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for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

VHF communications

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European VLF time and frequency standard transmissions from DCF 77 for control of coherent transmissions

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Contents

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Bernhard Kokot
Dieter Schwarzenau

A Home-made RF-Millivoltmeter

The authors have designed a RF-millivoltmeter for measuring low to medium RF-voltages. The frequency range, input impedance, and accuracy should be comparable to those of commercially manufactured RF-millivoltmeters.

Since simple rectifier circuits, and rectifier diodes provided with a bias current, are not able to fulfill these demands, the principle of compensation measurement was used.
Fig. 1 shows a photograph of the author's prototype.

1. BLOCK DIAGRAM

The operation of the unit can be seen in the block diagram given in **Figure 2.**

The actual compensation takes place in the probe. It is equipped with a differential rectifier, which also receives an internally generated, low-frequency alternating voltage. The output DC-voltage is proportional to the diffe-

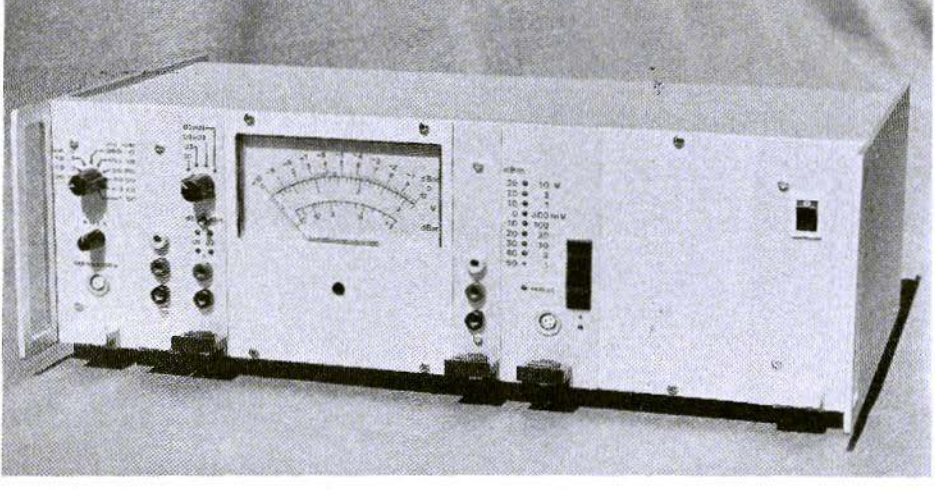


Fig. 1: A photograph of the author's prototype using two AF-modules and a logarithmic amplifier

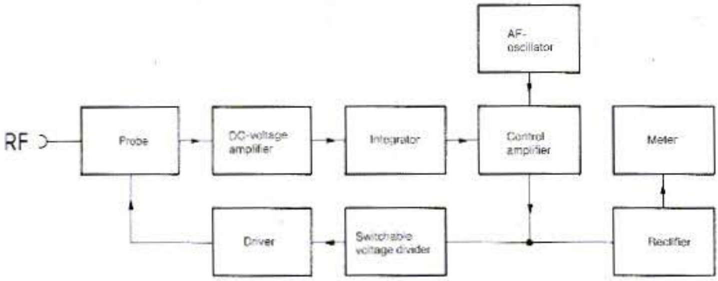


Fig. 2: Block diagram of the RF-millivoltmeter

rence of the peak values of the two alternating voltages.

It is amplified and fed to an integrator, which controls a control amplifier. The control amplifier varies the amplitude of the internally generated AC-voltage. This is fed via a voltage divider to the probe.

In the stationary state, the peak values of the voltage to be measured and the controlled voltage are identical, since the output voltage of the differential rectifier is then zero, and the signal at the output of the integrator will not change.

The internally generated voltage is, however, higher to the value of the division factor of the switchable voltage divider and is constant with respect to frequency. Furthermore, the output impedance of the control amplifier is sufficiently low so that the rectifier connected here can be constructed relatively simply and only needs to be optimized with respect to linearity.

The DC-voltage at the output of this rectifier is then proportional to the peak value of the voltage to be measured. The meter is connected to this position.

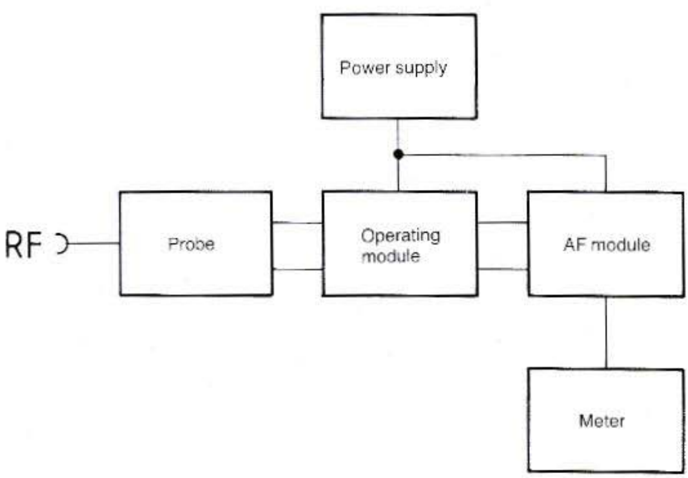


Fig. 3: Individual modules of the millivoltmeter

2. THE INDIVIDUAL MODULES OF THE UNIT

The basic version of the RF-millivoltmeter comprises five modules. This can be seen in **Figure 3.** With the exception of the probe and the meter, all modules are constructed on PC-boards of European standard size.

