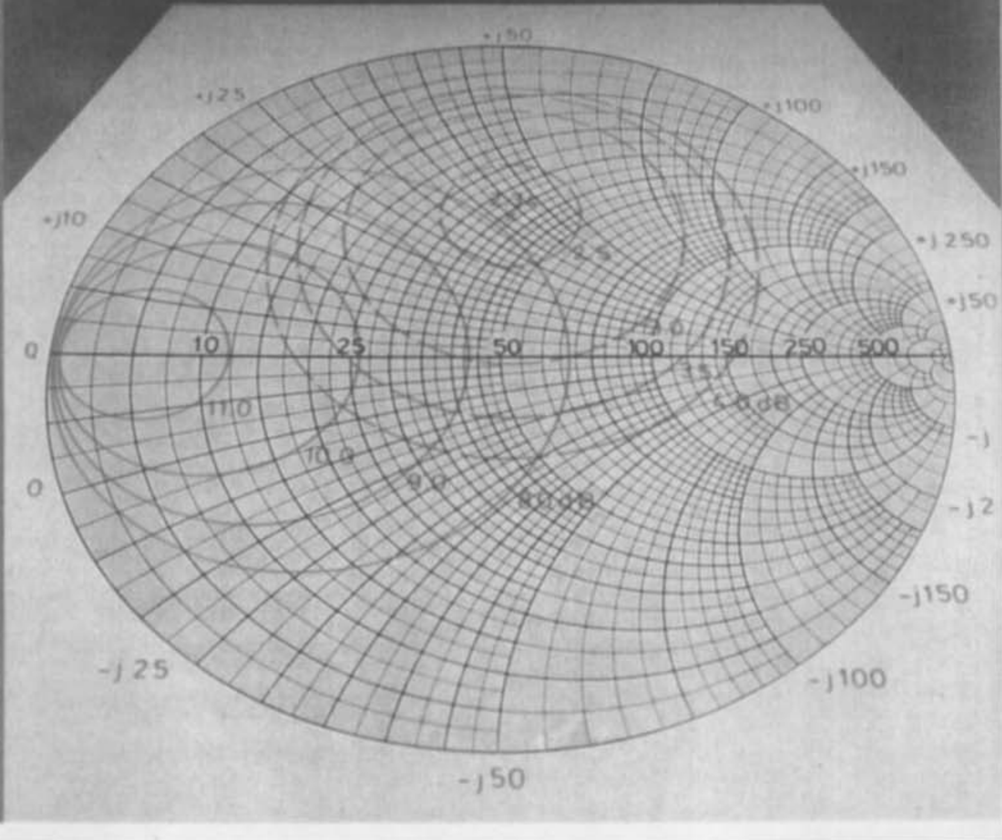
J 20 419 F

A Publication
for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 12 · Spring · 1/1980 · DM 5.00



As you will see, we are now printing VHF COMMUNICATIONS in a new look with two columns. This allows us to increase the amount of text that can be accommodated in the magazine to some degree. We hope that you will enjoy reading VHF COMMUNICATIONS, and do not forget to recommend the magazine to your friends.

Published by: Verlag UKW-BERICHTE, Hans J. Dohlus oHG, Jahnstrasse 14, D-8523 BAIERSDORF, Fed. Rep. of Germany, Telephones (0 91 33) 855, 856.

Publishers: T. Bittan, H. Dohlus

Editors: Terry D. Bittan, G 3 JVQ/DJ 0 BQ, responsible for the text

Robert E. Lentz, DL 3 WR, responsible for the technical contents

Advertising manager: T. Bittan

VHF COMMUNICATIONS the international edition of the German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is published in Spring, Summer, Autumn, and Winter. The subscription price is DM 18.00 or national equivalent per year. Individual copies are available at DM 5.00 or equivalent, each. Subscriptions, orders of individual copies, purchase of PC-boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative.

Printed in the Fed. Rep. of Germany by R. Reichenbach KG. Krelingstr. 39 · 8500 Nuernberg.

We would be grateful if you would address your orders and queries to your representative.

VERTRETUNGEN · REPRESENTATIVES :

Austria
Hans J. Dohlus, Creditanstalt Bankverein WIEN
Kto.17-90.599, PSchKto. WIEN 1.169.146

Australia
WIA PO Box 150, TOORAK, VIC. 3142, Tel.24-8652

Belgium
Stereohouse, Brusselsesteenweg 416, B-9218 GENT,
PCR 000-1014257 CCP, Tel. (091) 31 21 11

Denmark
E.Halskov, OZ 7 LX, Sigersted Gl. Skole,
DK-4100 RINGSTED, Postgiro 7 296 800, Tel. 03-616162

France
Christiane Michel, F 5 SM, F-89 PARLY, Les Pillés

Finland
Timo Nopanen, OH 3 YW, Metsäter 10, SF-15610 LAHTI

Germany
Verlag UKW-BERICHTE, H.Dohlus oHG, Jahnstr. 14,
D-8523 BAIERSDORF, Tel. (09133) 855, 856.
Konten: Postscheckkonto Nürnberg 304 55-858,
Commerzbank Erlangen 820-1154

Holland
MECOM, PA 0 AER, PO Box 40, Coendersstraat 24,
NL-9780 AA BEDUM, Tel.05900-4390, Postgiro 39 86 163

Israel
Z. Pomer, 4 X 4 KT, PO Box 222, K. MOZKIN
26 100, Tel. 974-4-714078

Italy
Franco Armenghi, I 4 LCK, Via Sigonio 2,
I-40137 BOLOGNA, Tel. (051) 34 56 97

Luxembourg
P. Wantz, LX 1 CW, Télévision, DUDELANGE,
Postscheckkonto 170 05

New Zealand
E. M. Zimmermann, ZL 1 AGQ, PO Box 85,
WELLSFORD, Tel. 8025

Norway
Henning Theg, LA 4 YG, Postboks 70,
N-1324 LYSAKER, Postgirokonto 3 16 00 09

South Africa
SA Publications, PO Box 2232, JOHANNES-
BURG 2000, Telephone 22-1496

Spain + Portugal
Julio A. Prieto Alonso, EA 4 CJ, MADRID-15,
Donoso Cortès 58 5°-B, Tel. 243.83.84

Sweden
POLY RADIO, Östervärnsgatan 10,
S-21218 MALMÖ, Bo Nilsson, SM 7 FJE,
Tel. 040-292420, Pg: 431 56 99-1, Bg: 317-2939

Switzerland
Hans J. Dohlus, Schweiz. Kreditanstalt ZÜRICH,
Kto. 469.253-41; PSchKto. ZÜRICH 80-54.849

UK North
SOTA Communication Systems Ltd., 26 Childwall
Lane, Bowring Park, LIVERPOOL L 14 6 TX

UK South
VHF COMMUNICATIONS, Dept. 802, 20 Wallington
Square, WALLINGTON Surrey SM 6 8 RG

Yugoslavia
Tito Cvrković, YU-56000 VINKOVCI, M. Gupca 27

VHF communications

Volume No. 12
Spring / Ed. 1
1980

A Publication for the Radio Amateur
Especially Covering VHF, UHF, and Microwaves

Two-Stage Low-Noise Preamplifiers for the Amateur Bands from 24 cm to 12 cm

by J. Grimm, DJ 6 PI

Microwave converters for the amateur bands in excess of 70 cm usually operate with diode mixers such as types 1 N 21 – 1 N 23 or hp 2800 – hp 2817 etc. According to construction and extent of the circuit (image frequency trap !), the noise figures will be in the order of 8 to 16 dB. VHF COMMUNICATIONS has published preamplifiers in the past equipped with transistors AF 139, AF 279, BFR 34, which exhibited noise figures in the order of 5 dB. Bipolar SHF-stripline transistors are now available that allow noise figures in the order of 2 dB to be realized. Such transistors are manufactured by NEC, HP, Microwave Associates, Motorola and some others, and are available at prices that are still acceptable to the radio amateur. The following article is to describe two-stage amplifiers equipped with NEC-transistors of type NE 645 35 and NE 578 35.

Such low-noise preamplifiers increase the signal-to-noise ratio by the same factor as when increasing the output power by four to ten times. This increase of transmit power would cost many times that of such a preamplifier.

These preamplifiers represent a further development of similar preamplifiers described in (1), and (2). The amplifiers are constructed in stripline technology on a printed circuit board since this is far easier to calculate and construct than coaxial amplifiers using cavities. The losses in the PC-board material are only in the order of a fraction of a dB and are therefore most certainly acceptable when considering the simple and reliable construction that results from this.

For those readers not interested in the theory and calculation of such amplifiers, it is now possible for them to jump the following sections and to continue with the practical information that commences with Section 4.

1. CALCULATION OF THE MATCHING NETWORKS

1.1. Transistor Impedances

In order to calculate the matching networks, one must know the input and output impedances of the transistor at the required frequency, and for the various operating points. The manufacturers provide the complex s-parameters of the transistors for this purpose. Since the determination of the transistor impedances from the s-parameters has not been described in detail in amateur literature, this is to be discussed in more detail here.

Two methods of calculation are to be used in the following. All amplifier versions have been calculated according to these methods, and subsequently constructed. The differences in the measured values are given in the table in Section 5.

The s-parameters are determined by the manufacturer by terminating the input or output of the transistor with 50 Ω. The parameters s_{11} and s_{22} are the input and output reflection coefficients for this mode. Parameters s_{12} and s_{21} are also a measure how great an impedance variation at the input will have an effect on the output impedance, and vice versa.

Table 1:
An example of the s-parameters of transistor NE 578 35

U_{CE} = 8 V, I_C = 10 mA.
1st Column: Amount, 2nd Column: Angle

Frequency	s_{11}		s_{21}		s_{12}		s_{22}	
100 MHz	0,65	− 32	19,20	152	0,02	72	0,90	− 15
200 MHz	0,55	− 63	15,20	138	0,03	63	0,78	− 24
500 MHz	0,40	−127	9,80	102	0,05	52	0,50	− 38
1 GHz	0,40	−160	7,70	82	0,07	52	0,35	− 33
2 GHz	0,50	158	3,00	57	0,11	52	0,24	− 60
4 GHz	0,63	128	1,34	18	0,16	33	0,22	−128

The s-parameters are given in the data sheets for various frequencies and operating points in the form of a table (and sometimes also in the form of Smith diagrams); since these are complex magnitudes (vectors), they are given with amount and angle.

1.1.1. Simplified Calculation Method

In practice, the input and output of the transistor is not terminated with 50 Ω. If a slight mismatch is acceptable, one can save a considerable amount of calculation when one only uses the input and output reflection coefficients s_{11} and s_{22} for the calculation of the transistor input and output impedance.

The forward and return transfer coefficients s_{21} and s_{12} are not taken into consideration.

The impedances are calculated from the reflection coefficients using the following equation:

$$Z_{in} = Z_0 \times \frac{1 + s_{11}}{1 - s_{11}} \qquad (1)$$

where Z_0 = 50 Ω.

Since the reflection coefficients are complex magnitudes, the real and reactive impedance is calculated using the following equation:

$$Z_{in} = \frac{(1 - |s_{11}|^2) \times 50}{1 + |s_{11}|^2 - 2|s_{11}| \times \cos \sphericalangle s_{11}} + j\,\frac{(2|s_{11}| \times \sin \sphericalangle s_{11}) \times 50}{1 + |s_{11}|^2 - 2|s_{11}| \times \cos \sphericalangle s_{11}} \qquad (2)$$

The output impedance Z_{out} is also calculated in the same manner from parameter s_{22}, or from $|s_{22}|$ and $\sphericalangle s_{22}$. Example from table 1:

Z_{in} at 2 GHz = 17.2 + j 8.6 Ω

Z_{out} at 2 GHz = 57.6 − j 25 Ω

For those readers that do not wish to calculate the input and output impedances from this equation, it is possible for them to be taken in an approximate manner direct from the Smith diagram. Unfortunately, these diagrams are usually so small in the data sheets so that it is only possible to extract the impedances with a certain degree of inaccuracy. It is easier, and more accurate, when the tabular values are firstly transferred into a large Smith diagram. It can then be of any required size. This is achieved by firstly drawing the angles as lines from the center point of the diagram. The positive angles are to be found in the upper, the negative angles in the lower half of the diagram. The lengths of the lines result from the radius of the Smith diagram used, multiplied by the magnitude factor from the table.

Coefficient data is referred to an impedance of 50 Ω, which means that the values taken from a standardized diagram should be multiplied by 50 Ω.

Figure 1 shows an example of a Smith diagram in which all values have already been multiplied by 50 Ω. It contains the s_{11} and s_{22} values for a common emitter circuit which has been taken from Table 1. The

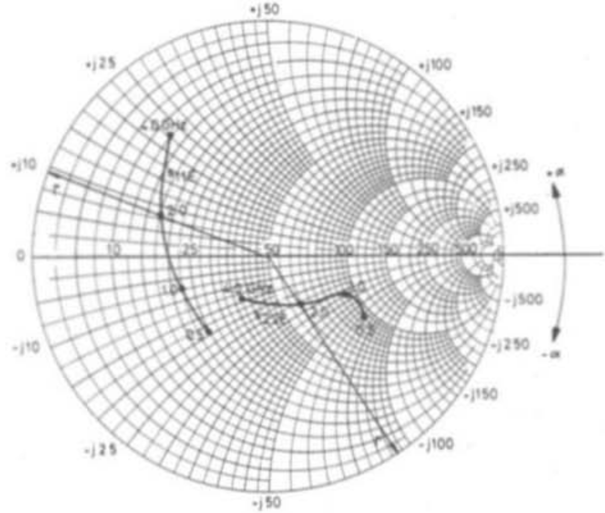


Fig. 1:
Smith diagram with the input and output reflection coefficients of transistor NE 578 35 at U_{CE} = 8 V and I_C = 10 mA

lines have been drawn for a frequency of 2 GHz: s_{11} = 158° and 0.5 of the radius of the diagram, and s_{22} = −60° and 0.24 for the radius.

The following impedances can be read off:
Z_{in} approx. 17 + j 9 Ω;
Z_{out} approx. 60 − j 26 Ω.

1.1.2. Exact Method of Calculation

This method takes the reaction within the transistor and the actual terminating impedances of the transistor input and output into consideration. One requires the four s-parameters of the transistor at the required frequency, and the required operating point for calculation. This can be obtained with the aid of an s-parameter meter such as the hp 8410, which is hardly accessible to radio amateurs. However, the previously mentioned tables are available from the semiconductor manufacturers in which the four s-parameters are given for the various operating points in a frequency spacing of 200 MHz. The values for any required frequency can be obtained by inter or extrapolation.

The calculation of the corrected coefficients s_{11}^* and s_{22}^* for simultaneous matching of the transistor at the input and output (and thus the corrected impedances) would, however, be too extensive for this article. A calculation program for the TI 59 was developed, and further details regarding this are to be given in Section 6.

2. DESIGN OF THE MATCHING NETWORKS

The PC-board material must guarantee low line losses at the required frequency and a high Q of the resonant circuit. Epoxy glassfibre boards can be used up to 1.3 GHz; at higher frequencies, PTFE glassfibre boards should be used. This article is not to discuss the width and length of stripline circuits here, since this has been described several times in (3) to (6). In our case, the striplines were calculated using a TI 59 calculation program developed by the author.

2.1. Matching Circuits

Optimum gain characteristics can only be achieved when the transistor is correctly matched. In the case of a complex impedance, matching conditions will be present when terminated with the conjugated complex value. In this case, the real impedance of the termination has the same value of that of the source; the reactive impedance of the termination should have the same value, but of opposite sign as that of the source **(Figure 2)**. This means that an inductance is compensated for with a capacitance of the same reactive impedance, and vice versa.

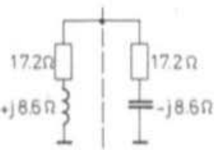


Fig. 2: Matching of the NE 578 35 input at 2 GHz

The matching of the reactive impedance can often be made directly at the transistor. In some cases, however, it must be carried out with the aid of a transformation link. This is to be discussed in detail. Since the real impedance of the transistor output will not coincide with the input of the following transistor, it is necessary for a transformation link to be provided. This is also the case between amplifier input or output and transistor input or output respectively.

There are several different methods of transformation possible. However, due to the simple calculation, only the λ/4 transformation is to be used here. This means that matching is only exhibited within a certain bandwidth on both sides of the required frequency.

2.1.1. Reactive Impedance Matching

If the reactive impedance component of the transistor impedance is positive, in other words inductive, it is possible for the matching to be made directly at the transistor with the aid of a capacitance. Due to the possible spread of the transistor specifications, a trimmer should be used, and this should be a ceramic tubular trimmer with a low minimum capacitance for UHF and SHF applications. A parallel circuit to ground is more suitable for alignment than a series circuit. For this reason, the transistor series impedance $Z_S = R_S \pm jX_S$ must be converted to an equivalent parallel impedance in order to calculate the required capacitance value. This is made with the aid of the following equations:

$$R_P = R_S + \frac{X_S^2}{R_S} \qquad (3a)$$

$$X_P = \frac{R_S \times R_P}{\pm X_S} \qquad (3b)$$

$$Z_P = R_P \pm jX_P \qquad (3c)$$

Example:
$Z_{P\,in}$ of the NE 578 35 at 2 GHz is 21.5 || + j 43 Ω. The required parallel capacitance is calculated according to equation 4 as follows:

$$C/pF = \frac{1}{2\pi f/MHz \times 10^6 \times X_P \times 10^{-12}} \qquad (4)$$

C = 1,85 pF für 2 GHz.

The whole transformation process is given in **Figure 3**.

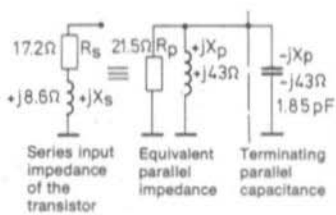


Fig. 3: Termination of the input reactive impedance of the NE 578 35 at 2 GHz

If the reactive impedance component of the transistor impedance is negative (capacitive), it is possible for matching to be made using an inductance directly at the transistor. Unfortunately variable inductances cannot be constructed in the SHF-range, which means that it is advisable to carry out a λ/4 transformation of the reactive impedance. The reactive impedance is transformed to an opposite value in a λ/4 line, which means that an inductance will be transformed into capacitance.

The transformed value can be calculated according to equation 5:

$$X_2 = \frac{-Z_0^2}{X_1} \qquad (5)$$

where
Z_0 = Impedance of the transformation line.

If the transformation line is in series with the transistor impedance, the series-reactive impedance of the transistor is inserted in equation 5 for X_1. **Figure 4** shows this transformation process in the form of a circuit diagram.

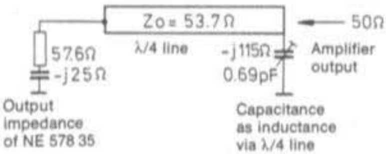


Fig. 4: Matching the output impedance of the NE 578 35 at 2 GHz and U_{CE} = 8 V / I_C = 10 mA into 50 Ω

One exception during the reactive impedance matching using variable capacitors is the output of the first amplifier stage. The reactive component of the output impedance of the transistors used is capacitive. Since this impedance must be transformed to the input impedance of the second stage, a fixed inductance is used at the collector of the first stage in order to provide a simple transformation path.

According to equation 6, an inductance can be realized by using a stripline circuit that is shorted at the end for radio frequencies. The impedance can be selected as required. Since the width of the stripline is mainly dependent on the impedance, a compromise must be made during the selection of the impedance:

The compensation line should be as narrow as possible in order not to have an effect on the λ/4 transformation line. On the other hand, in order to maintain the required impedance, it must be wide enough that the width can be accurately realized within ± 0.5 mm in the design drawing. In practice, a well-proved compromise is for the width of the compensating line to be w ≦ the width of the λ/4 transformation line.

In professional applications, such stubs are usually made in the form of two partial stubs in order to have as little effect as possible on the transformation line. This method was not used in the described amplifiers.

The following equation provides information for calculating a stripline as an inductance (end short-circuited !):

$$X_L = Z_0 \tan \Theta$$
$$[\,\Theta < 90° \triangleq \lambda/4\,] \qquad (6)$$
$$\Theta\,[\,l/\lambda_0\,] = \frac{\arctan X_L/Z_0}{360}$$
$$[\,0, \text{ to } \lambda_0\,]$$

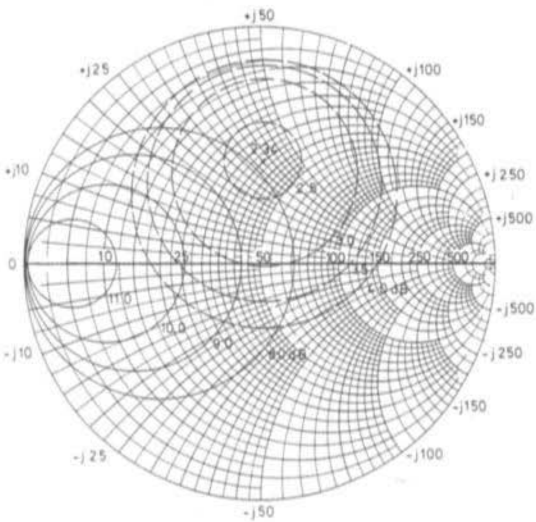


Fig. 5: Noise figure and gain curves of the NE 578 35 at 2 GHz and U_{CE} = 8 V and I_C = 3 mA

2.1.2. Real Impedance Matching

Only the real components of the complex impedances are used for calculation of the impedance of the λ/4 transformation line. The impedance results as given in equation 7 :

$$Z_0 = \sqrt{Z_1 \times Z_2} \qquad (7)$$

If a reactive element is connected in parallel with the transistor, it is necessary for the equivalent real, parallel impedance to be inserted as Z_1 or Z_2.

