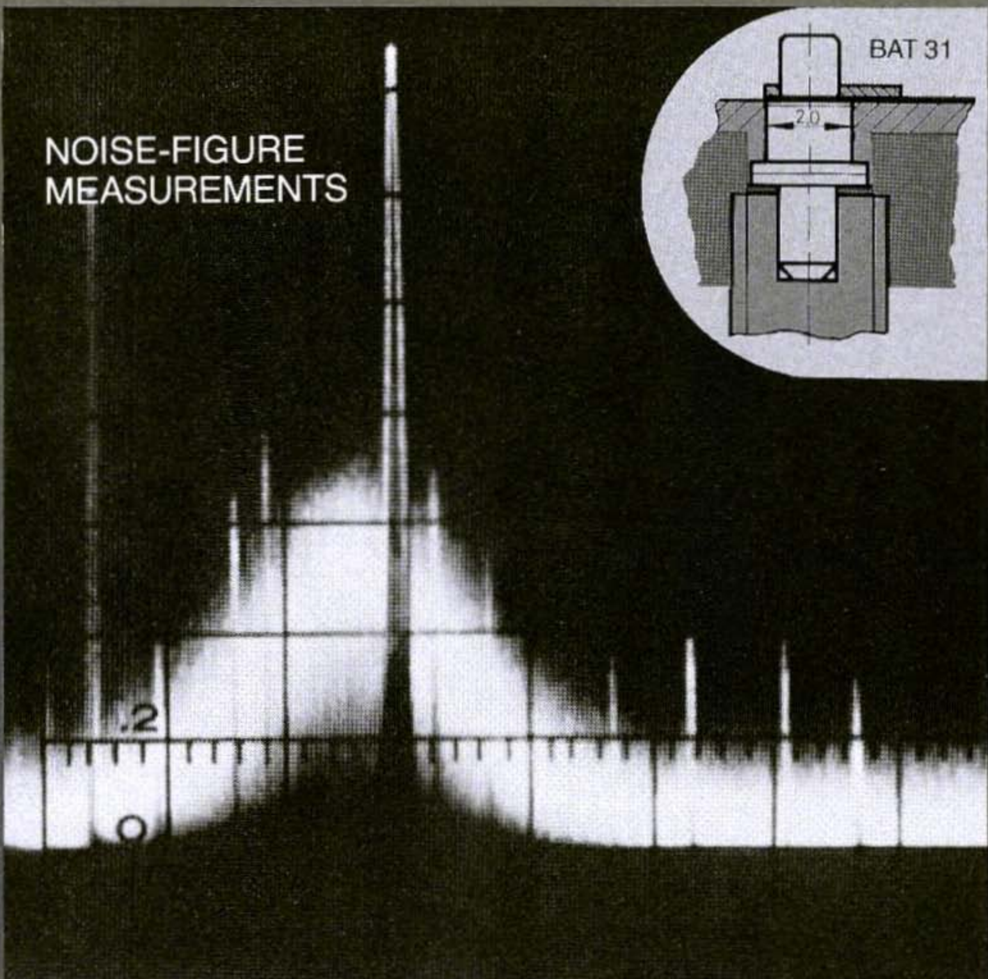
J 20 419 F

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

VHF communications

Volume No. 16 · Autumn · 3/1984 · DM 6.50



VHF communications

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

Volume No. 16 · Autumn · Edition 3/1984

Published by: Verlag UKW-BERICHTE,
Terry Bittan
Jahnstrasse 14
D-8523 BAIERSDORF
Fed. Rep. of Germany
Telephones (9133) 855, 856.

Publisher: Terry Bittan, DJ 0 BQ

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: Terry 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 1983 subscription price is DM 22.00 or national equivalent per year. Individual copies are available at DM 6,50 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.

Representatives

Austria
Verlag UKW-BERICHTE, Terry D. Bittan
POB 80, D-8523 Baiersdorf/W. Germany
Creditanstalt Bankverein, WIEN Kto. 17-90.599;
PSchKto WIEN 1.169.146

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

Belgium
COMELEC, D. Wilmus, Rue de Juifs, 26,
7000 MONS, Tel. 065/31 60 97
Stereohouse, Brusselsesteenweg 416, B-9218 GENT,
PCR 000-1014257 CCP, Tel. (091) 31 21 11

Denmark
Halskov Electronic, OZ 7 LX, Sigersted gamle Skole,
DK-4100 Ringsted. Telf. 03-616162. Giro 7296800.

France
Christiane Michel, F 5 SM, SM Electronic,
20 bis, Avenue des Clairions, F-89000 AUXERRE
Tel. (86) 469659

Finland
Erkki Hohenthal, SF-31400 SOMERO
Joensuuntie 6, Tel. 924-46311

Holland
MECOM, PA 0 AER, PO Box 40, Coendersstraat 24,
NL-9780 AA BEDUM, Tel.05900-14390.
Postgiro 3986163

Israel
Z. Pomer, 4X4KT, PO Box 222, K. MOZKIN 26114
Tel. 00972-4714078

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

Luxembourg
TELECO, Jos. Faber, LX 1 DE, 5-9, Rue de la fontaine,
ESCH-SUR-ALZETTE, Tel. 53752

New Zealand
E. M. Zimmermann, ZL 1 AGQ, PO Box 31-261
Milford, AUCKLAND 9, Phone 492-744

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
Carl-Oscar Biese, SM0HVL
Gjuteribacken 12 B
S-17239 SUNDBYBERG

Switzerland
Terry Bittan, Schweiz.Kreditanstalt ZÜRICH,
Kto.469.253-41; PSchKto.ZURICH 80-54.849

USA
Data Service Co., KØRC
3110 Evelyn Street, ROSEVILLE MI55113, Tel. (612) 636-9469

Contents

Josef Grimm, DJ 6 PI	**A Low-Noise Preamplifier for Weather Satellite Reception at 1.7 GHz**	130–136
Harald Fleckner, DC 8 UG	**A Noise Generator with Defined Noise Power for Applications up in the Microwave Range**	137–145
Bernhard Kokot, and Dieter Schwarzenau	**A Home-Made RF-Millivoltmeter Second, concluding part**	146–160
Hans Ulrich Schmidt, DJ 6 TA	**Introduction into Spread Spectrum Technology Article based on a Lecture at the Weinheim VHF-Convention 1982. Second, concluding part**	161–164
Friedrich Krug, DJ 3 RV	**An Optimum Crystal Filter for Coherent Telegraphy (CCW)**	165–170
Wilhelm Schürings, DK 4 TJ	**A VSWR-Indicator**	171–180
Willy van Driessche, ON 6 VD	**Suggestion for Standardizing SSTV and FAX Transmissions**	181
Werner Hanschke, DC Ø RZ	**C-MOS Frequency Counter for 10 Hz to 1 GHz**	182–187
T. Bittan, G3JVQ/DJØBQ	**Satellite News**	188–189

UKW**berichte** Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133-855. For Representatives see cover page 2

Josef Grimm, DJ6PI

A Low-Noise Preamplifier for Weather Satellite Reception at 1.7 GHz

Several hundred radio amateurs are now operating weather satellite reception stations for the geostationary weather satellites, and most of these are home-made stations. Until now, PC-board DJ6PI 010 has been used for constructing the 1.7 GHz preamplifier. This preamplifier was described in (1) and can, although designed for 2.3 GHz, also be aligned for 1.7 GHz. However, the price for PTFE PC-boards has increased so considerably that a new, inexpensive preamplifier was required. New preamplifiers for 1.3 GHz and 2.3 GHz were published in (2).

In the following a two-stage preamplifier is to be described which is equipped with the bipolar transistors NE 645 35 and NE 578 35, and uses the inexpensive epoxy PC-board material with a thickness of 0.79 mm. This preamplifier is able to provide the same specifications as the original PTFE preamplifier DJ6PI 010: **Noise Figure NF = 2.4 dB and Gain G = 23 dB.** These values make it very suitable for use in conjunction with the interdigital filter converter described in (3).

In addition to this, a single-stage preamplifier equipped with a GaAs-field effect transistor was designed using the more expensive RT/duroid 5870 PC-board material also with a thickness of 0.79 mm. Due to the expensive PC-board material, it was necessary to keep the preamplifier as small as possible. How-

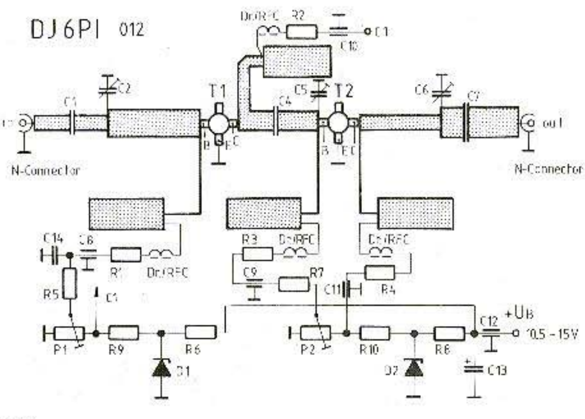


Fig. 1:
Two-stage preamplifier for 1.7 GHz equipped with bipolar transistors. NF = 2.4 dB and G = 23 dB

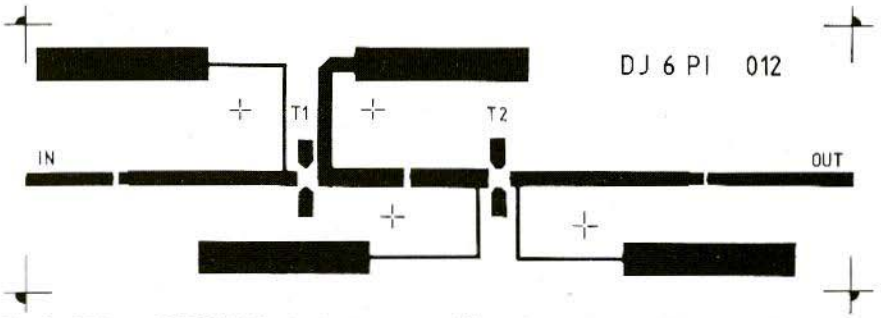


Fig. 2: PC-board DJ6PI 012 for the 2-stage preamplifier using a 0.79 mm thick epoxy PC-board

ever, the excellent characteristics of the GaAs-FET and the low-loss material RT/duroid can clearly be seen from the specifications of this preamplifier: **Noise Figure NF = 1 dB, and Gain G = 14 dB.** This preamplifier is not designed for driving the passive mixer of the interdigital filter converter (NF = 7 to 10 dB), since its gain is too low for this application. However, it is ideal for mounting directly at the antenna, where it can be used for eliminating cable noise (= loss), and to obtain more reserve of sensitivity.

1. TWO-STAGE PREAMPLIFIER DJ6PI 012

The striplines were calculated with the aid of a computer program using the transistor specifications (1). The input transistor NE 645 35 is aligned for minimum noise figure, and the second transistor NE 578 35 for maximum gain.

According to the data book, a noise figure of 1.5 dB is to be expected. However, the epoxy PC-board material and the use of lossy plastic foil trimmers causes the relatively small deterioration of 0.9 dB.

1.1 Construction

The circuit shown in **Figure 1** is virtually identical to that described in (1); the differences are only in the striplines.

The following components are required:

T 1:	NE 645 35 (NEC)
T 2:	NE 578 35 (NEC)
D 1, D 2:	C9V1 zener diodes
C 1, C 4, C 7:	20 – 100 pF chip capacitors
C 2, C 5, C 6:	0.7 – 5 pF miniature plastic foil trimmers SKY (green)
C 8 _ C 12:	470 pF – 1 nF feedthrough capacitors
C 13:	10 μF/25 V tantalum electrolytic
C 14:	47 nF ceramic capacitor
R 1 _ R 4:	10 – 22 Ω (uncritical)
R 5:	18 kΩ
R 6:	220 Ω
R 7:	12 kΩ
R 8, R 9:	150 Ω
R 10:	100 Ω
P 1:	25 kΩ
P 2:	10 kΩ

4 pcs. chokes: one ferrite bead on each connection wire of R 1 – R 4.

1 metal box 37 x 111 x 30 mm
2 coaxial connectors suitable for SHF, 50 Ω

Figure 2 shows PC-board DJ6PI 012 which uses 0.79 mm thick epoxy glasfibre material. The lower side of the board possesses a continuous ground surface, and the dimensions are suitable for mounting the above mentioned metal box.

The connection surfaces for the emitter must be through-contacted to the ground surface (lower side) in a low-inductive manner. This is made by sawing along the inside of the sur-

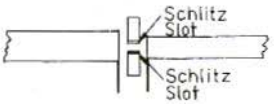


Fig. 3: It is necessary to provide a through-contact for the emitter connections

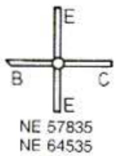


Fig. 5: Connections of the NEC-transistors

face with a fretsaw so that slots are made in the PC-board **(see Figure 3).** An approximately 2 mm wide strip of copper foil is placed through the slots and soldered to both sides.

The holes for the four feedthrough capacitors in the PC-board for base and collector voltages are made in the vicinity of the transition from the narrow to wide choke line so that the resistors R 1 – R 4 have enough room together with their ferrite beads. These positions are marked with crosses on the PC-board.

Only high-quality flange connectors such as N, SMA, SMC etc. can be used for the SHF-input. It is possible to use BNC for the output connector. According to the connector used, cutouts must be provided on the PC-board to accommodate them.

The position of the holes for the ground connections of trimmers C 2, C 5, and C 6 depend on their dimensions. In the case of the recommended SKY-trimmers, the hot connections are bent by 90°, as shown in **Figure 4,** and are soldered to the ends of the associated striplines. It is then possible for the three chip capacitors to be soldered into position vertically at the interruptions of the stripline. The SHF-connectors should be soldered in front of the board into the metal frame. This is done by firstly shortening the inner conductor to a suitable value (approx. 2 mm) and fitting the board into the case with a spacing of 10 mm between

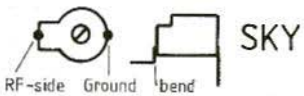


Fig. 4: The hot connection of the SKY-plastic foil trimmers is bent up, and a hole is drilled for the ground connection

its ground surface and the lower edge of the case. In some cases, it may be necessary for the flange to be sawn off at one side so that the cover on the ground side of the board can be closed. The inner conductors of the connectors should lie directly on the striplines, when the PC-board has been fitted in and is then soldered to the metal box.

It is now possible for the feedthrough capacitor C 12 to be mounted at the output end of the ground side of the box, after which the whole DC-voltage network for supplying the transistors is mounted into place (self-supporting) on this side of the board.

The last components to be mounted are the two transistors whose connections are given in **Figure 5.** The NE 645 35 is the transistor marked with **K**u (small letter variable), and the NE 578 35 is marked with **C**x (small letter is variable).

The photographs of the SHF and ground side of the author's prototype are given in **Figures 6 and 7.**

1.2. Alignment of DJ 6 PI 012

Firstly adjust potentiometers P 1 und P 2 for alignment of the transistor operating points to their ground stop. Terminate the input and output of the amplifier with 50 Ω (attenuators or 20 m RG-58/U with 50 Ω at the far end); this is to avoid any tendency to self-oscillation during the alignment process. Connect a mA-meter into the collector line of T 1, and T 2, respectively. Connect the operating voltage and adjust the collector current of T 1 with the aid of P 1 to 7 mA, and that of T 2 to 10 mA with the aid of P 2. After this, remove the mA-meter.

In order to align the three trimmer capacitors, one will require an operational receiver (converter) and an input signal. Firstly, align the

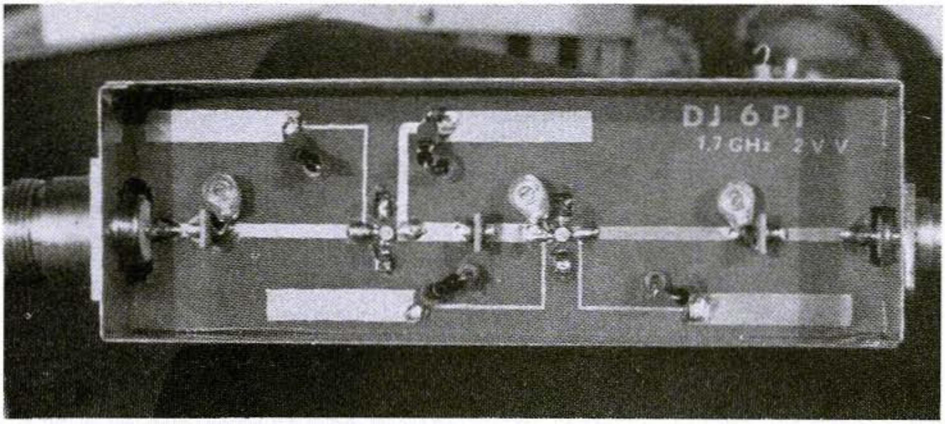


Fig. 6: The N-connector is the input side of the author's prototype

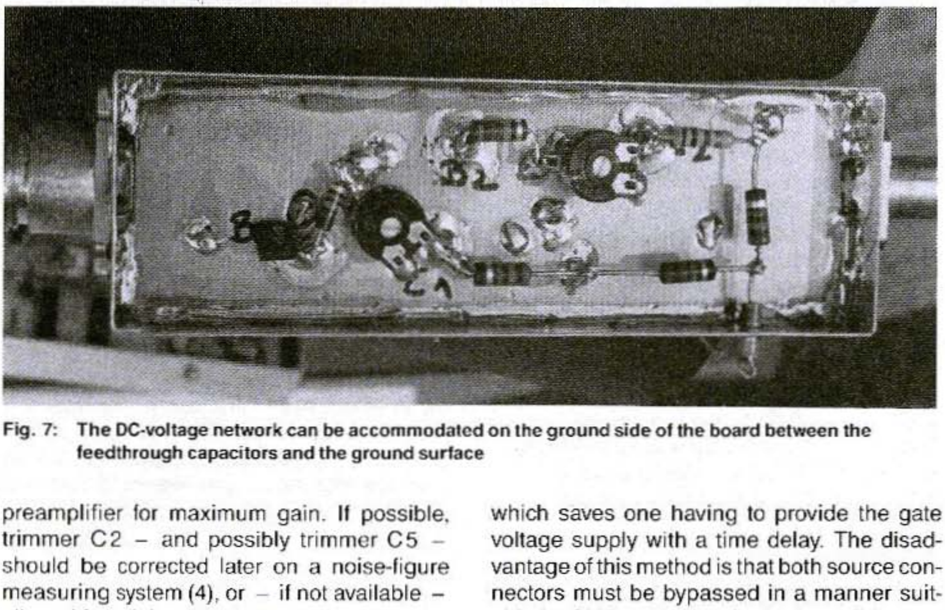


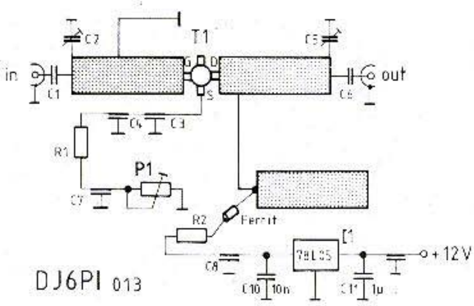
Fig. 7: The DC-voltage network can be accommodated on the ground side of the board between the feedthrough capacitors and the ground surface

preamplifier for maximum gain. If possible, trimmer C2 – and possibly trimmer C5 – should be corrected later on a noise-figure measuring system (4), or – if not available – aligned for minimum noise.

2. SINGLE-STAGE GaAs-FET PREAMPLIFIER DJ6PI 013

As can be seen in **Figure 8,** the source of the FET is connected to provide the bias voltage, which saves one having to provide the gate voltage supply with a time delay. The disadvantage of this method is that both source connectors must be bypassed in a manner suitable for SHF and in the shortest possible way.

The striplines on the low-loss RT/duroid 5870 material were calculated for use with the Mitsubishi GaAs-FET MGF 1400. However, the tuning range of the two trimmers is sufficiently large to allow other GaAs-FETs to be used. In this manner, one can use the even better GaAs-FETs MGF 1402, MGF 1412, manufactured by Mitsubishi, or the Siemens GaAs-FETs CFY 11 – CFY 19. However, it is not only



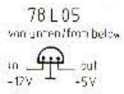


Fig. 8:
Single-stage GaAs-FET preamplifier for 1.7 GHz with NF = 1 dB and G = 14 dB.

the transistor itself, that determines the noise figure, but also the microwave trimmer used at the input, the disk capacitors and their construction, the quality of the connectors, as well as finally the actual alignment. For this reason, the following sections are very important.

2.1. Components for DJ 6 PI 013

T 1:	MGF 1400 (Mitsubishi)
C 1, C 6:	10–40 pF ceramic disk capacitor (ATC)
C 2:	0.5–2.5 pF microwave trimmer Johanson 9402-0
C 5:	as C 2, or miniature plastic foil trimmer SKY 0.7–5 pF (green)
C 3, C 4:	approx. 1 nF ceramic disk capacitor
C 7__C 9:	approx. 1 nF ceramic feedthrough capacitor for solder mounting
C 10:	10 nF ceramic capacitor
C 11:	1 µF/16 V tantalum electrolytic
R 1:	27–47 Ω (value uncritical)
R 2:	27–47 Ω with ferrite bead
P 1:	100 Ω trimmer potentiometer
I 1:	78L05 5 V-voltage stabilizer

2 pcs. flange connectors (or 1 plug, 1 connector) N, SMA, SMC according to antenna connector or plug.
1 metal box 37 x 74 x 30 mm

2.2. Construction of the GaAs-FET Preamplifier

PC-board DJ6PI 013 is constructed from RT/duroid 5870 material and possesses a thickness of 0.79 mm. The PC-board layout is shown in **Figure 9.** This material is very soft, which means that a very sharp drill must be used at low speed for the holes of C 3, C 4, C 7, and C 8. The diameter of these holes depends on the trimmers available.

The two disk, bypass capacitors C 3 and C 4 should be connected as near as possible to the striplines so that the source lines have the lowest possible length after soldering (preferably zero). The drawing in **Figure 10** can be used for information and **Figure 11** shows a photograph of the author's prototype.

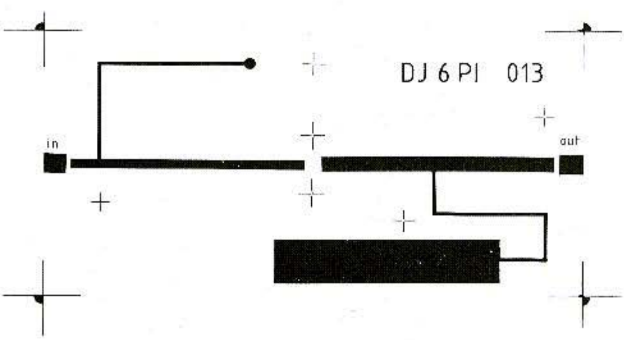


Fig. 9:
The RT/duroid PC-board DJ 6 PI 013 for the single-stage FET preamplifier

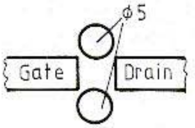


Fig. 10:
Position of the source capacitors

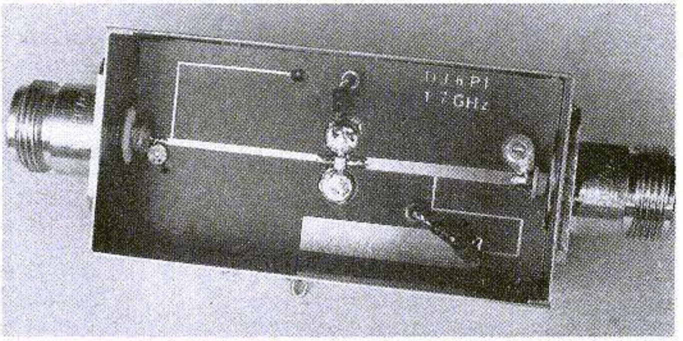


Fig. 11: A few, high-quality components are required for the 1 dB preamplifier

The ground connection of these two capacitors is made to the continuous ground surface by placing a piece of copper foil through the holes on the PC-board. After this, the disk capacitors are pressed into the holes on the PC-board and soldered to the lower, copper surface.

The hole for the ground connection of the two tuning trimmers C 2 and C 5 depends on the dimensions of the trimmer used. A low-loss trimmer Johanson 9402-0 should always be used at the input. The ground connection is marked with a red point.