If a 19 inch cabinet is used, it is possible for the unit to be extended easily. A logarithmic amplifier is available as a sixth module. It can be connected between the meter and two completely constructed basic versions of the RF-millivoltmeter. This allows gain and attenuation values to be measured directly in dB.

Furthermore, the logarithmic amplifier is required if the voltage to be measured is to be indicated lineary in dBm.

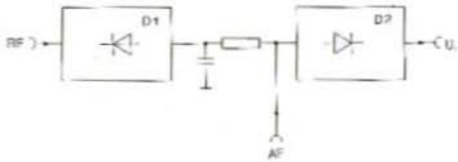


Fig. 4: Principle of the test probes

3. CIRCUIT DESCRIPTION

3.1. Probe

The probe is equipped with two rectifiers that are coupled together. **Figure 4** shows the principle of this circuit. The rectifier D 1 is fed with the AC-voltage to be measured, and charges the capacitor. This DC-voltage is superim-

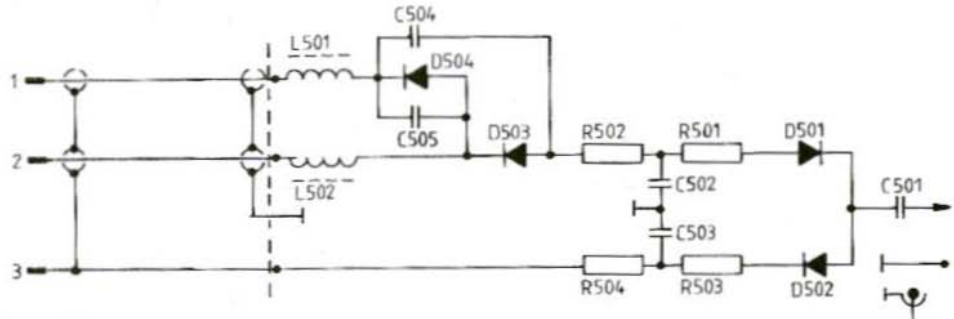


Fig. 5a: Circuit diagram of the test probe DL 0 HV 005

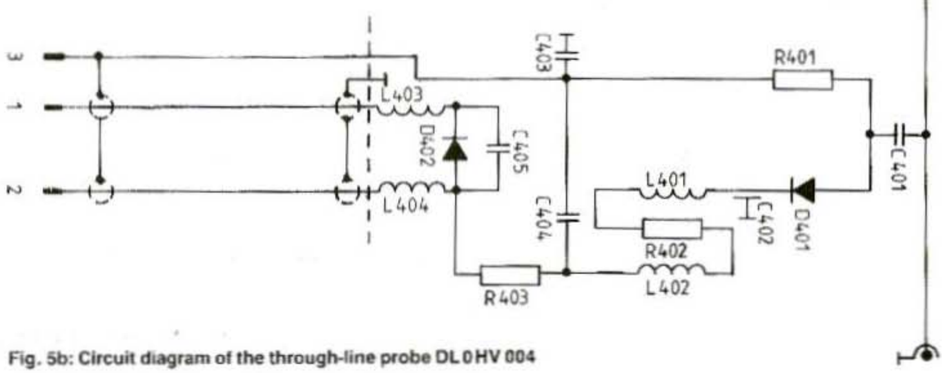


Fig. 5b: Circuit diagram of the through-line probe DL 0 HV 004

posed with the additional AF-voltage and is fed to rectifier D 2. If the peak voltage of the AF-signal is just as great as the DC-voltage, the output voltage of D 2 will be zero.

In all other cases, an offset voltage will result. The operation is therefore independent of the magnitude of the two AC-voltages if the characteristics of the two rectifiers are identical. The shape of the characteristic is primarily unimportant. However, if very low RF-voltages are to be measured, attention should be paid that the rectified voltages are greater than the noise.

Special diodes designed for this application (zero-bias and point-contact diodes) are, however, very expensive. For this reason, experiments were made with conventional germanium diodes. The best results were obtained using cheap germanium point contact diodes, type AA 119. It is possible with these diodes to increase the measuring range of the RF-millivoltmeter down to approximately 1 mV.

In order to avoid thermal voltages from the object to be measured, it is necessary for the rectifier D 1 to be capacitively coupled to the object to be measured. Furthermore, attention should be paid in order to obtain a good, thermal stability so that the characteristic curves do not shift with respect to another due to the difference in temperature.

The complete circuit diagram of the probe is given in **Figure 5a. Figure 5b** shows the circuit diagram of a through-line probe. The construction of both probes is to be described in Section 4.1.

The upper frequency limit of the probes is mainly determined by the capacitance and inductance of the RF-diodes, or their connections. With the described construction and germanium diodes of type AA 119, the cutoff frequency is approx. 1500 MHz (−3 dB).

This could be increased by using special point contact diodes, or zero-bias diodes.

3.2. Control Module

The control module consists of a changeover switch for up to 12 measuring ranges, an input DC-voltage amplifier, and an impedance converter for the AF-supply of the probe. These are available in two versions:

In the case of the first version, the switching is made mechanically using a switch. This switch has three wafers with 12 changeover contacts. Since such switches are relatively expensive, an additional version was designed that uses digitally controlled FETs. Both circuits are to be described.

3.2.1. Control Module with Mechanical Swithing

Figure 6 shows the circuit diagram of the control module with mechanical switching. One will see a DC-voltage amplifier with five digital gain values in the upper part of the circuit diagram. Since the signal coming from the probe can be very low, a chopper amplifier type ICL 7650 was used for this. This integrated circuit behaves exactly as a normal operational amplifier. Actually, it is two amplifiers between which a continuous switching is made.