2.2. Noise Matching of the Input Stage

The lowest noise figure of an amplifier stage will not coincide with its optimum (power) matching, but with a certain degree of mismatch. The complex source-reflection coefficient Γ_0 is given by the manufacturers for alignment of an amplifier stage for minimum noise figure. The required source impedance can be obtained from this value with the aid of equation 2 or from the Smith diagram. The values for Γ_0 are not given in tabular form, but in the form of a Smith diagram.

Figure 5 shows such a diagram with circular contours. The gain curves (continuous lines) intersect the noise figure curves (with dashed lines). The given noise figure can be obtained within a noise figure curve. The matching gain value results from the dB-value of the intersected gain curve. For example: NF = 4 dB at G_p = 10 to 11 dB, or NF = 2.34 dB at G_p = 8 to 9 dB. It is thus possible to determine whether exaggerated values are given by the manufacturers of preamplifiers and converters by determining whether the given noise figure and gain values match those given for the transistor used.

This complicated method is only advisable when an exact calculation of the interactions within the transistor is to be taken into consideration. This method of calculation is part of the previously mentioned TI 59-program. The impedance obtained from s_{11} is used for a simplified noise matching.

In order to align the amplifier for minimum noise figure by measurement or by ear, the input network should be in the form of a variable Pi-filter that can be varied within wide limits. A Pi-filter is also advisable when no accurate s-parameters are available for the exact calculation, but only inter or extrapolated values. This is the case, for instance, for the 2.3 GHz version to be described later.

3. FURTHER CIRCUIT DETAILS

3.1. Operating Voltage Supply

The base and collector voltages must be fed to the transistors in such a manner that they do not affect the matching networks. This is achieved using extremely narrow λ/4 lines that are short-circuited for SHF at the ends. This short circuit is transformed by the λ/4 line to an infinite value so that it does not interfere with the matching lines at the transistor connections.

An impedance of 150 Ω was selected for the voltage supply lines. This means that they are only 0.1 to 0.4 mm wide.

The SHF-short circuit at the end of the lines is also achieved with the aid of λ/4 lines that are connected to the narrow voltage supply lines and are open at the other end. An impedance of 25 Ω was selected for these lines. Due to their relatively large area, these lines also represent capacitors and form an additional short circuit for such high frequencies.

The voltage is fed to the intersection of both lines via a small resistor.

3.2. Number of Amplifier Stages

According to well-known formulas, the overall noise figure of the receive system will have a low value if the noise figure of the preamplifier is considerably lower than that of the mixer, and when the preamplifier also possesses a high gain. For example, for a preamplifier with a noise figure of 3 dB and a gain of 10 dB in front of a mixer with a noise figure of 12 dB (for example a hybrid ring mixer), an overall noise figure of 5.4 dB would result. On the other hand, a preamplifier with the same noise figure but 20 dB gain in front of the same mixer would provide an overall noise figure of approximately 3.3 dB.

Since the gain of a transistor stage aligned for minimum noise figure is only listed as being 8 to 11 dB in the data sheets, it is necessary to use two amplifier stages in order to ensure that the low noise figure of the transistor is utilized to the full in the receive system. In this case, the first stage is aligned for minimum noise, and the second stage for maximum gain.

The large-signal behaviour has not been taken into consideration, and this is still not important on the SHF-amateur bands.

3.3. Operating Points

The operating points (U_{CE}, I_C) for minimum noise and maximum gain are given in the data sheets of the SHF-transistors. These voltages must be stabilized with respect to temperature and voltage fluctuations. The circuit given in (7) and (8) has been found to be very favorable for this and is used in the described amplifiers.

The stabilization circuit uses a voltage feedback to the base in conjunction with a base-constant current source. A voltage stabilization using a zener diode is used for stabilizing voltage fluctuations that can occur especially during portable operation. **Figure 6** shows the circuit diagram of this DC-circuit.

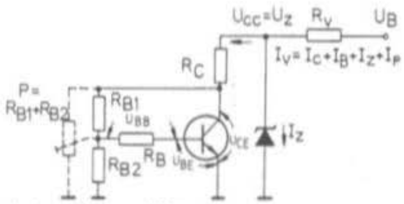


Fig. 6: Operating point adjustment and stabilization

The zener voltage should be approximately 1 V greater than the given collector-emitter voltage. The value of the dropper resistor R_d is selected so that twice the collector current flows through R_d at the nominal operating voltage of 12 V.

Example:

U_{CE} = 8 V; U_Z = 9.1 V; I_C = 10 mA;

U_B = 12 V.

$R_d = \frac{2.9\ V}{0.02\ A} = 145\ \Omega$ (stand. value 150 Ω)

The fundamental considerations in the previously mentioned articles are not to be repeated here. However, the equations for Figure 6 are to be given:

$$R_B = \frac{h_{FE}\,(U_{BB} - U_{BE})}{I_C} \quad (8)$$

$$R_{B2} = \frac{h_{FE} \times U_{BB}}{5\,I_C} \quad (9)$$

$$R_{B1} = \frac{h_{FE}\,(U_{CE} - U_{BB})}{6\,I_C} \quad (10)$$

$$R_C = \frac{h_{FE}\,(U_{CC} - U_{CE})}{(h_{FE} + 6) \times I_C} \quad (11)$$

$U_{CC} = U_Z$;

U_{CE} and I_C from data sheet;

U_{BE} = 0.7 V for silicon transistors;

U_{BB} = 2 V for silicon transistors;

h_{FE} from the data sheet (assume h_{FE} = 100 if not listed).

For alignment of the operating point, the individual resistors R_{B1} and R_{B2} are replaced by a single trimmer potentiometer.

This temperature stabilization operates very satisfactorily: The collector current only varies by approximately ± 1 mA from the value selected at + 20°C throughout the temperature range of − 20°C and + 60°C !

The operating voltage may drop down to 10.5 V without altering the collector current. The voltage stabilization is especially advantageous since the maximum permissible collector-emitter voltage of the given NEC transistors only amounts to 11 and 12 V, respectively.

4. CONSTRUCTION OF THE AMPLIFIER

For the previously mentioned reasons, two-stage amplifiers were constructed for the 23 cm and 13 cm band equipped with tran-

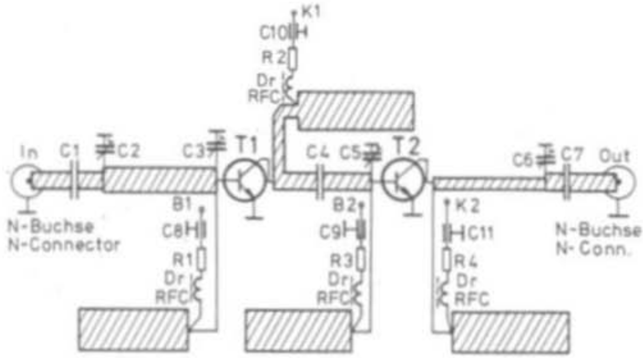


Fig. 7:
SHF-circuit diagram for the described amplifiers

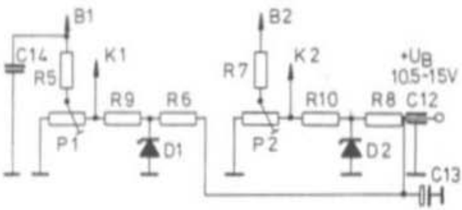


Fig. 8: Circuit diagram of the operating point adjustment and stabilization

sistor type NE 645 35 in the input and transistor NE 578 35 as second stage. The amplifier was constructed using a PTFE glassfibre board (1.57 mm thick). In addition to this, a cheaper version was developed for the 23 cm band equipped with two NE 578 35 using an epoxy glassfibre board, also 1.57 mm thick. **Figure 7 and 8** give the circuit diagram valid for all three amplifiers, separately for SHF and DC.

Components:

Band	T 1	T 2
23 cm inexpensive version	NE 578 35	NE 578 35
23 cm	NE 645 35	NE 578 35
13 cm	NE 645 35	NE 578 35

Band	R 5	R 6	R 7	R 8	R 9	R 10	P 1	P 2
23 cm inexpensive version	47 k	270	12 k	150	330	100	50 k	10 k
23 cm + 13 cm	18 k	220	12 k	150	150	100	25 k	10 k

D 1, D 2:	Zener diode C 9 V 1
RFC :	Ferrite bead on resistor lead
C 1, C 4, C 7:	Chip capacitor 20-100 pF
C 2, C 3, C 5:	0.5 - 3 pF ceramic miniature tubular trimmer
(C 3 deleted for 23 cm amplifier)	
C 6:	0.5 - 6 pF ceramic miniature tubular trimmer
C 8, C 12:	Feedthrough capacitor 470 - 1000 pF
C 13:	10 µF / 25 V tantalum electrolytic
C 14:	47 nF ceramic capacitor
R 1 - R 4:	10 Ω (value uncritical)

Figures 9, 10 and 11 show the conductor side of the 23 cm preamplifier constructed on a PTFE board (DJ 6 PI 008) and of epoxy material (DJ 6 PI 009), as well as the 13 cm preamplifier on a PTFE board (DJ 6 PI 010). Each of the three versions is shown from both sides in the photographs given in **Figures 12 to 17.**

The ground surface of the boards is not etched. Holes of 3.2 mm diameter are drilled for the feedthrough capacitors C 8 - C 11 at four positions on the board marked with rings.

Fig. 9:
PC-board of the 23 cm preamplifier using PTFE material

Fig. 12 and 13:
Photograph of the 23 cm preamplifier (PTFE) showing both sides

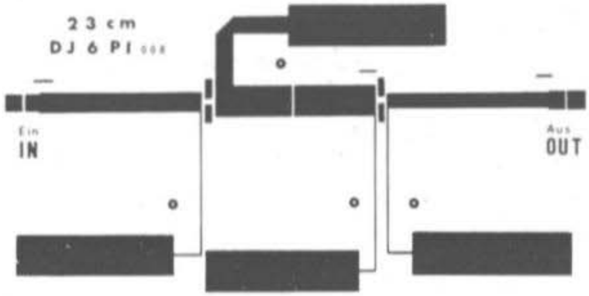


Fig. 10:
PC-board for the 23 cm preamplifier using epoxy material

Fig. 14 and 15:
Photograph of the 23 cm preamplifier (epoxy) from both sides

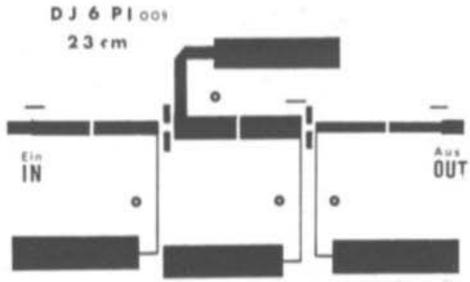
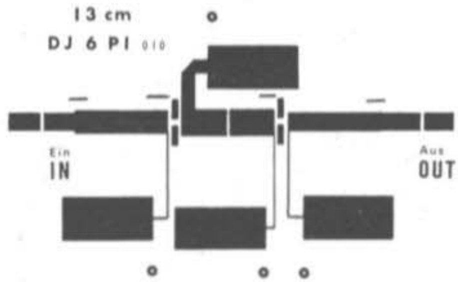


Fig. 11:
PC-board for the 13 cm preamplifier using PTFE material

Fig. 16 and 17:
Photograph of the 13 cm preamplifier



Some tendency to oscillation was observed with the first prototypes when the two emitter connections of the transistors were bent down and placed through 1 mm holes to the ground side of the board. For this reason, only small contact surfaces are provided for the emitter connections, which must be provided with low-inductive contact to the ground surface. This is achieved by sawing a small cut along the narrow side of this surface and placing an approximately 2 mm wide copper foil strip through this slot. The copper foil is then soldered to both sides of the board. **Figure 18** shows this in the form of a drawing.

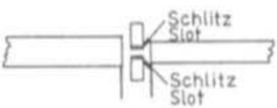


**Fig. 18:
It is necessary for through-contacts to be made for the emitter connection surfaces**

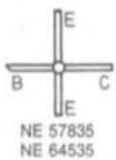
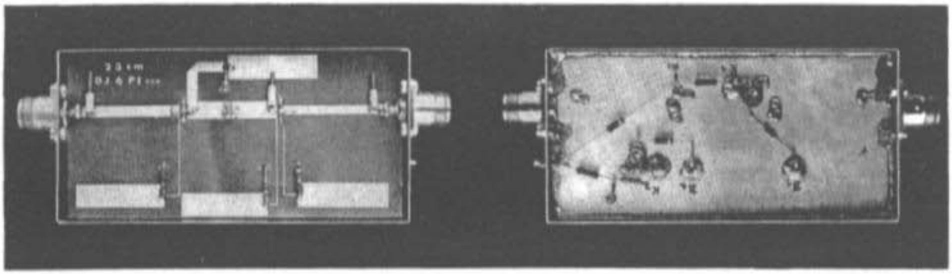
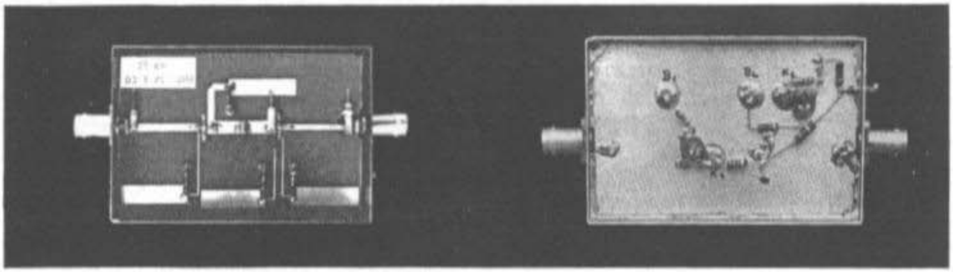
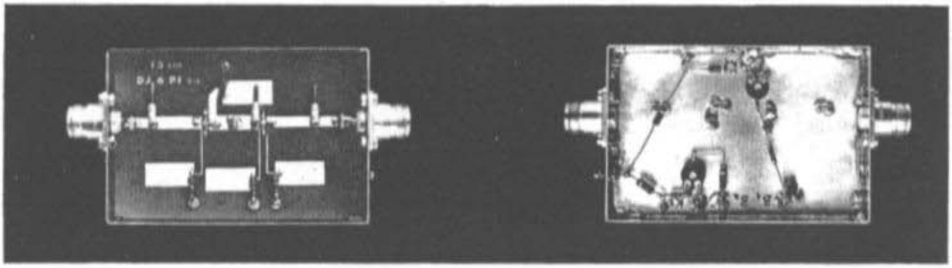


**Fig. 19:
Connection diagram for the NEC transistors**







After this, the transistors are soldered to the striplines without shortening the connection leads. In the case of the NEC transistors, the diagonally cut connection is the base (see **Figure 19**).

Due to their minute dimensions, the transistors can only be provided with a coded type designation: this is C_g (or another small letter) in the case of the NE 578 35, and Kz (or another small letter) for the NE 645 35.

The trapezoidal chip capacitors C 1, C 4, and C 7 are soldered into position vertically at the interruptions of the striplines. The ground connection of the ceramic tubular trimmers depends on the type used. In the case of the subminiature trimmers type RT 23 (Stettner), slots are cut into the board at the positions marked with lines, and the ground connection of the trimmer is then placed through the slot and soldered. In this manner, it is possible to achieve very short connections. The capacitance surface of the trimmers is then soldered to the striplines.

If trimmers without connection leads are to be used, a small bracket made from silver-plated wire should be bent around the ground connection and the two ends of the wire placed through 1 mm diameter holes in the board and soldered both to the ground surface and the trimmer. The capacitance surface is soldered to the stripline.

The feedthrough capacitors are soldered to the ground surface. The low-value resistors for the base and collector voltages are soldered with short connections to the feedthrough capacitors. Ferrite beads are placed over the wire ends at the other side of the resistors after which they are soldered into position at the intersection of the wide and narrow lines.

The PC-board should be soldered into a case which allows the base and cover to be removed. It is, however, also sufficient when a screening panel constructed from thin brass, copper or tin plate is provided around the edge of the board. Holes of approximately 4 mm diameter should be provided for aligning the trimmers.

A small cutout must be provided at the input and output of the amplifier to fit the protruding insulation between the inner condcutor and flange of the BNC (N is better) connector. The inner conductor of the connector is directly soldered to the end of the striplines. This can be seen in the photographs.

The dropper resistors R 5 - R 10, the two trimmer potentiometers, the two zener diodes and the bypass capacitor C 14 are accommodated in the ground side of the case as shown in the photographs. The common 12 V connection is made via feedthrough capacitor C 12 in the screening panel of the case.

5. ALIGNMENT AND MEASURED VALUES

Before connecting the operating voltage for the first time, the trimmer resistors for the base voltage should be adjusted to their ground end stop. A mA-meter is now connected into the collector line of the transistor to be adjusted. The trimmers are now adjusted so that the following collector currents result:

Band	First stage	Second stage
23 cm inexpensive version	3 mA	10 mA
23 cm + 13 cm	7 mA	10 mA

A signal generator, beacon signal or a signal from another radio station is required for alignment of the trimmer capacitors. It is assumed that the mixer, oscillator and IF-amplifier have already been aligned correctly. In most cases, the signal will still be heard weakly inspite of the fact that the preamplifier has still not been aligned correctly. The trimmers are then aligned one after another in several processes for maximum gain. Finally, the input trimmer or trimmers should be aligned with the aid of a noise generator, or by ear, for best signal-to-noise ratio.

The bandwidth of these amplifiers is so great that the relatively wide amateur bands including any change of the 13 cm band are amplified without deterioration.

5.1. Measured Values

The gain was measured with a HP signal generator, multiplier and spectrum analyzer. The noise figure was measured with a AIL noise generator together with an interdigital converter as described by DC 0 DA and included preamplifiers subsequent to the preamplifiers to be measured. Since a narrow-band mixer is used, the measured noise figures are single-sideband noise figures of the preamplifier including a small contribution from the subsequent unit.

The PC-boards described here are designed for transistors type NE 578 35 and NE 645 35. If other transistor types are used, one must expect inferior values for gain and noise figure if the striplines are not recalculated.

If strong signals are present in the neighbouring frequency ranges, it may be necessary to provide a narrow-band, low-loss filter. Of course, the insertion loss of the filter will deteriorate the noise figure. Provision of a filter is also advisable when the noise component at the image frequency of a wideband mixer is to be eliminated. In

	Overall noise figure		Gain	
Band	Amplifier with simple calculation	Amplifier with exact calculation	Amplifier with simple calculation	Amplifier with exact calculation
23 cm inexpensive version	3.2 dB	2.5 dB	20 dB	26 dB
23 cm	2.5 dB	1.8 dB	23 dB	26 dB
13 cm	3.6 dB	2.5 dB	18 dB	21 dB

Table 2: Measured values of the three preamplifier modules

order to ensure that the noise figure of the preamplifier is not considerably deteriorated, the filter should be connected between preamplifier and mixer in this case. High-quality coaxial or interdigital filters should be used in all cases.

A similar preamplifier for the 9 cm band will be constructed using RT-duroid material as soon as activity warrants this. This will then be described in a future article.

6. COMPUTER PROGRAMS

As previously mentioned, programs have been developed for the miniature computer TI-59 which is programmed with magnetic cards. This allows the following to be computed:

- The output reflection coefficient s_{22}^* for noise matching of the first stage
- The corrected input and output reflection coefficients s_{11}^* and s_{22}^* for simultaneous input and output matching (both taking the forward and return transfer coefficients s_{21} and s_{12} into consideration)
- The transistor impedances from the reflection coefficients
- The stability behaviour and the calculated gain
- Matching networks (length and width of the striplines, dimensions of the reactive impedance elements)
- Resistance values for alignment of the operating points

The programs are available from the author at a cost of DM 100.—. They include three recorded TI 59 magnetic cards, the program list, input instructions, and a complete calculation example.

7. REFERENCES

(1) P. Shuch, WA 6 UAM: Microstripline Preamplifiers for 1296 MHz
ham radio magazine
April 1975, pages 12 - 27 and October 1975, pages 42 - 46

(2) P. Shuch, WA 6 UAM: Solid-state Microwave Amplifier Design
ham radio magazine
October 1976, pages 40 - 47

(3) K. Hupfer: Striplines for VHF and UHF
VHF COMMUNICATIONS 3, Edition 4/1971, pages 207 - 216

(4) W. Schuhmacher: Dimensioning of Microstripline Circuits
VHF COMMUNICATIONS 4, Edition 3/1972, pages 130 - 143

(5) J. Fisk, W 1 HR: Simple Formula for Microstrip Impedance
ham radio magazine
December 1977, pages 72 - 73

(6) J. Fisk, HR:
Microstrip Transmission Line
ham radio magazine
January 1978, pages 28 - 37

(7) K. Richter: Design-DC-Stability Into Your Transistor Circuits
Microwaves, December 1973, pages 40 - 46

(8) J. Fisk, W 1 HR:
Low-Noise 30 MHz Preamplifier
ham radio magazine
October 1978, pages 38 - 41

A System for reception and display of METEOSAT Images Part 3

by R. Tellert, DC 3 NT

A local oscillator module is required to operate the VHF-receiver described in part 2 of this article. The oscillator to be described in this article can be constructed easily.

The image processing electronics comprise eight PC-boards and have been further developed in the meantime; (it is often a long and difficult path between an operating prototype and a system that is reliable enough to be published !). At present, the latest, third prototype is being constructed. The latest state-of-art is also to be mentioned to show the extent of the circuitry and the possibilities.

4. THE LOCAL OSCILLATOR

The reasons why the VHF-receiver could not be fed to a crystal-controlled frequency of 137.5 MHz, was already discussed in previous editions: Drift of the METEOSAT transmit frequency due to aging (according to the specifications a maximum of ± 20 kHz in 3 years), as well as variations of the local oscillator frequency in the converter due to temperature fluctuations and aging. If ± 20 kHz are also assumed for this, this will result in a required tuning range of ± 40 kHz for the VHF-receiver.