The same is valid for the selection, position, and installation of the SHF-sockets as was mentioned in Section 1.1. Also, the installation of the DC-voltage supply is very similar. The potentiometer in the source line, and the voltage stabilizer with its bypass capacitors are mounted on the ground side of the board, as can be seen in the photograph of the author's prototype in **Figure 12.**

The gate end of the printed choke line is connected through the board to the ground surface using a short piece of wire. The source bias resistor R 1 is soldered to one of the two source connections and fed to the ground side of the board via a feedthrough capacitor. The drain bias resistor R 2 is connected in the same manner after providing a ferrite bead.

The last piece of construction work is soldering the transistor onto the striplines, and providing the bypass capacitors. Protective measures must be observed here against damaging the transistor due to static charge. It is necessary to use a soldering iron with galvanic separation from the power line, or one should at least disconnect the soldering iron

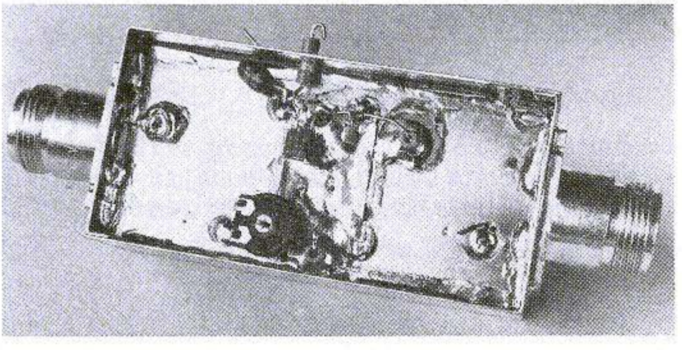


Fig. 12:
The DC-voltage network is accommodated in the 10 mm high ground-side of the case

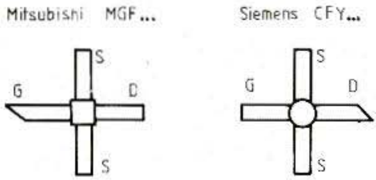


Fig. 13: Shows how sensible the manufacturers provide their connection configurations!

from the power line just before soldering. It is also necessary for the constructor to ground himself by connecting his wrist to a grounded object. The transistor is placed into position with the aid of tweezers and soldered as quickly as possible. In order to ensure that "Murphy's Law" does not become valid, attention should be paid to the correct connections, as shown in **Figure 13.** It will be seen that the various manufacturers use different connection configurations!

2.3. Alignment of DJ 6 PI 013

The input and output of this preamplifier are also terminated with 50 Ω, and potentiometer P 1 is adjusted to its **free** end. A mA-meter is connected in the drain line, and the drain current is adjusted to 10 mA after connecting the operating voltage.

The two alignment trimmers are firstly set for maximum gain, and later corrected for minimum noise figure. A special alignment tool is necessary for the microwave trimmer manufactured by Johanson. This is a short plastic pin with a square tip. If such a tool is not available, it is possible to use a watchmaker's screwdriver whose tip fits into the square diagonally. Due to the detuning caused by this metal tool, it is necessary for the alignment to be made in steps.

3. REFERENCES

(1) J. Grimm, DJ6PI:
Two-Stage Low-Noise Preamplifiers
for the Amateur Bands from 24 to 12 cm
VHF COMMUNICATIONS 12,
Edition 1/1980, Pages 2–13

(2) J. Grimm, DJ6PI:
Rauscharme Vorverstärker in Kammer- und Platinen-Bauweise für 23 und 13 cm mit GaAs-FETs
UHF Unterlage III, Pages 451–458

(3) B. Rößle, DJ 1 JZ:
A Receive Converter
for the Geostationary Weather Satellites
METEOSAT-GOES-GMS
VHF COMMUNICATIONS 13,
Edition 4/1981, Pages 207–213
VHF COMMUNICATIONS 13,
Edition 1/1982, Pages 24–30

(4) M. Dohlus:
A Home-Made Automatic Noise-Figure Measuring System
VHF COMMUNICATIONS 15,
Edition 1/1983, Pages 2–11
VHF COMMUNICATIONS 15,
Edition 2/1983, Pages 66–83

Subscription price for Vol. 17 (1985) of VHF COMMUNICATIONS

Unfortunately, the prices of paper, printing and binding, as well as nearly everything else, have increased considerably over the last two years. This forces us to increase the price for the **1985 subscription** to **DM 24.00.**

We are sure that you will agree that VHF COMMUNICATIONS is well worth this amount.

Harald Fleckner, DC 8 UG

A Noise Generator with Defined Noise Power for Applications up in the Microwave Range

In recent years, several constructional articles have been published for noise generators (1), (2), (3), for a noise measuring system (4), as well as information regarding this measuring technology (5), (6).

The semiconductor noise generators described in (1) and (2) use the base-emitter diode of an RF-transistor operated in the backward direction. The avalanche effect required for wideband noise is not very prevalent due to the low breakdown voltages of the RF-transistors BFR 34a and BFR 96 of < 5 V. It is mainly a tunnel effect that appears here, which means that the noise spectrum drops off considerably towards higher frequencies (> 1 GHz), (7).

The described noise generator operates with a silicon avalanche diode type BAT 31 which provides a wideband noise spectrum from < 10 Hz to > 18 GHz with a typical noise power (ENR = Excess Noise Ratio) of > 34 dB. This allows manual or automatic noise measurements to be made up to the satellite TV-band of 12.4 GHz with sufficient accuracy even for radio amateurs.

The following article describes the construction of the source and provides technical specifications of the noise diode, as well as describing the construction of a power supply with switching amplifier for connection to an automatic noise measuring system, and gives further information regarding operational experience.

1. NOISE DIODE

The noise diode used is a BAT 31, which is manufactured by Philips/Mullard in a microwave case type SO 86 **(Figure 1)**. This case is

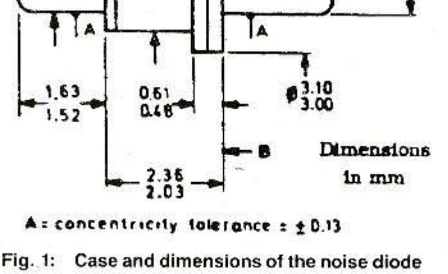


Fig. 1: Case and dimensions of the noise diode BAT 31

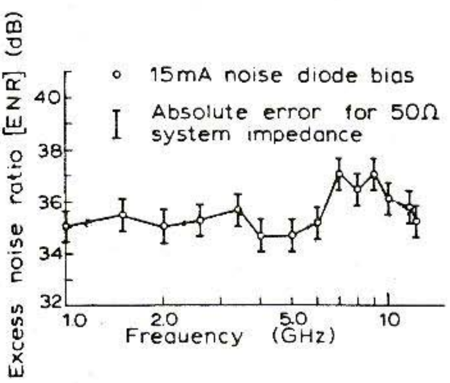


Fig. 2:
Noise power as a function of frequency when installed in a 50 Ω system

known from the varactor and Gunn diode technology.

The noise power as a function of frequency in a 50 Ω system is shown in **Figure 2.** The fluctuations of the ENR in the amateur bands in excess of 1 GHz amount to ± 0.5 dB, which corresponds to an absolute value of 1 dB when referred to a noise power of 35 dB; this is a very low value with respect to the large frequency range.

Figure 3 gives the ENR as a function of diode current, and of the frequency. One will see clearly that a current of 13–17 mA is required for wideband operation. The breakdown voltage is between 17 and 22 V.

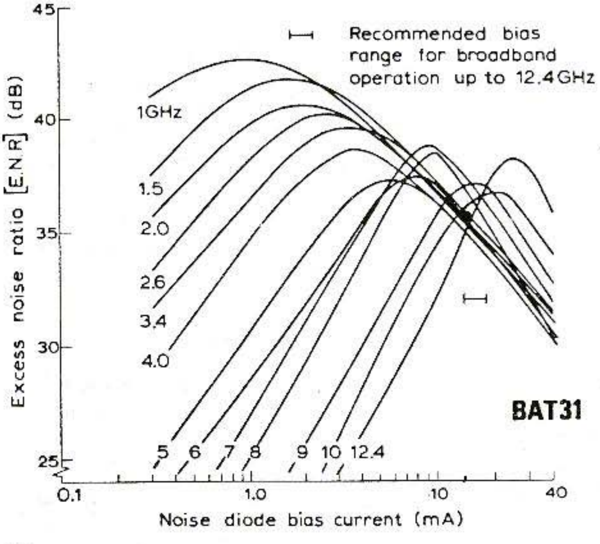


Fig. 3:
Noise power as a function of avalanche current with frequency as parameter when installed in a 50 Ω system

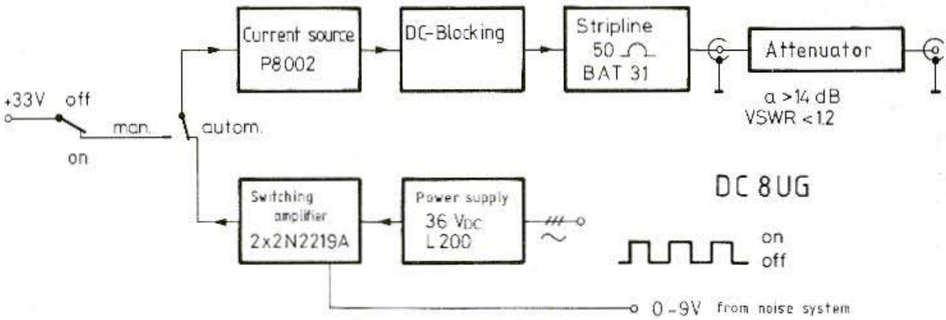


Fig. 4: **Block diagram of the noise source and its power supply with switching amplifier**

2. CIRCUIT

When used as noise source in switched operation (manual or automatic), the component should exhibit a source impedance of 50 Ω at all times.

In order to obtain this, the diode is built into a 50 Ω coaxial or stripline system and terminated with a wideband attenuator of > 14 dB, VSWR < 1.2 (see **Figure 4**). The value of the return loss has a decisive effect on the quality of the measurement (see Section 5). In this circuit, it is exclusively determined by the attenuator. A value of 20 dB with a VSWR ≦ 1.15 up to 4 GHz and VSWR ≦ 1.25 from 4–12.4 GHz has been found to be sufficiently good (Weinschel, HP, Radiall, or Greenpar). Home-made attenuators are not recommended here! With an attenuation of 20 dB, the ENR will amount to approximately 15 dB and its frequency-dependent fluctuations will now be determined by the diode and attenuator. Good coaxial attenuators have a tolerance of ± 0.5 dB.

Most noise measuring systems are calibrated for use with an **E**xcess **N**oise **R**atio ENR of 15.2 dB. This means that the source is virtually compatible.

In the author's prototype, the diode operates in a 50 Ω stripline system as shown in **Figures 5 and 7.** The switch or operating voltage is fed to the stripline via three, series-connected 150 Ω metal-film chip resistors. The position, dimensions, and quality of the resistors have an effect on the wideband characteristics of the decoupling.

The diode current is adjusted with the aid of a FET-constant current source, which determines the noise power. Two parallel-connected ATC-100 chip capacitors decouple the swit-

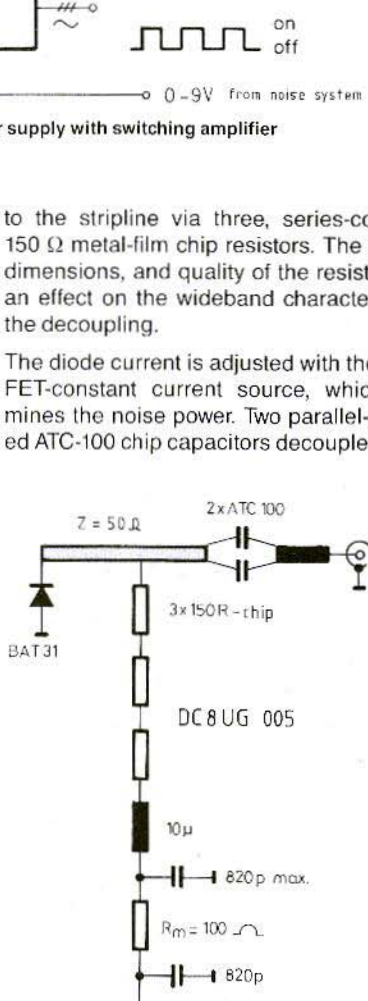


Fig. 5: **50 Ω stripline system with DC-voltage blocking at the output connector, and wideband choke in the supply line**

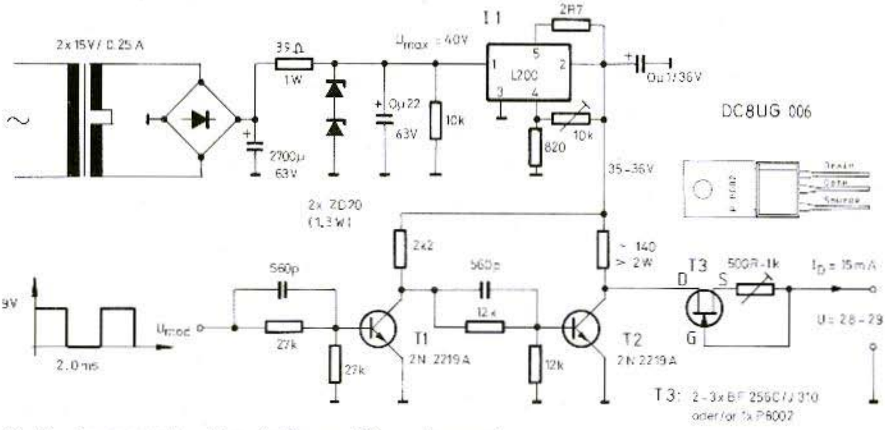


Fig. 6: **Power supply with switching amplifier and current source**

ching voltage from the attenuator. The switching voltage should amount to at least 33 V, which is generated most favorably using a seperate power supply. **Figure 6** shows the circuit diagram of the power supply with switching amplifier and current source. If the amplifier input is provided with a modulator pulse of, for instance, 9 V, 500 Hz squarewave-voltage, a current and voltage-controllable pulse of the same frequency and sufficient power will be available at the output. This allows the source to be used together with the

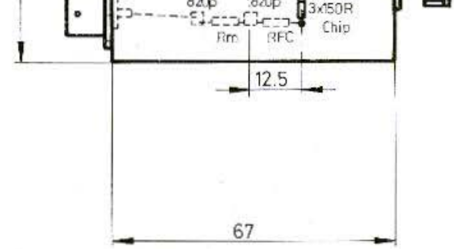


Fig. 7: **Layout of a 0.79 mm RT/duroid board**

noise figure measuring system described in (4), or with older noise figure measuring systems (eg. the GM 551 A).

3. CONSTRUCTION

The diode system comprises mainly a double-coated, PTFE-board (RT-duroid 0.79 mm) onto which a 50 Ω line has been provided. The coupling of the diode to the line must be made extremely accurately, since it determines the value of the available noise power at the upper frequency limit.

Further details regarding this are given in **Figure 8.** The diode mount comprises a gold-plated M4-screw with which the diode is coupled to the stripline on the ground-side of the board via a M4-nut. In order to achieve this, the line is provided with a 1.5 mm hole at the end, which is carefully extended to 2 mm diameter. The centrical position of the nut is guaranteed by use of a dummy instead of the noise diode (such as a defective varactor or Gunn diode). After this, a small gold plate is prepared with a width of 2.3 mm, length approx. 3 mm, and a 1.6 mm hole, and is placed over the dummy and soldered to the

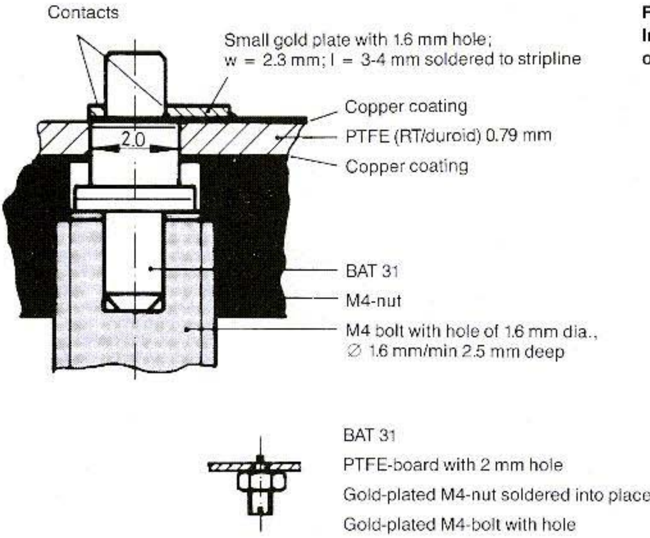


Fig. 8: Installation details of the noise diode

stripline. The small gold plate can be obtained from a connection strip of a power transistor. The dummy is now removed, and the hole is widened from the rear side through the M4-nut with the aid of a 2.1 mm drill, so that one can guarantee that the diode is coupled to the stripline with the aid of the lower cathode connection via the gold plate. With a careful construction, one will achieve a good stripline-coaxial transition.

After this, the PC-board is soldered into a metal box. The DC-switching voltage is now fed in via a BNC-connector, and the stripline should be terminated with the lowest discontinuity with an SMA or N-connector (preferably N-plug). One should only use precision N-connectors (clean thread, gold pin, maybe using a small flange). If one expects to obtain reproducible measuring results, it is important that no movable coaxial connections are provided. For this reason, BNC should not be used!

Finally, the chip capacitors, chip resistors, and the choke with the precision resistor, and the two disk capacitors on the ground side, are installed.

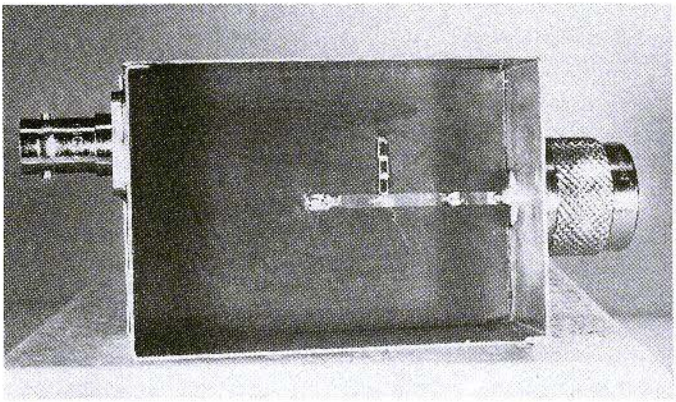
On inserting the diode, check for correct polarity (blocked operation) using an Ω-meter. **Figure 9** shows the author's prototype from both sides.

The power supply with switching amplifier and constant current source can be accommodated on the epoxy PC-board shown in **Figure 10.** If manual operation is to be used exclusively, one will not require the switching amplifier, which means it can be deleted from the board.

4. OPERATION

Check the power supply before connecting it to the noise source. This is achieved by replacing the noise diode by a 20 V zener diode and connecting it via a dropper resistor of 470 Ω to the power supply. At an output voltage of 36 V from the stabilizer L 200, it should be possible to adjust a current of 15 mA with the aid of the trimmer potentiometer of the current source. CAUTION: T 2 must block!

Furthermore, the switching amplifier should be driven with the modulator pulse from the noise measuring system, and the waveform at



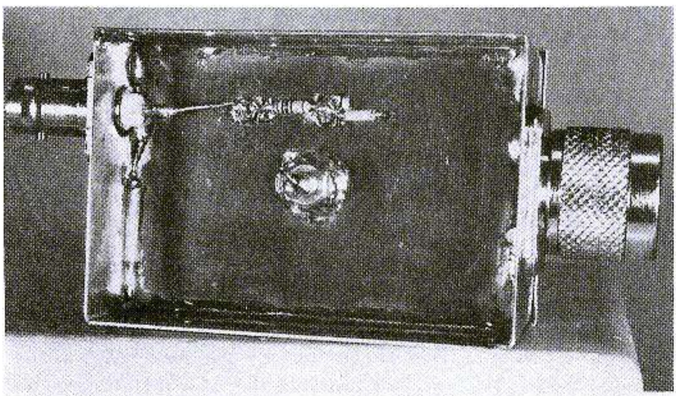


Fig. 9: On the upper side of the noise source (without attenuator), one will see the blocking capacitors, the three chip resistors, and the built-in diode. On the lower side, one can see the diode mount and the precision resistor soldered into place between the two disk capacitors.

the output checked with the aid of an oscilloscope.

If everything is working correctly, the source can be connected to the power supply. The current value should be aligned using the trimmer of the current source. This can be checked by measuring the voltage drop across the precision resistor R_m = 100 Ω. Preliminary, it is sufficient to adjust the current to 15 mA ≙ 1.5 V voltage drop. In the case of a pulsed diode, one can measure this best with the aid of a RMS-voltmeter, or oscilloscope.

For relative noise measurements, or comparison measurements for optimizing the receive sensitivity of amplifiers etc., it is not necessary for the ENR to be calibrated exactly. However, for absolute measurements, it is absolutely necessary for the source and attenuator to be compared to a calibrated noise source over the whole frequency range of interest.

5. OPERATIONAL EXPERIENCE

The author has been using a BAT 31-noise source with great success for more than three years. It was firstly used in manual operation for measuring the Y-factor, and later used in conjunction with a noise measuring system, type 551 A, manufactured by General Microwave.

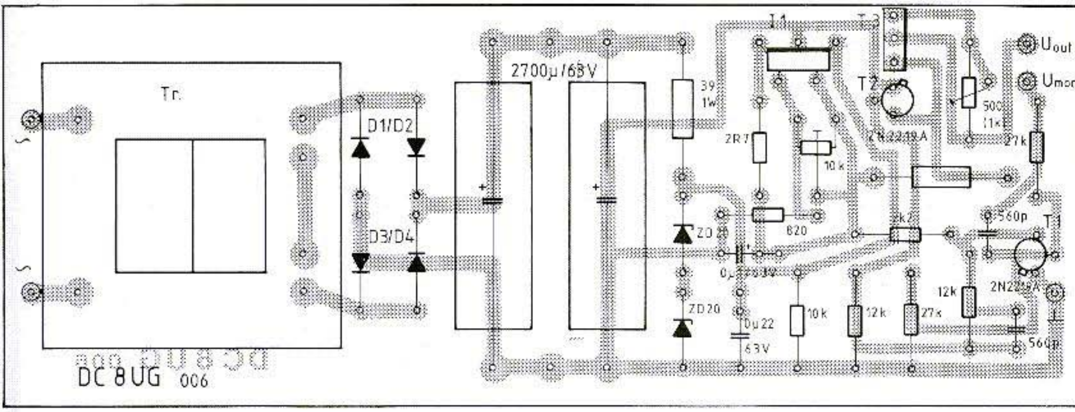


Fig. 10: **Component locations on the power supply/switching amplifier board DC 8 UG 006**

For absolute measurements, it is necessary to know the ENR for the amateur radio bands. Comparisons to AIL-noise systems using HP-sources showed that the source described here, used in conjunction with a 20 dB attenuator, exhibited the following ENR-values:

144	432	1296	2320	3456	5760	10368 MHz
15.8	15.4	15.2	15.2	15.0	16.9	14.0 dB

Sufficient noise power is also available in the shortwave range. It is even possible to carry out alignment for maximum sensitivity in the 80 m band. However, exact noise power values were not determined.

The principle of noise measurements is not to be described here, since sufficient articles have been published regarding this subject (1), (2), (6).

However, the author would like to give some information regarding the measuring accuracy, especially in conjunction with automatic noise measuring systems.

Among experts, there is no doubt that all automatic noise measurements have a considerable uncertainty, especially in the range below 3 dB. The following errors are added to the fluctuations of the source-ENR as a function of frequency:

1. Accuracy of the noise meter
   e.g. AIL 7514 ± 0.15 dB, GM 551 A ± 0.25 dB, HP 340 A ± 0.50 dB, all in the lowest range.
2. Temperature error
   ± 0.1 dB with a ± 10 degree deviation from 290 Kelvin
3. Image noise power
   up to + 3 dB, this can be avoided by using suitable circuitry.
4. Mismatch source – test object
   This error is not considered carefully enough during noise measurements in the range below 3 dB. For this reason, **Figure 11** shows this effect graphically.