The amplifier that is not in operation, is always automatically aligned. The offset voltage of the integrated circuit is therefore kept very low.

The field effect transistors T 201 and T 202 serve as over-voltage protection and have a low leakage current. The zero-alignment can be carried out with the aid of P 201. The operational amplifier I 202 serves as impedance converter for the AF-voltage divider, which is switched with the aid of wafer 'c' of the switch. This impedance converter supplies the probe via a transformer.

The transformer is necessary in order to ensure that a sufficiently high AF-amplitude is available for the large measuring range. With wafer 'a' of the switch, a $\frac{1}{\sqrt{10}}$ (10 dB) switching is carried out. This works directly into the AF-module.

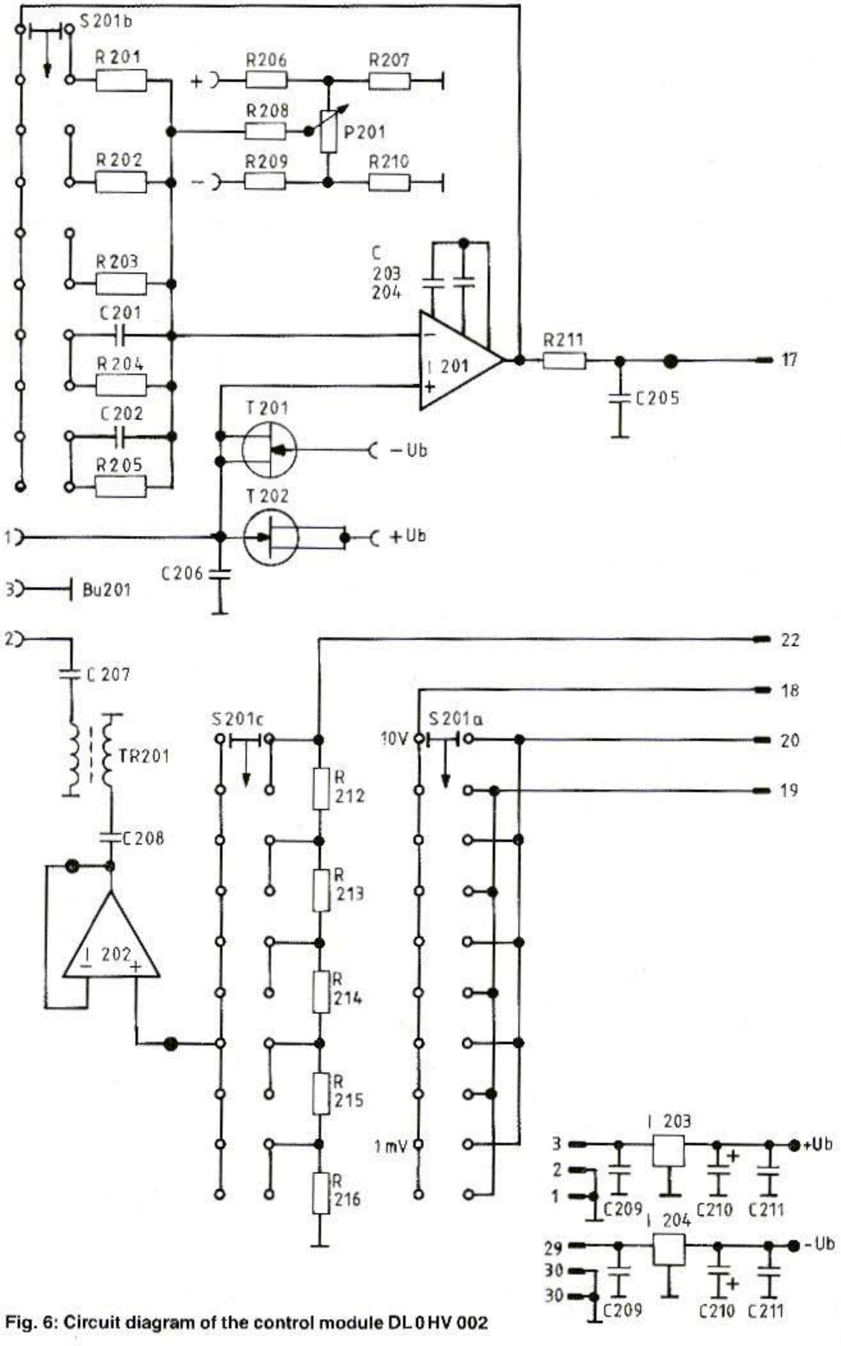


Fig. 6: Circuit diagram of the control module DL0HV 002

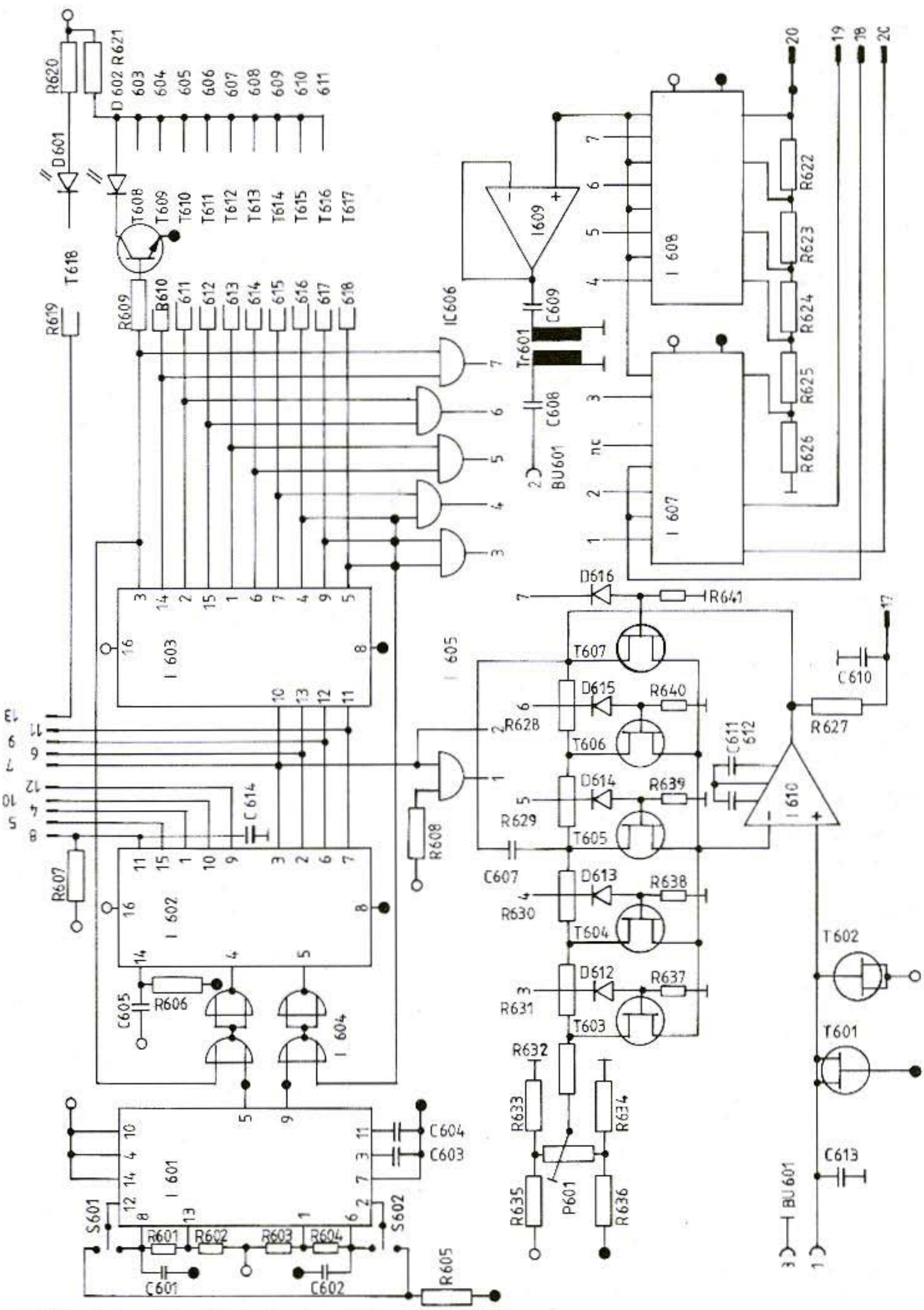


Fig. 7: Control module with electronic switching, DL 0 HV 006

3.2.2. Control Module with Electronic Switching

As can be seen in the circuit diagram given in **Figure 7,** all switching contacts of the mechanical switching are now made with the aid of FETs or MOSFET-switches (I 607, I 608).

These are controlled with the aid of a digital circuit, which is mainly formed by I 602, a binary 4-bit forward-backward counter. The counting impulses drives I 601 that is connected as a dual start-stop-oscillator.

In order to obtain an individual control line for each measuring range, the BCD-signal from I 602 is decoded by I 603, a BCD to decimal decoder. The measuring range is indicated by LEDs D 602-D 611 which are driven by transistors T 608-T 617.

In order to avoid a jumping from the highest to the lowest measuring range, and vice versa, the actual counting input of I 602 is blocked with the aid of NOR-gate (I 604) on reaching the maximum measuring ranges.