In order to keep the circuitry as simple as possible, an LC-oscillator is used that oscillates at the required final frequency (receive frequency minus IF), and can be tuned using a varactor diode (VCO). Of course, its stability is not sufficient without an automatic frequency control (AFC) loop which tunes the frequency – and thus the IF – to the center frequency of the FM-discriminator. The capture range of such an AFC-circuit is as great as the IF-bandwidth, in our case, ± 15 kHz. However, it must also be necessary to capture other signals that are not within the IF-bandwidth e.g. in periods of non-operation or when changing frequency. A scanning circuit is used here that sweeps the oscillator ± 250 kHz around the selected center frequency. As soon as a signal is received, the fine tuning is made automatically. The change-over switching from the scanning to AFC mode is controlled by the squelch of the receiver: as soon as a signal appears, the squelch will open, and the oscillator module will be switched to the AFC mode.

The wide tuning range of this voltage controlled LC-oscillator offers a further advantage: It is possible to delete the second crystal in the METEOSAT converter, and to save the switching system and control line to the converter, since it is possible for the local oscillator of the VHF-receiver to be switched to a frequency that is 3.5 MHz higher. In this case, one will only require the crystal for channel 2 (1691 MHz, which uses a crystal frequency of 86.306 MHz) and will then receive according to the following frequency plan:

1691.0 MHz – 137.5 MHz
1694.5 MHz – 141.0 MHz

The VHF-receiver module DC 3 NT 003 is then tuned to the frequency range of 136 to 138 MHz (VHF satellite band); the sensitivity at 141 MHz is more than sufficient when used in conjunction with the previously mentioned microwave converter.

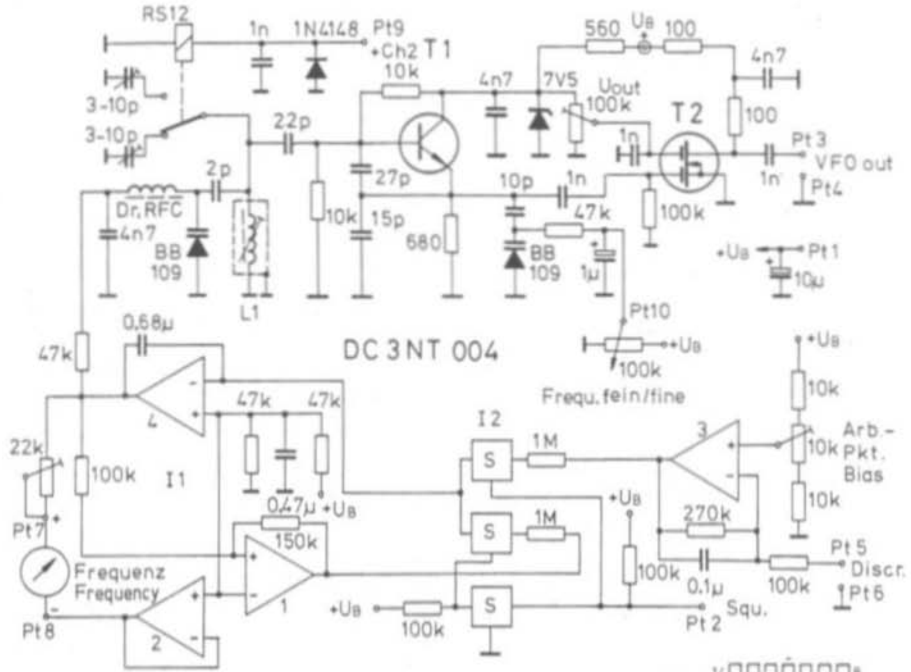
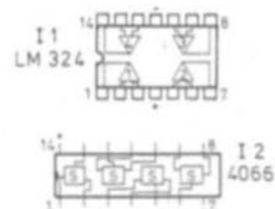


Fig. 21: Circuit diagram of the scanning local oscillator



4.1. Circuit Details

The complete circuit diagram of the oscillator is given in **Figure 21**. The actual oscillator is equipped with transistor T 1 in a Colpitts circuit and is followed by a buffer equipped with the DG-MOSFET T 2 (BF 900). The output power available between Pt 3 and Pt 4 is variable in the range of approx. + 3 dBm (2 mW) and – 10 dBm (0.1 mW), which allows it to be matched to the requirements of the mixer. This is achieved by varying the gate-2 voltage of transistor T 2. This process has little effect on the previous circuits, and the output power is very stable, since gate 2 is fed with the stabilized voltage from the oscillator.

A miniature relay is provided on the board that allows two different trimmers to be connected in parallel with the oscillator coil. This allows the frequency shift of 3.5 MHz to be set for reception of the two METEOSAT channels:

The oscillator frequency is switched from 126.8 MHz to 130.3 MHz. If this is not required, it is possible for the relay and one of the trimmers to be deleted and to connect the remaining trimmer capacitor to inductance L 1 with the aid of a bridge.

The frequency can be fine-tuned with the aid of the varactor diode provided at the emitter of the oscillator transistor. It is possible with this to tune the receive frequency from 137.5 MHz (NOAA) to 137.3 MHz (METEOR), or to 137.62 MHz (TIROS N).

A quad amplifier I 1 (LM 324) and three of four electronic switches (I 2. 4066) are to be seen in the lower part of this circuit diagram. One amplifier (No. 4) is connected as

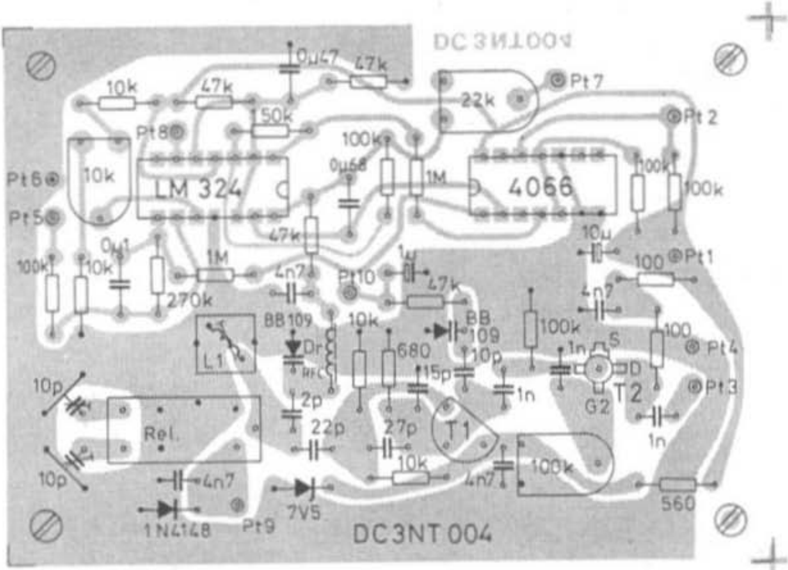


Fig. 22: PC-board DC 3 NT 004

an integrator, and a further (No. 1) as an amplifier with hysteresis behaviour. When the squelch is closed, they are fed back via an electronic switch, which results in a triangular-wave output voltage whose reversal points correspond to the hysteresis switching threshold. The triangular-wave voltage now sweeps the oscillator frequency with the aid of the second varactor diode that is loosely coupled to inductance L 1. The scanning range is dependent on the value of the coupling capacitor, in our case 2 pF.

If a signal is received, the squelch will open, and Pt 13 of module DC 3 NT 003 will activate a further electronic switch via Pt 2 of the receiver module which will then switch the integrator via a matching stage (amplifier No. 3) to the discriminator of the receiver. This means that it is switched to the AFC-mode.

The electronic switches are in C-MOS technology, and the type 4066 used possesses a buffer amplifier. Although type 4016 has the same pin connections, it does not have these buffers, and since the squelch can provide voltages in the range of 0 V up to the value of the operating voltage, this can cause unreliable switching conditions if a buffer is not used.

4.2. Construction

A single-coated PC-board DC 3 NT 004 was developed for accommodation of the described scanning oscillator (see **Fig. 22**). The dimensions of this board are 100 mm x 70 mm. The local oscillator should be accommodated in a screened metal case into which holes have been provided for alignment (for the two trimmer capacitors of L 1, and the three trimmer potentiometers).

4.2.1. Components

T 1: BF 199 or similar UHF transistor
T 2: BF 900 or similar DG-MOSFET

I 1: LM 324 (Nat. Semiconductor)
I 2: 4066 (RCA, Motorola, Philips etc.)

2 BB 109 (Siemens)
1 C 7 V 5 zener diode
1 any silicon diode

L 1: ready-wound inductance (Neosid 005118)

RFC: 18 turns of 0.3 mm dia., enamelled copper wire wound on a coil core

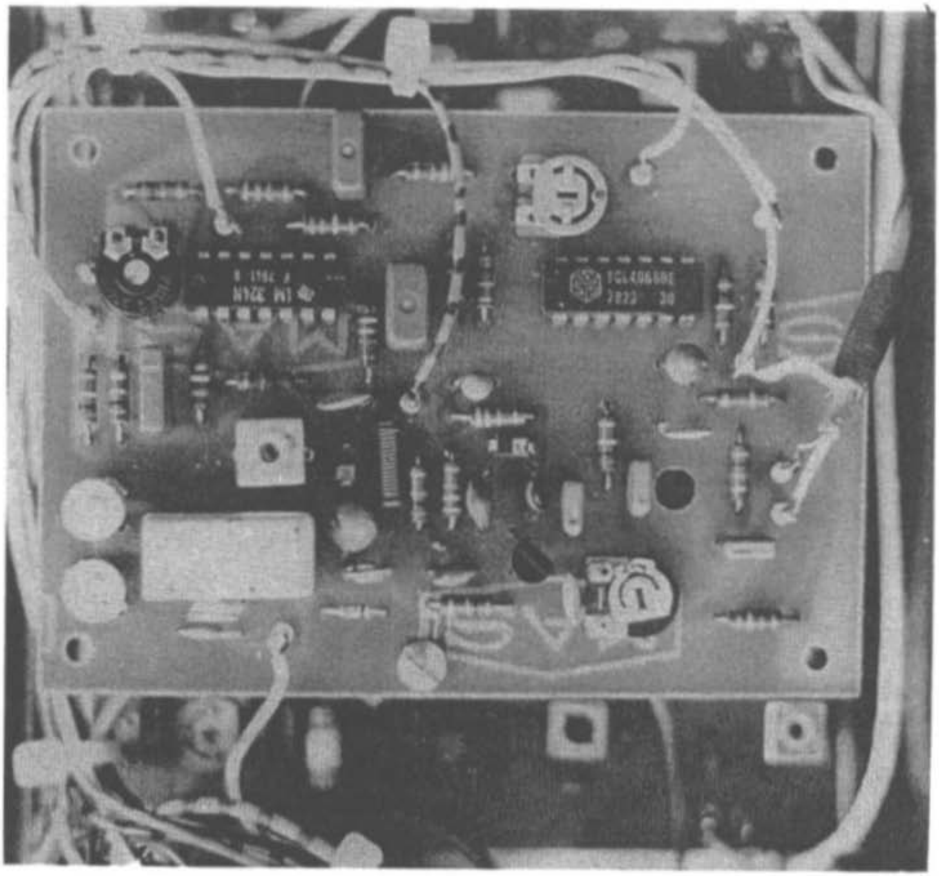


Fig. 23: Photograph of the author's prototype oscillator module, matching VHF receiver DC 3 NT 003

2 plastic-foil trimmers, 10 pF, 7.5 mm dia. (Philips)

A spacing of 5 mm is available for the various ceramic disk capacitors between 2 pF and 4.7 nF.

A spacing of 7.5 mm is provided for the three plastic-foil capacitors of 0.1 µF, 0.47 µF and 0.68 µF.

The three trimmer potentiometers are types for horizontal mounting with a spacing of 10/5 mm.

4.3. Connection and Alignment

The following interconnections should be made between the receiver module DC 3 NT 003 and the described oscillator DC 3 NT 004 :

DC 3 NT 003	DC 3 NT 004	Connection
Pt 3	Pt 1	+ 12 V
Pt 2/ground	Pt 3/Pt 4	VFO out
Pt 13	Pt 2	Squelch
Pt 7	Pt 5	Discrim.
Pt 17	Pt 9	via a switch

It is also necessary to connect the front-panel potentiometer for fine tuning as shown in the circuit diagram and to connect the discriminator meter to connection points Pt 7 and Pt 8, although it is no longer so advantageous when using this oscillator. It will now show the state of the scanning process; and the position at which the meter stops will give an idea of the frequency.

Figure 23 shows a photograph of the ready-to-operate oscillator module. A frequency counter is very helpful during the alignment process. The frequency is now set with the aid of the coil core and trimmer capacitor to the required receive frequency minus 10.7 MHz. The fine-tuning potentiometer should be in its center position, and the squelch open.

The trimmer potentiometer for the operating point is now adjusted with the squelch open so that the frequency indication (meter between Pt 7 and Pt 8) slowly moves in one of the two directions.

To check operation, it is necessary to receive a signal. The frequency indication will then stop, and half the operating voltage (i.e. approx. 6 V) should be present at pin 2 of I 3 of module DC 3 NT 003, when measured with a high-impedance DC-voltmeter. This voltage value will only appear when the received frequency is tuned exactly symmetrical to the discriminator characteristic.

Under normal conditions, the squelch should only just close, so that it opens quickly enough as soon as a signal is received.

5. IMAGE PROCESSING

The overall circuit and operating possibilities are now to be discussed with the aid of a block diagram and front panel drawing. This system does not only allow one to choose between two methods of image processing (Photographically using a monitor tube or with a FAX-machine) but also allows one to select the various signal sources such as METEOSAT converter, VHF-receiver, LW or SW receiver, or tape recorder. It also includes the circuits for the various operating speeds of 240 to 48 lines per minute, as well as the evaluation circuits for the various start-stop signals.

This allows the constructor to decide which modes of operation are required before commencing construction. It is, of course, possible that only one or two of these possibilities are required. In this case, the other parts of the circuit can be deleted. The PC-boards contain all individual circuits; however, the components that can be deleted when one or more of these functions are not required will always be mentioned in the detailed descriptions.

After completing the third prototype, all of the electronic circuitry is now accommodated on eight PC-board modules. Three of these are located in the vicinity of the monitor tube: The high-tension supply DC 3 NT 010, the deflection output stage DC 3 NT 011 and the power supply for the camera motor DC 3 NT 012. The other five PC-board modules DC 3 NT 005 to 009 have standard European sizes and are accommodated in an individual case. The front panel drawing of the image processing module given in **Figure 24** shows the possibilities provided by this unit.

In the first module from the left, one will find the change-over switches for the various AF-signal sources, for the sub-carrier modulation (AM/FM), and for the recording speed, as well as the on/off switch.

The second module contains the start-stop logic: two adjustable AF-filters between 200 and 500 Hz are provided complete with LED-indicators. For instance, these filters should be set to 300 Hz and 450 Hz in the case of METEOSAT to activate the evaluation circuitry. The start pulse can be either positive or negative going.

The other controls on this module are for image commencements in the case of orbiting weather satellites such as TIROS N or NOAA 6. These satellites transmit their APC-images in the 137 MHz range continuously, i.e. without start tone, synchronizing impulses, and stop tone. However, a tone burst is transmitted at the commence-

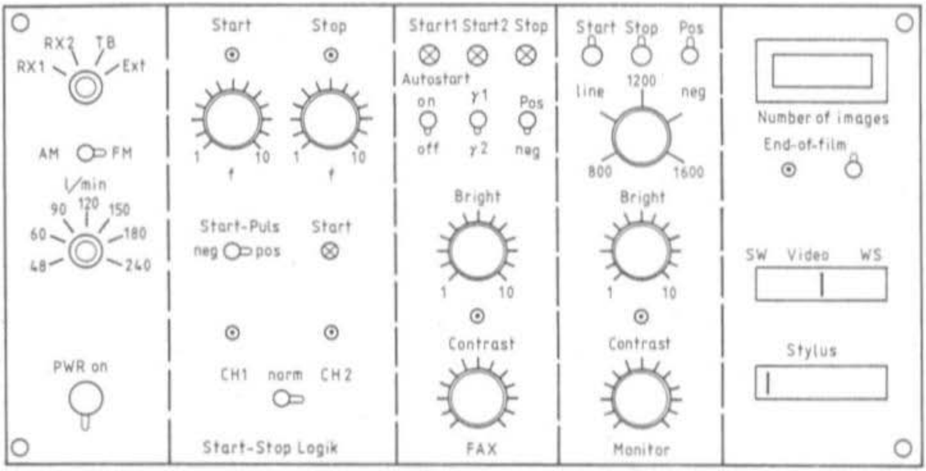


Fig. 24: Front panel of the image processing unit

ment of each line. This burst is used for starting the image processor, and a push-button switch is provided on the front panel for this. If a certain image segment is required, it is necessary to know the satellite path and the time of the equator crossing. It is then possible to determine the commencement of the image with the aid of a stop watch.

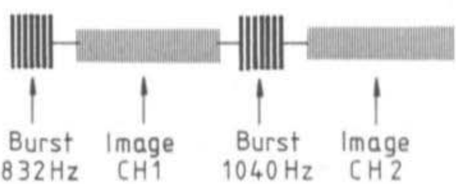


Fig. 25: A video signal line of TIROS N or NOAA 6

A special feature of the TIROS N and NOAA 6 satellite should be mentioned: Both of these satellites transmit the image information simultaneously in two channels (e.g. visible and infrared). These two channels are written side by side using two different bursts, as shown in **Figure 25**. It is now possible for either both images to be recorded side-by-side by setting the line speed to 120 lines per minute, or to record with 240 lines per minute and to blank the non-required channel with the aid of the burst frequency. The lowest change-over switch with the two control lights on the second module is provided for this.

The third module from the left contains the control elements for the faximile machine:

Start 1 will commence in synchronism if the push-button is depressed during the synchronizing pulses; start 2 will commence immediately as long as the stop switch is not selected. The stop switch will disable the start circuit. To stop the faximile machine, the switch need only be switched off momentarily. Another switch allows the image to be recorded positively or negatively, and a further switch is provided to switch between two different gamma-equalizing circuits, i.e. grey-scale characteristics. Finally, controls are provided for brightness and contrast; the latter control can be varied between 0 and infinite (= black/white). The operation of the machine is indicated with the aid of a control lamp.

The fourth module from the left is very similar to the third and is provided for operating the monitor tube. An instantaneous start-switch, stop-switch and a positive-negative changeover switch are provided. If the images are to be projected in a slide projector after development, one should select position »POS«. In the case of paper images (enlargements), the images should be recorded in position »NEG«.

The potentiometer »Lines« should be set, to 800 for METEOSAT, and to 1600 for TIROS N and NOAA 6 (only half the image will be visible due to the blanking of one channel), and to between 900 to 1800 in the case of weather maps on LW or SW.

The brightness and contrast controls are independent of those of the faximile machine.

The last module from the left finally contains a numerical switch for selecting the number of images, an end-of-film indication and switch-off, a meter for video drive to aid adjustment of brightness and contrast, as well as a meter for indicating the stylus current of the faximile machine.

5.1. Block diagram

The block diagram of the image processor that has been discussed in the previous section is given in **Figure 26**. As has been previously mentioned, it contains the five modules DC 3 NT 005 to 009. The following information is to be given to provide some understanding of the concept and to allow a decision as to the individual requirements.

DC 3 NT 005 is the so-called video board. This module is provided with the subcarrier via the AF-source switch. Trimmer potentiometers provided on the rear panel allow all levels to be adjusted to a certain value. It is possible to select either an FM or AM demodulator (METEOSAT). The demodulator is followed by a low-pass filter to reduce the noise bandwidth. The video signal can be recorded at this point on one channel of a stereo tape recorder. A synchronizing tone of 1350 Hz from module DC 3 NT 008 is recorded on the other channel simultaneously. This is used as reference frequency during playback and controls the whole system with the aid of a PLL-circuit.

Module DC 3 NT 005 also contains the filter for the start pulses and the pulse-width modulator for the stylus current of the FAX recorder. The stylus is driven with a voltage of 100 V, which is fed from the modulator to a connector where an oscilloscope can be connected for monitoring purposes.

DC 3 NT 006: The power supply board provides a stabilized voltage of 15 V for the electronic circuits, as well as 100 V for the stylus of the faximile machine. It also contains the two output stages for the motor voltage of the faximile machine. This is a 6 W synchronous motor which requires two sinusoidal AC-voltages phase-shifted by 90°.

DC 3 NT 007 contains the evaluation circuit for the start and stop signals. The AF-filter in each of these channels is followed by a delay line and a monoflop circuit; these are provided for the image processing of METEOSAT signals. In the case of TIROS N and NOAA 6 reception, the AF-filters are followed by a selection circuit for the two individual channels.

DC 3 NT 008 is the frequency generator module. It is equipped with a crystal oscillator and a frequency divider to provide a frequency of 1350 Hz. This frequency is used as reference for a PLL-circuit either directly or via the tape recorder. The VCO of the PLL-circuit oscillates at 86.4 kHz. This frequency is divided in a variable divider to the required recording speed. The selected output frequency controls the X-deflection of the monitor tube.

In the case of the output signal for the faximile machine, the output impulses of the switchable divider are in the form of a sinusoidal voltage. A 90° phase-shifted cosine voltage is derived from this sinusoidal voltage; both of these signals are then fed to the output stage on the power supply board.

DC 3 NT 009: The monitor module contains the sawtooth generators (integrators) for the horizontal (X) and vertical (Y) deflection. The frequency of the horizontal deflection is controlled by the switchable frequency generator, whereas the image time is set with the aid of a potentiometer. Both circuits have an output of between 5 and 10 V.