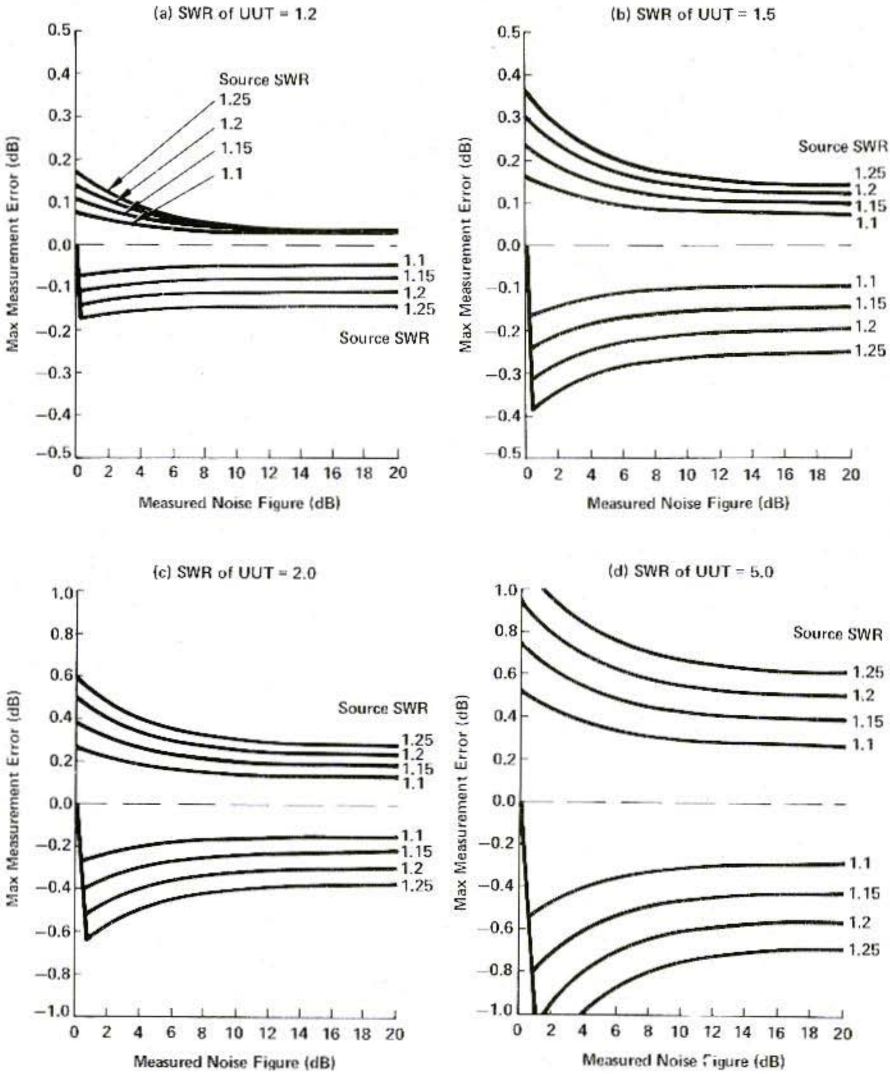


Fig. 11: Maximum measuring error as a function of the measured noise figure, as well as of the VSWR-values of the noise source and test object (UUT, Hewlett-Packard).

Example: In the case of a determined noise figure of 2 dB, a source-VSWR of 1.2, and a test object-VSWR of 2.0, the measuring uncertainty will be in the order of ± 0.5 dB!

As can be seen further, the return loss of the noise source cannot be large enough in order to provide tolerable measurements in the range below 2 dB, especially when the test objects possess a high standing wave ratio.

If these errors are added in the correct sign, one will arrive in a tolerance range that makes the decimal dB-values seem ridiculous. Even if one assumes that those can appear with opposite signs, and the actual accuracy of the measuring system is greater than the tolerance value, one should note that differences of 0.5 dB must be accepted, when using the same preamplifier at different times with different measuring set-ups on the same measuring system.

This just is to demonstrate how questionable most noise figure values under 1 dB are if they are not measured using an exactly calibrated hot/cold standard, but have been automatically measured, and that tenths-of-a-dB-values can be ignored.

To summarize, it can be said that the possibility of comparison measurements in the SHF-range represents an important aid for the home constructor and critical amateur consumer, especially considering the increasing activity. Absolute values should be of second importance to radio amateurs, since the optimization of every receive system must be made including the antenna, and this is achieved best using solar noise (8).

6. REFERENCES

(1) Michael Ulbricht, DB2GM
A Noise Generator for VHF and UHF
VHF COMMUNICATIONS 13,
Edition 1/1982, Pages 38–43

(2) Dr. R. Waxweiler:
Rauschgenerator mit definierter Rauschleistung
cq-DL 12/1981, Pages 585–587

(3) H. Fleckner:
Noise Generator with BAT 31
Dubus Info 3/1983, Pages 188–189

(4) Martin Dohlus:
A Home-Made Automatic Noise-Figure Measuring System
VHF COMMUNICATIONS 15,
Edition 1/1983, Pages 2–11
VHF COMMUNICATIONS 15,
Edition 2/1983, Pages 66–83

(5) Gannaway/Holmes:
Some Pitfalls in Noise-Figure Measurements
VHF COMMUNICATIONS 14,
Edition 1/1982, Pages 44–48

(6) B. Stein:
Automatic Noise-Figure Measurements Facts and Fancy
Ham Radio Magazine 1978,
August, Pages 40–54

(7) Unger/Schultz:
Elektronische Bauelemente und Netzwerke I
Vieweg-Verlag Braunschweig,
ISBN 3-5281-3505-0

(8) G. Hoch, DL6WU:
Determining the Sensitivity of Receive Systems with the Aid of Solar Noise
VHF COMMUNICATIONS 12,
Edition 2/1980, Pages 66–72

Bernhard Kokot,
and Dieter Schwarzenau

A Home-Made RF-Millivoltmeter Second, concluding part

4.2.2. Mechanically-switched Control Module

This double-coated PC-board has been designated DL0HV 002. **Figure 18** shows the component locations on this board and **Figure 19** a photograph of the author's prototype. **Table 4** gives all required components. The two inch wide front panel is drilled as shown in **Figure 20.**

4.2.3. Electronically-switched Control Module

This module comprises the electronic board DL0HV 006a (Euroboard) and a control board DL0HV 006b (see **Figure 22**), whose dimensions are 40 mm x 100 mm. It is mounted behind the two inch front panel as shown in **Figure 23.** A photograph of the author's prototype is given in **Figure 24.**

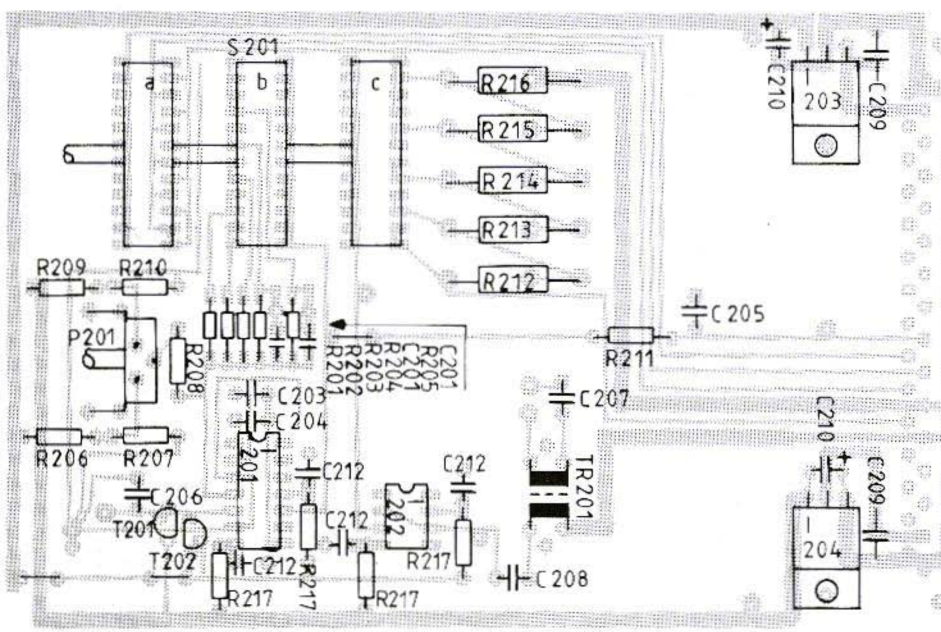


Fig. 18: Components of the control module (mechanical) DL0HV 002

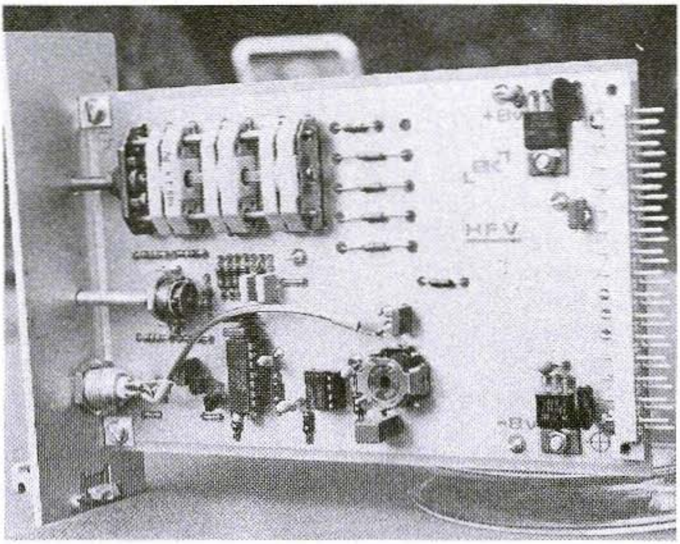
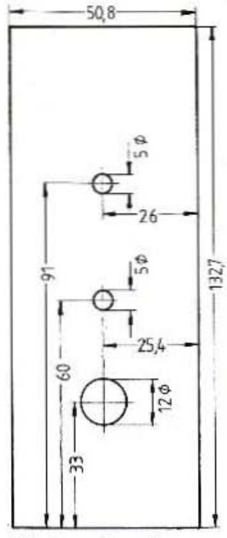


Fig. 19: Photograph of the author's prototype control module with mechanical switching



All dimensions in mm

Fig. 20:
Front panel for the mechanically-switched control module

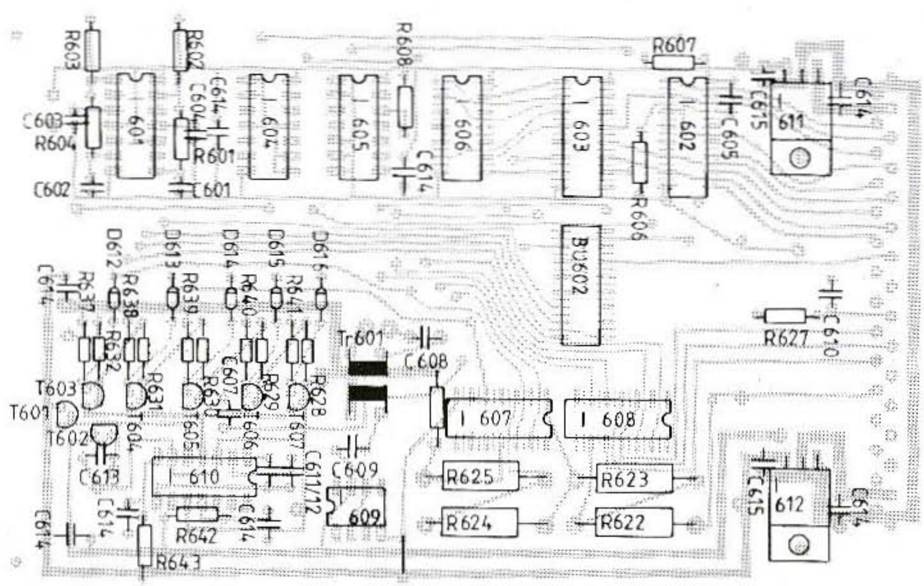


Fig. 21: Components of the electronic board DL 0 HV 006a for the electronically-switched control module

Quant.	Designation	Component
1	I 201	ICL 7650 CPD (Intersil)
1	I 202	TL 071 CP
1	I 203	MC 7808 CT
1	I 204	MC 7908 CT
2	T 201, T 202	BF 256
2	R 207, R 210	Resistor 1 Ω
1	R 216	Resistor 10 Ω
1	R 215	Resistor 90 Ω
4	R 217	Resistor 100 Ω
1	R 214	Resistor 900 Ω
2	R 201, R 208	Resistor 1 kΩ
1	R 213	Resistor 9 kΩ
2	R 202, R 211	Resistor 10 kΩ
1	R 212	Resistor 90 kΩ
1	R 203	Resistor 100 kΩ
3	R 204, R 206 R 209	Resistor 1 MΩ
1	R 205	Resistor 10 MΩ
1	C 208	Capacitor 1 μF / MKS
2	C 201, C 202	Capacitor 1 nF / FKC
2	C 206, C 207	Capacitor 22 nF / MKS
2	C 203, C 204	Capacitor 100 nF / MKS
7	C 205, C 209 C 212	Capacitor 100 nF / ceramic
2	C 210	Capacitor 10 μF / tantalum
1	P 201	Potentiometer 100 Ω
1	TR 201	Potted core 14 x 8, Al 2100 (Siemens) with coil former and mounting plate 1.) 300 turns; 2.) 75 turns enamelled copper wire 0.1 mm dia.
1	S 201	Switch SBL 11.3 x 12
1	BU 201	Connector
1	–	31-pin connector DIN 41617
1	–	Front panel 2 inch / Vero F2V1F
1	–	Knob for 6 mm dia. shaft
2	–	Mica disks for TO 220
–	–	Screws and nuts
–	–	Solder pins
1	–	PC-board DL0HV 002, 100 x 160 mm, double-coated

Table 4: Components list for the mechanically-switched control module

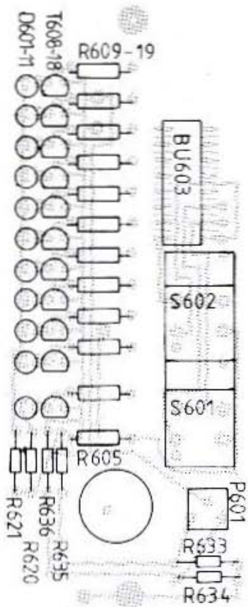


Fig. 22: Component locations of the control board DL0HV 006b for the electronically-switched control module

All dimensions in mm

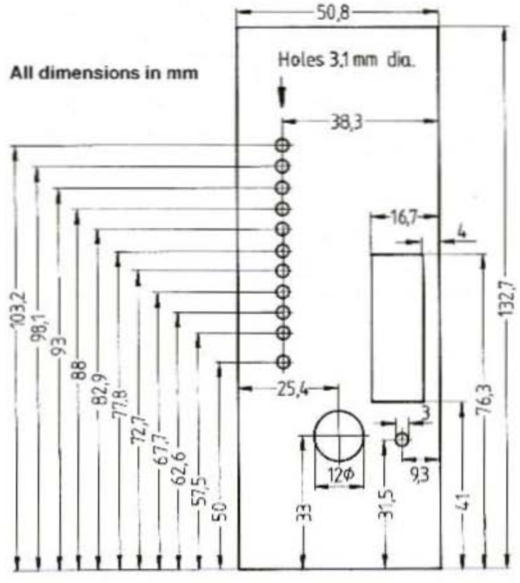


Fig. 23:
Front panel for the electronically-switched control module

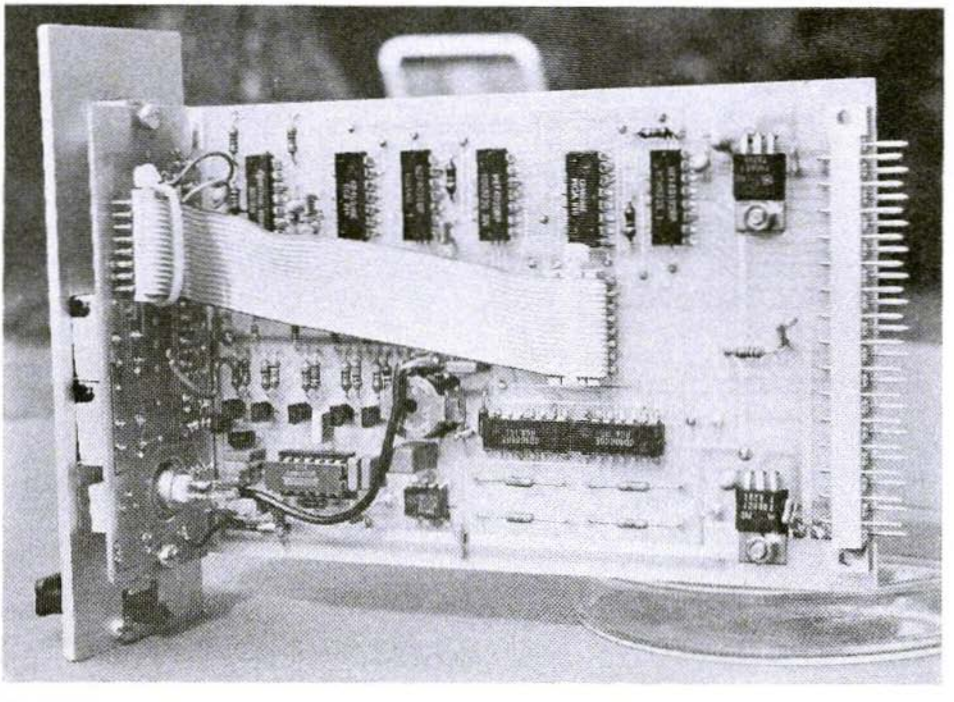


Fig. 24: Photograph of the electronically-switched control module DL 0 HV 006a + b

Two aluminium blocks of 10 mm per side are provided with a threaded hole of 3 mm diameter and are used for mounting the control board DL0HV 006b behind the front panel. It is fixed to the rear side of the front panel with the aid of a metal adhesive. The electrical connections to the electronic board DL0HV 006a are made via a 16-pole flat cable with DIL connectors at each end, and several additional wires. **Table 5** gives all required components:

4.2.4. Power Supply

As previously mentioned, the PC-board for the power supply is only single-coated and has been designated DL0HV 007. **Figure 25** shows the components on this PC-board; in the photograph of the open power supply **(Figure 26)** one will see the toroid-core transformer, as well as the temporarily used toroid-core choke, and the charge capacitors C 701 and C 702 formed from two parallel-connected capacitors. All required parts are given in **Table 6.**

Quant.	Designation	Component
11	D 601 – D 611	LED LD 350-4
5	D 612 – D 616	Diode 1 N 4148
2	T 601, T 602	FET BF 256
5	T 603 – T 607	FET BF 245 C
10	T 608 – T 617	Transistor BC 237 B
1	I 601	NE 556 N
1	I 602	HEF 40192 BP
1	I 603	HEF 4028 BP
1	I 604	HEF 4001 BP
2	I 605, I 606	HEF 4030 BP
2	I 607, I 608	CD 4066 BE
1	I 609	TL 071 CP
1	I 610	ICL 7650 CPD
1	I 611	MC 7808 CT
1	I 612	MC 7908 CT
2	R 633, R 634	Resistor 3.9 Ω
1	R 626	Resistor 10 Ω
1	R 625	Resistor 90 Ω
2	R 620, R 621	Resistor 820 Ω
1	R 624	Resistor 900 Ω
1	R 632	Resistor 1 kΩ
1	R 623	Resistor 9 kΩ
8	R 601, R 604 – R 608 R 627 – R 631	Resistor 10 kΩ
16	R 609 – R 619 R 637 – R 641	Resistor 47 kΩ
2	R 602, R 603	Resistor 82 kΩ
1	R 622	Resistor 90 kΩ
1	R 630	Resistor 100 kΩ

3	R 629, R 635 R 636	Resistor 1 MΩ
1	R 628	Resistor 10 MΩ
1	C 607	Capacitor 1 nF / FKC
2	C 608, C 613	Capacitor 22 nF / MKS
2	C 611, C 612	Capacitor 100 nF / MKS
9	C 610, C 614	Capacitor 100 nF / ceramic
1	C 609	Capacitor 1 μF / MKS
4	C 601 – C 604	Capacitor 2.2 μF, 25 V / tantalum
2	C 607, C 615	Capacitor 10 μF, 16 V / tantalum
1	Tr 601	Potted core 14x8, AL 2100 with coil former and mounting support pr.: 75 turns; sec.: 300 turns 0.1 mm dia. enammelled copper wire
1	Bu 601	connector
2	Tast S 601, Tast S 602	Digitaster, black
2	–	DIL sockets 16-pole
2	–	DIL connectors, 16-pole
–	–	flat cable, 16-core
1	–	31-pin connector DIN 41617
1	–	Front panel 2" Vero F2V1F
2	–	Aluminium block 10 x 10 x 10 mm
1	–	PC-board DL 0 HV 006, 100 x 160 mm, double-coated
1	–	PC-board 40 x 100 mm, double-coated
2	–	Mica disks for TO 220
–	–	Solder pins
–	–	Screws and nuts M3

Table 5: Components list for the electronically-switched control module

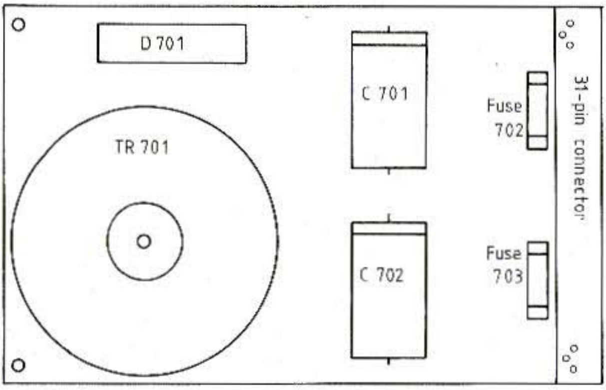


Fig. 25: Component locations on the power supply board DL 0 HV 007

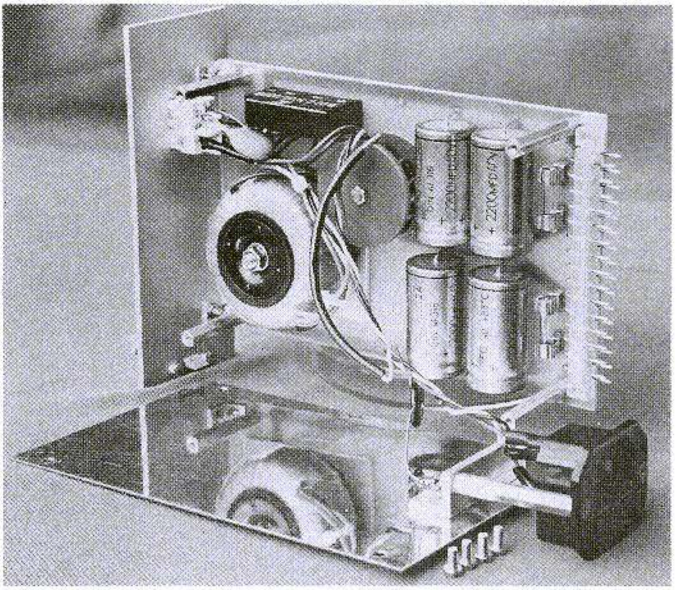
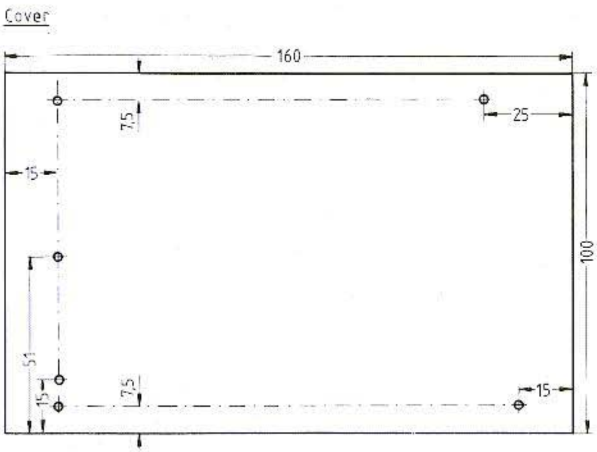


Fig. 26:
Photograph of the open power supply module

Quant.	Designation	Component
1	–	PC-board DL0HV 007, 100 x 160 mm single-coated
1	Tr 701	Toroid core transformator 2 x 12 V / 20 VA
2	C 701, C 702	Elektrolytic capacitor 4700 μF / 40 V
2	–	Fuse holder for PC-board mounting
2	Si 702, Si 703	Fuses 1 A, inert
1	Gl 701	Rectifier B40C3200
1	–	Two-inch front panel Vero F2V1F
1	S 701	Power switch with lamp
1	–	Power connector with fuse
1	Si 701	Fuse 0.1 A, quick-action
1	–	31-pin connector DIN 41617
1	–	Cover 100 x 160 mm
1	–	Aluminium bracket 10x15x45 mm
2	–	Spacers M3 x 40
4	–	Spacers M3 x 37
–	–	Screws M3 and spring washers
–	–	Insulating tubing

Table 6:
Components list for the power supply



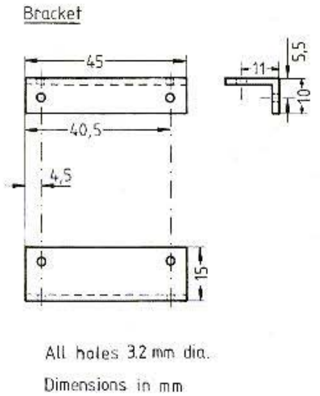


Fig. 27: Mechanical parts for constructing the power supply module

A completely insulated construction was used for feeding in the power supply voltage. The author constructed a cover plate and a bracket as shown in **Figure 27.** The cover plate was connected to the board using four spacers of 37 mm in length. The power connector with built-in fuse is connected to the bracket using two further, 40 mm long spacers. After screwing the bracket to the cover plate, it is possible for the interconnection to be made between the power supply and power connector. If the interconnection points are covered with insulating tubing, sufficient attention will have been paid to avoid electrical shock.