Since the lowest measuring range can amount to 3 mV, 1 mV, or 0.3 mV, according to the selection of the diodes in the probe, the possibility was provided of connecting the blocking line for the downward counting to three different outputs of I 603.

Capacitor C 605 together with R 606 provide a reset signal for I 602 after switching on. It is then automatically switched to the highest measuring range.

In addition to the described operation, the integrated counting circuit also allows the possibility to load a counter state directly. The required lines are provided on the connection strip of the module (connections 4-13). This makes it possible to remote-control the unit, or for it to be used together with a computer interface. The remote-control mode is indicated on the control module with the aid of the LED D 601.

CMOS-ICs have been selected for all integrated digital circuits. This deletes the problem of matching the levels to the field effect transistors, and of an additional voltage stabilization. The following should be noted with respect to the operating voltage: Various manufacturers give a maximum operating voltage of 15 V for CMOS-circuits. Philips (Valvo) allow a maximum voltage of 18 V for their circuits! However, no problems were encountered using circuits from other manufacturers.

3.3. AF-Module

The operation of the AF-module can be seen in **Figure 8.**

The DC-voltage signal from the control module is fed to the converting input of I 101. This operational amplifier is switched as an integrator. Its output must be able to be driven down to the negative operating voltage, since the amplitude of the AF cannot be controlled down to zero. For this reason, an operational amplifier type CA 3140 was used here.

In the case of I 102, a so-called OTA (Operational Transconductance Amplifier) is used. This is an operational amplifier whose gain can be adjusted by providing a current in the control input. This control current supplied by the output of the integrator via R 102. It controls the amplitude of the AF, which is generated in an RC-oscillator comprising I 103.