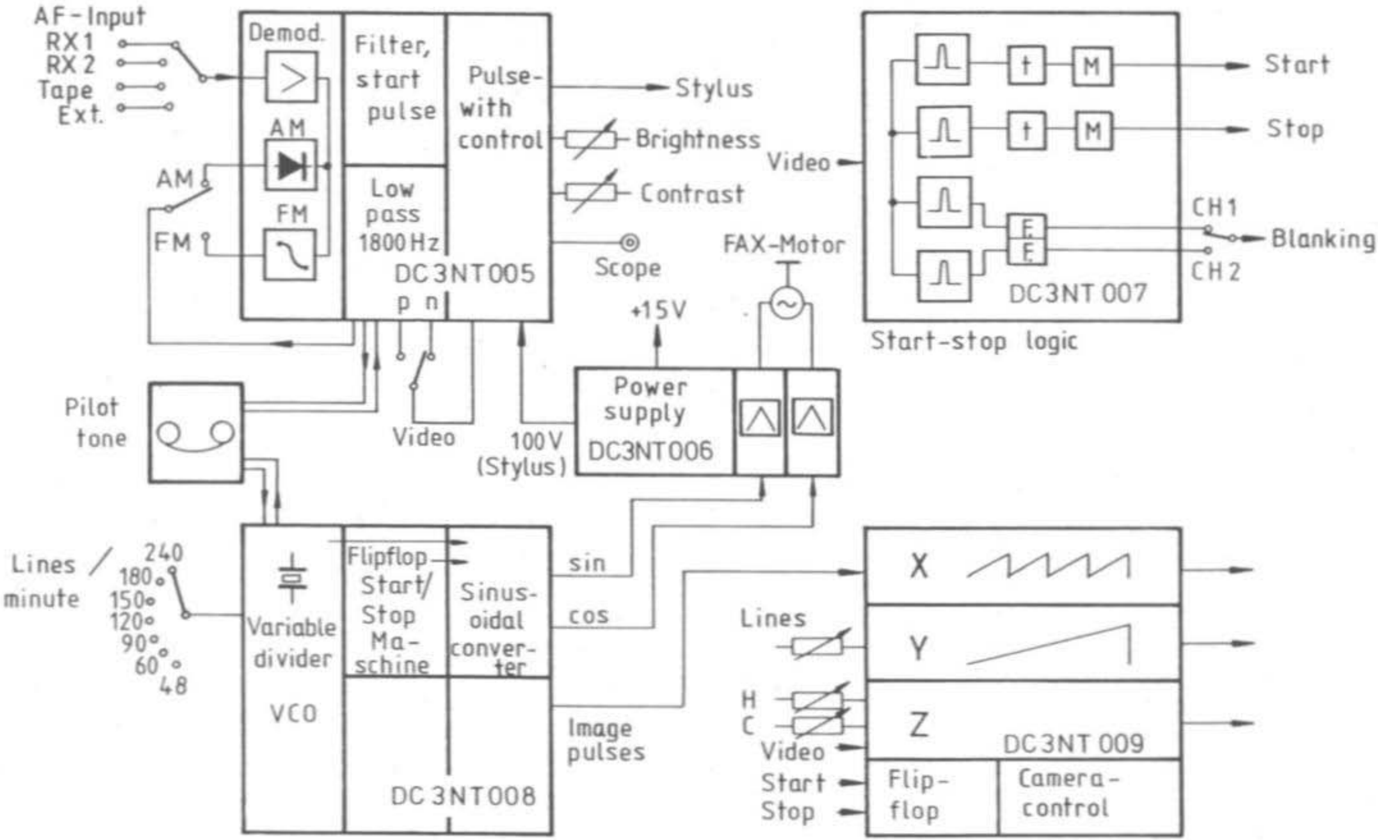


Fig. 26: Block diagram of the image processor

Z is the designation of the video output amplifier in module DC 3 NT 009. This stage modulates the beam current of the monitor tube, and thus the brightness. When no signal is present, the output voltage amounts to + 1 V, and the beam is then blanked; a value of a approximately + 7.5 V corresponds to grey.

This module also contains the camera control circuit. It is necessary for the shutter to be opened and closed, and the film transport motor to be switched on and switched off after a film transport of 26 mm. The image processing circuit should be switched off at the end of the film. This will be indicated by a control lamp on the front panel.

It is hoped that this provides some idea of the operation of the image processor. The following articles are to discuss the individual modules in detail together with circuit diagrams and component location plans. An interconnection diagram as well as construction and alignment details are also to be provided. The author and editors have still not decided whether the mechanical details of the faximile machine should also be published. A handy constructor having a small lathe should be able to construct such a machine without problems. The editors would be interested to hear the opinions of our readers.

REFERENCES TO PART III

R. Tellert: A system for reception and display of METEOSAT images – Part 2
VHF COMMUNICATIONS 11,
Edition 4/1979, pages 194 - 202

Simplified Inductance Calculation for Small Air-Spaced Coils

by H. Rathke, DC 1 OP

The readers of VHF COMMUNICATIONS have often requested articles describing an inductance meter. However, in the case of simple inductance meters, it is very difficult to measure inductances up into the VHF-range due to the effect of parasitic reactive impedances. The diagram given in **Figure 1**, however, allows small inductances from approximately 2 nH up to the mH-range to be calculated with sufficient accuracy in a theoretical manner. Required are only the dimensions: length and diameter of the inductance, and diameter of the wire used.

The curves given in the diagram give the specific inductance value L_0 (in nH) for a single-layer air-spaced coil in the range of 2 to 50 mm mean coil diameter, and from 2 to 50 mm coil length.

The inductance L of a coil is determined by:

$$L = L_0 n^2 \qquad (1)$$

where n is the number of turns.

The curves are derived from the equation:

$$L = \frac{n^2 r^2}{9r + 10l} \qquad (2)$$

where

r = radius of the coil (D = 2 x r)

l = length of the coil

Example 1:

Required is an inductance of 200 nH. The length of the coil is fixed at 20 mm, the diameter at 6 mm. When using 1 mm diameter wire, the mean diameter of the coil will be 7 mm. A value of $L_0 \approx 2$ nH can be taken from the diagram for a coil length of 20 mm, and a mean diameter of 7 mm.

This results in the following number of turns when derived from equation (1):

$$n = \sqrt{\frac{L}{L_0}} = \sqrt{\frac{200}{2}} = 10 \text{ turns}$$

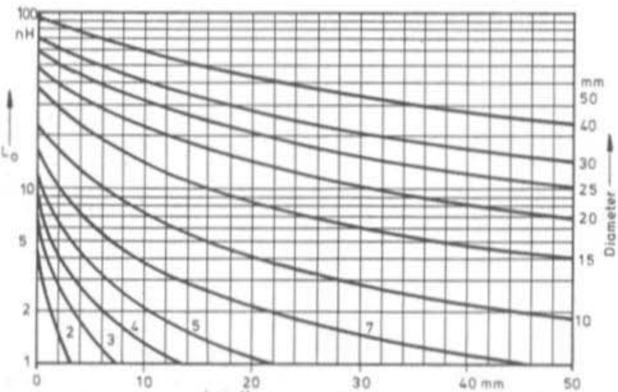


Fig. 1: Specific inductance of air-spaced coils as a function of length; parameter: coil diameter

Example 2:

Required is the inductance of a ready-wound coil having n = 10 turns. The external dimensions are:
Coil length = 20 mm, inner diameter = 6 mm, and a wire of 1 mm diameter is used. The mean diameter is thus 7 mm. A value of $L_0 \approx 2$ nH is taken from the diagram; the following results when derived from equation (1):

$$L = L_0 n^2 = 2 \times (10)^2 = 200 \text{ nH}$$

Before concluding, a few notes are to be given regarding practical application:

Inductances that were determined according to this method were usually found to have the calculated value, especially then when the spacing between turns was equal to the diameter of the coil wire used. The variance of the determined number of turns was approximately ± 1 turn, which means that long expanded coils should be reduced by one turn, and very close-wound coils should be increased by one turn.

Even though this article will not replace an inductance meter, it provides the designer with a tool to indicate the correct order of magnitude when constructing inductances, and will at least shorten the time-consuming method of »dipping« the frequency with the aid of a dip-meter.

REFERENCES

M. Mann: Inductance calculation simplified for small air-wound coils
Electronic Design 21, Oct. 11, 1974

Simple Design of λ/4 Stripline Circuits

by W. Lerche, DC 3 CL

This article is to describe a simple method for rapidly calculating the length, or shortening capacitances of λ/4 circuits. The curves were plotted with the aid of a computer type CDC 7600.

1. DESCRIPTION AND APPLICATION OF THE CURVES

Four diagrams were drawn:

Fig. 1: Suitable for striplines with $\varepsilon = 5$ (epoxy resin), with a thickness of d = 1.5 mm

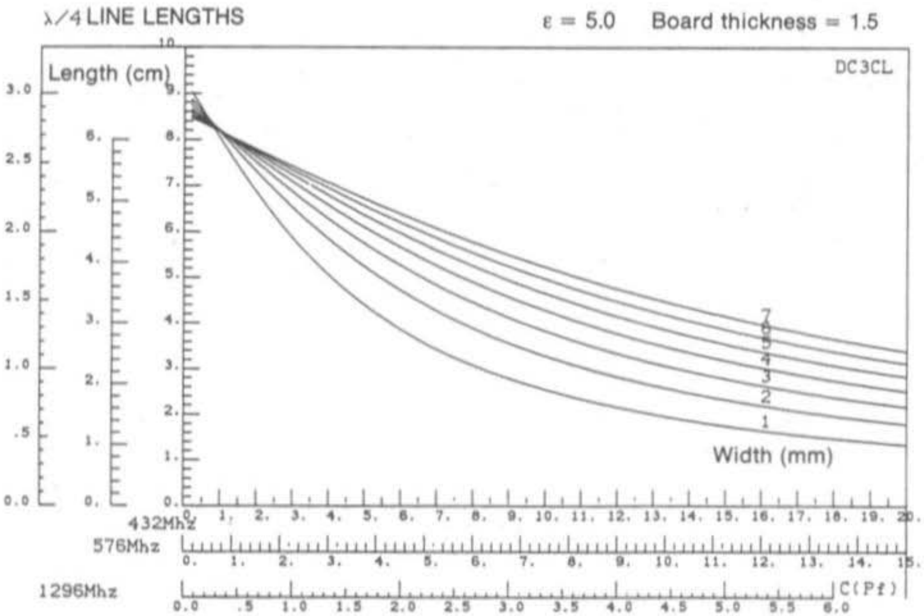


Fig. 1: Length of λ/4 stripline circuits when using 1.5 mm thick epoxy boards

Fig. 2: Suitable for striplines with $\varepsilon = 2.3$ (PTFE), board thickness d = 1.6 mm

Fig. 3: Suitable for striplines with $\varepsilon = 2.3$ (PTFE), board thickness d = 0.8 mm

Fig. 4: Suitable for air-spaced resonant circuits such as cavities etc.

These curves show which capacitance, length, and width one requires at a given frequency to obtain λ/4 resonance; the following experimental equations are used:

$$\varepsilon_{eff} = \frac{\varepsilon + 1}{2} + \frac{\varepsilon - 1}{2} \; (1 + 10\,h/w)$$

(h = thickness / w = width)

$$Z = \frac{1}{\sqrt{\varepsilon_{eff}}} \; 59.952 \ln (8\,h/w + w/4\,h)$$

with $0 \leqq w/h \leqq 1$

$$Z = \frac{119.904\,\pi}{(w/h) + 2.42 - 0.44\,(h/w) + (1 - h/w)^6} \times \frac{1}{\sqrt{\varepsilon_{eff}}}$$

with $w/h \geqq 1$

$$c = \frac{1}{\omega Z} \cot 2\pi \, l_{eff}/\lambda$$

According to (1) the equations should obtain an accuracy of 0.25 %.

The required line length is determined as follows:
The capacitance value that is to be used should be found on the corresponding frequency axis, and the intersection point with the curve having the required stripline width then gives the required length. Of course, it is necessary to use the axis valid for the same frequency. Attention should be paid that the bandwidth will decrease the larger the capacitance and the shorter the length of the stripline !

The capacitance value required for a given line length is obtained in the opposite manner.

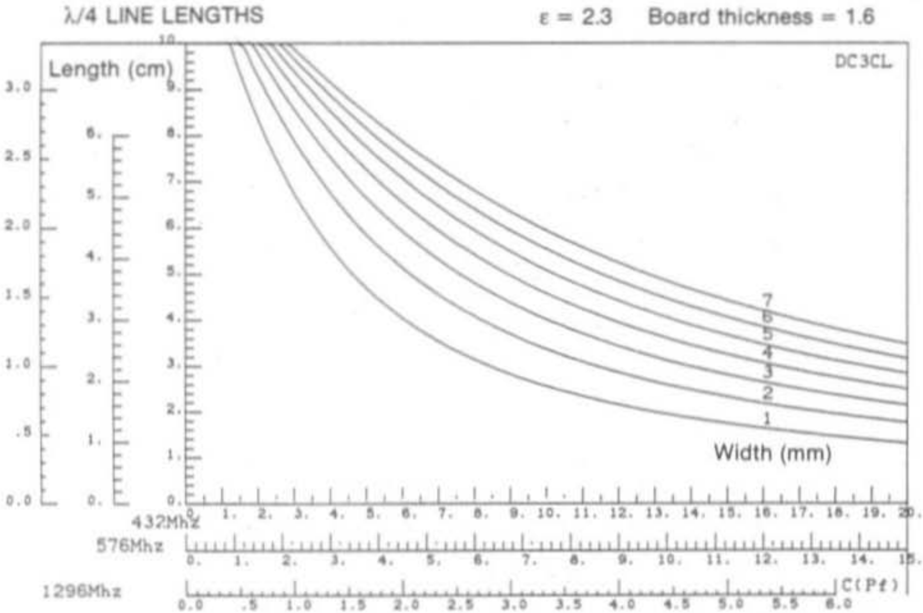


Fig. 2: Length of λ/4 stripline circuits when using 1.5 mm thick PTFE boards

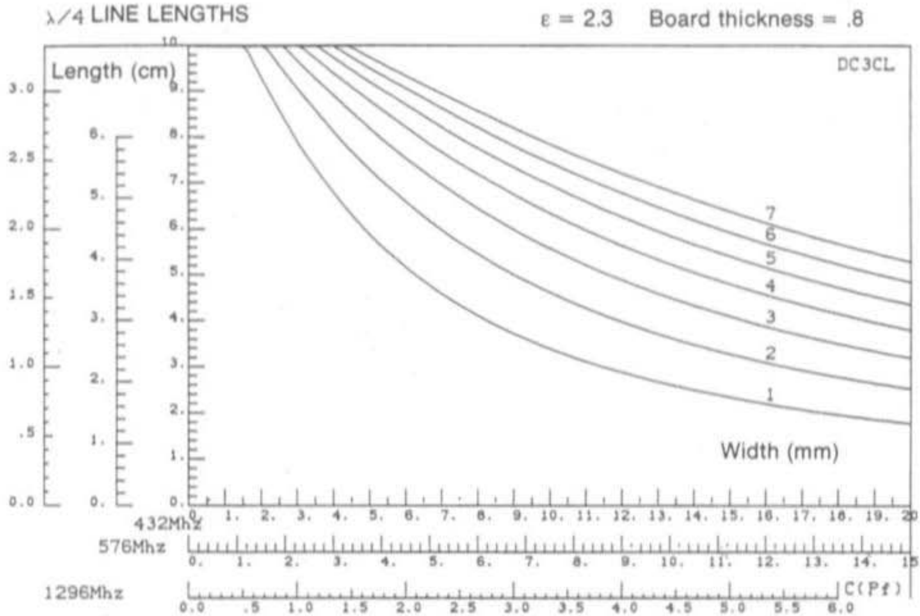


Fig. 3: Length of λ/4 stripline circuits when using 0.8 mm thick PTFE boards

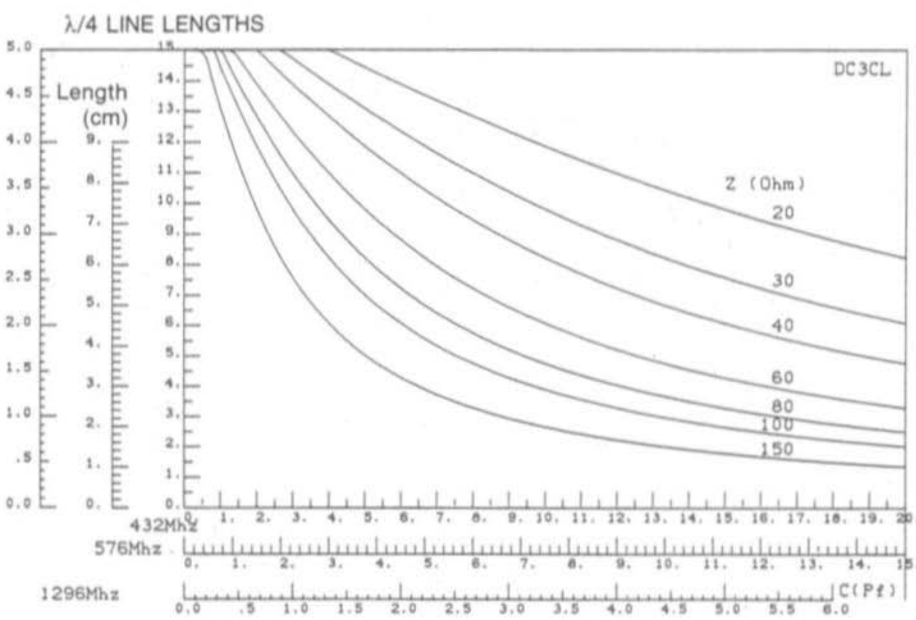


Fig. 4: Length of λ/4 resonant circuits with air dielectric

In order to obtain other frequencies, the capacitance axis should be divided in a similar manner to the frequency axis using the same factor; for this reason, a frequency of 576 MHz was selected for the central axis. It is then possible by doubling the values to obtain 288 MHz, by halving the values 1152 MHz and by dividing by four to obtain 2304 MHz (see example).

In **Figure 4**, the curve parameter is not the width of the line but the impedance Z. This is very useful for use in conjunction with cavity circuits.

2. EXAMPLES

2.1. A $\lambda/4$ circuit is to be calculated when using a 1.5 mm thick, epoxy-glassfibre board. The frequency is f = 432 MHz, the line width w = 3 mm and the capacitance C = 6 pF.
(see **Figure 5**).
One obtains a length L = 5.2 cm from diagram 1.

2.2. Which capacitance is required for a stripline with a copper-coated PTFE of h = 0.8 mm, where L = 2.5 cm, w = 3 mm and f = 1296 MHz ?

A value of C = 2 pF can be taken from diagram 3.

2.3. A resonant circuit for f = 2304 MHz is to be designed on a copper-coated PTFE board of h = 0.8 mm thickness, with w = 4 mm and L = 1.5 cm.

Diagram 3 is used; 2304/576 = 4, which means that the 576 MHz axes are divided by four. This means that one commences at L = 6 cm: this results in 4.8 pF. The required value is thus 1.2 pF at 2304 MHz.

2.4. A cavity circuit as shown in **Figure 6** with a round inner conductor of d = 5 mm and square outer conductor of D = 2.5 cm is to be constructed for a wavelength of 23 cm: With a ratio D/d = 5, one will obtain an impedance Z of approximately 100 Ω. A shortening capacitance of C = 0.5 pF = L = 4.5 cm is obtained from diagram 4.

3. PRACTICAL DETAILS

Although the curves are very accurate, large deviations can occur in practice since the direct environment, the line lengths to and within the trimmers, as well as the ground conditions at the cold end have a great effect. Also, one may not neglect the capacitances of a connected transistor !

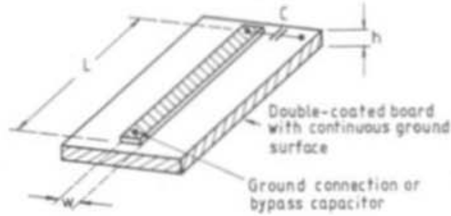


Fig. 5: Important dimensions of stripline circuits

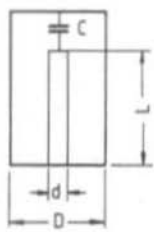


Fig. 6: Cavity circuit

For this reason, one should select a 10 to 20 % shorter length and take the line length of the trimmer, bypass capacitance (if provided), and the capacitances of the connected circuits into consideration.

The bypass capacitor should be selected that it is in series resonance together with its connection leads. This means:

approx.	100-200 pF (ceramic)	for 70 cm
approx.	20- 30 pF (chip)	for 23 cm
approx.	10 pF (chip)	for 12 cm

Figure 7 shows the installation of such capacitors.

Fig. 7: Mounting trimmer and bypass capacitors

(1) Gunston, M. A. R.:
Microwave Transmission-Line Impedance Data page 42-43
Van Nostrand Reinhold Co.
New York 1972

VHF communications

Index
for volume 11 (1979)

8. SATELLITE RECEPTION AND COMMUNICATIONS

A Remote Polarization Switching Unit for Crossed-Yagi Antennas

by H. Stoll, DF 7 SO

Polarization switching of cross-Yagi antennas was discussed in detail in (1), and (2). For this reason, it is not necessary to discuss these fundamentals here. A polarization switch which allows selection of the required polarization manually was described in (2) and has been available on the market for some time.

The remote polarization switching system to be described in this article possesses the following advantages:

- The polarization switching can be made remotely, and allows the use of a preamplifier located directly at the antenna. This means that cable losses can be compensated for in the receive mode.
- The matching is exact in each polarization mode. It is also possible to use the described system at high output power levels, since the individual dipole and coaxial relays used for polarization switching are only loaded with half the output power. The output power from the transmitter is distributed evenly to both dipoles in all polarization modes.
- The phase difference can be realized very accurately, since it is determined by the electrical length of the coaxial cables and it is not necessary to take the delay time of the relays or plugs into consideration as long as they are identical in each branch.