In order to ensure that the power connector is accessible from behind, it is necessary for a cutout to be provided in the rear panel of the unit, whose dimensions are given in **Figure 28.** The 2 inch wide front panel for the power supply plug-in is finally shown in **Figure 29.**

4.3. Installation into the Cabinet

31-pole connectors (DIN 41617) are installed into the plug-in cabinet at the required positions. At the same time, solder tags are mounted simultaneously with the screws, which are then connected to connections 1 and 31 in order to provide the required ground connections.

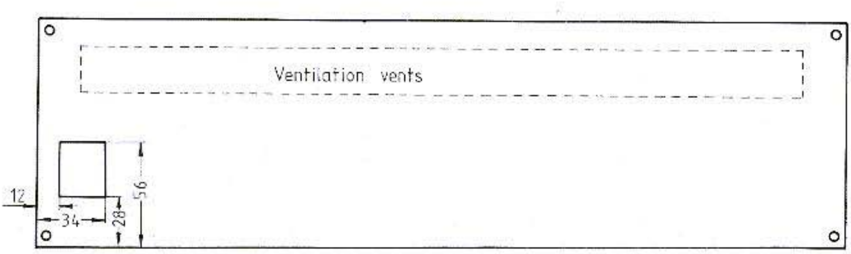


Fig. 28: Cutout on the rear panel of the cabinet

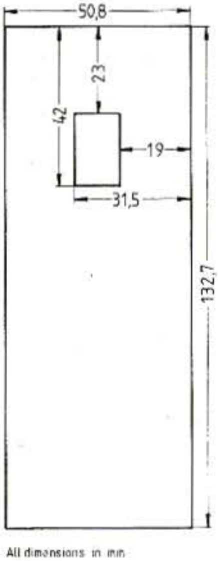


Fig. 29: Dimensional drawing for the front panel of the power supply

Finally, the power supply lines are provided in the cabinet, and soldered to connections 3 and 29 of each connector strip.

The following interconnections are required between the control and AF-module: Connections 17, 18, 19, 20, and 22.

The output of the AF-module – either connection 24 or 25 according to the position of S 101 – is connected to the corresponding input of the logarithmic circuit.

5. ALIGNMENT

The alignment can be made after the described modules have been constructed and wired together.

Firstly, check whether the RC-oscillator of the AF-module is operating correctly. An AC-voltage of approx. 1.5 V RMS should be present at test point TP 102 with a frequency of 7.5 kHz. If this voltage is too low, it may be necessary to change the values of capacitors C 103 and C 104. The frequency of 7.5 kHz is not critical.

For the actual alignment, the circuits should be connected as shown in **Figure 30.** Any available transmitter can be used as signal source. It should only be able to provide a signal of sufficient, and constant amplitude.

The attenuator, or output voltage of the signal source should be adjusted so that a DC-voltage of 4.47 V is measured after the Schottky diode.

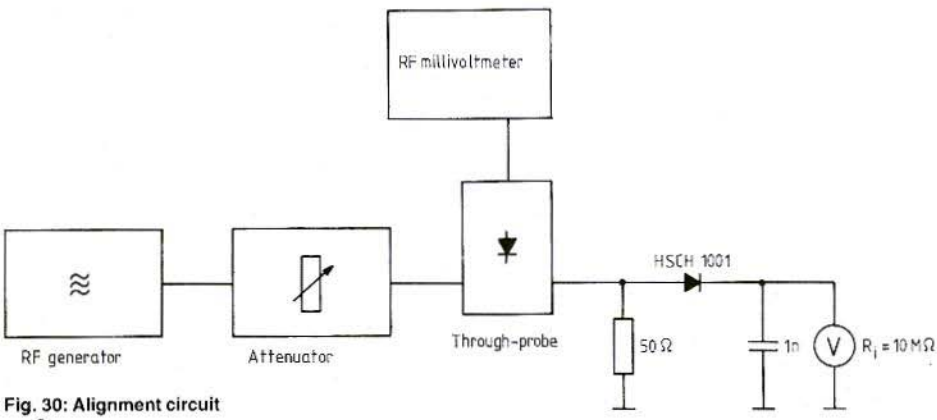


Fig. 30: Alignment circuit

Set the RF-millivoltmeter to its 3 V range and connect a meter to testpoint A of the AF-module. Align P 103 to obtain a voltage of 1 V at this testpoint (full scale).

Finally, an output voltage of 0.316 V is aligned with the aid of P 101 in the 10 V measuring range. If the same meter is used for alignment that is to be used in the indicator module, and if it possesses two scales, this indication will correspond to the same value, but on the other scale.

The 10 V measuring range is also selected for alignment of the digital voltmeter output. P 102 is aligned so that the same voltage is present at testpoint D as at testpoint A.

6. EXTENSIONS

6.1. Logarithmic Circuit

The logarithmic circuit to be described represents a useful extension to the RF-millivoltmeter. It is possible using this module to read off voltages or attenuation values directly in dBm or dB on the linear scale of the meter. It is only necessary for the meter to be marked in the opposite direction from −10 to 0 for the dBm-indication.

The logarithmic circuit is connected in front of the meter. It possesses two inputs between which one can switch (voltage measurement), or which are both used for measurements of voltage ratios (attenuation or gain).

6.1.1. Principle of the Logarithmic Circuit

Most known logarithmic circuits use the exponential characteristics of diodes or transistors with the aid of operational amplifiers. The disadvantages of this principle are the complicated alignment, and the large temperature drift. For this reason, a different method was used here, which utilizes the discharge curve of a capacitor to obtain the exponentional characteristic.

Figure 31 shows the principle used together

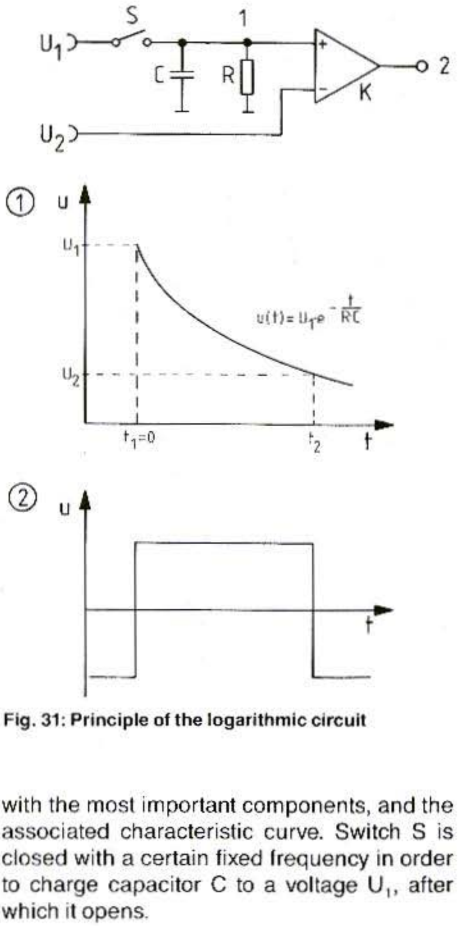


Fig. 31: Principle of the logarithmic circuit

with the most important components, and the associated characteristic curve. Switch S is closed with a certain fixed frequency in order to charge capacitor C to a voltage U_1, after which it opens.

Since U_1 is greater than U_2, the output of comparator K will become positive at the same time. The capacitor will now discharge itself via resistor R. The voltage can now be calculated according to the following equation:

$$u(t) = U_1 \times e^{-\frac{t}{R \times C}}$$

When the value of U_2 is reached, the comparator will switch, and the voltage at its output will be negative again.

If the corresponding values for this condition are placed into the equation, and if it is resolv-

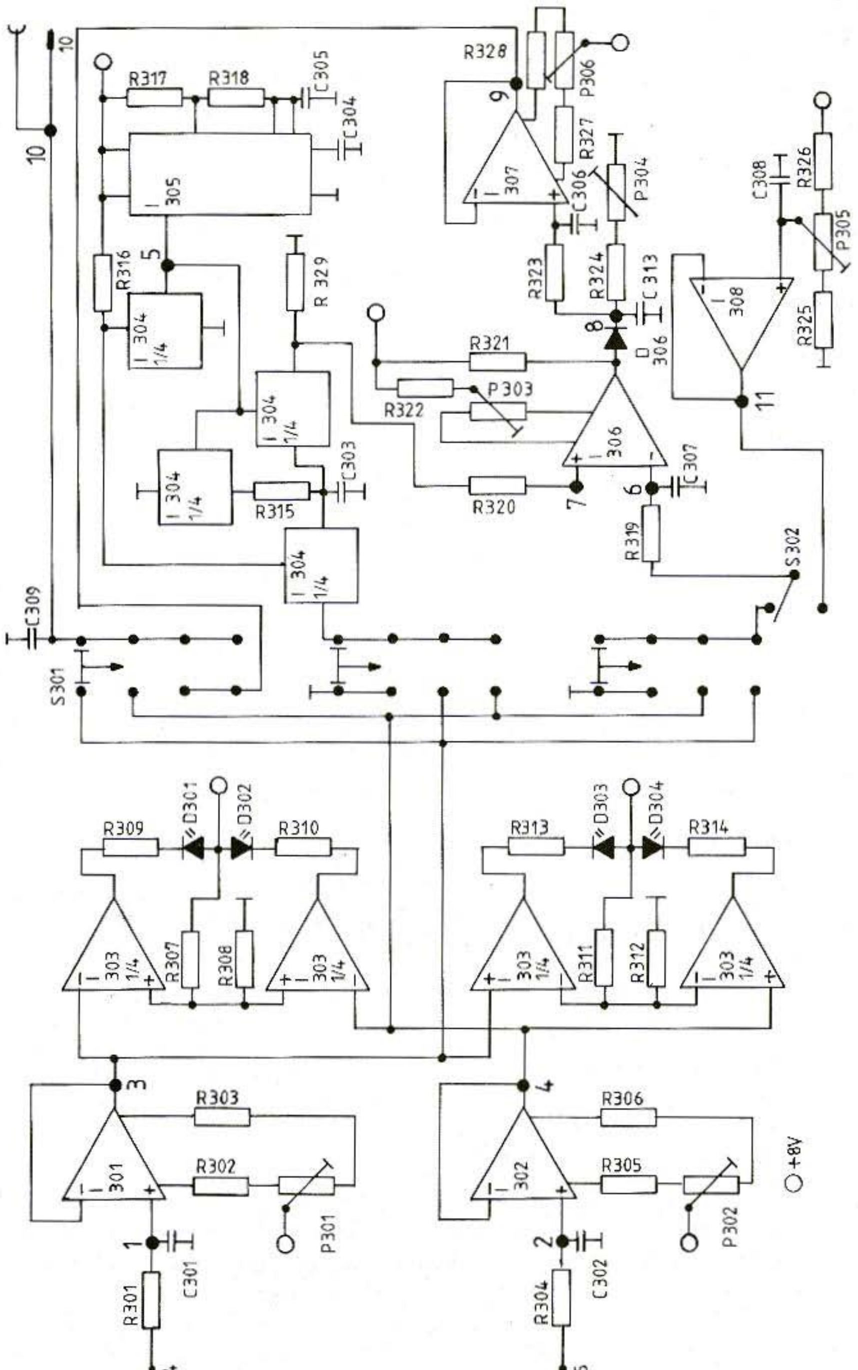


Fig. 32: The logarithmic circuit

ed according to the delay time, the following will result:

$$t_2 = R \times C \times \ln \frac{U_1}{U_2}$$

The pulse width of the voltage at the output of the comparator is therefore a measure of the logarithmic relationship between the two input voltages. A corresponding DC-voltage is obtained for the meter by forming the mean value of the signal.

6.1.2. Logarithmic Circuit

The circuit of the logarithmic module is given in **Figure 32.**

The two inputs are buffered with the aid of operational amplifiers (I 301 and I 302). The LEDs connected to the four comparators (I 303) indicate when the voltages are too high or too low, and that another measuring range should be selected.

It is possible for the various operating modes to be selected with the aid of the four-position switch S 301. In the two upper positions, one of the input voltages is directly indicated on the meter connected to connection 10. In the other positions, these voltages are passed via the logarithmic circuit.

The switch position should be selected at which the lower of the two voltages is present at the inverting input of the comparator (I 306). With the aid of S 302, it is possible for this voltage to be replaced by a fixed voltage as provided by I 308. This is necessary, if individual voltages are to be measured in dBm.

The actual logarithmic circuit differs from that described in the previous section in that it possesses one additional switch, each, in the discharge path, and subsequent to the capacitor. These are necessary because the MOS-FET switches used here exhibit a resistance which cannot be neglected when they are conducting, since this would form a voltage divider in conjunction with the discharge resistor. Furthermore, this means that the charge time for the capacitor cannot be kept as short as required, which means that the comparator must be disconnected from the capacitor during this period.

The switches are controlled from a timer IC (I 305), which is connected as multivibrator. It generates a square-wave signal with a keying ratio of 4.5:1 at a frequency of approx. 750 Hz.

The output signal of the comparator is fed via

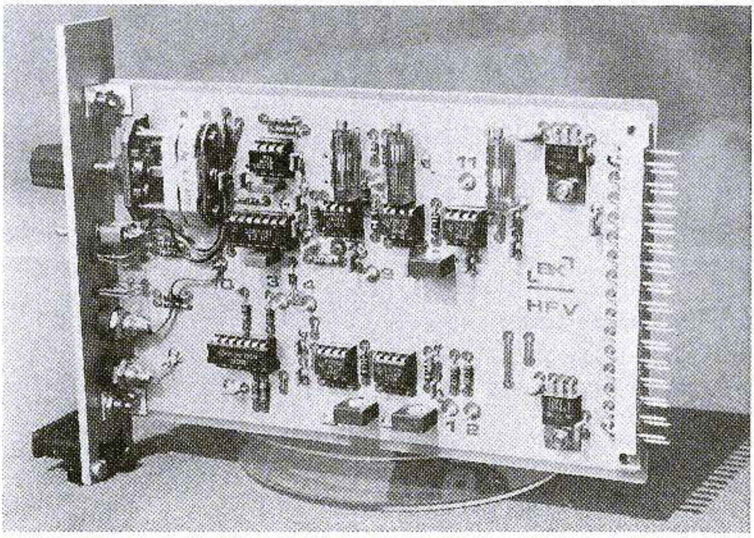


Fig. 33: Photograph of the author's prototype logarithmic module

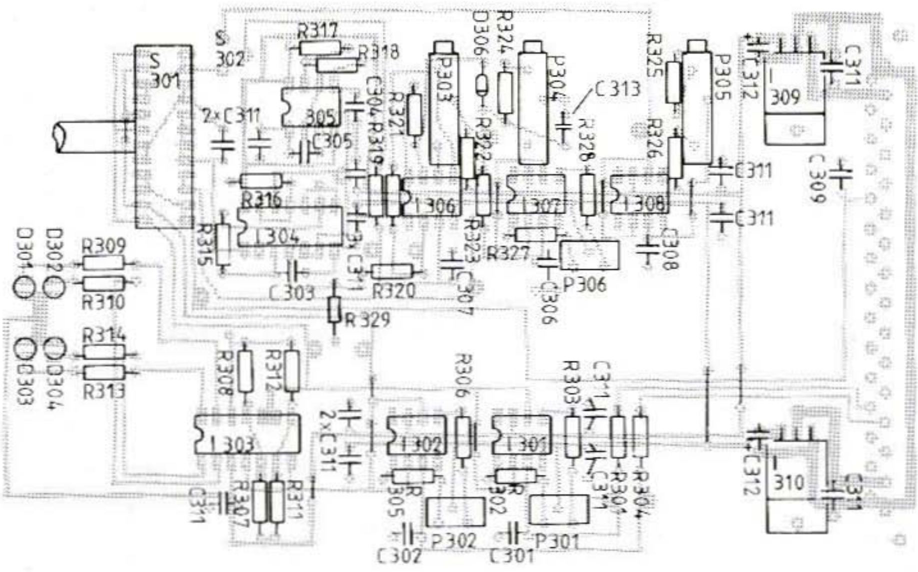


Fig. 34: Components on the logarithmic board DL 0 HV 003

a diode (D 306) to a low-pass filter for forming the mean value.

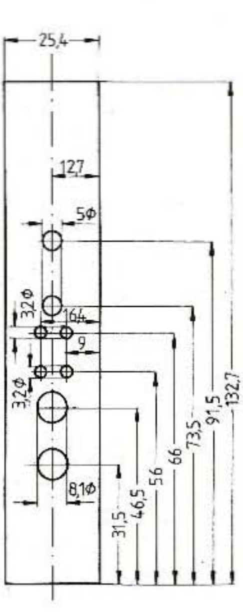
The required DC-voltage signal is available at the output of I 307 and is fed to a meter.

6.1.3. Construction of the Logarithmic Circuit

A double-coated PC-board (Eurocard size) has been developed for accommodating the logarithmic circuit **(Figure 33).** This board is designated DL0HV 003 and should be equipped according to the component location plan given in **Figure 34.** The components are listed in **Table 7:**

6.1.4. Alignment of the Logarithmic Circuit

For alignment of the logarithmic circuit, the two inputs should be connected together and grounded. The offset of the two input amplifiers can be aligned with the aid of trimmer potentiometers P 301 and P 302 and read off with the aid of a voltmeter connected to points 3 and 4.



All dimensions in mm

Fig. 35: Front panel of the logarithmic module

Quant.	Designation	Component
4	D 301 – D 304	LED LD 350-4
1	D 306	Diode HSCH 1001
4	I 301, I 302 I 307, I 308	TL 071 CP
1	I 303	LM 339 N
1	I 304	CD 4066 BE
1	I 305	NE 555
1	I 306	LM 311 N
1	I 309	MC 7808 CT
1	I 310	MC 7908 CT
1	R 312	Resistor 62 Ω
4	R 301, R 304 R 319, R 320	Resistor 100 Ω
1	R 325	Resistor 220 Ω
4	R 309, R 310 R 313, R 314	Resistor 470 Ω
1	R 324	Resistor 1.2 kΩ
1	R 318	Resistor 3.3 kΩ
1	R 308	Resistor 1.8 kΩ
1	R 322	Resistor 3 kΩ
1	R 315	Resistor 3.9 kΩ
1	R 321	Resistor 3.6 kΩ
10	R 307, R 311, R 316 R 326, R 302, R 303 R 305, R 306, R 327 R 328	Resistor 10 kΩ
1	R 317	Resistor 12 kΩ
1	R 323	Resistor 100 kΩ
1	R 329	Resistor 1 MΩ
1	C 305	Capacitor 100 nF / MKS
19	C 301, C 302, C 304 C 307, C 308, C 309 C 310, C 311	Capacitor 100 nF / ceramic
1	C 303	Capacitor 150 nF / MKS
1	C 306	Capacitor 1 µF / MKS
2	C 312	Capacitor 10 µF, 16 V / tantalum
1	C 313	Capacitor 3.3 nF / FKC
3	P 301, P 302, P 306	Potentiometer 100 kΩ
1	P 303	Potentiometer 2.2 kΩ
1	P 304	Potentiometer 470 Ω
1	P 305	Potentiometer 100 Ω
1	S 301	Switch SBL 11/1/4x3
1	S 302	Toggle change-over switch M90-3A
1	–	31-pole connector DIN 41 617
1	–	Front panel 1 inch Vero F1V1F
2	–	Telephone connector, insulated
1	–	Knob for 6 mm dia. shaft
2	–	Mica disks for TO 220
–	–	Solder pins
–	–	Screws and nuts
1	–	PC-board DL0HV 003, 100 x 160 mm, double-coated

Table 7: Components list for the logarithmic module

Testpoint 8 is grounded for the offset alignment of I 307; P 306 is then aligned so that the voltage at testpoint 9 is set to 0 V.

The next step is to test whether a square-wave signal with a keying ratio of 4.5:1 and a frequency of approx. 750 Hz is available at the output of I 305 (TP 5).

In switch position 3 or 4 (S 301), it is necessary for the offset of the comparator to be aligned with the aid of potentiometer P 303. This is made by connecting a DC-voltage of 1 V to connection 24, and a voltage of 0.5 V at connection 25. This represents a gain of 6 dB and a logarithmic output voltage of 0.4 V. Any deviation can be corrected with the aid of P 304. After this, connection 24 is provided with a DC-voltage of 300 mV and a voltage of 150 mV fed to connection 25. This also corresponds to a gain of 6 dB. P 303 should now be aligned for an output voltage at TP 9 of 0.4 V.

This procedure should be repealed alternately until no deviations are observed. Finally, it is only necessary for the output voltage of I 308 (TP 11) to be aligned with the aid of P 305 to 0.2236 V (0 dB).

Comparative experiments made with this logarithmic circuit and another constructed according to Tietze/Schenk (page 209, Fig. 11.22) show that high accuracies were achieved with both types.

However, the described circuit exhibited better temperature drift characteristics that were at least one order of magnitude better. Furthermore, the alignment is considerably easier.

7. SPECIFICATIONS

Power Supply: 220 V/50 Hz

Voltages for the modules: ± 12 V unstabilized
± 8 V stabilized

Measuring ranges: 1 mV to 10 V in nine ranges

Maximum DC-voltage component at the measuring probe: 40 V.

RMS-value indication in the case of sinewave voltages

Frequency range:
Through-probe: 1 MHz – 1.3 GHz
Test probe: 10 kHz – 1.3 GHz

Error limits:
Operating module with AF-module: <1%
Logarithmic circuit: <0.1 dB

Through-probe
(AA 119): <10% up to 500 MHz
Test probe
(AA 119): <10% up to 500 MHz

Impedances:
Through-probe: 60 Ω
Test probe, 10 MHz: >80 kΩ/2.5 pF

DC-voltage output:
For analog meter: 0...+1 V
For digital meter: 0...+0.316 V/1 V

Transcient time: 1 s

8. REFERENCES

(1) U. Tietze; Ch. Schenk:
Halbleiterschaltungstechnik, 5. Aufl.
Berlin, Heidelberg,
New York: Springer 1980

(2) RCA: Linear Integrated Circuits
USA: RCA Corporation 1977

(3) Valvo: Integrierte Digital-Schaltungen
LOCMOS-Reihe
Hamburg:
Verlag Boysen + Masch 1980/81

(4) Texas Instruments: Pocket Guide
Band 2 "professional linear"
Texas Instruments
Deutschland GmbH: 1977

Hans Ulrich Schmidt, DJ6TA

Introduction into Spread Spectrum Technology

Article based on a Lecture at the Weinheim VHF-Convention 1982.

Second, concluding part

4.4. Synchronization

One of the greatest problems in spread spectrum technology is the synchronization between transmitter and receiver. The method obtains its enormous processing gain, of course, from the fact that practically everything is known about the transmit signal at the receive end, with the exception of the actual information itself.