The AF-voltage proportional to the RF-voltage to be measured, is present at the output of the OTA under lock-in conditions. This is fed via C 112 to a very linear rectifier, which is followed by a RC-lowpass filter (R 132/C 116), and an impedance converter. A suitable meter can then be connected to test point TPA.

The rectifier circuit seems to be very extensive at first, however, it offers a great advantage over other circuits in that its operation is independent of the impedance of the meter, and a pure DC-voltage is available at the output. This means that one can also carry out measurements using more accurate, external meters, if a suitable connection is provided.

The controlled AF-signal is fed via C 105 to the amplifier stage equipped with I 104. Its gain can be switched, using the range switch on the control module, from 1 to $\sqrt{10}$ (10 dB switching). This allows one to obtain a suitable range switching.

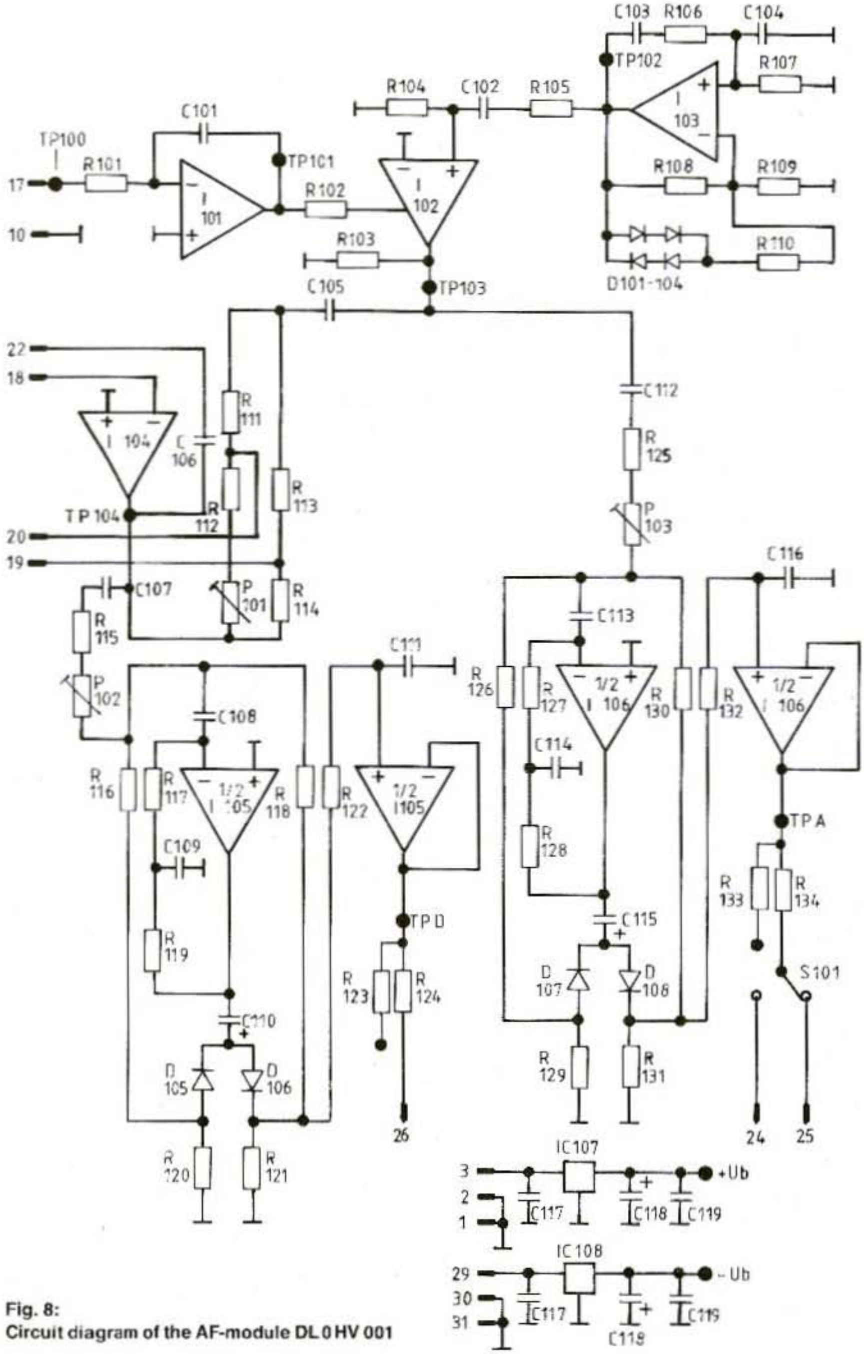


Fig. 8:
Circuit diagram of the AF-module DL 0 HV 001

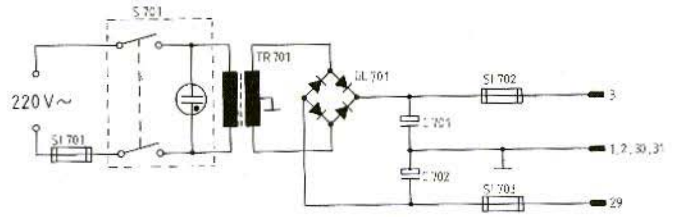


Fig. 9: Circuit diagram of the power supply DL 0 HV 007

A second rectifier is connected to the output of this amplifier stage that is identical to the previously mentioned one. This is provided for the connection of a digital voltmeter. The rectifier equipped with I 106 always provides the same voltage for the full-scale value in each range, independent of the selected gain of I 104, and therefore only two different scales with full-scale values of 1 and 3.16 are required for the analog meter.

On the other hand, a DC-voltage of the corresponding value is required for the connection of a digital voltmeter.