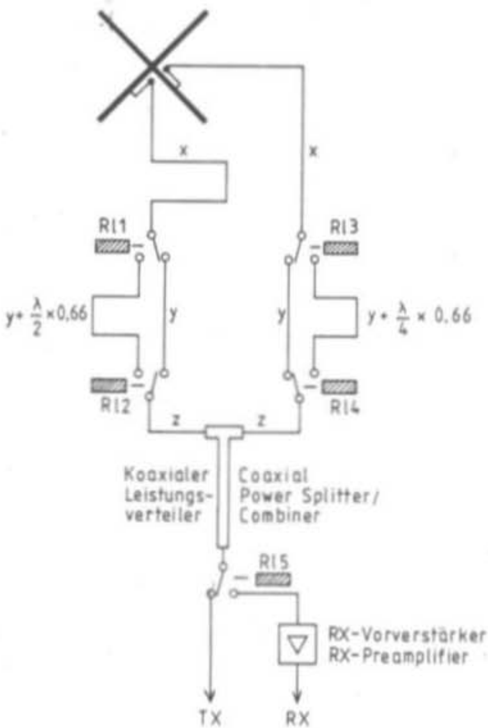


Fig. 1:
Circuit diagram of the polarization switching unit. Cross-Yagi mounted as »X«, and dipoles as seen from the reflector

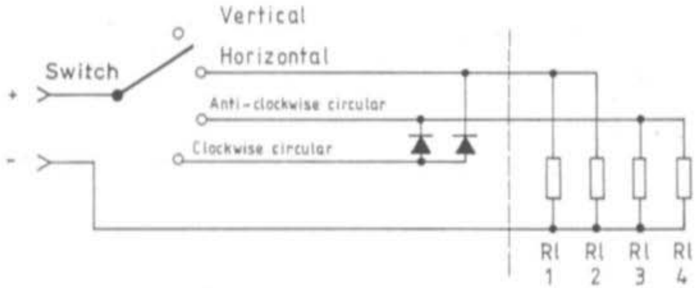


Fig. 2: Remote control of four relays from the station

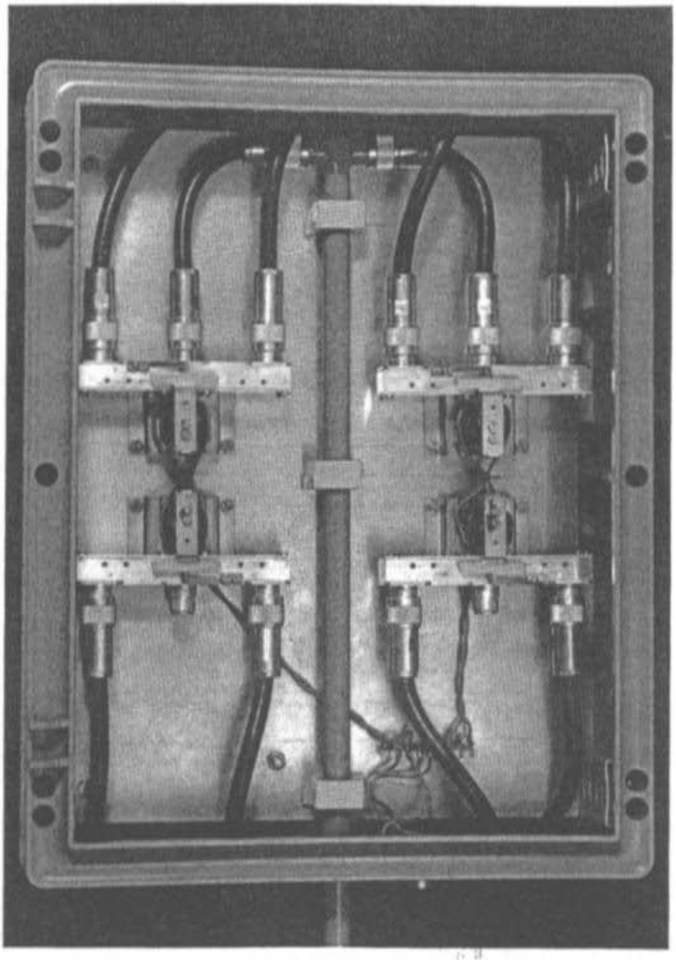


Fig. 3: Author's prototype of the polarization switching unit. The quarter-wave matching transformer (UKW-TECHNIK) is to be seen in the center and the coaxial relays of each branch on the left and right-hand side. The delay lines are mainly to be found behind the rear panel.

The circuit diagram of the remote polarization switching unit is given in **Figure 1**. The lengths x, y and z are not critical but must have exactly the same length in each branch. The velocity factor of 0.66 is valid for coaxial cable type RG 213/U (Ed. See editorial note below).

A quarter-wave matching transformer available from UKW-TECHNIK, was used to match the individual dipoles. This power divider distributes the power to the left and right-hand branch shown in Figure 1. If all four relays are not energized, the cable lengths in each of the branches will be equal, and this will result in vertical polarization when the dipoles are mounted as shown.

If relays Rl 1 and Rl 2 are energized, the length of the left-hand branch will be increased by $\lambda/2$, and the 180° phase difference between the dipoles will then result in horizontal polarization.

If relays Rl 3 and Rl 4 in the right-hand branch are energized, the additional $\lambda/4$ delay of the signal in the right branch will generate circular polarization.

The direction of the circular polarization is determined by whether the $\lambda/2$ delay in the left-hand branch is switched on or off. The following table gives the polarization modes that can be selected, and **Figure 2** gives the method of remotely controlling the relays from the station.

Polarization	Relays energized
Vertical	–
Horizontal	Rl 1, Rl 2
Anti-clockwise circular	Rl 3, Rl 4
Clockwise, circular	Rl 1, Rl 2, Rl 3, Rl 4

The described system was built into a weather-proof box, and has proved itself for some time when used for satellite communications.

Editorial note:

In our experiences, it has been found that the velocity factor varies from cable manufacturer to cable manufacturer and from roll to roll. We therefore recommend that the actual velocity factor be determined by measuring the resonant lengths of a quarter-wave stub before using these in the described switching units or constructing any other form of phasing systems. We have found that the velocity factor of RG 213 tables can vary between 0.62, and 0.66.

REFERENCES

(1) T. Bittan: Circular Polarization on 2 m
VHF COMMUNICATIONS 5,
Edition 2/1973, pages 104 - 109

(2) T. Bittan, DJ 0 BQ: Antenna Notebook
VHF COMMUNICATIONS 6,
Edition 1/1974, pages 38 - 41

A Noise Blanker for Large-Signal Conditions Suitable for Shortwave and VHF Receivers Having a Large Dynamic Range Part 1

by M. Martin, DJ 7 VY

A large number of different pulse type interference is present during radio communication in the shortwave and VHF range. The task of this article is to find ways to suppress such interference. Since the sources and type of interference differ greatly, it is very often impossible to obtain a satisfactory blanking of such interference. Part 1 of this article is to discuss fundamentals of interference suppression and to discuss the difficulties encountered in achieving this. Part 2 is to introduce a noise-blanker circuit, which uses various trigger possibilities that can be matched optimally to the interference in question. The circuit is compatible with the input circuit for a 2 m receiver described in (7).

1. THE VARIOUS TYPES OF INTERFERENCE PULSES

Firstly, the various interference pulses are to be discussed together with a few fundamentals of impulse technology, wherein the pulse processes are to be displayed both in the time (oscilloscope) and frequency (spectrum analyzer) domain, so that the effects appearing during the noise-blanking process can be seen more clearly.

1.1. Switching Clicks

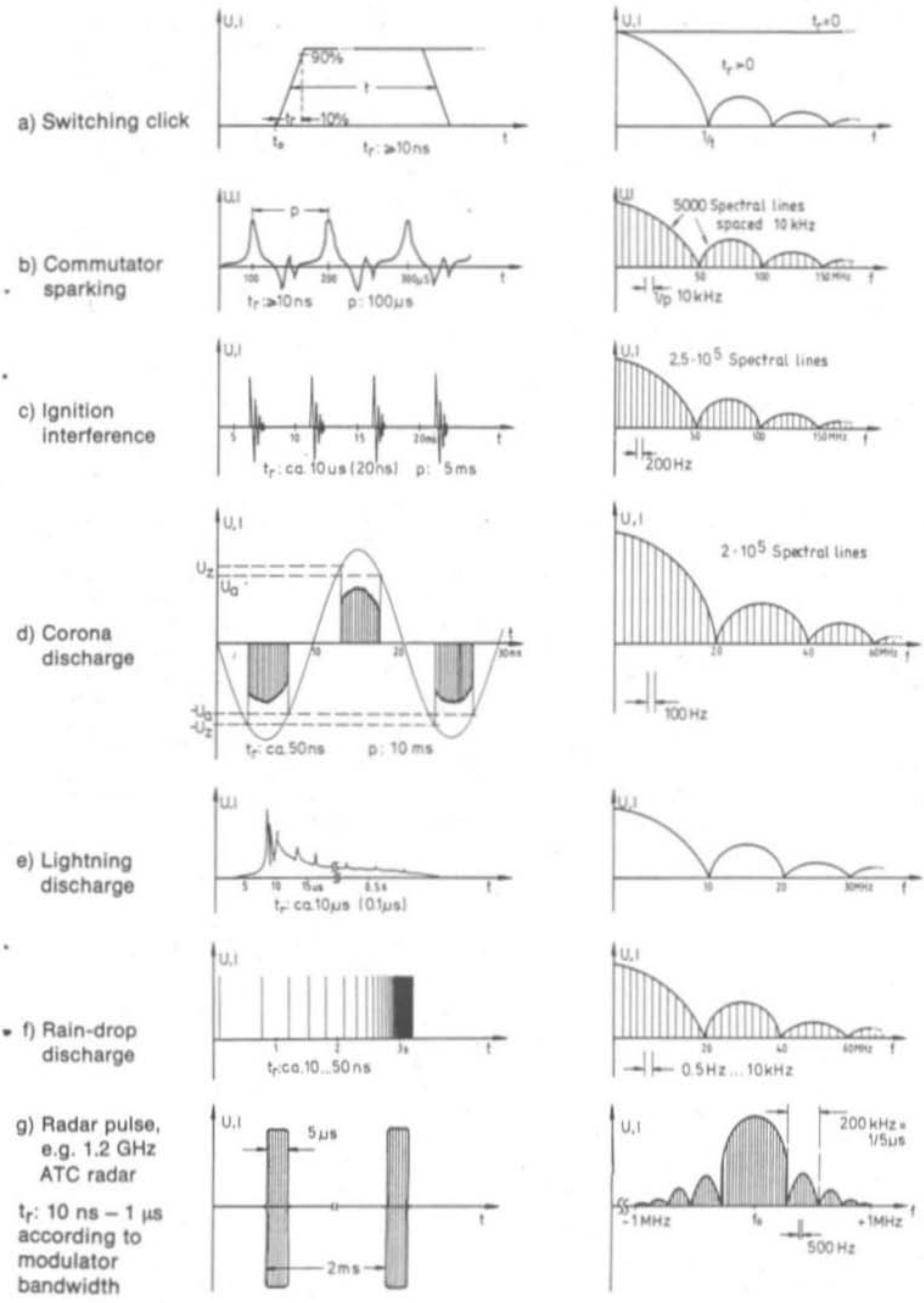
The clicks that appear on switching a current on or off represent the most simplest form of interference pulse **(Fig.1a)**. In this case, the energy of the interference spectrum is dependent on the amplitude (V, A) together with the rise time (slew rate) (V/s), the width of the rise time (t_r), measured between 10 and 90 % and the density of the repetition frequency. For wideband systems with a lowpass characteristic, the approximation formula $t_r = 0.3/\Delta f$ will be valid, where Δf = bandwidth in Hz and t_r = rise time in s.

In the case of a theoretical, vertical step with $t_r = 0$, i.e. infinite rise time, the spectrum will extend to infinite: The click will be heard at **all** frequencies ! With a finite rise time, the amplitude characteristic of the spectrum will have a Si-function and with increasing t_r the minima in the spectrum will be shifted towards lower frequencies. In practice, t_r will be at least 10 ns due to the effects of line inductance. This means that the main spectrum components of such a click interference are below 30 MHz, and their amplitudes are constantly decreasing above this frequency.

1.2. Commutator Flashover

Commutator flashover represents a special case of the switching interference shown in Figure 1a, in that a series of individual switching pulses are generated according to the type of machine, number of poles, and speed **(Figure 1 b)**. Repetition frequencies up to several kHz can be present. This results in a higher spectral energy density being generated according to this frequency factor than would be the case with a single switch. This means that sometimes the »rotating noise« of the machine can be heard even at 30 MHz. The spectral amplitudes can increase several orders of magnitude when arching occurs between the commutator and brushes. However, this does not often occur since it will quickly destroy the collector.

Fig. 1: Various types of impulses displayed in time scale and frequency range



1.3. Ignition Interference

Ignition interference is a high-voltage discharge in pressurized gas, and has a spectrum similar to that shown in Figure 1 b and a repetition frequency of less than 1 kHz **(Figure 1 c)**. In the case of conventional circuit-breaker ignition systems, the voltage amplitude is highest and the rise time shortest at low engine speeds. The spark gaps together with the high-tension cables connected between spark plugs and distributor form resonant circuits that act like a spark transmitter as were used in the early years of radio communications.

The oscillations decrease according to an e-function.

1.4. Corona Discharge

On a defective insulator of a 50 Hz high-tension line, a discharge can be ignited, which according to the »quality« of the insulator will commence more or less below the peak voltage. It will last until the voltage is reduced below a certain threshold, which is always lower than the starting voltage. This process will be repeated during the opposite halfwave of the AC-voltage **(Figure 1 d)**. In the case of a poor three-phase line, or in the case of fog (high humidity), an arching of all three phases could cause an overlap of the burning times since they occur every 3.33 ms !

1.5. Lightning Discharge

In relatively flat country, 85 % of all lightning strokes are in a »downward« direction, that means that the cloud is negatively charged and a current surge with up to 20 kA/μs goes to ground. According to (1), 80 % of these lightning flashes have maximum currents up to 50 kA, however it is also possible that currents amount to more than 200 kA when several clouds discharge via the same ionization path **(Figure 1 e)**. The duration of a lightning discharge can be between 0.1 ms and 0.5 s, depending on how many subsequent surges are induced by the primary discharge.

1.6. Rain-Drop Discharge

The charge transfer between a grounded antenna and statically charged rain-drops cause a click-type interference which can increase in frequency from an individual click up to a uniform white noise during heavy showers. Due to the very steep slope of this discharge, very high interference amplitudes are generated in the resonant ranges of the antenna. Although the capacity of a rain-drop only has several pF and is thus very rapidly discharged, the high potential and the large repetition frequency have the effect of a considerable energy density **(Figure 1 f)**.

1.7. Radar Pulses

In the case of radar transmitters in the GHz-range, as well as the shortwave over-the-horizon radar, a generator is modulated with the aid of a pulse modulator so that the reflected signal will return, according to the range, between the transmit pulses **(Figure 1 g)**. Depending on the keying mode with either a square wave or Gauss-type pulse, modulation sidebands will be generated of different amplitude and extent. As in the case of DC-voltage pulses, the spectral line spacing of RF-pulses will be determined by the repetition frequency. The frequency spacing of points in the spectrum zeros will result from the reciprocal pulse length.

2. EFFECTS OF PULSES ON NARROW-BAND SYSTEMS

Independent of the actual rise time given in section 1 for the various pulse interferences, a narrow-band system equipped with resonant circuit amplifiers will only allow those frequency components of the interference spectrum to pass that fall within its passband range. An individual resonant circuit will be excited to oscillation at its resonant frequency by a pulse slope, where the transient time is dependent on its Q, i.e. on its bandwidth. In the case of a multi-stage amplifier, or multipole crystal filter, the delay of a signal will be dependent on the group delay t_g. This will increase linearly with the number of resonators, and is calculated from switch-on time t_0 to the time at which the output signal has risen to 50 % of its amplitude. According to (2), the approximation formula gives $t_g = 0.35\ N \times 1/\Delta f$, with N = number of resonant circuits, Δf = bandwidth (Hz), and $t_r = 1/\Delta f$. If stages of differing bandwidth are connected in series, the rise time will be determined mainly by the narrowest filter, and the group delay will result from the sum of the individual delay times.

If RF-pulses having a very short rise time are fed to an amplifier with a far longer rise time, three different output signals can appear ([2], page 144): If the input pulse with the duration t_p is longer than the transient time t_{rv}, the output signal will achieve the full amplitude ($U_a = V \times U_{in}$) and will maintain the same for the duration of the drive time minus transient time $t_p - t_{rv}$. If t_p is equal to t_{rv}, the output voltage will obtain the value $V \times U_{in}$ during time t_{rv}, and will commence immediately afterwards with the down slope to zero for the duration of t_{rv}. If the pulse is shorter than the transient time, the rise and fall time process will be superimposed which means that the output signal can only increase to a portion of $V \times U_{in}$: the shorter the pulse, the less will be passed via the amplifier. In other words: Important spectral components of the input signal are not within the bandwidth of the amplifier and are therefore not present at the output for addition to the full amplitude.

In most systems (not FM), an AGC-circuit is usually provided to ensure that a large range of input signal voltages are brought to the same level at the demodulator output. Since the AGC-voltage generation is made subsequent to an IF-filter with a bandwidth of say 2.4 kHz, it will possess transient times in the ms-range. This means that steep pulses can already drive several amplifier stages into saturation, especially in the case of multi-conversion superhets where the selectivity is usually made at the lower end, before a reduction in gain is caused. This means that the audible interference amplitudes are then several times

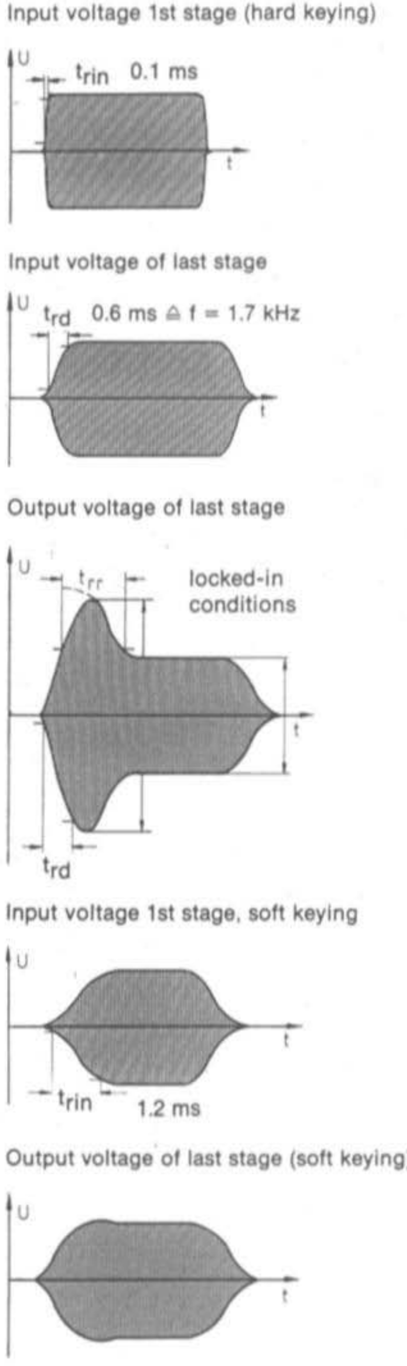


Fig. 2: Pulse drive of a RF-amplifier with AGC

louder than the required-signal level after AVC-leveling ! This effect is often worsened by the decoupling (with low-pass characteristic) between the AVC-amplifier and the IF-stages. It is only when the rise time of the input pulse is longer than the transient time of the AVC (e.g. during telegraphy with »soft« keying), that the output amplitude will not overshoot **(Figure 2)**. An effective method of limiting the maximum drive of the demodulator and thus to obtain more favorable conditions for the listener, is to clip the signal just in front of this with anti-phase diodes to ensure that the IF-amplifier is not able to provide more than 0.7 V (peak-to-peak value). It is necessary, however, for the AVC-voltage diode to be biased with approximately 0.4 V in order to ensure that the clipping process will not interfere with the AVC-voltage generation.

3. METHODS OF SUPPRESSING THE INTERFERING PULSES

All interference suppression methods are based on the principle of »opposite modulation«. This means that a stage in the required-signal path is modulated so that the signal path is blanked during the amplitude-modulation process for the duration of the interference. When using the frequency modulation method, the signal path is shifted into a different frequency range. The latter was described in (3) and utilizes the attenuation overlapping of the IF-filters in a double-conversion superhet. The second oscillator is swept several kHz from the nominal frequency during the duration of the interference so that the filters of the second IF are shifted with respect to the first IF which means that the gain is reduced down to the value of the ultimate selectivity according to the slope of the filter curves. This method is especially advantageous since the switching spikes on switching off the signal should not be noticeable. However, when using an FM-modulator having a high speed, i.e. large bandwidth, spectral components that are within the bandwidth of the second IF can appear du to modulation. The most stringent limitation of its application is, however, that two virtually identical, narrow-band filters with different frequencies must be provided together with an intermediate mixer. This means that the concept is limited to a double-conversion superhet with variable first oscillator. When using the amplidute modulation method, two types of processing are possible:

(A) The interference signal is tapped off in parallel at the input of the system and increased to the trigger level of the blanker by an interference-channel amplifier having a passband which is far different from the signal path. This method is only effective against very wideband interference since noticeable interference energy components must fall into the passband range in order to allow triggering. This method will not be effective in the case of narrow-band interference such as radar pulses within or directly adjacent to the frequency range to be received.