For this reason, the receiver must know the following values exactly, in addition to the exact transmit frequency, in order to decode the transmit signal:

- The spread code sequence,
- the clock rate of the code sequence
- the commencement time of the code sequence (phase shift) to an accuracy of ± 0.5 Bit.

In this case, one considers a synchronous condition. The synchronization must be maintained under all conditions, such as doubler shift ("tracking").

Three methods are usually used to fulfill these demands:

- Transmission of the reference code by the transmitter
- Frequency and phase synchronization using a standard frequency transmitter
- Synchronization by only evaluating the spread spectrum signal.

In the case of the method that transmits the reference signal **(Figure 11)**, virtually all synchronization problems have been solved.

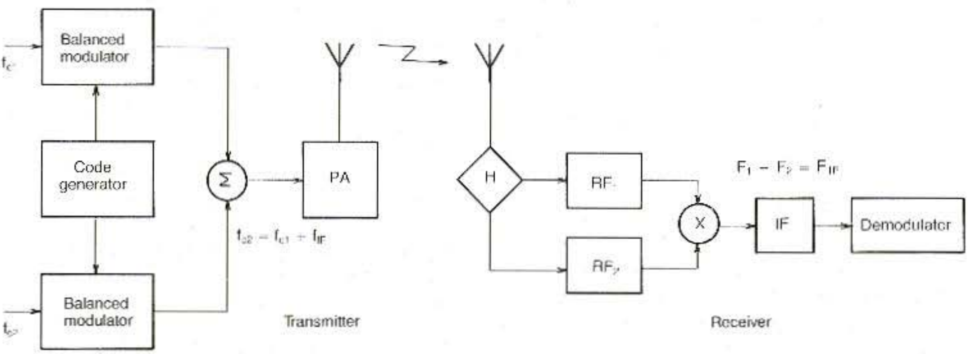


Fig. 11: Synchronization method using a transmitted reference signal

For this reason, the first experiments with spread spectrum technology were carried out in this manner. In this case, a modulated carrier f_1 is spread with the aid of a PN-code and transmitted, and an unmodulated carrier f_2 is also spread using the same code, and transmitted simultaneously.

At the receive end, both signals are received and fed to the two RF-inputs of a mixer. The unmodulated, spread carrier is now used as receive oscillator for the modulated one, and the intermediate frequency $f_2 - f_1$ will appear unspread at the IF-output of the mixer used as correlator.

This method can be used both with DS and FH-signals and is possibly the simplest method for radio amateurs for such communications in the microwave bands. Naturally, the security of the transmission will suffer, if it is received on a non-authorized receiver, since the transmitted reference and its frequency spacing can be determined.

In the case of the second method, all transmit and clock frequencies at both ends of the transmission link are derived from the signal of a standard frequency and time transmitter (e.g. DCF77WWV). As unknown constant, there only remains the phase shift due to the distance between transmitter, receiver, and standard frequency transmitter, as well as any possible phase response due to propagation.

As long as the stations are not mobile, and defined propagation conditions exist without variations in reflection and deflection, it should be possible for a manual adjustment of the phase shift to be sufficient for preliminary experiments. If this is not the case, it is necessary for the required phase shift to be forecast using doubler measurements made on the standard frequency, or to be compensated for using a phase synchronization.

If the transmitter and receiver can not be synchronized with the aid of a third signal, extremely high demands are placed on the frequency stability of the RF-systems, and automatic phase synchronization circuits will be necessary in all cases.

These circuits comprise a detector that measures the auto-correlation value at the output of the correlator (e.g. IF-output voltage of the receiver) as a function of the phase shift, a phase-shifter for the PN-code sequence that is controlled by the detector signal, as well as amplifiers and low or bandbass filters that form a phase control loop. In the case of the simplest FH-systems, the detector evaluates only the amplitude, and sophisticated demodulators evaluate the auto-correlation function according to amplitude and phase.

At the commencement of a synchronization process, it is necessary for the phase to firstly lock-in. This is achieved by shifting the phase of the receiver code sequence with respect to the transmit phase continuously, or in steps, until coincidence is achieved, which is when the output amplitude of the correlator is maximum (shift correlator). Since the shifting can only be carried out very slowly with respect to the code clock sequence, very long search times result with the long codes required for data security.

The length of the codes used for distance measurements will not allow this search process. These problems can be solved by using shorter codes that are only used for synchronization, or synchronization to parts of the codes in non-maximum code sequences. The disadvantage of such short synchronization codes is, however, the loss of interference signal suppression in comparison to that existing under synchronized conditions.

It would exceed the range of this article to discuss this in more detail and to discuss specific synchronization circuits. The greatest number of articles published during recent years have been discussing this problem of spread spectrum technology.

5. SPREAD SPECTRUM TECHNOLOGY AND APPLICATIONS

This article has shown that the spread spectrum technology possesses a number of technical characteristics that allow conventional telecommunications to be improved when used individually or as a combination.

The high security due to the low spectral power density, and the **high rejection of interference** have made the spread spectrum technology mainly interesting for military communications. For this reason, there are already a large number of military ground-to-satellite and ground-to-air communications, as well as missile control systems that are based on this technology, which are very difficult to jam with any electronic counter measures (ECM) available. One example described in (2) is the ground-to-air communication of the AWACS reconnaissance aircraft. They operate a DS-system over the whole of the L-band (915 – 1215 MHz) with a data rate of R = 56 kHz and a spread bandwidth of W = 300 MHz. This corresponds to a processing gain of $G_p' = 37$ dB. In addition to the high rejection to jamming, the use of an occupied frequency band is decisive.

The British manufacturer RACAL has developed a frequency-hopping system for tactical VHF-communications in the range of 30 to 88 MHz. In this case, frequency hopping is made between a total of 2320 channels ($G_p' =$ 34 dB). Due to the highly integrated circuits used in this equipment, these stations are not larger than a conventional VHF-transceiver, and this system can be used for portable and mobile communications.

In addition to the security against willful jamming, spread spectrum technology can also be used for the suppression of propagation-dependent interference. For instance, the interference due to multipath propagation (delay effects) can be eliminated well in the case of VHF/SHF over-the-horizon communications (Troposcatter). It is also possible for it to be used in the shortwave range (selective fading), however, the characteristics of the ionosphere limit the spread bandwidth. Quite recently, this technology has been used for experiments on carrier-frequency transmission systems on power lines, where great improvements are expected over existing systems.

A very interesting aspect, even for radio amateurs, is the most **effective use of the frequency spectrum**, and the possibility of **multiple use of frequency bands.** As has been mentioned previously, it is possible for the same frequency range to be used by a number of spread spectrum communication links using different codes, and more systems can be accommodated than when using conventional narrow-band systems. Furthermore, these transmissions can be used together with conventional systems, which means that already occupied frequency bands can be used "once again".

In the example of the AWACS aircraft, this has already been realized; in this case, the DS-system is operating in the same frequency range as the TACAN and IFF-equipment of military aircraft, and the DME and transponder of civil aircraft. In the USA, recommendations have been made and experiments are being carried out to use the, relatively speaking, enormously wide TV-bands for additional spread spectrum communications. This could solve the problems of VHF-mobile communications. There are already plans to reorganize the complete mobile telecommunication system based on spread spectrum technology and for it to be redistributed.

It is also possible for additional channels to be obtained in carrier-frequency systems on cable and microwave networks. Measuring equipment is available on the market that allows measurements to be made on fully operational systems using spread spectrum signals.

Further advantages of the spread spectrum technology result from the possibility of **multiple access** to telecommunication networks and from the addressability. In the case of a code multiplex system, it is possible to simultaneously transmit to all stations, and for confidential transmissions to be made between individual stations of the network. In this case, individual stations are addressed with the aid of a certain PN-code. Recommendations already exist for car-telephone, mobile communications, and automatic perimetry systems.

Another very important characteristic is the possibility of carrying out **exact distance measurements** with the aid of spread spectrum signals. Interplanetary space flights are unthinkable without this technology, as are the orbital parameter measurements of satellites (OSCAR 10), as well as the satellite navigation systems for aeronautical and marine use.

6. RADIO AMATEUR APPLICATIONS

As can be seen in US-American radio amateur publications (4 – 6), one is becoming more and more interested in this new transmission technology, and it is in the interest of the experimental nature of amateur radio that radio amateurs should experiment with this. In addition to the general interest in this technology, it is also possible for it to be used in the future for new band-plans for repeater stations, VHF-communication channels, linear transponders, and for amateur satellite communications. In the shortwave range, it is possible for interference caused by the overfilled frequency bands and unauthorized transmissions to be solved with the aid of a frequency-hopping system. Since personal computers are becoming more and more popular, it would be possible for even code multiplex communication networks with addressable stations to be built up on the UHF/SHF-bands, and the "package radio"-experiments of American radio amateurs are a first step in this direction.

After considering all these new "wideband" ideas, one might get the false impression that all previously used narrow-band systems are suddenly obsolete. The opposite is the case, since the new technology obtains its processing gain partly in that the receiver portion subsequent to the correlator carries out a very narrow-band processing of the signal.

The signal-to-noise ratio of this part, and thus the whole processing gain can still be improved by a further reduction of the IF-bandwidth (for example: using matched filters, synchronous demodulation etc.).

7. REFERENCES

(1) R. C. Dixon: Spread Spectrum Systems
John Wiley & Sons, Inc. (1976)

(2) R. C. Dixon (Edit.):
Spread Spectrum Techniques
IEEE Press (1976)
with many reprints from IRE- and IEEE Publications, e. g. Shannon: Communications in the Presence of Noise (1949)
Costas: Poisson, Shannon and the Radio Amateur (1959)

(3) A. J. Viterbi:
Spread Spectrum Communications – Myths and Realities
IEEE Communications Magazine, May 1979

(4) P. L. Rinaldo, W4RI: Spread Spectrum and the Radio Amateur
QST, Nov. 1980

(5) QST, May 1981, Happenings:
AMRAD gets special waiver for Spread Spectrum Experiments

(6) W. D. Clift, WA3NLO:
Spread Spectrum proposed for some VHF Amateur Bands
QST, Dec. 1981

(7) K. Mainzer, DJ4ZC:
Die digitale Nachrichtentechnik der AMSAT Phase-3-Satelliten
cq-DL 10/1978, S. 447

Friedrich Krug, DJ3RV

An Optimum Crystal Filter for Coherent Telegraphy (CCW)

The following article is to describe the practical construction of a narrow-band 9 MHz crystal filter for coherent telegraphy (CCW), and represents the second part of an article by Bernd Neubig, DK1AG, regarding "Optimum IF-Selectivity for Coherent Telegraphy (CCW)" in Edition 3/1982 of VHF COMMUNICATIONS.

The described filter is designed for pulse-modulated signals having a low bandwidth and is to be used by the author in a RTTY-receiver for a transmission speed of 50 Baud. It can, however, also be used in the IF-circuit described by the author in Edition 2/1982 of VHF COMMUNICATIONS instead of the described crystal filter module DJ3RV 001.

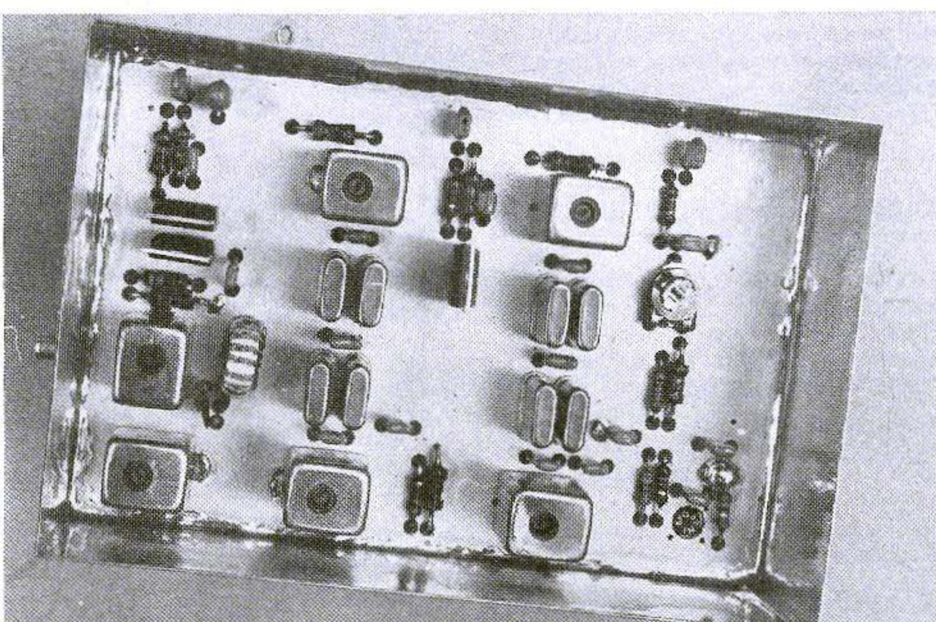


Fig. 1: Photograph of the author's prototype CCW-crystal filter for use as a module in the IF-circuit designed by DJ3RV (VHF COMMUNICATIONS 14, Ed. 4/82, Pages 239–252)

A filter suitable for good pulse transmissions should have a good transcient behaviour without overshoot. At the same time, it should also possess a good selectivity with a high stop-band rejection outside of the passband range. This is only possible using special types of filters.

Since such filters are not readily available and must be constructed especially, a crystal filter was calculated according to the previously mentioned article (1), published by B. Neubig; it was in the form of two series-connected, identical four-pole filters as described by Ulbrich/Piloty for home-made construction.

This circuit was built up using two series-connected, four-pole crystal filters interconnected with the aid of an amplifier stage, which was constructed using large-signal, and low-reactive amplifiers. **Figure 1** shows a photograph of the author's prototype.

The filters are decoupled using the low-reactive amplifiers, and are simultaneously matched to the required terminating and source impedances. This allows a sufficiently good selectivity, and good pulse behaviour to be obtained.

1. FOUR-POLE CRYSTAL FILTER

The four-pole crystal filter calculated by Bernd Neubig is constructed from two half-bridges **(Figure 2)**, and designed for an input and output impedance of 50 Ω. The nominal specifications for the filter are as follows:

Center frequency $f_0 = 9$ MHz
$\Delta f = \pm 75$ Hz (-3 dB)
and ± 325 Hz (-60 dB)

The following values result for the components:

$$L1 = L2 = 2.82\,\mu H$$
$$C1 = C3 = 270\,pF$$
$$C2 = 833\,pF$$

The series resonance frequencies of the cry-

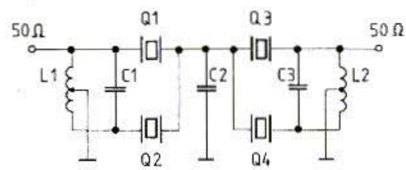


Fig. 2: A four-pole crystal filter as described by DK 1 AG (1)

stals are:

$$f_1 = 8999.757\text{ kHz}$$
$$f_2 = 9000.051\text{ kHz}$$
$$f_3 = 8999.840\text{ kHz}$$
$$f_4 = 8999.964\text{ kHz}$$

with the following equivalent specifications of the crystal:

$C = 9$ fF $\pm 10\%$ ($L = 34.7$ mH $\pm 10\%$)

$R =$ max. 13 Ω ($Q > 150000$)

$C_0 =$ max. 3 pF

In order to keep the spread of the filter and the

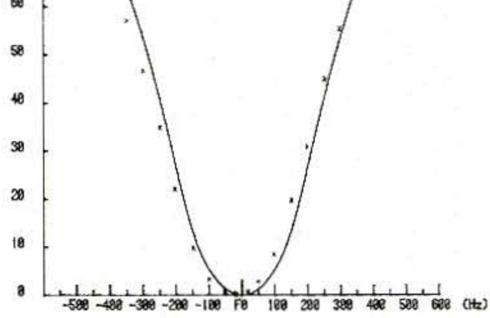


Fig. 3: Theoretical passband curve of the CCW-filter; the values marked with "x" are the measured values of the home-made filter

extent of alignments due to tolerances of the components to a minimum, crystals were required that had an alignment tolerance of $\pm 3 \times 10^{-6}$, and a temperature coeffcient of the frequency TC_f of $\pm 3 \times 10^{-6}$ between 0°C and 50°C.

The theoretical selectivity curve is given in **Figure 3** (Figure 5a of (1)) and the measured values of the author's prototype have also been added. The deviation of the measured values is caused by a center frequency that is 17 Hz lower, which is less than 2×10^{-6} at 9 MHz, and due to the reduction in ultimate selectivity. The latter is dependent on the construction, and can be improved by providing intermediate panels between the stages.

If such a low deviation of the center frequency of the passband range from the nominal frequency is permissible, this means that no alignment of the crystals and the bridge balance will be required. When using selected capacitors with a TC = NPO, it will be sufficient to align the filter only with the aid of the inductances. For this reason, filter kits are used for the input and output transformers instead of the usual ferrite toroid cores.

2. CIRCUIT DESCRIPTION

The circuit of the overall filter module is shown in **Figure 4.** The two, four-pole filters are extended by the buffer amplifier, and by one input and output amplifier each, which are used to compensate for the filter loss of 6 dB per filter. Well-proven circuits taken from the IF-amplifier described by the author (2) were used for these amplifiers. This ensures an input and output impedance of the module of 50 Ω, which is a great simplification for alignment, and for measurements. This also guarantees the use of this CCW-filter in the IF-module instead of DJ3RV 001.

The circuit is now to be described briefly. A detailed description and calculation of the circuits was given in (2).

The input bandpass filter constructed according to (3) is followed by a transformer Tr for matching, which is in turn followed by a large-signal amplifier equipped with two parallel-connected power FETs. This stage determines the intercept point of the module with IP_3 = 29

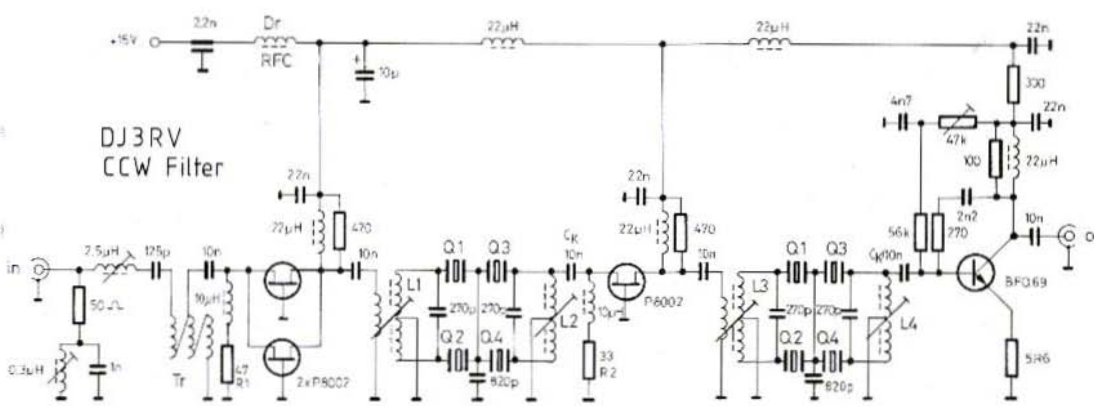


Fig. 4: Circuit diagram of the whole CCW-filter module

dBm, since the intercept point of the filter amounts to approximately 42 dBm with the coil formers used.

As shown in (2), the load resistance of the FETs represents a compromise between gain and intercept point, and is given by the input impedance of the filter and that required for matching to the source impedance. The 470 Ω resistor forms this source impedance, which is transformed to the calculated crystal filter impedance 50 Ω with the aid of L1. The input amplifier thus has a power gain of 6.5 dB.

The buffer amplifier following the first crystal filter is equipped with the FET P8002 in a common-gate circuit, which possesses a sufficiently low reaction in order to decouple the two filters. Furthermore, the transistor with its slope of 20 mS in a common-gate circuit has an input impedance of 50 Ω, which means that the first filter is correctly terminated.

With the selected drain resistor of 470 Ω, this stage has a power gain of 3.5 dB, and the matching to the second four-pole filter is made with L 3 in a similar manner to that of the first filter.

Since the load impedance of the filter amounts to 50 Ω, it would be possible for the output coupling to be made directly. However, the insertion loss of the filter is still not completely compensated for using the two amplifiers, and for this reason a further stage equipped with the low-noise BFQ 69 is added. The transistor will have an input impedance of 50 Ω using the current and voltage feedback shown, and a load impedance of 50 Ω, which means that the second filter is also terminated.

It should be mentioned here that the input impedance of a transistor with voltage feedback is very dependent on the load impedance. In order to obtain a good termination of the filter, it is necessary for a load impedance of as near as possible to 50 Ω to be provided at the output of the module.

The gain of the BFQ 69 is set to be 12 dB, which means that an overall gain of 10 dB results.

3. CONSTRUCTION AND ALIGNMENT

Special attention must be paid during construction of the circuit to obtain a good decoupling between the individual stages. The required ultimate selectivity will, otherwise, not be obtained. In the case of the author's prototype shown in Figure 1, an attenuation of 96 dB at ±1 kHz from the center frequency was obtained without screening panels. In the case of another prototype provided with such intermediate panels, as can be seen in (2) in the DJ3RV 001 module, it was possible to obtain an attenuation of more than 110 dB at ±2 kHz from the center frequency.

The circuit will operate without problems if the usual care is taken during construction. It is only the BFQ 69 that tends to self-oscillation if a composite carbon resistor with too high an inductive component is used as emitter feedback resistor. This effect was not noticed when using metal-layer resistors.

Filter kits D41-2165 (orange) (new designation 10x12-514 0500000) are used for the alignable inductances, and the specifications were obtained with the following windings:

L 1, L 3:	2 x 7 turns of approx. 0.2 mm dia. enamelled copper wire; 2.8 μH 21 turns of approx. 0.2 mm dia. enamelled copper wire
L 2, L 4:	2 x 7 turns of approx. 0.2 mm dia. enamelled copper wire 2.8 μH.
2.5-μH-coil:	15 turns of approx. 0.2 mm dia. enamelled copper wire
0.3-μH-coil:	5 turns of approx. 0.35 mm dia. enamelled copper wire
Tr:	3 x 12 turns of approx. 0.35 mm dia. enamelled copper wire wound on a toroid core, Siemens R10 N30

It is important that the 2 x 7 turns in L 1 to L 4 are bifilar wound (two wires together) and are wound evenly as lower layer. Attention should

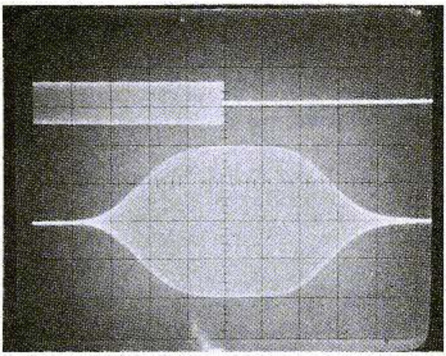


Fig. 5a: Pulse behaviour of the CCW-filter module with input and output pulse displayed linearly. Pulse duration: 12.5 ms; time scale: 2.5 ms/division

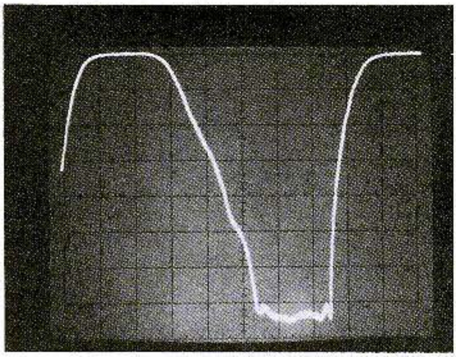


Fig. 5b: Output pulses of the CCW-filter module in a logarithmic display of the amplitude. Pulse duration: 12.5 ms; pulse repetition frequency: 25 Hz; amplitude: 10 dB/division; time scale: 5 ms/division

be paid to good balance during the wiring. Any error will have a great effect on the ultimate selectivity of the filter, which means that the stopband attenuation will be reduced.

It should be mentioned here that the values of the source resistors R 1 and R 2 are only valid for the author's prototype. These must be selected so that the transistor provides matching to the input impedance. The selection of the transistors is also described in (2).

The alignment is relatively simple since the circuit concept is designed so that the individual stages can be measured separately.