The AF-signal is fed from I 104 via C 106 to the voltage divider and driver stage of the control module.

3.4. Power supply

All described modules are equipped with their own voltage stabilizer circuit. This allows any interference on the supply lines to be effectively suppressed.

In addition to this, a relatively simple circuit can be used as power supply **(Figure 9)**, and no problems are encountered with heat dissipation.

The power supply module only consists of a transformer, rectifier, as well as the associated smoothing capacitors for the positive and negative voltage.

Two fuses are provided for overload protection.

3.5. Meter

The indicator module comprises only a meter, whose full-scale deflection is adjusted to 1 V with the aid of a dropper resistor. It should be provided with two scales having a full-scale value of 10 and $\sqrt{10} = 3.16$.

If the measured voltage is also to be indicated in dBm, this will require a further scale.

If a logarithmic amplifier is available, the indication of dB and dBm will be linear. To read off the voltage in dBm, it is necessary for the upper scale to be additionally provided with a marking from −10 to 0 in the opposite direction. These scales can be seen in **Figure 10.**

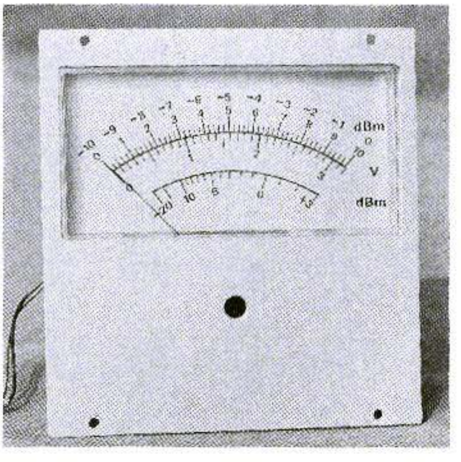


Fig. 10: Meter scales of the author's prototype

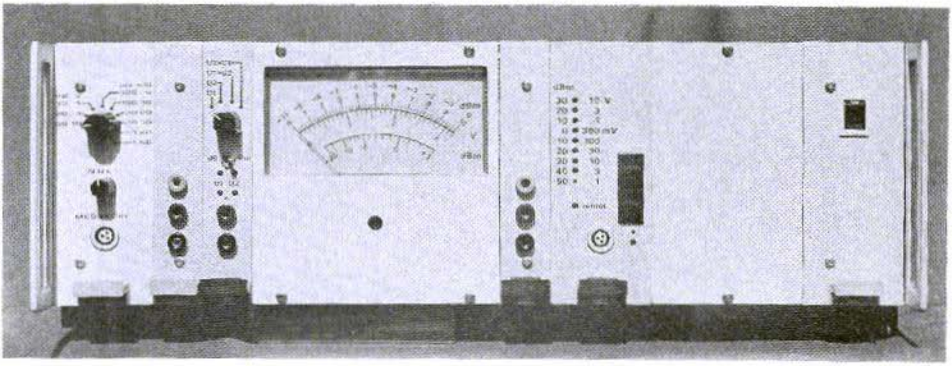


Fig. 11: The completed RF-millivoltmeter in a 19 inch cabinet together with one dummy panel

4. CONSTRUCTION

With the exception of the probes and the indicator module, all other boards of the measuring system are built up using double-coated PC-boards of the standard European format.

All modules are provided with a 31-pin connector strip. This allows the unit to be accommodated in a 19 inch cabinet.

Enough room is provided in the cabinet for extending the instruments. The selected construction allows all modules to be easily accessible. As in the case of the author's prototype (see **Figure 11**), all non-used portions of the cabinet can be covered with a dummy panel.

4.1. Construction of the Probes

Figure 12 shows the two through-line probes, as well as a test probe.

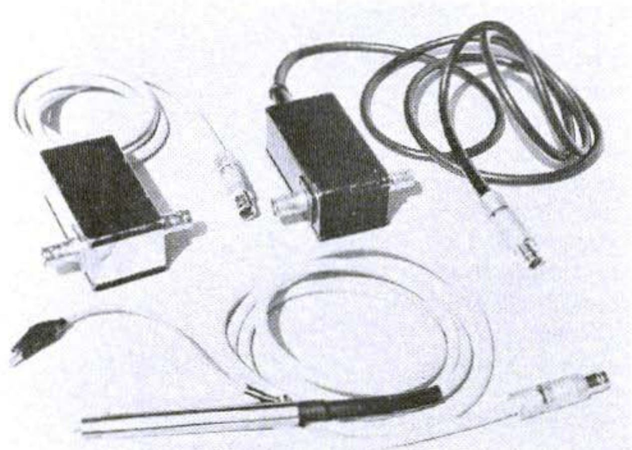


Fig. 12:
Test probes for the RF-millivoltmeter

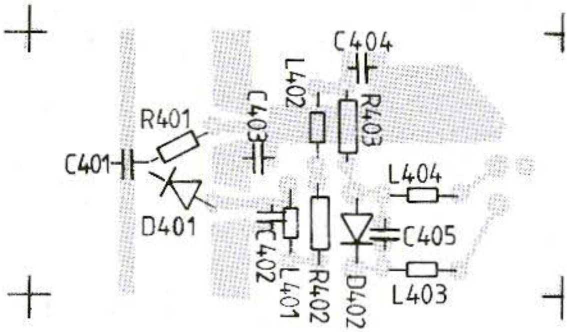


Fig. 13:
Component location plan of PC-board DL0HV 004 (through-line probe)

4.1.1. Through-Line Probe

The through-line probe is accommodated in a metal box, whose dimensions are 37 x 74 x 30 mm. One BNC-connector is provided on each of the wide sides. The cable is fed in to the bottom of the box as shown in Figure 12. The components of the probe are accommodated on a PC-board of 71.5 mm x 34.7 mm; this board is double-coated, and one side is provided as a continuous ground surface **(Figure 13.)** This board is designated DL0HV 004.