(B) The interference signal is tapped off from the required-signal channel directly subsequent to the mixer (4, 5) and fed to a fixed-frequency, second IF-amplifier where it is amplified up to the triggering level. However, since there is a danger of cross-talk from the interference channel to the required-signal amplifier, it is advisable to use a frequency conversion in the interference channel so that the main amplification is made at a different frequency than that of the signal IF. Furthermore, attention must be paid when using this method that no switching spikes generated during the blanking process can be fed back to the interference channel tap-off points, since there is a danger of pulse feedback. The return attenuation must therefore exceed the gain in the interference channel between the tapping point and the blanker.

The blanker must be placed ahead of the narrowest IF-filter in the signal path, and must be able to blank before larger components of the transient have passed this filter. To ensure this in the case of a small group delay in the interference channel, its bandwidth has to be sufficiently great, and too many subsequent resonant circuits should not be used. It is most effective to insert a delay between the interference-channel tap-off point and the signal path blanker so that there exists sufficient time

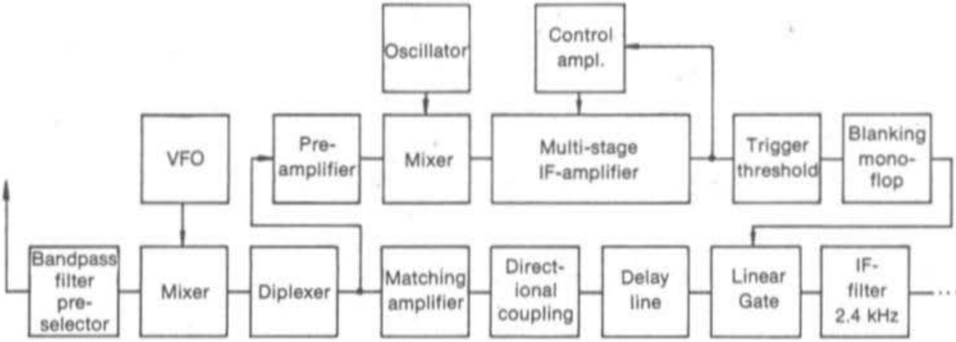


Fig. 3: Block diagram of an interference blanker

for processing the interference signal. In this case it is not necessary to make the interference-channel bandwidth excessively wide, and it will be possible to suppress the residual peak.

The operation of an effective interference blanker in a single-conversion superhet receiver is shown in the block diagram given in **Figure 3** and in the pulse diagram given in **Figure 4** (6). In this case, radar pulse interference is fed to the receiver input, whose required signal is an unmodulated carrier (a, b). After the mixer, the rise slope of the pulse is lengthened by the transient time of the preamplifier circuit and the diplexer (c) and its group delay adds to $t_{g1} = 1\ \mu s$. At the output of the interference-channel circuit, it is additionally delayed by the group delay $t_{gst} = 4\ \mu s$ (d). If blanking takes place now (e), the interference pulse will already have passed the linear gate for the duration of $t_{g2} = t_g1 + t_{gst} = 5\ \mu s$ and the subsequent IF-amplifier will commence its transient process. This means that a residual interference peak of approximately 3 µs remains (f). It is true that a lowering of the trigger threshold as well as an increase of the bandwidth of the interference channel with resulting reduction of t_{g2} will reduce this further, however, both measures have their limits: If the threshold is too low, it is possible that the required signal is processed as interference so that the linear gate will remain blocked. If the bandwidth of the interference channel is too wide, strong signals in the vicinity of the required frequency may override the interference. After inserting the delay $t_{gd} > t_{g2}$ into the required-signal path, (g) the residual peak can be suppressed (h). In the case, the value of the interference suppression coincides exactly with the blanking attenuation of the linear gate.

4. DEMANDS ON THE INTERFERENCE BLANKING CIRCUIT

Since the insertion of an interference blanking circuit into the signal path of the receiver should not cause any reduction in its dynamic range, stringent requirements have to be placed on the individual components. Very linear receiver input circuits with a dynamic range of more than 90 dB and intercept point values up to 30 dBm on the shortwave bands and up to 10 dBm in the VHF-range as described in (6, 7) were therefore modified.

4.1. Linear Gate

In the beginning, a so-called linear gate was examined as modulator which was supposed to work in an interference blanker accessory for a receiver-frequency range of 1.5 to 30 MHz. The optimized

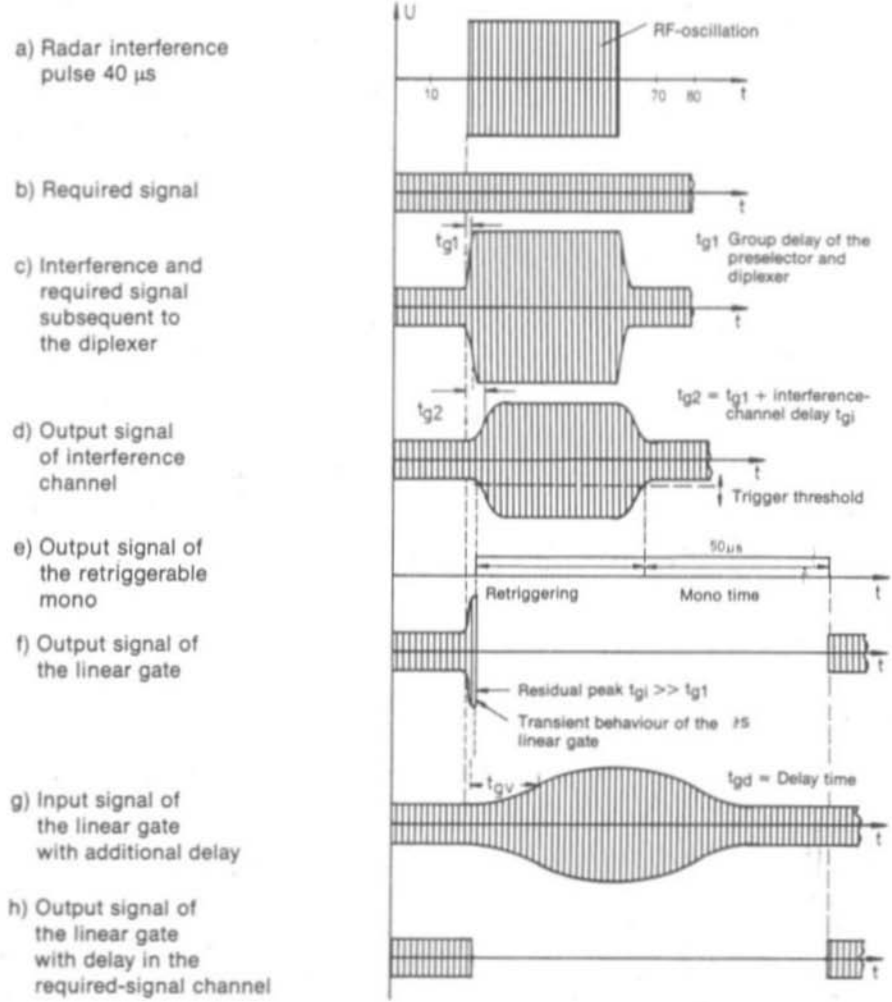


Fig. 4: Operation of the impulse blanker shown in Figure 3

circuit **(Figure 5)** using four diodes 1 N 4148 consists of balanced wideband transformers, together with a low-pass filter which is provided to improve the stopband and passband behaviour at higher frequencies. When balanced in amplitude and phase, it will exhibit a very good suppression of the switching spikes, which means that only very little interference will be fed to the subsequent stages of the receiver even when the blanking frequencies are in the kHz-range.

In order to ensure that the balance and thus the suppression of the switching peaks are independent of the ambient temperature, the diodes should be selected to have the same characteristics, and be coupled thermally to another.

Distortions caused in the linear gate were measured using the two-tone method and this resulted in an intercept point of IP = 31 dBm at a diode current of 15 mA per branch. At 10 mA each, it amounted to only

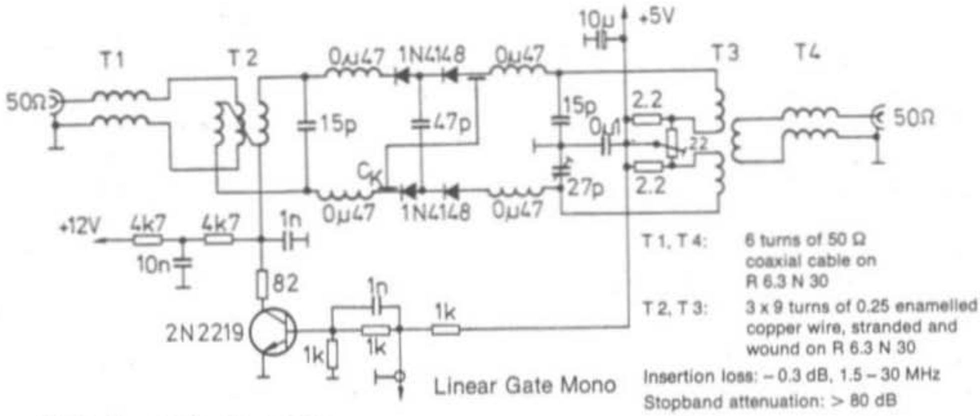


Fig. 5: Linear gate 1.5 - 30 MHz

20 dBm, which shows quite easily the large-signal handling capabilities of many professional blankers, which operate with only a few mA total diode current. The wideband blanking suppression of the circuit of more than 80 dB is achieved with the aid of bridge compensation of the diode junction capacitance of approx. 0.1 to 0.5 pF.

4.2. Directional Coupling and Matching Amplifier

The return attenuation necessary to avoid pulse feedback from the blanking circuit must be higher than the gain in the interference-channel amplifier. The better the balance of the linear gate, the lower the attenuation required. At an interference-signal gain of more than 100 dB and a common mode attenuation of more than 40 dB, the return attenuation must amount to approximately 60 dB. A favorable directional coupling circuit with a return attenuation of more than 80 dB is formed by common-gate circuits equipped with the high-current field effect transistors type P 8000, as shown in **Figure 6**. At a transconductance of 20 mA/V they exhibit a real input impedance of 50 Ω at a noise figure of approximately 3.6 dB. Their intercept point amounts to more than 30 dBm when the output lead impedance and supply voltage are optimum ! They are therefore very suitable for use as termination for high-power Schottky mixers that require a real 50 Ω termination for all frequencies at the IF-port if their maximum IP-values are to be achieved [see also (6)]. 50 Ω terminations with a bandwidth of several 100 MHz can be obtained using a diplexer and P 8000 amplifier, and at 9 MHz IF frequency it is possible to keep the diplexer attenuation below 0.4 dB (7).

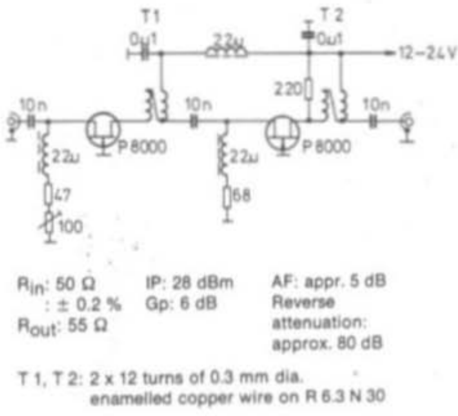


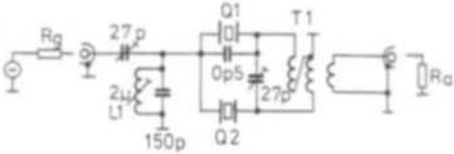
Fig. 6: Directional coupling with FETs

4.3. Delay Line

Several different types of signal delay were examined for suppression of the residual peak: Delay lines using cables with intermediate amplifiers were too extensive and large. When using n-stage bandpass filters aligned to the intermediate frequency, it was possible to obtain a delay of n-times the group delay, but the insertion loss increased with each further circuit and became too high.

A two-pole crystal filter with an insertion loss of 2 to 3 dB and a delay time of 20 µs at a bandwidth of approx. 20 kHz was tried as alternative. A simple half-lattice filter provided the required result, however, unfortunately exhibited an intercept point of only 16 dBm. A closer examination of the intermodulation behaviour of the filter showed distortion in the ferrite differential transformer. This was confirmed with the aid of measurements which showed that third-order intermodulation products can easily be caused by high-Q resonant circuits on small toroids due to the very high magnetizing flux in the cores. Information regarding mathematical evaluation of intermodulation in crystal filters was found in (8) and (9). After making several modifications it was possible to increase the IP-value of the filter to 30 dBm, which coincides approximately with the value calculated.

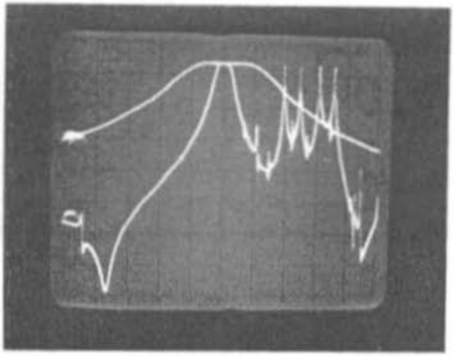
Fig. 7: Two-pole crystal filter with high intermodulation rejection when used as delay line



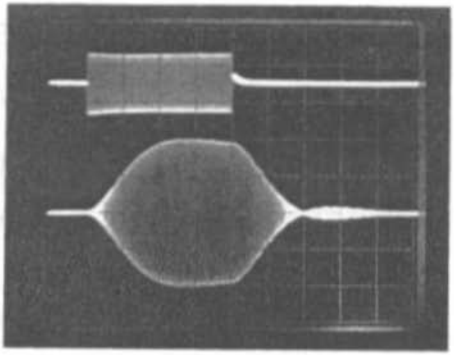
L 1:	12 turns of 12 x 0.04 mm dia. stranded wire wound in special coil sets
T 1:	2 x 11 turns of 0.3 mm dia. enam. copper wire on R 10 K 1 + 2 coupling turns
R_{in}:	50 Ω
R_{out}:	50 Ω
Gp:	– 2 dB
Δf:	20 kHz
Q 1:	9008.5 kHz or CB crystal 27.025 MHz
Q 2:	8991.5 kHz or CB crystal 26.975 MHz
IP:	30 dBm

Transient time:	t_r = 45 µs
Delay:	t_g = 20 µs
Transient time to 5 %:	5 µs
Ripple:	< 0.5 dB

Figure 7 shows the circuit diagram. The obtained frequency response and the transient time are given in **Figure 8**. A positive side effect of this filter is that all subsequent stages up to the actual SSB-filter with a bandwidth of 2.4 kHz are not loaded by signals that are more than ± 10 kHz from the receive frequency.



Horizontal:	10 kHz/Div. 50 kHz/Div.
Vertical:	10 dB/Div.
Center frequ.:	9 MHz



Horizontal:	20 µs/Div.
above:	Input signal at diplexer
below:	Output signal of the crystal filter

Fig. 8: Frequency response and transient time of the filter shown in Figure 7

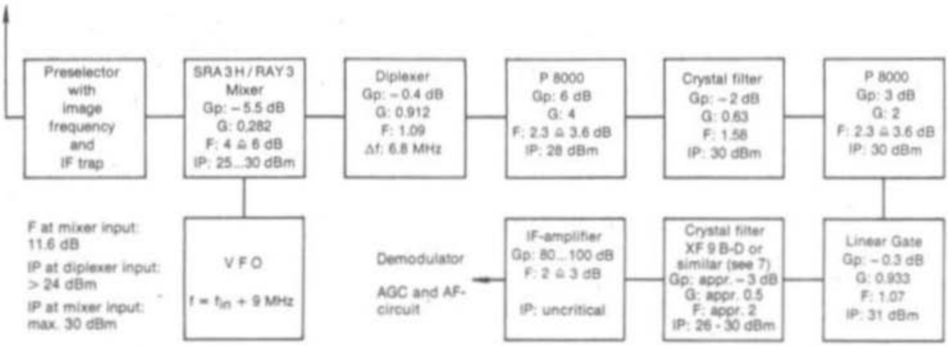


Fig. 9: Block diagram with level plan of the receiver input circuit when using an optimized interference blanker

This means that their IP-value is virtually increased by the value of the actual attenuation of the delay filter. This also has an effect on IF-amplifiers equipped with crystal filters of the XF 9-series (KVG) at the input. The IP of such filters ranges between 26 and 30 dBm according to the bandwidth. The narrow-band filters have the lower IP-values.

Practical circuits and a PC-board are to be described in Part 2.

4.4. Level Plan

A level plan using the described components for the required-signal path of the receiver is given in the block diagram **Figure 9**. According to the calculations given in (7), the noise figure at the input of the mixer was calculated to be 11.6 dB. Preamplifier and selectivity will alter this value according to the well-known sum formula. The intercept point value for a two-tone test with two carriers having a spacing of more than 10 kHz exhibited values up to 30 dBm.

A description of the different trigger circuits and boards is to follow in part 2 of this article.

5. REFERENCES

(1) K. Berger and E. Vogelsanger:
Messungen und Resultate der Blitzforschung der Jahre 1955-63 auf dem Monte Salvatore
Bulletin SEV 56 (1965) 1, January 9, pages 2 - 22

(2) R. Feldtkeller:
Rundfunksiebschaltungen
Hirzel-Verlag Leipzig 1945, 3rd edition

(3) R. T. Hart:
Blank Noise Effectively with FM
Electronic Design No. 18,
1. September 1978, pages 130 - 132

(4) M. Martin, DJ 7 VY:
Die Störaustastung
cq-DL 11/1973, pages 658 - 663

(5) M. Martin, DJ 7 VY:
Moderner Störaustaster mit hoher Intermodulationsfestigkeit
cq-DL 7/1978, pages 300 - 302

(6) M. Martin, DJ 7 VY:
Empfängereingangsteil mit großem Dynamikbereich
cq-DL 6/1975, pages 326 - 336

(7) M. Martin, DJ 7 VY:
A Modern Receive Converter for 2 m Receivers having a large Dynamic Range and Low Intermodulation Distortion
VHF COMMUNICATIONS 10, Edition 4/1978, pages 218 - 229

(8) Robert C. Smythe:
Some Recent Advances in Integrated Crystal Filters
Proceedings IEEE Vo. 67, No. 1, Jan. 1979, pages 119 - 129

(9) W.H. Horton and R.C. Smythe:
Experimental Investigations of Intermodulation in Monolithic Crystal Filters
Proceedings of the 27th Annual Frequency Control Symposium 1973, pages 243 - 245

Receive Mixer for the 6 cm Band

by R. Heidemann, DC 3 QS

A relatively simple receive mixer is to be described as a contribution to amateur activity in the virtually neglected 6 cm band. Usually, interdigital filter-type mixers (1) are used in the lower microwave bands of 23 cm, 13 cm and 9 cm. The author obtained a single-sideband noise figure of 11 dB in the 9 cm band when using an interdigital filter converter, but a single-sideband noise figure of only 7 dB when using a waveguide mixer. This improvement of 4 dB led to the construction of a waveguide mixer for the 6 cm band.

A further disadvantage are the small dimensions of the interdigital filter converters for the 6 cm and 3 cm bands, and thus the problems of installing the mechanically large standard mixer diodes type 1 N 23 and BAW 95. Mixers constructed from normal profiles waveguides are not suitable for the 23 cm and 13 cm bands due to their large mechanical dimensions, however, the dimensions of waveguide mixers for the 9 cm, 6 cm, and 3 cm bands allow construction of portable transverters.

PRINCIPLE OF OPERATION

Up to now, microwave receivers using waveguide mixers have mainly used the principle shown in **Figure 1**. The input

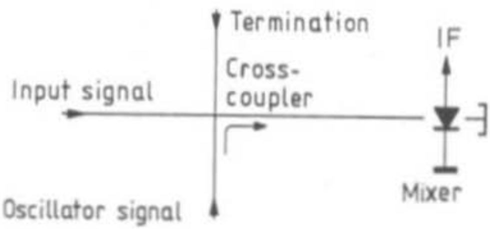


Fig. 1: Conventional microwave waveguide mixer

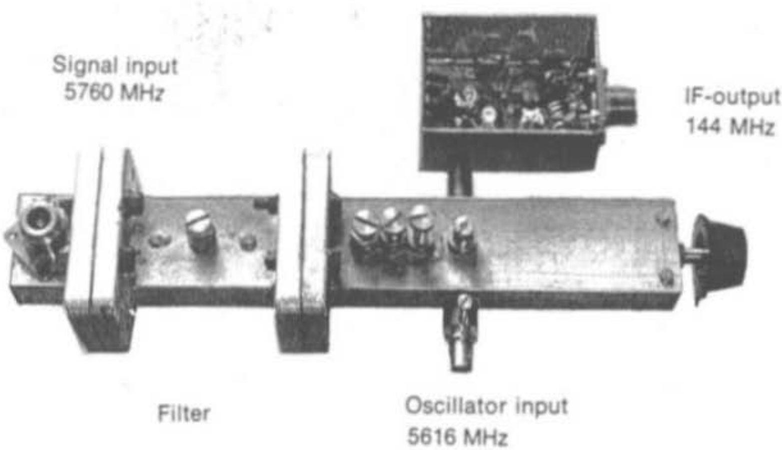


Fig. 2: The author's prototype with IF-preamplifier for 144 MHz

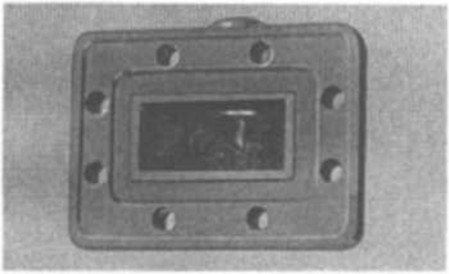


Fig. 3: Waveguide/coaxial transition

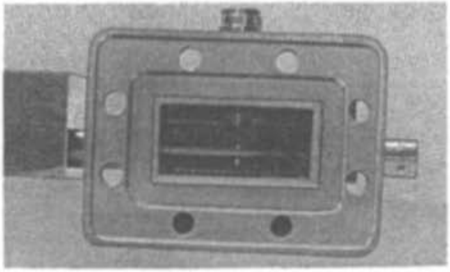


Fig. 4: Mixer with flange and short-circuit plunger

signal is fed directly to the mixer diode, and the oscillator signal via a directional coupler. A cross-coupler is suitable as directional coupler, since it can be constructed simply, and secondly because wideband characteristics are not necessary for amateur radio applications. A considerable disadvantage of this method is the relatively large coupling loss of the cross-coupler of, for instance, 20 dB, which means that an oscillator power of 50 to 100 mW will be required to obtain sufficient diode current.