The greatest problem will probably be found in the provision of suitable measuring equipment. One will require a signal generator that is sufficiently stable in the frequency and which is able to be pulse-amplitude modulated. Furthermore, one will require an indicator circuit that is able to display both linearly and logarithmically. This means that one will require a signal generator, an oscilloscope, and – as used by the author – a spectrum analyzer.

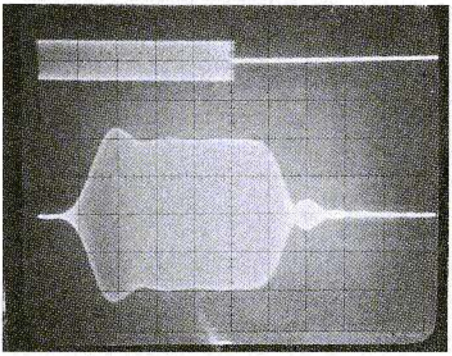


Fig. 6a: Pulse behaviour of a filter with Tschebyscheff characteristic (XF-9NB). Input and output pulse displayed linearly. Scales as in Figure 5a.

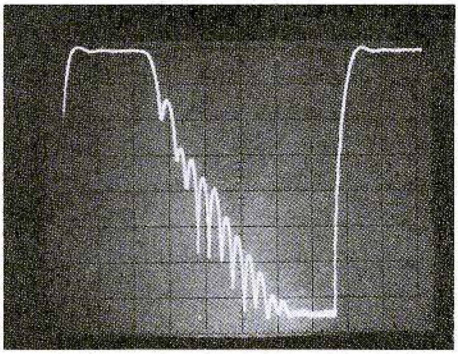


Fig. 6b: Output pulse of a XF-9NB filter with the amplitude displayed logarithmically. Pulse duration: 12.5 ms; pulse repetition frequency: 25 Hz; scales as shown in Fig. 5b.

If the two coupling capacitors C_K are removed, it is possible for the three parts of the circuits to be measured separately at an input and output impedance of 50 Ω.

The alignment of the filters is made in conjunction with a pulse-modulated signal to obtain the best pulse behaviour while varying the carrier frequency by ±500 Hz. This is made in order to ensure that a similar behaviour is achieved on both filter slopes.

One will then automatically obtain a good selectivity curve if the filter bridges are constructed symmetrically

The results obtained with the author's prototype are shown in the photographs given in **Figures 5 and 6. Figure 5a** shows the input pulse and the output pulse displayed linearly with a pulse duration τ = 12.5 ms and a pulse repetition frequency of 25 Hz, after passing through the whole filter. The time scale is 2.5 ms per division, so that the group delay of t = 6.7 ms can be read off easily.

Figure 5b shows the output pulse sequence in a logarithmic display. The amplitude decays from −6 dB to −70 dB in 12 ms; it is not possible to determine more due to the noise limit of the equipment. A photograph trace shows, however, that the filter can also be used for higher pulse repetition frequencies.

For comparison, the pulse behaviour of a filter with Tschebyscheff characteristic was examined. **Figures 6a and b** show the values measured in conjunction with module DJ 3 RV 001a together with a XF-9NB crystal filter. The filter is aligned for a good selectivity curve, without taking the input behaviour into consideration.

The results show that it is possible to carry out home-made construction of narrow-band CCW-filters if suitable alignment equipment is available. It should be mentioned, however, that the results could only be obtained with the assistance of Bernd Neubig, DK1AG, and the author would like to take this opportunity of thanking him once again.

The author would also like to thank KVG for the provision of very-close-tolerance crystals, which were manufactured especially.

4. REFERENCES

(1) B. Neubig, DK1AG: The Optimum IF Selectivity for Coherent Telegraphy (CCW) VHF COMMUNICATIONS 14, Edition 3/1982, Pages 163–171

(2) F. Krug, DJ3RV: A Versatile IF-Module Suitable for 2 m Receivers, or as an IF-Module for the SHF Bands
Part 2: Matching Stage for the Crystal Filter
VHF COMMUNICATIONS 14, Edition 2/1982, Pages 112–124
Part 4: Construction of the Crystal Filter Module DJ3RV 001
VHF COMMUNICATIONS 14, Edition 4/1982, Pages 239–252

(3) J. Kestler, DK1OF: Matching Circuits for Schottky Ring Mixers
VHF COMMUNICATIONS 8, Edition 1/1976, Pages 13–18

Wilhelm Schürings, DK 4 TJ

A VSWR-Indicator

After reading this title, many readers will believe that a conventional type of standing wave meter is to be described. However, inspite of its name, the operation of this unit is completely different. Standing wave bridges, as we call them, are power indicators, that measure the forward and reflected power on feeder cables. It is possible to calculate the VSWR from these power readings.

In contrast to this, the VSWR-indicator is actually that: An indicator nothing more! The measurement of standing wave ratios, for instance, is only possible in conjunction with a directional coupler or slotted line. However, the VSWR-indicator can do much more:

Measurements of attenuation and gain; for instance, the gain of a preamplifier or converter, the attenuation of cables, fixed and variable attenuators, filters, the cross-talk attenuation of coaxial relays, etc. It is also possible to trace the passband curve of a filter, and even to measure the antenna gain in comparison to a reference antenna. Together with a directional coupler, or return-loss bridge, a large number of measurements of matching are possible; for instance, VSWR of amplifier stages, receiver inputs, antennas, determination of the best matching of inductive and capacitive voltage dividers, etc.

Inspite of the multitude of applications, the VSWR-indicator is a relatively simple piece of equipment. It is virtually a sensitive, selective, and **uncalibrated** AC-voltmeter, which can be switched in 10 dB-steps.

Since the indicator is not calibrated – it is only an indicator as the name implies – no calibrations must be made to any RF, AF, or DC-voltage standards during construction and operation.

1. PRINCIPLE

The accuracy of the measurement is based on the accuracy of the square characteristics of a diode demodulator, which is required for all measurements. The square range is the portion of a diode characteristic in which the output **voltage** is proportional to the input **power**. The dynamic range of the square characteristic of a diode amounts in theory to approximately 38 dB. A range of 30 dB can surely be achieved in conjunction with a home-made demodulator probe using normal germanium diodes such as the OA 90, or AA 119. This range of approximately 30 dB represents the directly usable measuring range of the indicator. The dynamic range of the measuring setup can be extended infinitely, in theory, by adding attenuators.

The disadvantage of this measuring method is that an amplitude-modulated test signal will always be required, since it is only the AF-voltage obtained during demodulation that is indicated. The required RF-level is, however, very low; it is in the order of 20 to 200 mV. Furthermore, very low demands are placed on the quality of the AM. It is sufficient to use a simple ON/OFF modulator. Since many of our readers will not possess a signal generator, a simple diode modulator with matching drive module is to be described. The RF-signal can be taken, for instance, from one of the many local oscillator modules described in this magazine. With suitable crystals, it is then easily possible to provide a signal source for each band, or for the application in question; the only disadvantage with respect to a signal generator is that the frequency is not variable.

2. CIRCUIT

The VSWR-indicator can be split into three different parts. As can be seen **Figure 1**, the first stage is a two-stage, DC-coupled AF-amplifier. The first transistor is operated with a collector current of approximately 200 μA in order to obtain the most favorable noise characteristics. The capacitor between the base of the second transistor and ground is used to short out higher-frequency noise components.

The second stage is an AF-filter for 1000 Hz. As far as we know, this type of AF-filter is not well known, and for this reason, calculation information is to be given in the appendix. The filter can be tuned by approximately $\pm$50 Hz, since the modulation frequency will not always be 1000 Hz. In contrast to other variable frequency networks, no double potentiometer is required here for tuning. The 3 dB bandwidth of this filter amounts to approximately 20 Hz, and the overall gain of the stage is approximately 30 dB.

The last stage is a linear rectifier comprising diodes in the feedback path of an operational amplifier. A sensitivity increase of 5 dB is possible by switching a resistor. This allows the readout accuracy to be increased, which is very low in the lower range of the scale, due to the logarithmic characteristic. A red LED will light, in order to ensure that this +5 dB is not forgotten when reading off the indication.

A RC-link is provided to adjust the readout to indicate the RMS-value of a sinewave signal.

Since the whole amplifier is very sensitive (G $\approx$ 100 dB, input sensitivity less than 1 μV), the meter is dampened with the aid of a RC-link (56 Ω/1 mF) having a large time constant in order to suppress meter fluctuations due to noise.

The resistors provided on the two range-switch wafers S1a and S1b represent a division ratio of 1:10. The ratio of 1:10 corresponds to 20 dB when referred to the input **voltage** of the indicator. However, since **power** ratios are measured in the square portion of the diode characteristic, this division ratio amounts to 10 dB.

The linear scale of the meter can be converted to a corresponding logarithmically calibrated scale using the following equation:

$$\text{Attenuation in dB} = -10 \times \log \frac{1}{I}$$

where I is is the relative current through the meter.

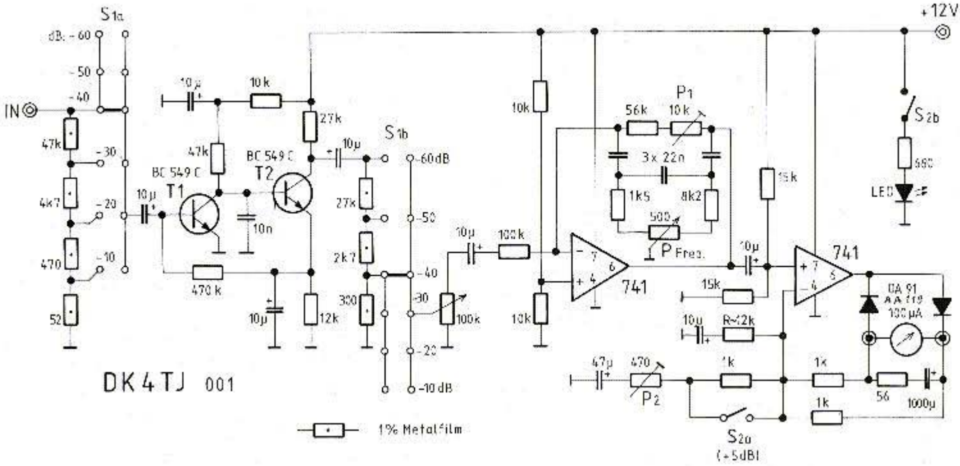


Fig. 1: The VSWR-indicator comprises AF-amplifier, 1000 Hz-filter and linear rectifier

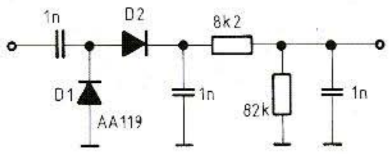


Fig. 2: The demodulator probe should be built up using good UHF-practice

The following calibration points and associated dB-values can, however, be noted, which could save having to calibrate a new scale:

I = 0.79 ≙ −1 dB
I = 0.5 ≙ −3 dB
I = 0.25 ≙ −6 dB
I = 0.2 ≙ −7 dB
I = 0.1 ≙ −10 dB

The potentiometer between the preamplifier stage and filter is used for fine adjustment. Due to the use of this potentiometer, no absolute measurements are possible.

The demodulator probe **(Figure 2)** does not have any special features with the exception of the voltage doubling circuit. This probe was originally designed for use with a swept frequency generator, but has been found to be suitable for use in conjunction with the VSWR-indicator. However, attention should be paid that the construction is suitable for UHF-applications.

3. CONSTRUCTION

With the exception of the range-switch, the potentiometers for frequency and fine adjustment, as well as the RC-link for the meter, the whole electronic circuit of the VSWR-indicator is accommodated on a PC-board. **Figure 3** shows the component locations on this board.

Resistor R is only installed after carrying out the alignment of the board. In order to avoid ground loops, only one common ground point has been provided. This ground point is the BNC-input connector. The PC-board is installed in an insulated manner and connected to the connector with the aid of a wire. The ground connection of the fine-adjustment potentiometer, and the two resistors of the voltage divider are also connected to this point. A screened AF-cable is used for the interconnection between the input connector, the rotating switch, and the electronic circuit.

If the electronic circuit is to be enclosed in a metal box, it will not be necessary for the board to be screened additionally. It is advisable to use the largest-size meter possible, since this will provide the highest readout accuracy.

No special demands are placed on the operating voltage. It should only be stable and hum-free.

3.1. Construction of the Demodulator

The demodulator circuit is shown in Figure 2. It is accommodated on an approximately 13 mm x 40 mm long, double-coated PC-board. This board can be made in the following manner:

Cut or etch one side of the PC-board to provide an approximately 6 mm wide conductor lane that is surrounded by the ground surface with the exception of one front side. This conductor lane can be interrupted at the required positions, for instance with the aid of a sharp screwdriver. This board can be accommodated in a piece of 18 mm diameter copper water pipe, which can be provided with a cap. The cap itself is provided with a hole in the center for accommodating a BNC-connector for single-hole mounting. Two bent solder tags are mounted at the same time as the connector.

These ground tags are soldered on both sides to the ground surface of the board, whereas the inner conductor of the connector is soldered to the "hot" conductor lane.

A disk or chip capacitor is used for coupling and it is placed in a slot in the PC-board. No holes are required on the board, since all components are soldered directly to the conductor lanes using short connections as required by UHF-applications.

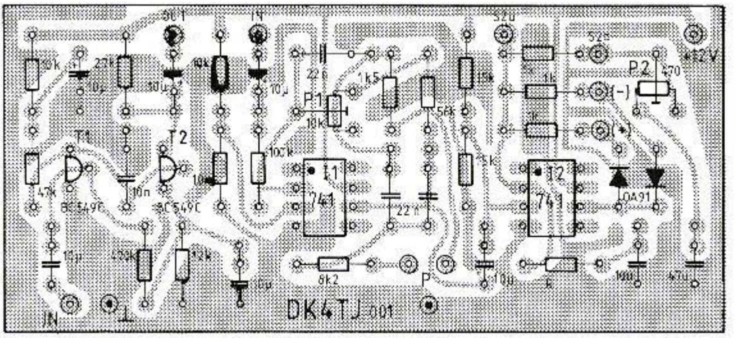


Fig. 3: PC-board and component location plan for the VSWR-indicator

RG-174 is used as interconnection cable to the indicator and is provided with the required BNC-connector. A piece of copper tubing is placed over the board up to the cap and fixed into place using a small self-tapping screw.

Only germanium RF-diodes should be used: AAY 27, AAY 21, 0A 90, AA 143, or AA 119. For this application without bias current, Schottky diodes will be approximately 20 dB less sensitive.

3.2. Alignment of the VSWR-Indicator

The alignment is limited to the filter and the +5 dB-position. This is carried out with the aid of a 1000 Hz signal of several 100 mV. A small meter reading is adjusted in the upper range with the aid of the fine-adjustment potentiometer. Align the frequency potentiometer and trimmer P_1 on the board alternately for maximum meter reading. If the meter reading is full scale, it will be necessary for the input level to be decreased. This procedure should be repeated several times since the two potentiometers interact. In some cases, it may be impossible to obtain a maximum with the aid of the frequency potentiometer due to the tolerances of the capacitors. In this case, the 1.5 kΩ resistor adjacent to the potentiometer should be changed by one standard value up or down. The potentiometer should be approximately in its center position at 1000 Hz.

Finally, the fine potentiometer and/or the AF-level is adjusted to obtain a reading of 0.316. Close switch S2 (+5 dB) and align trimmer potentiometer P_2 to full scale. This should also be repeated several times. This is followed by measuring the total resistance of the 1 kΩ resistor and trimmer P_2 with the aid of an ohmmeter. The value of the missing resistor R is obtained from the measured resistance $R = (1\ k\Omega + P_2) \times 9.03$.

The exact resistance value is not very critical; the nearest standard resistance is satisfactory, and should be installed. The switch position +5 dB should now be rechecked and corrected, if necessary.

The multi-position switch is now provided with the markings from −10 dB to −60 dB. This calibration is not per chance, but has been taken from professional equipment (for instance HP 415B, E).

4. MODULATOR

In order to help those readers that do not have a modulated signal generator, a diode modulator is now to be described that originates from a publication of Hewlett Packard given in (1). The advantage of this series-parallel circuit, as shown in **Figure 4**, is that the input and output VSWR only change slightly when switching between the two states: RF-ON and RF-OFF. This means that a good matching is guaranteed at all times for the signal generator, and test object.

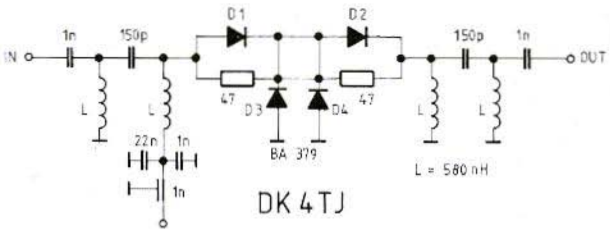


Fig. 4:
Wideband modulator using 4 pin-diodes

The three-pole high-pass filters at the input and output suppress the pulses from the driver so that they do not reach the signal generator or test object. It is a Tschebyscheff filter with a ripple of 0.01 dB within the passband range. When using the given Neosid inductances (BV5036, blue-orange, 580 nH), and the series capacitor of 150 pF, the lower cutoff frequency will be approximately 22 MHz.

The driver circuit (see **Figure 5)**, comprises a 1000 Hz-multivibrator equipped with a fast operational amplifier, which drives diodes D 1/D 2 or D 3/D 4 alternately via transistors T 1 and T 2. This generates a RF-signal that is switched on and off at a frequency of 1000 Hz.

The current flowing through the diode is limited to approximately 20 mA with the aid of trimmer potentiometer P 2. It is possible using this potentiometer to adjust the balance of the modulation signal between positive and negative amplitude. This unbalance is caused by the difference in the current gain factors of transistors T 1 and T 2.

4.1. Construction of the Modulator

The modulator is accommodated on a 35 mm x 72 mm double-coated PC-board. The component locations are shown in **Figure 6**, and it is built up very similar to the layout in the circuit diagram.

The non-required pins of the Neosid inductances are cut off as near to the coil as possible, and the ground tags are bent out by 90°. The components are mounted on the ground side of the board, and the ground tags of the screening cans of the coil are directly grounded to the surface.

After completion, the PC-board can be fitted into a metal box of 35 x 72 x 30 mm. The holes for the two BNC-flange connectors should be drilled at the appropriate positions in the case. CAUTION: The center point of the holes should not be in the center of the front panel, and should have a spacing of at least 16 mm from the upper edge of the case, otherwise it will not be possible to fit on the cover.

The board is now filed so that the conductor lane connections for the connectors are directly adjacent to the center pins of these connectors, after which they can be soldered into position. The connectors can be screwed, or soldered into place. The PC-board is now soldered to the case all around the edge on both sides of the board. This ensures a very good ground connection. The connection for the driver circuit is made via a feedthrough capacitor.

The alignment of the inductances is limited to inserting the cores to maximum inductivity.

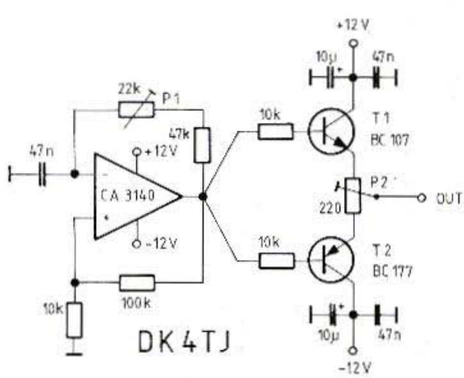


Fig. 5: 1000 Hz drive circuit for the modulator shown in Figure 4

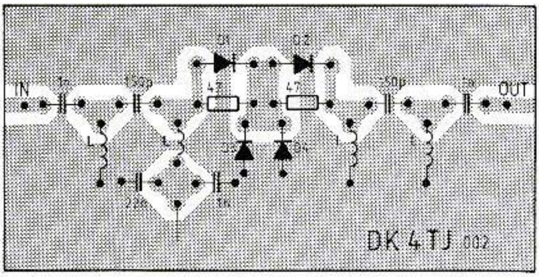


Fig. 6:
PC-board and component location plan for the modulator shown in Figure 4

4.2. Construction of the Control Circuit

The control circuit is very simple and can be easily constructed using a piece of Veroboard. Attention should be paid that the control voltage is satisfactorily bypassed, since the rise and fall slopes of the square-wave pulses are very steep.

4.3. Alignment of the Drive Circuit

A frequency counter is connected to the output and a frequency of 1000 Hz is selected with the aid of trimmer potentiometer P 1. The modulator is now connected and trimmer potentiometer P_2 is aligned to balance the positive and negative amplitude at the input of the modulator with the aid of a DC-coupled oscilloscope. If no oscilloscope is available for this alignment, the wiper of the potentiometer is set approximately to a value of 1/3 to 2/3 in the direction of T 1.

5. APPLICATIONS

When using the measuring setup comprising modulated signal generator, demodulator probe, and VSWR-indicator, it is now possible for all types of attenuation and gain measurements to be made. It should be noted , however, that the RF-part of the measuring setup must always be the same (for instance 50 Ω), and it may be necessary to improve the matching by providing additional attenuators. The required RF-level is then, of course, higher. Since the demodulator probe also does not possess a terminating resistor, a correct termination is provided using a 50 Ω resistor in a BNC-connector, which is connected via a BNC T-adapter to the demodulator. The system impedance is now fixed at 50 Ω. **Figure 7** shows a block diagram of this measuring circuit.

Before using the measuring system for the first time, one must determine at which RF-input voltage the square range of the demodulator ceases. This range commences in the noise, and will cease at the transition into the linear range, where the output voltage is proportional to the input **voltage.**

In order to carry out this test, one will require an additional 3 dB attenuator (fixed or variable). Firstly, connect the measuring setup as shown in Figure 7, preferably using a frequency of less than 100 MHz.

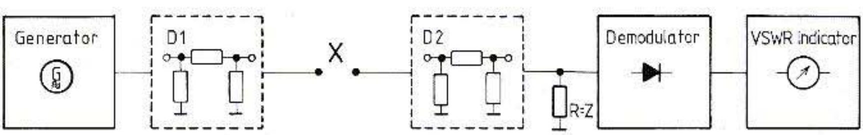


Fig. 7: Measuring setup for use with the VSWR-indicator

The test should be made as follows:

1. Set the fine-adjustment potentiometer to maximum
2. Set the range switch to −30 dB
3. Adjust the RF-level for full scale
4. Connect the 3 dB-attenuator at "X".

The meter reading should now drop by 50%. If this value is not indicated, the input level will probably be too high. In this case, reduce the input signal, for instance, to a value of 0.5 in the −30 dB-position (= −33 dB), or even to full scale in the −40 dB-position. Repeat the test again.

In this manner, it is possible to determine at which input level the square range of the demodulator ceases. This range should **never** be exceeded, otherwise no accurate measurements will be possible.

The selection of the −30 dB-point is also taken from professional measuring equipment. However, this point need not be defined exactly. It is firstly dependent on the gain of the VSWR-indicator (the gain of the two input transistors T1 and T2, and thus the gain of this stage is not defined), and secondly on the diodes and their dynamic load in the demodulator.

As an example, an attenuation measurement is to be carried out on a home-made 20 dB-attenuator. The measuring setup is as shown in Figure 7.

1. Set the level by adjusting the RF-signal and potentiometer to, for instance, −30 dB
2. Interconnect the attenuator at "X".

The potentiometer may not be varied during the following measuring procedure, since this would alter the reference level and make the measurement invalid.

3. Readout: The readout will drop to less than 0.1; Switch to −40 dB range, readout still less than 0.1; Select −50 dB-range, readout indicates 0.96 (−0.2 dB).
4. Test result:
   2 x −10 dB + −0.2 dB
   = −20.2 dB attenuation.

The examined attenuator thus has an attenuation of 20.2 dB at the selected measuring frequency.

Due to the relative spread of the scale in the upper half, especially between values of 0.79 (−1 dB), and 1.0 (0 dB), it would even be possible to evaluate 0.1 dB steps. This is very useful, for instance, when carrying out measurements of the insertion loss of filters.

As second example, let us carry out an attenuation measurement of the characteristic curve of a bandpass filter. Since the dynamic range of the measuring setup only amounts to approximately 30 dB, whereas the filter is to be examined at frequencies far from its center frequency, it was decided to use the attenuator measured in example 1 in the measuring setup shown in Figure 7.