Attention should be paid when constructing this module that the chip capacitor C 401 is soldered into place directly on the board. The connections of the BNC-connectors are also directly soldered to the conductor lane.

A screening panel should be provided behind components R 401 and D 401, in order to isolate the AF-side from the RF-circuit. The installation of the other components should be made according to the component location plan given in Figure 13.

Capacitors C 402 to C 404 should, if possible, also be chip-capacitors. The interconnection cable should have three cores two of which must be screened. The components for the through-line probe are given in **Table 1.**

4.1.2. Test Probe

A 9 mm x 68 mm narrow board was developed for accommodating the components of the

Quantity	Designation	Component
2	D 401, D 402	Diode AA 119 (BAT 16, GD 741)
1	R 402	Resistor 47 Ω
1	R 401	Resistor 4.7 kΩ
1	R 403	Resistor 47 kΩ
1	C 405	Capacitor 100 pF / ceramic
4	C 401 – C 404	Capacitor 1 nF / ceramic
2	L 401, L 402	Fixed inductance 1 μH
2	L 403, L 404	Fixed inductance 10 μH
1	–	Metal box 30 x 37 x 74 mm
2	–	BNC connectors
–	–	Cable 4-core, individually screened
1	–	Double-coated board DLØHV 004, 34.7 x 71.5 mm

Table 1:
List of components for the through-line probe

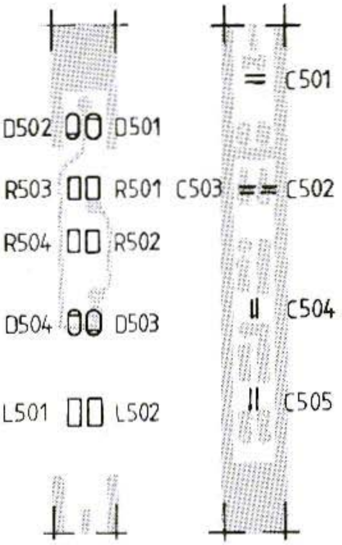


Fig. 14:
PC-board DL0HV 005 for the test probe

test probe, (see **Figure 14**). It is double-coated and has been designated DL0HV 005. It can be mounted in any suitable metal box. Attention should be paid that the interconnection of the test probe to the first rectifier is kept as short and as low-capacitance as possible.

In the case of the author's prototype, the completed PC-board was built into a brass tube having an inner diameter of 9 mm. Special attention was paid to obtain a simple construction.

As was already mentioned in the operational description, it is very important with respect to the measuring accuracy that the characteristic curves of the two diodes used, coincide as well as possible. The selected diodes should therefore be from the same production run. If this is not possible, it is necessary for the diodes to be selected – this is also valid for the through-line probe!

Of course, it is possible for diodes specially developed for RF-measuring equipment to be used instead of the diodes type AA 119. Such diodes are the BAT 16 or GD 741 manufactured by Siemens. These are suitable for higher frequencies, but also have a far higher price.

Table 2 contains a list of the components for this test probe.

4.2. Construction of the other Modules

As has been previously mentioned, all PC-board modules with the exception of the probe boards, are built in standard European size, and equipped with their own 31-pin connector strip. Double-coated PC-boards are used for the AF-module, control module, and logarithmic amplifier. A single-coated PC-board was designed for the power supply. Each board is provided with its own front panel, whose dimensions are 25.4 or 50.8 mm in width.

Quantity	Designation	Component
4	D 501 – D 504	Diode AA 119 (BAT 16, GD 741)
2	R 501, R 503	Resistor 180 Ω
2	R 502, R 504	Resistor 10 kΩ
1	C 505	Chip capacitor 100 pF
3	C 501 – C 503	Chip capacitor 1 nF
1	C 504	Chip capacitor 22 nF
2	L 501, L 502	Fixed inductance 10 μH
1	–	Double-coated board DLØHV 005, 9 x 68 mm
–	–	Cable 4-core, individually screened
1	–	Casing