In the case of the mixer described here, no directional coupler is required and very little oscillator power will be required due to the direct capacitive injection of the oscillator signal (< 1 mW). A filter for suppressing the image frequency is integrated in the mixer module. The input signal is fed in either via an N-connector or a R 70 standard waveguide flange. Conventional mixer diodes of the 1 N 23- or BAW 95 family are used.

CONSTRUCTION AND SPECIFICATIONS

A photograph of the author's prototype is given in **Figure 2**. In order to optimize the individual modules: filter, mixer and waveguide/coaxial transition **(Figure 3)**, they have been equipped with waveguide flanges. Furthermore, the diode matching can be adjusted with the aid of a short-circuit plunger, as can be seen in **Figure 4**.

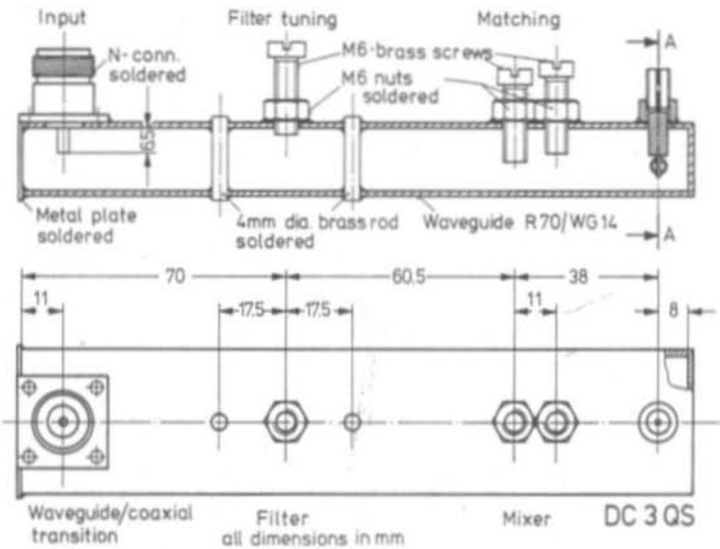


Fig. 5: A waveguide mixer for 5760 MHz

Since both the flanges and the short-circuit plunger represent a considerable extent of fine metalwork, and their use will not bring any distinct advantages during construction, it is possible for them to be deleted. It is possible to use a 176.5 mm long piece of R 70 (WG 14, WR 137, RG 50/U) waveguide.

Figure 5 shows the constructional diagram of the mixer as seen from the side and above. The signal frequency input at 5760

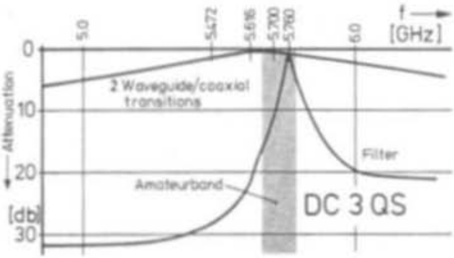


Fig. 6: Passband curves of two transitions and a simple filter

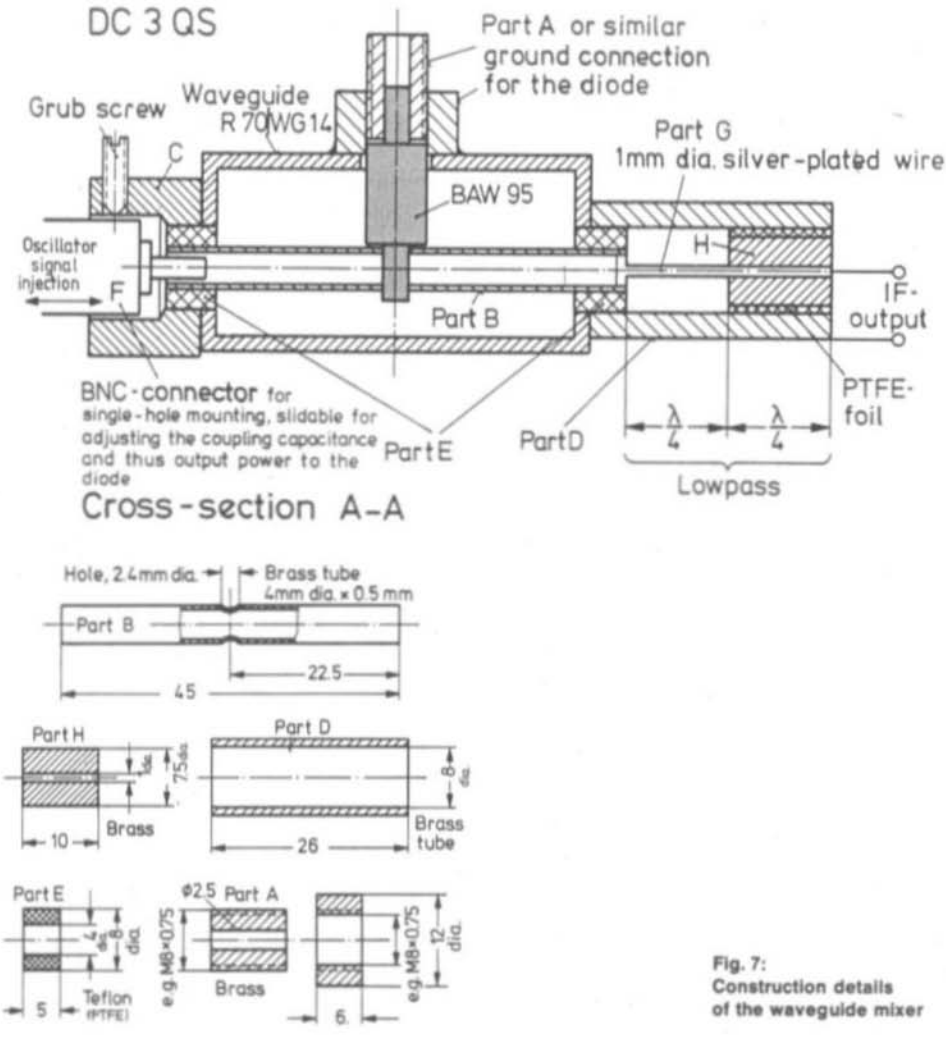


Fig. 7: Construction details of the waveguide mixer

MHz is on the left-hand side. It is possible for the coaxial input (N-connector UG-58 A/U) to be exchanged for a waveguide flange (no modification is necessary to the inner conductor). The waveguide/coaxial transition is followed by a single-stage waveguide filter of simple construction. The passband curve of this filter and of two waveguide/coaxial transitions is given in **Figure 6**. It can be seen here that the transitions provide negligible loss within the 6 cm amateur band. The following attenuation values are exhibited by the filter:

	Frequency	Attenuation
Input frequency	5760 MHz	1 dB
Oscillator frequency	5616 MHz	22 dB
Image (IF 144 MHz)	5472 MHz	29 dB
Image (IF 28 MHz)	5700 MHz	14 dB (!)

3 dB bandwidth: 25 MHz

The adjustment screws of the filter for matching the signal and image frequency to the diode are followed by the mixer, which is shown in detail in **Figure 7** (cross-section A-A in Figure 5). The mixer diode is clamped between ground connection A and the center conductor B. Part B is held at both ends in PTFE collars (part E). The lathed part C is soldered into place centrically to tube B on one side panel of the waveguide. The threading should be lathed from a BNC-connector for single-hole mounting F (UG-1094/U), after which it can be shifted axially in part C. This is the input connector for the local oscillator signal of 5616 MHz. The coupling capacitance between B and the inner conductor of connector F can be varied by sliding this connector (F), and it is thus possible to adjust the oscillator power at the mixer diode.

The IF-signal (144 MHz) can be taken from the opposite side of the waveguide via the lowpass filter comprising parts D, G, and H. The filter is constructed as follows: Tube D is soldered into place centrically to part B, and ring E (PTFE) pushed into place. Part G (1 mm dia. silver-plated copper wire) should be soldered into B, after which part H should be placed over G and soldered together. Part H should be covered with PTFE foil (0.05 mm) and parts B, G, and H pushed into the waveguide. Insert a defective mixer diode to center, and lock into place with the ground connection A. Connect the IF-preamplifier directly, without intermediate cables (impedance matching !).

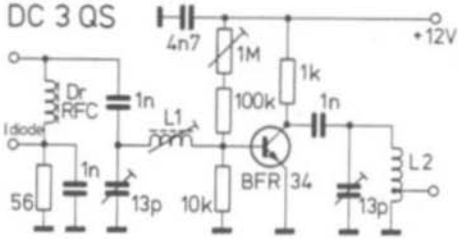


Fig. 8:

144 MHz IF-preamplifier matching the 6 cm converter

L 1: 4 turns of 1 mm dia. silverplated copper wire wound on a 4 mm coil former with VHF core

L 2: 5 turns of 1 mm dia. silverplated copper wire wound on a 6 mm former, self-supporting, coil tap: 1 turn from the cold end

The standard circuit is given in **Figure 8** and a prototype construction shown in **Figure 9**. When using other circuits, attention should be paid that the input impedance of the preamplifier should be transformed to an impedance of $Z \approx 300\ \Omega + j\ 0\ \Omega$, which is the most favorable impedance of the IF-circuit for diodes type 1 N 23 or BAW 95.

CONNECTION AND ALIGNMENT

For preliminary alignment of the mixer, a local oscillator signal should be fed to the BNC-connector F and a 6 cm signal, or harmonic of a sufficiently strong 1152 MHz signal fed to the input. This is followed by aligning the input filter, the tuning screws, and the IF-preamplifier (L 1, C 1, L 2) for maximum signal. In the case of diodes of the 1 N 23 family, the diode current should be in the order of 0.2 and 0.5 mA, or of 1.0 mA in the case of Schottky diodes

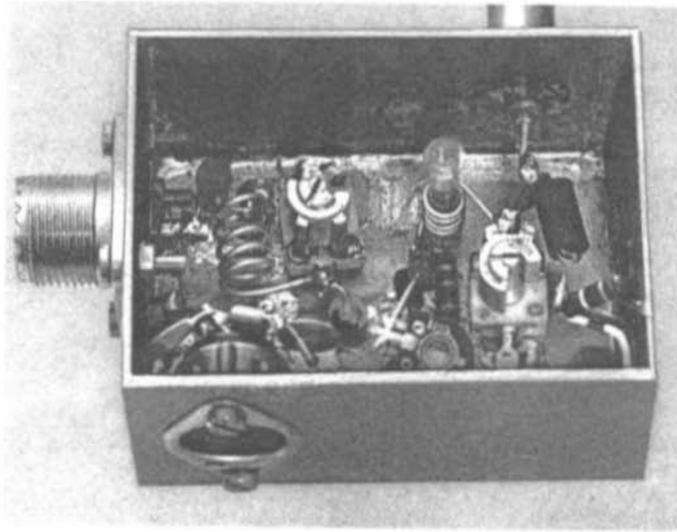


Fig. 9:
Photograph of the author's prototype 144 MHz preamplifier

similar to the BAW 95. The diode current is adjusted by varying the oscillator power and/or the coupling with the aid of connector F. The author obtained the minimum single-sideband noise figure in the order of 9 dB in conjunction with a noise generator to obtain the most favorable signal and image frequency matching to the diode.

REFERENCES

(1) R. E. Fisher, W 2 CQH :
Interdigital Converters for 1296 and 2304 MHz
QST, Jan. 1974, pages 11 - 15

SSB on the 10 GHz Band

Information regarding a future description in VHF COMMUNICATIONS

by H. Fleckner, DC 8 UG
and G. Börs, DB 1 PM

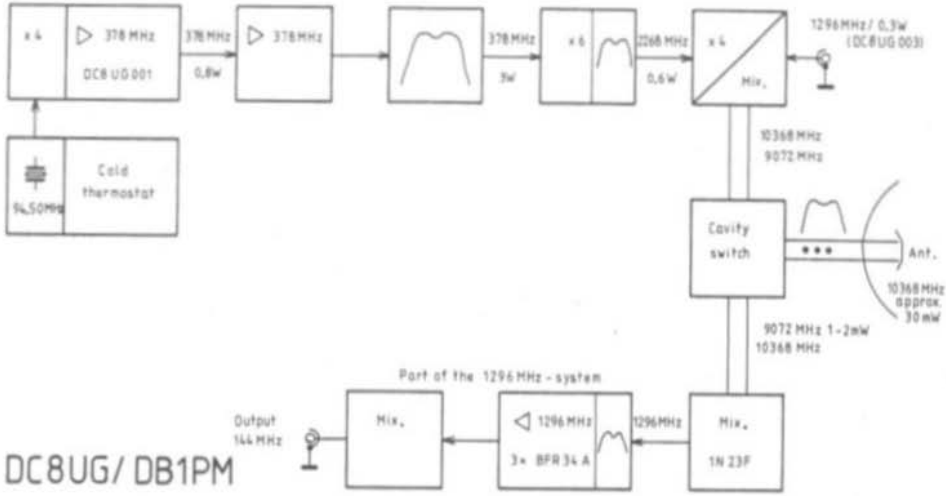
Virtually all communications on the 10 GHz band are still made using wideband FM systems equipped with Gunn-oscillators, which were described in detail in (1), (2), (3). Articles regarding narrow-band systems have been relatively rare up to now (4), (5), since the technical complexity is considerably greater than when using Gunn-oscillator circuits.

A comparison between systems shows that a narrow-band system for SSB or CW will offer a reduction of 14 dB of the noise power in the receive channel. This is achieved by the reduction of bandwidth and suppression of the first IF-image, but does not take any further improvements, for instance, due to the use of a low-noise preamplifier (6), (7) into consideration.

A sensitivity increase in the order of 14 dB allows communication under non-line-of-sight conditions (for instance via reflections (6)) and over far larger distances and under unfavorable weather conditions. This means that it is extremely worthwile for the microwave amateur to consider the higher technical requirements of narrow-band technology, especially in conjunction with fixed operation.

The authors have constructed two SSB systems for the 10 GHz band which are to be described in the next editions of VHF COMMUNICATIONS. The block diagram of such a system is shown in **Figure 1**. The frequency plan corresponds to the recommendations given in (4), however, the varactor types, the individual stages, and their interconnections have been partly, or completely changed. This results in a unit which has been simplified to make it especially suitable for portable operation. During the first few weeks of operation, it was possible using the systems for SSB-communications to be made over distances of 20 and 37 km, as well as one-way communication over 25 km under non-line-of-sight conditions.

The authors plan to describe such SSB equipment for the three intermediate frequencies of 144 MHz, 432 MHz, and 1296 MHz, as well as to describe a transceiver without circulator or cavity switch. The task of these constructional articles is to activate the narrow-band technology at 10 GHz, and to provide fundamentals for discussions and further developments.



DC8UG/DB1PM

REFERENCES

(1) Dr. D. Evans: Getting Started on 10 GHz
VHF COMMUNICATIONS 9,
Edition 1/1977, pages 19-29

(2) B. Heubusch, Dr. Hock, H. Knauf:
Introduction to Microwave Techniques.
A Description of a 10 GHz Transceiver
VHF COMMUNICATIONS 9,
Edition 2/1977, pages 66 - 70

(3) J. Reithofer:
A Transceiver for the 10 GHz Band
VHF COMMUNICATIONS 11,
Edition 4/1979, pages 208 - 215

(4) C. Neye: Transverter for Transmitting
and Receiving 10368-10370 MHz/
1296-1298 MHz Narrowband
DUBUS, VHF-UHF Technik

(5) R. Griek and M. Münich:
A Frequency Multiplier for Narrow Band
3 cm Band Communications
VHF COMMUNICATIONS 11,
Editions 2/1979, pages 66 - 73

(6) D. Vollhardt: The 10 GHz Amateur Band
Consideration of Present and Future
Technologies
VHF COMMUNICATIONS 10,
Edition 4/1978, pages 244 - 251

A Microcomputer for Amateur Radio Applications
Part 1: Introduction

by W. Kurz, DK 2 RY

As may be already known, AMSAT intend to soon launch the first Phase III OSCAR satellite. This amateur transponder satellite will possess an elliptical orbit in contrast to the circular low altitude orbits of previous satellites such as OSCAR 7, 8 and the Soviet satellites RS. This means that it will be more difficult to calculate the antenna tracking, and will make such aids as the OSCALATOR obsolete. There are two methods of calculating the azimuth and elevation angles for the antenna:

1. Calculation using a programmable calculator
2. Calculation using a microcomputer

Possibility 1 has the advantage of having simple programmability, however, the actual control of the rotators into the calculated position must still be made manually. It is more elegant for this to be made in a microcomputer. The disadvantage of this is that a suitable program (assembler) must be developed. This has probably been the reason why such computers have not become available. Even though the expense and effort are higher, a micro-computer can be utilized for a multitude of additional tasks.

DESCRIPTION OF A MICROCOMPUTER SYSTEM

A microcomputer operates according to the same basic principles as its larger brothers. However, they are faster and are even more »intelligent«. The principle of the microcomputer is as follows:

The heart of the system is the central processing unit CPU where all operations take place. This processing unit requires a memory, which comprises a number of binary storages called random access memory RAM. Of course one must tell the CPU what to do and how. This is achieved using a program which is contained in a separate memory, usually a fixed program storage. Such a storage is called a read-only memory ROM or programmable read-only memory PROM.

Finally it is necessary for the microcomputer to communicate, which requires periphery units (input and output modules). It is now necessary for these units to be interconnected, which is made with the aid of the system bus. **Figure 1** shows such an arrangement.

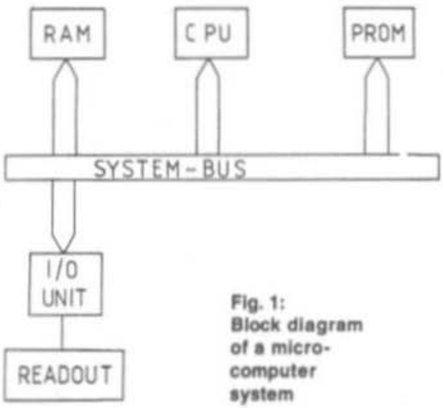


Fig. 1: Block diagram of a microcomputer system

How is a microcomputer programmed ?
There are several ways of doing this:

1. Programming in a higher language such as BASIC, FORTRAN etc. In this case one will require compilers or interpreters to translate the commands into the machine code.
   Advantage: Easy to program.
   Disadvantage: Very expensive.
2. Programming in the assembler (machine-orientated language).
   Advantage: Cheaper than 1, however, one requires an assembler.
   Disadvantage: Difficult to program.
3. Programming in machine language.
   Advantage: Very cheap.
   Disadvantage: Extremely difficult to program.
4. Purchase of a prepared program.
   Advantage: Can be used even without programming experience.
   Disadvantage: Can only be used for the actual application.

Possibility 4 will no doubt be the most favourable for radio amateurs who only require the microcomputer to assist them in their hobby. For this reason, programs will be offered later in the form of PROMs for the described applications.

APPLICATIONS

The first program is to be developed for the main application: Control of a rotator system in both the azimuth and elevation. This is to allow simple operation for communications via satellites, especially those with elliptical orbits. Our well-known author DK 1 OF already described a digital rotator controller with BCD-inputs in VHF COMMUNICATIONS No. 4/1979.

In addition to this, further applications are to be considered such as coder and decoder for Morse code, especially designed for meteor-scatter applications. A program is also to be developed for RTTY. Finally, we have also thought of the contest operator and his difficulties in logbook-keeping. A program is also to be developed that will not only keep the log during the contest, but will also check whether the station in question has already been worked. If not, it will output the contest number and distance to the station. Other programs will be developed as required to round off the system; such programs could be, for instance, in the form of monitor programs for TV-receivers as PROMs.

DEVELOPMENT

The microcomputer system comprises the following modules, which are to be described one after another in the following editions of VHF COMMUNICATIONS:

1. CPU board with buffered data inputs and outputs, and buffered address outputs.
2. Fixed-program board with a maximum of 48 K storage, as well as a Random-access memory with a maximum of 48 K storage. The bus board.
3. Input/Output unit with three parallel positions of two channels eachs.
4. Clock, real-time clock, number-crunching processor.
5. Power supply.
6. TV-interface.
7. Mechanical part.

The microcomputer is designed that it can be extended to have a maximum of 64 K storage capacity if required.

ANTENNA NOTEBOOK

Use of 75 Ω Low-Loss CATV-Coaxial Cables in 50 Ω Antenna Systems

by T. Bittan, G 3 JVQ / DJ 0 BQ

1. INTRODUCTION

Professional and amateur radio antenna systems are virtually standardized at 50 Ω throughout the world. North American radio amateurs have the advantage that the same impedance is also used for consumer electronics systems: Community antenna systems and CATV. This means that low-loss CATV cables are available readily and inexpensively on the market in the USA. Unfortunately, this is not the case in Europe where consumer coaxial systems have been standardized at an impedance of 75 Ω. Since far larger quantities of CATV-cables are required, they are far cheaper than similar quality cables for an impedance of 50 Ω.