The reference level is now set at the center frequency of the filter, for instance, once again to −30 dB. The frequency of the signal generator is now varied in suitable steps, and the resulting attenuation values are noted. When the measuring limit is reached at −60 dB, it is possible for the attenuator to be removed, and the range switch of the VSWR-indicator can be switched up to −40 dB.

The same meter reading should now be adjusted. This allows the available dynamic range to be increased by the value of this 20 dB (or 20.2 dB) of the attenuator. The usable range now amounts to approximately 50 dB. Of course, one should not forget this value of 20.2 dB when evaluating the test results.

It should be mentioned, of course, that the amplitude and the degree of modulation of the signal generator must remain constant on varying the frequency.

An example of a gain measurement is now to be given by measuring the gain of a preamplifier. The measuring setup is as shown in Figure 7.

1. Set the RF-level and potentiometer to, for instance, −60 dB. This corresponds to a meter reading of 0.1 in the −50 dB range.
2. Connect the preamplifier to position "X"

3. Readout: Meter reading at full scale, switch up to the next highest range (−40 dB), meter still at full scale, switch up to −30 dB range, meter reading is approx. 0.2.

In order to improve the readout accuracy, the +5 dB switch is selected: The meter reading is now 0.66 (−1.8 dB).

4. Test result:

$$\underset{\text{switch}}{2 \times +10\,\text{dB}} + \underset{\text{scale}}{10\,\text{dB} - 1.8\,\text{dB}} - \underset{\text{switch}}{5\,\text{dB}} =$$

$$+30\,\text{dB} - 6.8\,\text{dB} = 23.2\,\text{dB gain}$$

Once again, attention must be paid that the potentiometer is not altered during the measuring process since otherwise one will lose the reference level.

Of course, everyone knows, that it is necessary, when carrying out measurements on active modules, for their operating voltage to be provided (important when carrying out matching measurements).

In the author's opinion, the greatest advantage of the VSWR-indicator is provided when carrying out measurements of matching. It is no longer necessary to determine the correct tapping point of an inductance, or the correct division ratio of a capacitive voltage divider experimentally: one can measure it! Of course, one will require a directional coupler, or a return-loss bridge as described in (2), and (3). Further details regarding the measuring principle were given in (3).

When using the measuring setup shown in **Figure 8,** it is possible to carry out matching measurements down to a VSWR of less than 1.05 if the quality of the directional coupler or return loss bridge is good enough. One does not measure the VSWR directly, but the return loss, from which the VSWR can be calculated.

As an example, let us determine the correct tapping point on a coil, or cavitiy.

1. Adjust the level with RF-signal and/or potentiometer, for instance, to −30 dB
2. Connect the test object to the bridge, **non potentiometer!**
3. Readout: Meter reading is less than 0.1; select −40 dB range, readout is 0.31 (−5 dB)
4. Test result:
   −10 dB + −5 dB = −15 dB return loss
   switch + scale

The resulting return loss can now be converted to VSWR in the following equation:

$$\text{rho} = 10^{-0.05\,\text{RL (dB)}}$$

$$\text{VSWR} = \frac{1+\text{rho}}{1-\text{rho}}$$

According to this, a return loss of −15 dB corresponds to a VSWR of 1.43.

Since this is still not satisfactory for our application, the VSWR is to be improved by changing the tapping point. It should be noted in practice that values of less than 1.2 (≙ return loss = −20 dB) are only academic.

A table with return loss and the associated VSWR-values is to be found in (4).

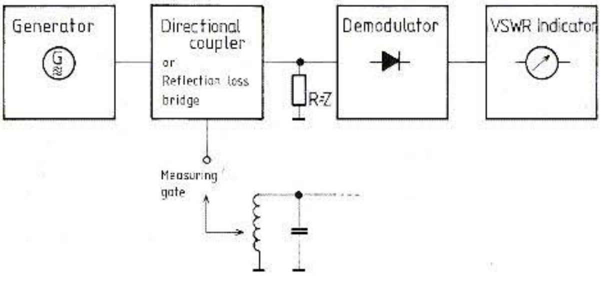


Fig. 8: Measuring setup for matching measurements

It should also be mentioned that the "frequency" potentiometer of the VSWR-indicator should be adjusted to the modulation frequency of the signal generator used, previous to each measurement.

Furthermore, the fine-adjustment potentiometer should always be within the the upper third of its adjustment range so that one cannot exceed the square range of the demodulator.

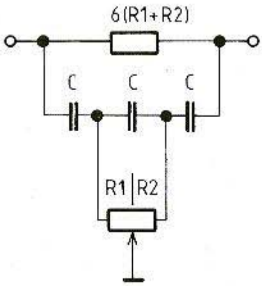


Fig. 9: The (Notch) filter of the VSWR-indicator

6. MEASUREMENT ERRORS AND EXPERIENCE

Anyone that carries out measurements regularly, will know the numerous measuring errors that can be made. On its own, (as AF-voltmeter), the VSWR-indicator is very accurate, although the −60 dB position is of limited use due to the effect of noise in the lower part of this range. However, during all measurements it is necessary for the whole system comprising signal generator, modulator, possibly attenuators, terminating resistor, demodulator, and VSWR-indicator, to be considered. Furthermore, the test object also has its effect. Each coaxial transition (or adapter) and each piece of cable have an effect on the test result (5). According to the author's experience, the measuring system can be used without limitation up to 300 MHz. The inaccuracies in excess of 300 MHz have not been examined with the available measuring aids.

In order to determine the quality of the modulator-demodulator part of the system, the author measured the matching of a terminating resistor (Bird Thermaline) with the aid of a directional coupler manufactured by EME.

In comparison to a VSWR-indicator HP 415 E, and a professional diode-detector 1N 21 (with built-in 50 Ω termination) in conjunction with a directly modulated RF-signal from professional signal generators (Wavetek 3001, R&S SDR), the values measured using the described circuit were virtually just as good.

For this reason, the author is willing to suggest that the described measuring system including modulator can be used up to approximately 1000 MHz. When using the VSWR-indicator together with professional equipment (generator and demodulator), the upper frequency limit is determined by these components. The technical upper limit is in excess of 10 GHz.

6.1. Measuring Equipment Used

Wavetek signal generator 3001 (520 MHz)
R&S signal generator SDR (1000 MHz) with 50 Ω matching network DAF
HP calibrated attenuator 355 C + 355 D
Directional coupler (EME, type 7020/30 A) (144 MHz – 1296 MHz)
Bird „Thermaline" terminating resistor 50 Ω, type 80 M
Philco 50 Ω crystal detector, type 148 (with 1N 21)
Minicircuit attenuators type CAT 3, 6, 10, 20 dB
HP 415 E VSWR-indicator.

7. APPENDIX

The AF-filter shown in **Figure 9** is actually a Notich filter, and is based on a development of General Radio. It is described in detail in (6). The basic design criteria is:

$$F_{zero} = \frac{1}{2\pi \times C \times \sqrt{3 R_1 R_2}}$$

with R in Ω, and C in Farad.

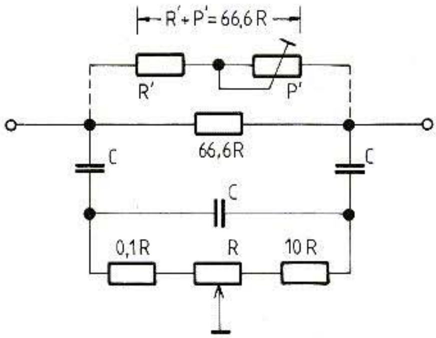


Fig. 10: Design of the filter for a tuning range of 1:3

When using the resistance values shown in **Figure 10**, one will obtain a tuning range of approximately 1:3. The calculation of the frequency range to be covered can be made as follows:

Lower cutoff frequency $F_l = \frac{27700}{C \times R}$

Upper cutoff frequency $F_{up} = 3 \times \frac{27700}{C \times R}$

with R in Ω and C in μF.

When used in the VSWR-indicator, the tuning range is limited to approximately ± 50 Hz at a center frequency of 1000 Hz by selection of a suitable small potentiometer. In order to obtain the lowest possible bandwidth, it is important that the three capacitors have exactly the same values (select, if necessary). Furthermore, the value of the upper resistor should have exactly six times the value of the sum of the three lower resistors. For this reason, this resistance value is best realized using a combination of a fixed resistor and trimmer potentiometer.

Altough originally a Notch filter, this filter will have a bandpass filter characteristic due to the invertion function of the operational amplifier.

8. REFERENCES

(1) H. Vifian/F. K. David/W. K. Frederik: „A Voltmeter for the Microwave Engineer" HP-Journal 11/1972, page 6

(2) Gerd Wolske: „Eine Stehwellen-meßbrücke für beliebige Frequenzen bis 500 MHz" cq-DL 10/1974, page 606

(3) Michael Martin: DJ7VY Wideband Directional Coupler for VSWR-Measurements on Receiver Systems VHF-COMMUNICATIONS 15, Ed. 3/1983, Pages 153–162

(4) Otto Frohsinn, DF7QF: A Home-Made UHF/SHF Power Meter VHF-COMMUNICATIONS 13, Ed. 4/1981, Pages 221–229

(5) J. Ehrlich: Meßtechnik im UHF/SHF-Bereich Tagungsheft VHF/UHF-Tagung München 1982, Page 58

(6) C. Hall: „RC-Tunable Notch Filter" Ham Radio Magazine 9/1975, Pages 16–20

Additional Information:

(a) Operation Manual HP 415 E VSWR Indicator Hewlett-Packard Co., Palo Alto

(b) Bob Melvin: The SWR-Meter Ham Radio Magazine 10/1970, Pages 6–14

(c) Tietze/Schenk: Halbleiterschaltungs-technik 3. Auflage, Pages 139, 258–261

(d) Fred Brown: Antenna Gain Measurements QST 11/1982, 12/1982, page 30

Willy van Driessche, ON6VD

Suggestion for Standardizing SSTV and FAX Transmissions

What is SSTV?

Slow-Scan Television (SSTV) is, basically speaking, not different from facsimile (FAX) transmissions with a speed of 960 lines per minute. However, one advantage of FAX-transmissions is not used: namely that is is not necessary for synchronizing pulses to be transmitted after the image has been "phased in". If no synchronizing pulses are needed, they cannot be missed, when transmission disturbances occur. Especially when the receive system is using a scrolling system such as is the case with the YU3UMV storage module, the stable line frequency is used to the full extent.

Comparison of the Line Frequencies Used

In SSTV, the following are used:

Europe: 60 ms incl. 5 ms sync.
≙ 16.666 Hz
USA: 66.7 ms incl. 5 ms sync.
≙ 15 Hz.

The line frequencies with FAX are:

60 lpm = 1 Hz
120 lpm = 2 Hz
240 lpm = 4 Hz
480 lpm = 8 Hz
960 lpm = 16 Hz

whereby the frequency deviation shall not exceed a value of 10^{-5}.

The two highest frequencies are still not used for FAX transmissions, and they have only been listed in order to indicate the similarities between FAX and SSTV. 480 lines per minute would be a good compromise between image quality and transmission time (32 seconds for an image). Such a transmission could be called FAX or SSTV.

Recommendation of a Suitable SSTV Standard

It would be advisable for SSTV to have a frequency of 16 Hz and a maximum stability of 10^{-5}. The 5 ms synchronizing pulse could be provided in each line. In this manner, SSTV equipment could "phase in" and thus be virtually fully compatible with FAX equipment. The stable line frequency of 16 Hz (8 Hz) would be just as suitable for FAX equipment such as the YU3UMV storage, as for SSTV equipment in Europe and the USA.

In the case of FAX transmissions, the 1200 Hz-synchronizing pulses could be used in the same manner for phasing in, as is the case on board YU3UMV 001 for the reception of the NOAA and METEOR weather satellites. In the case of SSTV, it would be possible for equipment available on the market to be used, which means that full compatibility would exist.

New SSTV equipment should have the possibility of using the synchronizing pulses in the same manner as with FAX (in other words only for phasing in), as well as a changeover switch to use normal SSTV line synchronization.

A publication in QST of August 1983 ("High-Resolution SSTV") indicated that there is a trend to more lines. In the case of images with 256 x 256 pixels, it would be very advisable when the standard was compatible with 480 lpm FAX.

Any comments would be gratefully received by the author: either direct or via the editors.

Werner Hanschke, DCØRZ

C-MOS Frequency Counter for 10 Hz to 1 GHz

Frequency counters represent probably one of the most popular pieces of measuring equipment for radio amateurs. The increasing integration density of digital C-MOS circuits, as well as advances made in input prescalers allow new concepts to be realized. This led to development of a small, highly integrated frequency counter that operates from the lowest audio frequencies up to the UHF-range

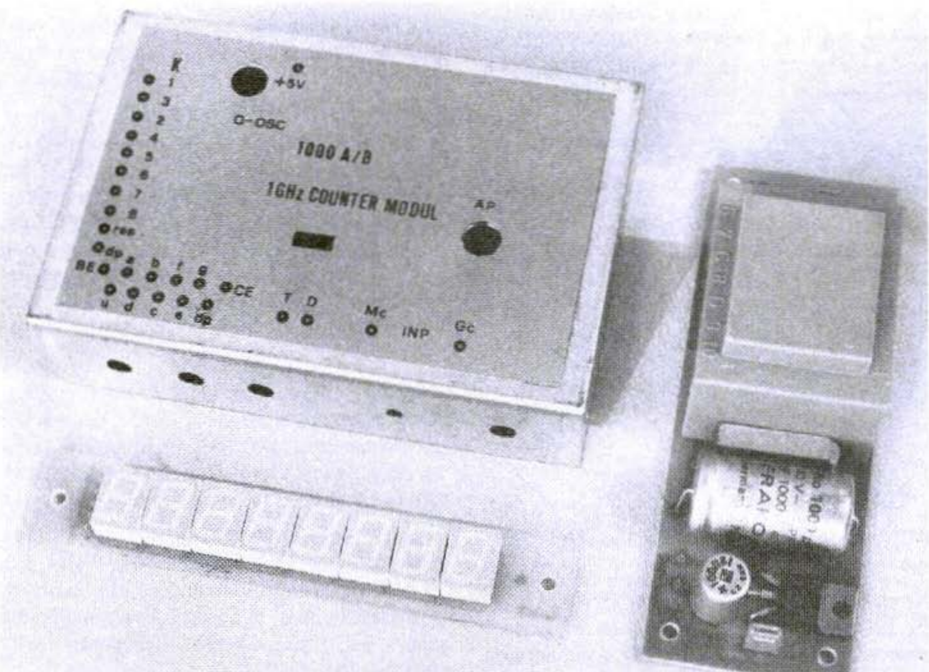


Fig. 1: Counter module, readout, and power supply before the connection

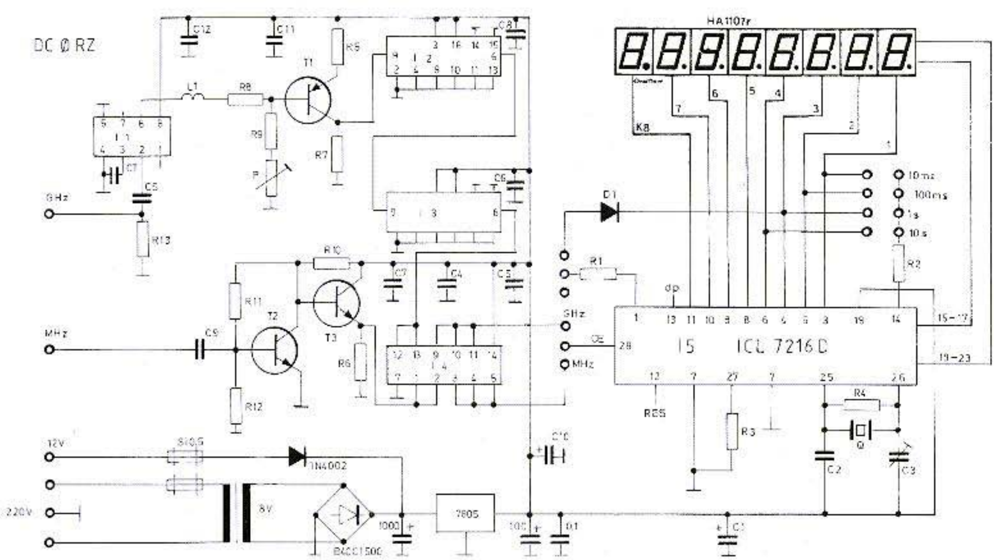


Fig. 2: Circuit diagram of the counter with eight-digit readout and power supply

The frequency counter has two measuring ranges: The first from 10 Hz to 10 MHz, and a further range from 20 MHz to 1 GHz. **Figure 1** shows the counter module (74 mm x 111 mm x 30 mm), the still not connected eight-digit readout, as well as the small power supply.

1. CIRCUIT DESCRIPTION

Figure 2 shows the circuit diagram of the frequency counter. The actual counter is within the integrated MOS-circuit type ICM 7216 D, manufactured by Intersil (1). This in turn drives eight, 10 mm high, 7-segment LED readouts without drivers. This is made in a Multiplex operation working with approximately 500 Hz. The counter-IC possesses a clock oscillator, an advanced control electronic for four gate times, measured-value indicator, overflow indicator (indicated by the first zero point of the display), automatic decimal point control, suppression of unwanted zeros, and finally the 8-digit BCD-counter with intermediate storage.

The integrated circuit can count frequencies up to 10 MHz (max. 14 MHz). A two-stage transistor amplifier (T 2, T 3) is provided for this direct frequency range. It provides sufficient gain for measurements between 10 Hz and maximum 14 MHz.

A prescaler is required for higher frequencies. A prescaler type U 664 B (2) manufactured by Telefunken was used here; it possesses a very unfavorable frequency division factor of 64, provides, however, on the other hand an excellent frequency range of 20 MHz to 1 GHz, and an integrated, highly sensitive preamplifier. Its sensitivity amounts to typically 10 mV. The correct division factor (decimal) of the UHF-prescaler is made with the aid of two, programmable TTL-dividers type SN74167. They divide by 5:4, each, so that a total division factor of 64:1 x 5:4 x 5:4 = 100:1 results.

Both signal paths: The direct 10 MHz-range and the UHF-range divided by 100:1, are fed to a TTL-trigger gate type SN74132. The trigger characteristics of this IC allow one to do without a complicated level control. The TTLsignals at the output of the SN74132 (I 4) are fed via the MHz/GHz bridge to the input of the counter.

In order to obtain a clear switching of the operating functions, it is advisable to use a six-pushbutton assembly having two changeover contacts, each. The ON/OFF-pushbutton and the MHz/GHz-button should be operated individually, whereas the gate time buttons should deactivate the unwanted ranges automatically.

The operating voltage of +5 V is provided by the power supply, which also allows 12 V operation, for instance, for mobile and portable applications.

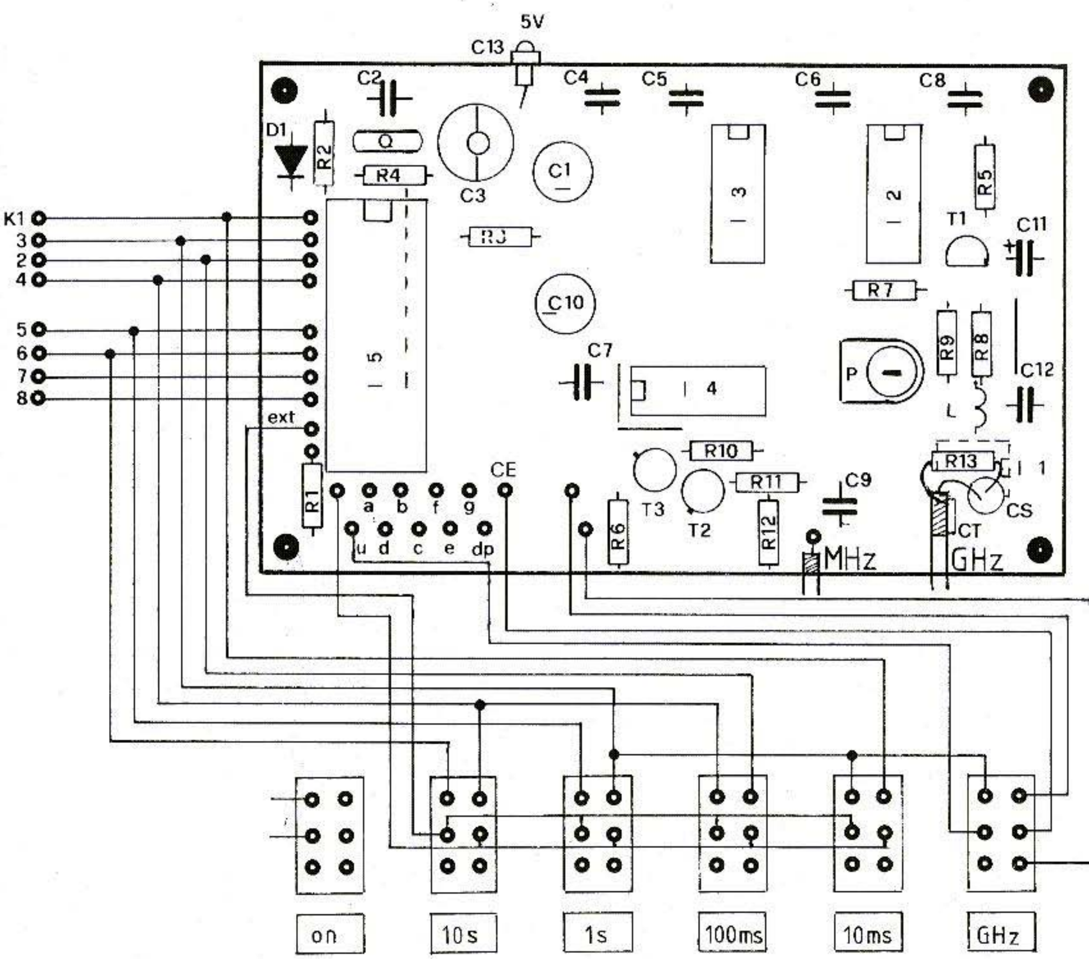


Fig. 3: Component and wiring plan for the counter board DCØRZ 002

2. SPECIFICATION

Frequency ranges:	10 Hz – 10 MHz, 20 MHz – 1 GHz
Sensitivity:	30 mV (typ.), 10 mV (typ.) with automatic triggering
Resolution:	0.1 Hz (max.), 10 Hz (max.)
Gate times:	10 s/ 1 s/ 100 ms/ 10 ms
Crystal time base:	10 MHz
Accuracy:	1×10^{-7} (typ.)
Readout:	eight digits, with unwanted zero suppression
Power supply:	220 V/5 V_{stab}/300 mA, as well as input 12 V-DC
Dimensions (mm):	74 x 111 x 30 (without readout and power supply)

Maximum test voltage at the input connector: 1 V into 50 Ω!