One may ask why 75 Ω was chosen for such CATV-applications in Europe rather than 50 Ω ? It is interesting to know that the loss of coaxial cables of the same quality and diameter is not the same for all impedances but has a minimum at a certain ratio of inner and outer conductor. Logically one would assume that the loss of the cable with the smaller inner conductor diameter (75 Ω) would be greater than at the larger diameter (50 Ω). However, this is not the case. It will be seen in **Figure 1** that the loss of a coaxial line is a minimum at a ratio of 3.6 between the inner diameter D of the outer conductor and the outer diameter d of the inner conductor (1).

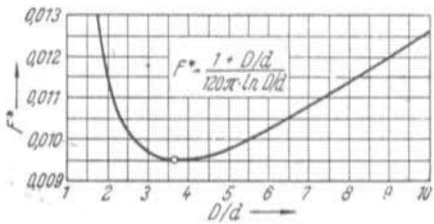


Fig. 1: An additional loss-factor F* of a coaxial line as a function of the D/d ratio

When using the equation:

$$Z = 138 \log_{10} \frac{D}{d}$$

this will correspond to an impedance of 77 Ω for air-spaced coaxial cables. It is therefore assumed that this was the reason for selecting an impedance of 75 Ω for CATV-applications in Europe.

In **Figure 2**, the loss characteristics of three diameters of CATV-cable are compared to the well-known cable types RG-58/U (UR-43) and RG-213/U. It will be seen clearly when comparing the loss of the RG-213/U cable to that of the same size CATV-cable (10.5 mm diameter) that the loss of the CATV-cable is considerably lower: only 14.3 dB/100 m at 1250 MHz compared with 26 dB / 100 m with RG-213/U, which cannot be ignored.

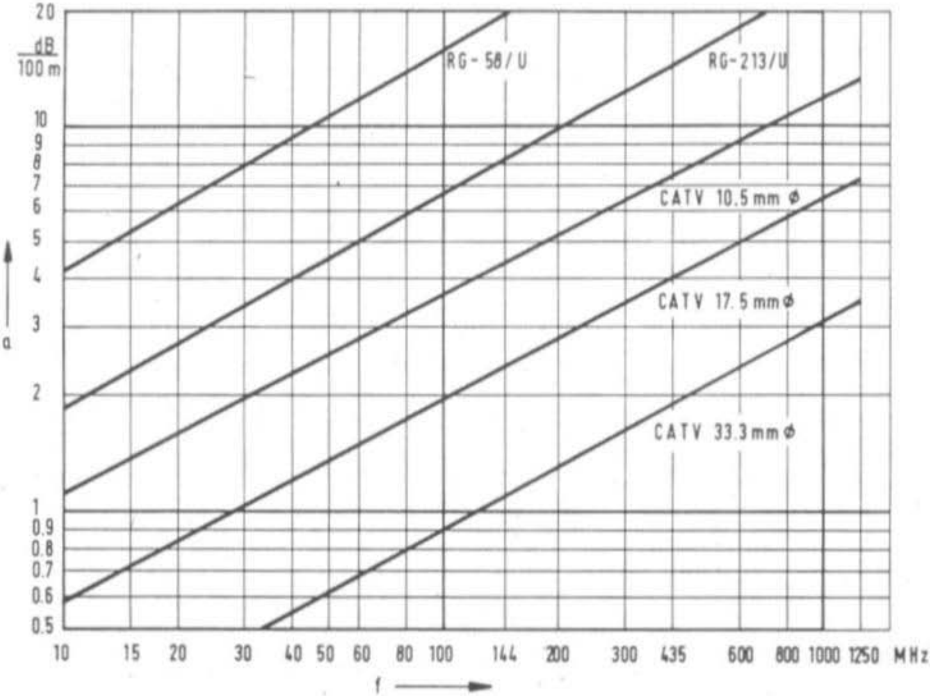


Fig. 2: Loss characteristics of several coaxial cables

Of course the difference in loss is not only due to the difference in impedance but mainly due to the construction of the CATV cables; the above mentioned CATV-cables are semi-airspaced types that are available in Europe.

2. THE EFFECT OF USING 75 Ω COAXIAL CABLES IN 50 Ω SYSTEMS

Many radio amateurs directly connect their 50 Ω impedance antennas to either surplus 60 Ω, or 75 Ω cables without further measures, assuming that the loss due to the VSWR of 1.2 or 1.5 is negligible, which is true to a certain extent. However, the length of the coaxial cable will become critical with **any** amount of VSWR, and will cause an increase or decrease of the actual mismatch according to the electrical length.

The worst case would be at an electrical feeder length of an odd multiple of λ/4 where the following equation is valid:

$$Z_{out} = \frac{Z_{cable}^2}{Z_{ant}}$$

or for instance $Z_{out} = \frac{75^2}{50} = 112.5\ \Omega$

Where: Z_{out} is the impedance at the end of the cable; Z_{cable} is the impedance of the coaxial cable used; and Z_{ant} is the impedance of the antenna.

It will be seen that the worst-case VSWR when using a 75 Ω feeder with a 50 Ω antenna will not be 1.5 as assumed, but 112.5 ÷ 50 = 2.25, which is too high to be neglected.

3. METHODS OF OBTAINING OPTIMUM MATCHING BETWEEN 50 Ω ANTENNAS AND 75 Ω CABLES

3.1. Avoiding Critical Cable Lengths

The simplest method of using 75 Ω coaxial cable in 50 Ω systems is to ensure that the cable length is a multiple of an electrical halfwave ($\lambda/2 \times VF$). In this case, the 50 Ω input impedance will be transformed back to an output impedance of 50 Ω whatever the impedance of the cable.

This has a number of disadvantages since it is not even possible to insert a reflectometer into the 75 Ω portion of the feeder without causing a considerable mismatch condition. This will result in a rather critical situation, especially with foam dielectric cables whose velocity factor can vary considerably along its length due to variations of the dielectric density.

3.2. Impedance Transformation using λ/4 Coaxial Lines

In the opinion of the author it is far better to transform the 50 Ω to 75 Ω using one or two λ/4 transformers. This not only ensures an uncritical feeder length but also optimum matching. The impedance of the required 50 Ω to 75 Ω matching transformer can be obtained using the following equation:

$$Z_{\lambda/4} = \sqrt{Z_{in} \times Z_{out}}$$

$$\text{or } Z_{\lambda/4} = \sqrt{50 \times 75} = 61.24\ \Omega$$

It will be seen that an electrical λ/4 of approximately 61 Ω is required.

The easiest way is to use an electrical quarterwave of 60 Ω high quality coaxial cable (not TV cable that can be anywhere between 60 Ω and 75 Ω). A 75 Ω N-connector should be provided at one end, and a 50 Ω N-connector at the other. Since such cable is not readily available on the market, a tubular matching transformer is to be described that provides excellent matching between 50 Ω loads and 75 Ω cables.

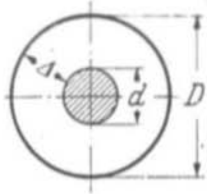
4. CONSTRUCTION

The impedance of a coaxial line with air dielectric is given by the following equation:

$$Z = 138 \log_{10} \frac{D}{d}$$

Where: D is the inner diameter of the outer conductor, and d the outer diameter of the inner conductor (**Figure 3**). As can be seen in 3.2., we require an impedance of approximately 61 Ω corresponding to a D/d-ratio of 2.77.

Fig. 3



Since standard tubing is to be used, it is necessary for the diameter of the outer conductor to be selected to suit the connectors used, and the wall thickness and inner conductor diameter selected to obtain the required D/d ratio of 2.77. The author chose the following tubing:

1. Outer conductor: 11 mm diameter, wall thickness 0.5 mm, corresponding to D = 10 mm.
2. Inner conductor: In order to obtain the required impedance, a tube of 3.5 mm diameter was selected. The wall thickness chosen was 0.75 mm so that it is possible to join the pin of the N-connector easily by placing a 2 mm diameter copper wire intermediately both into the inner conductor and center pin of the N-connector.

Frequency	Inner conductor length	Outer conductor length according to connector
145 MHz	489 mm	482 mm
432 MHz	165 mm	158 mm
435 MHz	163 mm	156 mm
1250 MHz	57 mm	50 mm
1296 MHz	55 mm	48 mm

Table 1: Mechanical dimensions when using N-connectors UG-218 and UG-94 A/U

These tubes provide a D/d ratio of 2.86 which results in Z = 63 Ω. This value is sufficiently close to the required value.

4.1. Mechanical Lengths

Due to the harmonic relationship between the 144 MHz, 432 MHz and 1296 MHz bands it is possible to use such a 144 MHz transformer for 432 MHz or even 1296 MHz. However, the bandwidth of the matching section will decrease with the number of $\lambda/4$ sections used, and any length error will be multiplied by three or nine times, respectively. It is therefore recommended that only one $\lambda/4$ section be used. The required mechanical lengths are given in **Table 1** for 145 MHz, 432 MHz, 435 MHz, 1250 MHz (ATV) and 1296 MHz.

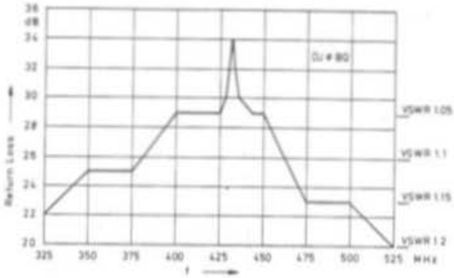
The length of the outer conductor should be shortened slightly in order to fit the N-connectors used. In the author's prototype using UG-21B (50 Ω) and UG-94A/U (75 Ω) N-connectors, it was necessary to shorten the outer conductor by 7 mm. The dimensions given in Table 2 are valid for these connectors.

4.2. Assembly

Firstly cut the inner conductor to the required length. Solder a piece of 2 mm diameter wire into each end of the 3.5 mm diameter inner conductor and cut this wire so that it fits into the hole of the N-connector pins, which are then soldered into place at each end of the inner conductor. The outer conductor is now placed around the inner conductor, the N-connector bodies placed over the outer conductor, and the center pins placed through the PTFE insulators in the center of the connectors. If the length of the outer conductor is correct, it is possible now for the N-connectors to be assembled completely with the rear feed-through nuts. After this, these nuts are soldered to the outer conductor tube.

5. MEASURED VALUES

The curves given in **Figures 4 and 5** show the passband characteristics of the 432 MHz version of the coaxial matching transformer constructed according to the dimensions given in Table 1. It will be seen that the bandwidth of the transformer is very wide, and it is not necessary to construct a separate unit for the communications and ATV portions of the band.



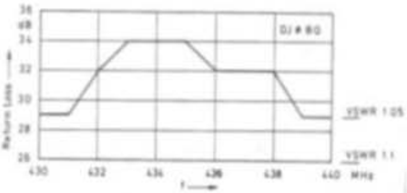


Fig. 4 and 5: Return loss of the 70 cm transformers vs frequency

6. SUMMARY

It is assumed that it is necessary to use two such transformers: one at the antenna, and one between the feeder and the 50 Ω station. It is, however, possible to modify the transmitter and receiver so that they match into 75 Ω. In this case it is also necessary to modify the internal cables of the equipment so that they are also 75 Ω (e.g. RG-59/U).

When one considers the difference in loss, for instance, between RG-213/U and the CATV-cable of the same diameter, a pair of such transformers could represent an inexpensive means of improving system performance together with a low-loss CATV coaxial cable.

7. REFERENCES

(1) H. Meinke, F. W. Gundlach: Taschenbuch der Hochfrequenztechnik 3. Edition, page 255.

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 1/1980 of VHF COMMUNICATIONS

DJ 6 PI 008	**LOW-NOISE 2-STAGE PREAMPLIFIER FOR 23 cm**		**Ed. 1/1980**
PC-board	DJ 6 PI 008	Double-coated PTFE	DM 68.—
Semiconductors	DJ 6 PI 008	1 each NE 64535, NE 57835, 2 zener diodes	DM 124.—
Minikit	DJ 6 PI 008	3 ferrite beads, 3 chip caps., 3 miniature tubular trimmers, 5 feedthru caps., 1 tantalum cap., 1 ceram. cap., 2 trimmer pot., 10 resistors, 1 case, 2 N-connectors	DM 41,—
Kit	**DJ 6 PI 008**	**complete with above parts**	**DM 228.—**
DJ 6 PI 009	**INEXPENSIVE TWO-STAGE PREAMPLIFIER for 23 cm**		**Ed. 1/1980**
PC-board	DJ 6 PI 009	Double-coated epoxy	DM 17.—
Semiconductors	DJ 6 PI 009	2 NE 57835, 2 zener diodes	DM 78.—
Minikit	DJ 6 PI 009	3 ferrite beads, 3 chip caps., 3 miniature tubular trimmers, 5 feedthru caps., 1 tantalum cap., 1 ceram. cap., 2 trimmer pots., 10 resistors, 1 case, 2 BNC-connectors	DM 34.50
Kit	**DJ 6 PI 009**	**complete with above parts**	**DM 129.—**
DJ 6 PI 010	**LOW-NOISE 2-STAGE PREAMPLIFIER for 13 cm**		**Ed. 1/1980**
PC-board	DJ 6 PI 010	Double-coated PTFE	DM 64.—
Semiconductors	DJ 6 PI 010	1 each NE 64535, NE 57835, 2 zener diodes	DM 124.—
Minikit	DJ 6 PI 010	3 ferrite beads, 3 chip caps., 3 miniature tubular trimmers, 5 feedthru caps., 1 tantalum cap., 1 ceram. cap., 2 trimmer pots., 10 resistors, 1 case, 2 N-connectors	DM 42.—
Kit	**DJ 6 PI 010**	**complete with above parts**	**DM 229.—**
DC 3 NT 004	**LOCAL OSCILLATOR MODULE FOR VHF-RECEIVER DC 3 NT 003**		**Ed. 1/1980**
PC-board	DC 3 NT 004	single-coated	DM 20.—
Semiconductors	DC 3 NT 004	2 transistors, 4 diodes, 2 IC's	DM 24.—
Minikit	DC 3 NT 004	1 coil set, 2 pl. foil trimmers, 2 tantalum caps., 12 ceramic caps., 3 pl. foil caps., 21 resistors, 3 trimmer pots., 1 relay	DM 38.—
Kit	**DC 3 NT 004**	**complete with above parts**	**DM 80.—**

Verlag UKW-BERICHTE, H. Dohlus oHG

Jahnstraße 14 – D-8523 BAIERSDORF

West-Germany · Telephone (0 91 91) 91 57 or (0 91 33) 855, 856

Bank accounts: Postscheck Nürnberg 30455-858 · Commerzbank Erlangen 820-1154 · Stadtsparkasse Erlangen 5-001.451 · Raiffeisenbank Erlangen 410.080

Professional Weather Satellite Reception Systems

We offer a complete system of inexpensive modules for professional applications. These are of special interest for meteological offices at smaller airports, for harbour and similar applications. They are also suitable for instruction at universities, and scientific institutes. A number of image processing systems and receivers are available to suit the application in question. Equipment is available, or under development for:

- The METEOSAT and GOES/GMS satellites in Geostationary orbit, or
- The NOAA and TIROS M satellites in polar orbits (136 - 138 MHz).

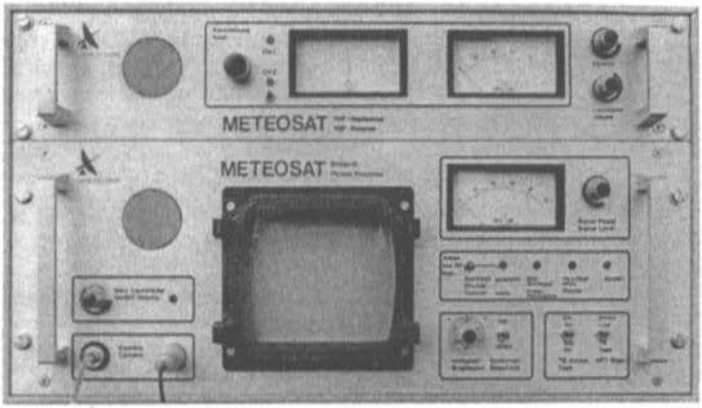
Technical specifications of the basic METEOSAT system:

PARABOLIC ANTENNA:
1.2 m dia., 24 dB gain

SHF-CONVERTER:
2-stage low-noise preamplifier with noise figure 3 dB

VHF-RECEIVER:
Noise figure 2 dB
ÄF-bandwidth:
800 - 4000 Hz
Subcarrier output:
2.4 kHz/1 V

VIDEO PROCESSOR:
Monitor tube and Polaroid camera



NEW GENERATION OF 23 cm TRANSVERTERS

About two years ago we brought out the first generation of 1296 MHz transverters on to the market. This is now to be replaced by a second generation of equipment of superior quality, and more modern technology.

Features:

- Double conversion in both the transmit and receive mode in order to obtain excellently high values of image and spurious rejection, and a virtually clean spectrum.
- Overall noise figure of the receive converter module: better than 4 dB
- Transverter available with an IF of either 144 or 28 MHz.
- Available with output power levels of either 1 W or 3 W.

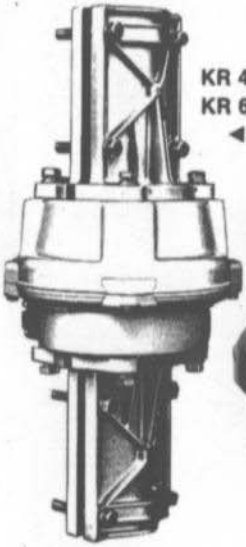
Available Models:

ST 1296/144 A:	144 MHz IF, 1 W output	**DM 585.00**
ST 1296/144 B:	144 MHz IF, 3 W output	**DM 685.00**
ST 1296/28:	28 MHz IF, 1 W output	**DM 695.00**
SP 1296:	3 W linear amplifier for ST 1296/144 A or ST 1296/28	**DM 138.00**

Available from our representatives or from the publishers.

U K W - T E C H N I K · Hans Dohlus oHG
D-8523 BAIERSDORF · Jahnstraße 14
Telephone (09133) - 855, 856 · Telex: 629 887
Bank accounts: Postscheck Nürnberg 30 455 - 858
Commerzbank Erlangen 820-1154

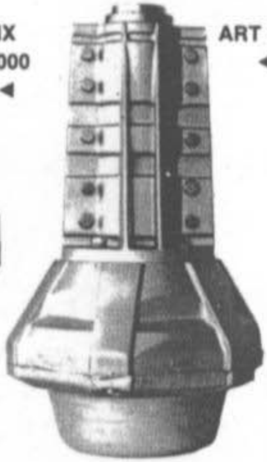
ANTENNA ROTATING SYSTEMS



KR 400
KR 600
◄

KR 2000
◄

MX
1000
◄



ART 8000
◄

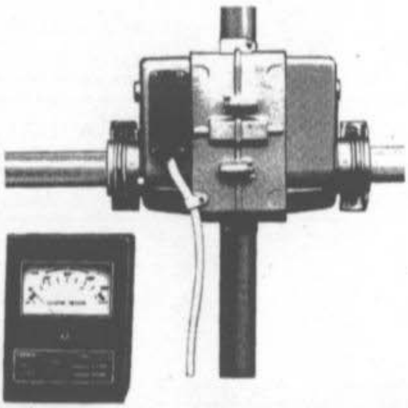
SPECIFICATIONS

Type of Rotator	KR 400	KR 600	KR 2000	MXX 1000	ART 8000	
Load	250	400	800	1000	2500	kg
Pending torque	800	1000	1600	1650	2450	Nm *)
Brake torque	200	400	1000	1200	1400	Nm *)
Rotation torque	40	60	150	180	250	Nm *)
Mast diameter	38 - 63	38 - 63	43 - 63	38 - 62	48 - 78	mm
Speed (1 rev.)	60	60	80	60	60	s
Rotation angle	370°	370°	370°	370°	370°	
Control cable	6	6	8	7	8	wires
Dimensions	270 x 180 ∅	270 x 180 ∅	345 x 225 ∅	425 x 205 ∅	460 x 300 ∅	mm
Weight	4.5	4.6	9.0	12.7	26.0	kg
Motor voltage	24	24	24	42	42	V
Line voltage	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	
	50	55	100	150	200	VA

*) 1 kpm ≙ 9.81 Nm

Getting ready for OSCAR 9 ?
Then you need the Vertical Rotor KR 500

Especially designed for vertical tilting of antennas for EME, OSCAR etc.



Type	KR 500
Load	ca. 250 kg
Brake torque	197 Nm *)
Rotation torque	40 Nm *)
Horiz. tube diam.	32 - 43 mm
Mast diameter	38 - 63 mm
Speed (1 rev.)	74 s
Rotation angle	180° (+ 5°)
Control cable	6 wires
Line voltage	220 V/50 Hz 30 VA
Weight	4.5 kg

UKW-TECHNIK / UKW-BERICHTE · Hans DOHLUS oHG · D-8523 BAIERSDORF ·

CRYSTAL FILTERS OSCILLATOR CRYSTALS

SYNONYMOUS FOR QUALITY AND ADVANCED TECHNOLOGY

NEW STANDARD FILTERS

CW-FILTER XF-9NB see table

SWITCHABLE SSB FILTERS

for a fixed carrier frequency of 9.000 MHz

XF-9B 01

8998.5 kHz for LSB

XF-9B 02

9001.5 kHz for USB

See XF-9B for all other specifications
The carrier crystal XF 900 is provided

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

XF-9A and XF-9B complete with XF 901, XF 902
XF-9NB complete with XF 903

KRISTALLVERARBEITUNG NECKARBISCHOFSHEIM GMBH
D 6924 Neckarbischofsheim · Postfach 7