3. PARTS LIST

PC-board DC0RZ 002
Metal box 74 x 111 x 30 mm

I 1:	U 664 B (Telefunken)
I 2, I 3:	SN74167 (TI)
I 4:	SN74132 (TI)
I 5:	ICM 7216 D (Intersil)
T 1:	BF 324 (PNP-RF transistor)
T 2, T 3:	2N 708 or similar switching transistor
D 1:	1N 4148 or similar switching diode
Q:	10 MHz crystal, HC-18/U
Cs:	1 nF ceramic disk
C_T:	1 nF chip capacitor
C 1, C 10:	100 µF/16 V, 5 mm spacing
C 2:	56 pF ceramic, spacing 5 mm
C 3:	Plastic foil trimmer, 10 mm dia., yellow
C 4 – C 6:	10 nF, ceramic, 2.5 mm spacing
C 7:	100 nF, ceramic, 5 mm spacing
C 8, C 12:	10 nF, ceramic, 2.5 mm spacing
C 11:	1 µF/16 V, 2.5 mm spacing
C 13:	1 nF feedthrough capacitor
L 1:	5 turns of 0.5 mm dia. enamelled copper wire wound on a 5 mm former, self-supporting
R 1, R 2:	10 kΩ (all 0207, 5%)
R 3:	100 kΩ
R 4:	22 MΩ
R 5:	22 Ω
R 6:	470 Ω
R 7:	330 Ω
R 8:	150 Ω
R 9:	1 kΩ
R 10:	2.2 kΩ
R 11:	15 kΩ
R 12:	22 kΩ
R 13:	51 Ω
P:	4.7 kΩ poti., spacing 10/5 mm

1 pc. 28-pin IC-socket

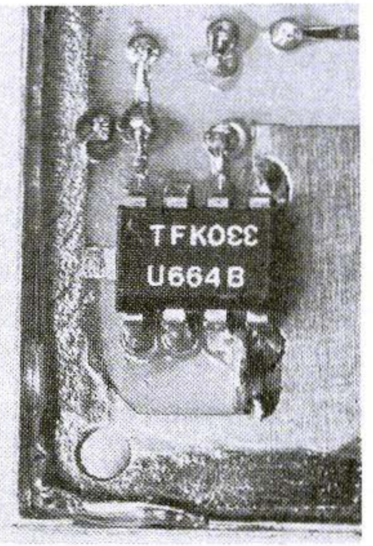


Fig. 4: When correctly installed for UHF, many U 664 B operate up to 1300 MHz

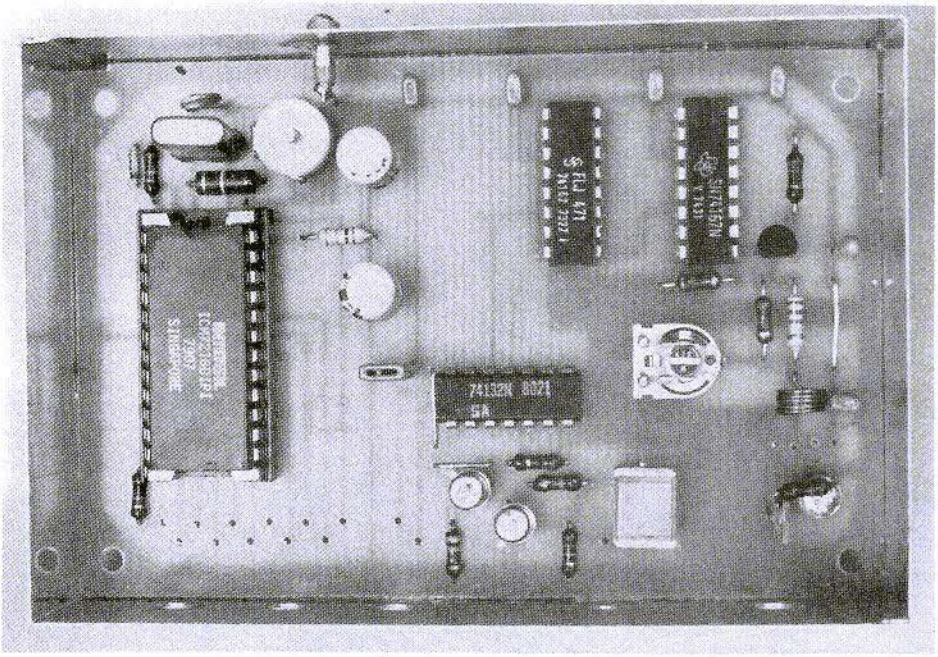


Fig. 5: One will see the input circuit of the U 664 B of the counter module

4. CONSTRUCTION

The frequency counter is accommodated on a 109 mm x 72 mm single-coated PC-board. **Figure 3** shows the component locations and the required external connections to switches, readout, power supply, and input connectors of this board, which has been designated DR0RZ 002. Before mounting the components, the PC-board is mounted into the metal box with a spacing of 5 mm from the base plate.

The next step is to cut a slot for the chip capacitor C_T between connections 3 and 4 of I 1 with the aid of a fret saw. Attention should be paid that I 1 must be soldered into position on the conductor side of the board, and that C_T is mounted on the component side directly between the pins of the IC. Before soldering into place, pins 1, 3, 5, 6, 7 and 8 of I 1 (U 664 B) should be shortened so that only pin 2 and 4 appear on the component side and can be bent down. **Figure 4** shows I 1 after being soldered into place.

The ceramic disk capacitor C_s is now soldered to the bent-down pin 2 on the component side. It is now possible for the 51 Ω resistor R 13 to be connected between this disk capacitor and the ground side of the chip capacitor. This can be seen on the lower right-hand side of **Figure 5**.

A 28-pin socket should be used for I 5, the integrated counter; before which one should not forget to solder the required bridge into position. Three further small bridges at other positions on the board should also not be forgotten. These bridges are made in order to save using a double-coated construction.

The three TTL-circuits can be soldered directly to the board without sockets, after which the other components can be mounted

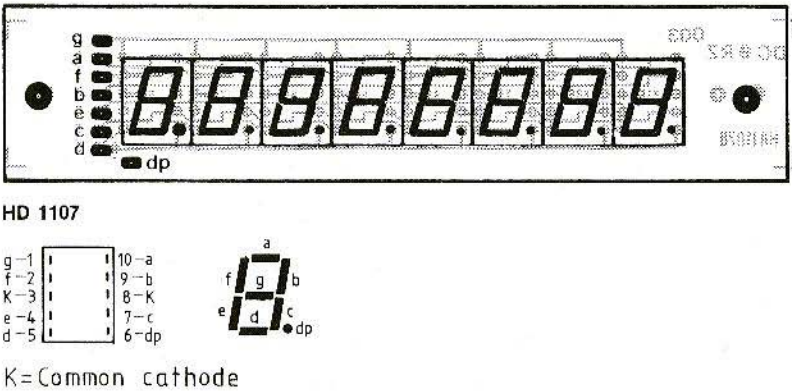


Fig. 6: Readout board DC Ø RZ 003 with connections for the 7-segment LED readouts

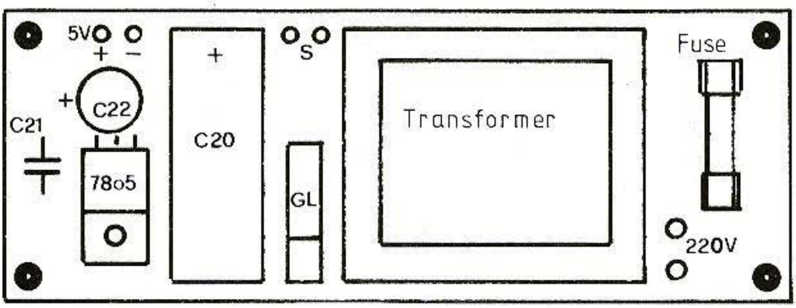


Fig. 7: Power supply board DC Ø RZ 004 is able to provide 300 mA at +5 V

into position. The UHF-prescaler U 664 B is installed last, after being provided with heat-conductive paste so that it can be cooled by the lower cover, or by a plate screwed into position in the counter.

5. READOUT AND POWER SUPPLY

The eight-digit, 7-segment readout can be assembled on the single-coated PC-board DC 0 RZ 003 shown in **Figure 6**. The PC-board is only 20 mm wide, and 109 mm long. The LED-readouts type HD 1107 r (Siemens) have 10 mm high, red digits. Each of their 7 segments is parallel-connected at the side of the PC-board where connections a – g are provided. The connections for the parallel-connected decimal points dp are also provided there. The cathodes, on the other hand, are connected individually from K 1 – K 8 to the counter module.

Figure 7 finally shows a recommendation for the power supply. This can be accommodated on the 109 mm x 40 mm single-coated PC-board DC 0 RZ 004. Of course, this board is only suitable when the matching transformer is available. All other components are standard types.

6. REFERENCES

(1) Data sheet Intersil ICM 7216 B

(2) Data sheet Telefunken U 664 B

T. Bittan, G3JVQ/DJØBQ

Satellite News

NEW METEOSAT 2 Dessemination Schedule.

A new dessemination schedule (timetable) becomes effective on September 1st, 1984. A copy of this schedule can be obtained from the publishers for DM 3.00, or from your national representative. A new booklet describing the METEOSAT operating system is now available in English for DM 1.00.

There are a number of changes in the Meteosat Schedule, since many of the times have changed. The main changes for European users are that the half-hourly sequence of D2/C02/C03 is now to commence at 10 minutes and 42 minutes past the hour; unfortunately, still not a 30 minute interval!

NOAA-8 has failed and NOAA-6 reactivated

NOAA-8 suddenly failed in orbit. According to our information this has been caused by a fault in the master oscillator that controls all gyro and stabilization systems. This is not only a loss to the meteorologists but also to aviation as a whole, since NOAA-8 was the first of the NOAA series to have a search and rescue system on board to locate crashed aircraft and other emergency transmissions worldwide on 121.5 MHz.

It was possible to reactivate NOAA-6 to obtain satellite imagery, and this satellite is now performing well in its morning descending (N → S), and evening ascending (S → N) orbits.

NOAA-9 to be launched in October 1984.

The launch of the next NOAA satellite, NOAA-9, is planned for October 23rd, 1984. However, the difficulties involved with NOAA-8 could possibly delay the launch until the problems have been determined and solved.

TV-satellites that can already be received in Europe.

Many readers will not know that a number of TV-satellites can already be received in Europe. Of course, this is not so easy as in North America where these satellites operate with normal standard NTSC (A5) transmissions that do not need to be recoded. The only such satellite that can be received in Europe is the Russian Gorizont satellite located in a geostationary orbit at 14°W. This satellite transmits the USSR first programme on 3875 MHz, and second USSR programme on 3825 MHz. The second programme is approximately 3 dB weaker. The publishers have been receiving just noise-free images with a 1.4 m parabolic antenna. The transmissions are right-hand (clockwise) circular-polarized.

Another Gorizont satellite has been received which is located at 53°E and is operating on 3675 and 3875 MHz.

All of the other communication satellites operating in the 3 to 4 GHz satellite band require extremely large dishes, which are usually too large for European gardens!

Now to the TV-satellites operating in the 11 to 12 GHz satellite band. These satellites are mainly using wideband-FM video and PCM-sound. Table 1 gives a list of the TV-satellite transponders that can be received with 1.5 to 2 m diameter parabolic antennas from Intelsat V (FO4), which is located at 27.5°W.

The only other satellite providing sufficiently strong signals for small antennas in Europe is ECS 1, which is located at 13°E.

Some of these channels are coded. Further information is given in Table 2.

In the meantime, the ESA has launched ECS-2, but no channel information is known at present.

We at VHF COMMUNICATIONS intend to study the TV-satellite reception technology further and hope to also bring constructional articles to allow our readers to construct their own TV-satellite reception systems. The editors would like to hear from amateurs that have designed and constructed such equipment and would be willing to pass on their experience to other readers.

Frequency /MHz)	Polarization	Operator
11.135	linear, horizontal	Screen Sport, UK
11.175	linear, horizontal	The Entertainment Network, UK
11.674	linear, horizontal	The Music Channel, UK

Table 1: Strong TV-transponders aboard Intelsat V (FO4)

Frequency (MHz)	Polarization	Operator
11.055	linear, horizontal	ZDF and ORF (Germany/Austria)
11.491	linear, horizontal	TV-5 (France)
11.650	linear, horizontal	Sky Channel (UK)
10.986	linear, vertical	Paysat (Switzerland)
11.158	linear	Esselte Video (Belgium)
11.507	linear, vertical	PKS-Studios (Germany)
11.674	linear, vertical	The Music Channel (UK)

Table 2: Strong TV-transponders aboard ECS-1

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 3/1984 of VHF COMMUNICATIONS

			Art. No.	
DJ6PI	**Low-Noise Preamplifiers for Weather Satellite Reception at 1.7 GHz**		Art. No.	**Ed. 3/1984**
	1. Two-stage bi-polar version (F = 2.4 dB; G = 23 dB)			
PC-board	DJ6PI 012	0.79 mm thick glass/epoxi board, double coated, silver-plated, undrilled	6865	DM 17,—
Components	DJ6PI 012	1 NE64535, 1 NE57835, 1 Z-diode, 3 chip, 5 feed-through, 1 ceram. capacitors, 3 foil trimmers SKY green, 1 tantalum electrolytic cap., 2 trimmer potentiometers, 10 resistors, 4 ferrite beads, 1 metal box, 1 N-flange conn., 1 BNC-flange connector	6866	DM 161,—
Kit	**DJ6PI 012**	**complete, with above parts**	**6867**	**DM 175,—**
	2. Single-stage GaAs-FET version (F = 1 dB; G = 14 dB)			
PC-board	DJ6PI 013	0.79 mm thick RT/Duroid 5870, double-coated, silver-plated, undrilled	6868	DM 31,—
Components	DJ6PI 013	2 ATC chip capacitors, 1 Johanson trimmer, 1 foil trimmer SKY green, 2 ceram. disk caps., 3 feedthrough, 1 ceramic, and 1 tantalum capacitor, 2 resistors, 1 trimmer potentiometer, 1 voltage stabiliser, 1 metal box, 1 ferrite bead, 1 N-flange plug, 1 N-flange socket	6869	DM 87,—
Kit	**DJ6PI 013**	**with above parts (without GaAs-FET!)**	**6870**	**DM 115,—**
GaAs-FETs for	DJ6PI 013	MWT 11 (Siemens)	9737	DM 79,—
		MGF 1402 (Mitsubishi)	9738	DM 130,—
		CFY 19 (Siemens)	9739	DM 94,—
DC8UG	**Noise Generator with defined noise power**			**Ed. 3/1984**
	Available as a ready-to-operate, tested noise generator module, together with PC-board DC8UG 006 (power supply/switching amplifier) with a kit of semiconductors.		0359	DM 398,—*
DJ3RV	**Optimum Crystal Filter for Coherent CW and Coherent RTTY**		Art. No.	**Ed. 3/1984**
	The publishers plan to offer this filter as a ready-to-operate module (aligned by the author) if sufficient demand arises. Since the specially manufactured crystals will be rather expensive, we would like to ask interested readers to Contact us (directly and non-obliging).		approx.	DM 1400,—*

* Available directly from the publishers.

DLØHV	**RF-Millivoltmeter with probes for 1500 MHz**		Art. No.	**Ed. 2+3/84**
PC-board	DLØHV 001	European Standard Size (ESS = 100 mm × 160 mm), double-coated, silver-plated, drilled	6851	DM 39,—
Components	DLØHV 001	2 voltage stab. ICs, 4 switching and 4 Schottky diodes, 6 ICs, 1 pair of 31-pin connectors (DIN 41617)	6852	DM 77,—
PC-board	DLØHV 002	ESS, double-coated, silver-plated, drilled	6853	DM 39,—
Components	DLØHV 002	2 voltage stab. ICs, 2 FETs, 2 ICs, 1 potted-core kit, 1 pair of 31-pin conn. (DIN 41617)	6854	DM 60,—

PC-board	DLØHV 003	ESS, double-coated, silver-plated, drilled	6855	DM 41,—
Components	DLØHV 003	4 red LEDs, 1 Schottky-diode, 8 ICs, 1 pair of 31-pole conn.	6856	DM 52,—
PC-board	DLØHV 004	34.7 mm × 71.5 mm, double-coated, silver-plated, undrilled	6857	DM 15,—
Components	DLØHV 004	2 Germanium diodes, 4 mini RF-chokes, 1 trapezoid and 3 chip capacitors, 1 tinned-metal case 37 × 74 × 30 (mm)	6858	DM 20,—
PC-board	DLØHV 005	9 mm × 68 mm, double-coated, silver-plated, undrilled	6859	DM 14,—
Components	DLØHV 005	4 Germanium diodes, 2 mini RF-chokes, 5 chip capacitors	6860	DM 14,—
PC-board	DLØHV 006a	ESS, double-coated, silver-plated, drilled	6861	DM 40,—
PC-board	DLØHV 006b	40 mm × 100 mm, double-coated, silver-plated, drilled	6862	DM 25,—
Components	DLØHV 006	11 small red LEDs, 5 switching diodes, 7 FETs, 11 AF transistors, 3 linear and 7 C-MOS ICs, 1 potted-core kit, 1 pair of 31-pin connectors (as above)	6863	DM 111,—
DK4TJ	**VSWR-Indikator**		Art. No.	**Ed. 3/1984**
PC-board	DK4TJ 001	single-coated, silver-plated, undrilled	6871	DM 27,—
Components	DK4TJ 001	2 transistors, 1 IC, 2 diodes, 1 LED, 2 trimmer potentiometers, 19 carbon and 7 metal film resistors, 4 ceram., 7 tantalum and 2 aluminium capacitors	6872	DM 29,—
PC-board	DK4TJ 002	double-coated, silver-plated, undrilled	6873	DM 16,—
Components	DK4TJ 002	4 PIN diodes, 4 Neosid inductances, 2 resistors, 6 ceram., 1 feedthrough capacitors, 1 tinned-metal box, 2 BNC flange conn.	6874	DM 39,—
Kit	**DK4TJ 001/002 complete, with all above parts**		**6875**	**DM 108,—**
DCØRZ	**C-MOS Frequency Counter for 10 Hz to 1 GHz**			**Ed. 3/1984**
PC-board	DCØRZ 002	single-coated, drilled	6876	DM 27,—
Mechan. parts	DCØRZ 002	PC-board in metal box already drilled, 28-pin IC-socket	6877	DM 62,—
Components	DCØRZ 002	1 prescaler IC, 3 TTL-ICs, 1 CMOS counter IC, 3 transistors, 1 diode, 1 crystal 10 MHz, 1 disc and 1 trapeze capacitor, 7 ceram. and 1 feedthrough cap., 1 foil trimmer, 1 tantalum and 2 aluminium electrolytics, 1 m of 0.5 mm enamelled copper wire, 1 trimmer potentiometer, 13 resistors, 8 off 7 segment displays	6878	DM 308,—
PC-board	DCØRZ 003	single-coated, drilled	6879	DM 19,—
PC-board	DCØRZ 004	single-coated, drilled, as well as		
Components	DCØRZ 004	1 mains transformer (220 V), 1 bridge rectifier, 1 electrol. cap., 1 voltage stabiliser	6881	DM 33,—
Kit	**DCØRZ 002 – 004 complete, with above parts**		**6882**	**DM 410,—**

A system for Reception and Display of Weather-Satellite Images using a digital scan converter/storage module

A) A complete system as kits

Description	Edition	Kit designation	Art. No.	Price DM
Parabolic antenna, 1.1 m diameter,	3/1979	**Set of 12 segments**	0098	**180.00**
12 segments to be screwed or riveted together,		**Riveting machine + rivets**	0105	**93.00**
3 plastic supports for radiator,		**1.7 GHz Cavity radiator kit**	0091	**90.00**
mast-mounting parts with elevation mechanism		**3 radiator supports**	0106	**29.00**
		Mast-mounting parts	0107	**85.00**
Low-noise amplifier for 1.7 GHz	3/1984	**DJ6PI 012**	6867	**175.00**
METEOSAT Converter, consisting of	4/1981	**DJ1JZ 003**	6705	**189.00**
two modules – Output first IF = 137.5 MHz)	1/1982	**DJ1JZ 004**	6714	**185.00**
VHF Receiver, frequency range:	4/1979	**DC3NT 003**	6141	**225.00**
136 – 138 MHz,	1/1980	**DC3NT 004**	6145	**80.00**
Output: 2400 Hz sub-carrier				
Digital scan converter	4/1982	**YU3UMV 001**	6736	**675.00**
(256 × 256 × 6 Bit)	1/1983	**YU3UMV 002**		
PAL-Color module with VHF modulator	2/1983	**YU3UMV 003**	6739	**150.00**

B) Aligned ready-to-operate PCB-modules and equipment

Cavity radiator for above parabolic antenna	0092	**150,–**
VHF receiver for 136 – 138 MHz, DC3NT 003	6731	**395,–**
Oscillator for VHF receiver, DC3NT 004	6732	**168,–**
Digital scan converter (256 × 256 × 6 Bit) YU3UMV 001 + 002	6734	**1150,–**
PAL-Color module with VHF oscillator YU3UMV 003	6738	**285,–**

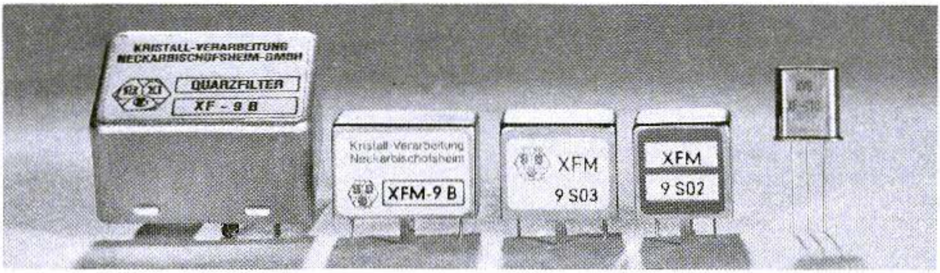
C) A complete system, ready-to-operate in cabinets

Parabolic antenna, 12 segments, riveting machine and rivets, cavity radiator, supports	0108	**510,–**
METEOSAT converter with GaAs-FET preamplifier and mixer, 2 channels, in casing	3026	**692,–**
Antenna for orbiting satellites, DJØBQ-137 (VHF COMMUNICATIONS 4/1981)	0101	**198,–**
Power combiner for above, AT-137	0306	**98,–**
6-channel VHF receiver in cabinet, programmed for: 137,130/137,300/137,400/137,500/137,620/137,850 MHz	3300	**1298,–**
Digital scan converter, 256 × 256 × 6 Bit, with control electronic **and PAL-Color module**/VHF oscillator in cabinet	6735	**1980,–**
Video monitor, black/white, with 31 cm C.R.T.	3301	**550,–**

All 10 editions of VHF COMMUNICATIONS containing information on weather satellite reception	6742	**49,–**
Dissemination Schedule of METEOSAT, incl. surface mail	005D	**3,–**
Audio Compact Cassette with 2 × 30 minutes of selected subcarrier recordings of METEOSAT and NOAA, resp.	6740	**25,80**

UKW**berichte** Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133-855. For Representatives see cover page 2

OUR GREATEST now with reduced dimensions !



Case: 1 15 14 13 17

DISCRETE CRYSTAL FILTER	Appli-cation	MONOLITHIC EQUIVALENT					
		with impedance transformation			without impedance transformation		
		Type	Termination	Case	Type	Termination	Case
XF-9A	SSB	**XFM-9A**	500 Ω ∥ 30 pF	15	**XFM-9S02**	1.8 kΩ ∥ 3 pF	13
XF-9B	SSB	**XFM-9B**	500 Ω ∥ 30 pF	15	**XFM-9S03**	1.8 kΩ ∥ 3 pF	14
XF-9C	AM	**XFM-9C**	500 Ω ∥ 30 pF	15	**XFM-9S04**	2.7 kΩ ∥ 2 pF	14
XF-9D	AM	**XFM-9D**	500 Ω ∥ 30 pF	15	**XFM-9S01**	3.3 kΩ ∥ 2 pF	14
XF-9E	FM	**XFM-9E**	1.2 kΩ ∥ 30 pF	15	**XFM-9S05**	8.2 kΩ ∥ 0 pF	14
XF-9B01	LSB	**XFM-9B01**	500 Ω ∥ 30 pF	15	**XFM-9S06**	1.8 kΩ ∥ 3 pF	14
XF-9B02	USB	**XFM-9B02**	500 Ω ∥ 30 pF	15	**XFM-9S07**	1.8 kΩ ∥ 3 pF	14
XF-9B10*	SSB	—	—	–	**XFM-9S08**	1.8 kΩ ∥ 3 pF	15

* **New: 10-Pole SSB-filter, shape factor 60 dB : 6 dB 1.5**

Dual (monolithic twopole) **XF-910;** Bandwidth 15 kHz, R_T = 6 kΩ, Case 17

Matched dual pair (four pole) **XF-920;** Bandwidth 15 kHz, R_T = 6 kΩ, Case 2 x 17

DISCRIMINATOR DUALS (see VHF COMMUNICATIONS 1/1979, page 45)

for NBFM **XF-909** Peak separation 28 kHz

for FSK/RTTY **XF-919** Peak separation 2 kHz

CW-Filters – still in discrete technology:

Type	6 dB Bandwidth	Crystals	Shape-Factor	Termination	Case
XF-9M	500 Hz	4	60 dB : 6 dB 4.4	500 Ω ∥ 30 pF	2
XF-9NB	500 Hz	8	60 dB : 6 dB 2.2	500 Ω ∥ 30 pF	1
XF-9P*	250 Hz	8	60 dB : 6 dB 2.2	500 Ω ∥ 30 pF	1

* **New !**

KRISTALLVERARBEITUNG NECKARBISCHOFSHEIM GMBH

D-6924 Neckarbischofsheim · Postfach 61 · Tel. 07263/6301