Table 2:
List of components for the test probe

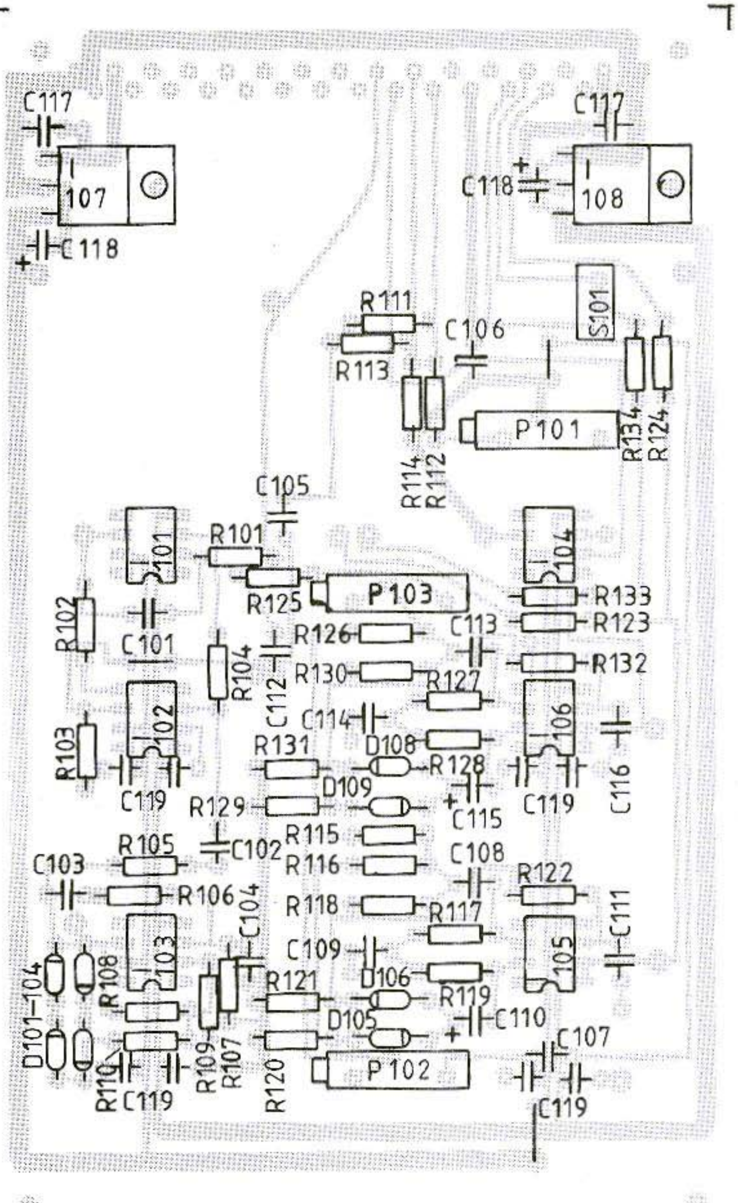


Fig. 15: Component locations of the AF-module DL0HV 001

4.2.1. AF-Module

Figure 15 shows the component locations on the double-coated PC-board DL0HV 001. The component side of this board is provided with a continuous ground surface, and this is drilled free for those connections that are not connected to ground. **Figure 16** shows a photograph of a completed board. The required components are given in **Table 3.** The rectifier circuit comprising I 105 need only be provided when an output for a digital voltmeter is required.

The AF-module is provided with a 1 inch-front panel, whose dimensions are given in **Figure 17.** It is connected to the PC-board with the aid of a bracket. It is not necessary to provide any markings on the front panel.

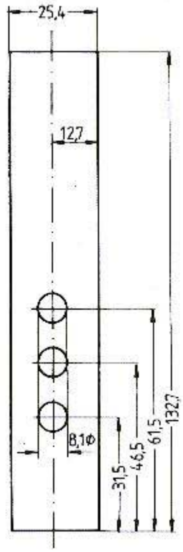


Fig. 17: Front panel of the AF-module

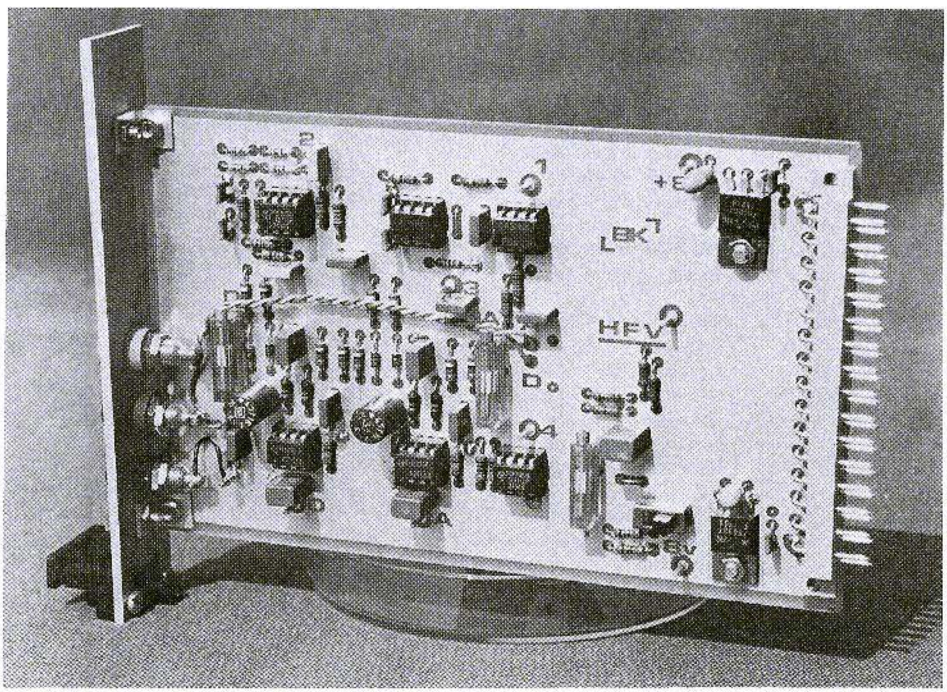


Fig. 16: Photograph of a completed AF-module

Quantity	Designation	Component
4	D 101 – D 104	Diode 1N 4148
4	D 105 – D 108	Diode HSCH 1001
1	I 101	CA 3140 E
1	I 102	CA 3080 E
2	I 103, I 104	TL 071 CP
2	I 105, I 106	TL 072 CP
1	I 107	MC 7808 CT
1	I 108	MC 7908 CT
4	R 123, R 124 R 133, R 134	Resistor 100 Ω
1	R 104	Resistor 150 Ω
1	R 103	Resistor 2.7 kΩ
2	R 106, R 107	Resistor 4.7 kΩ
1	R 102	Resistor 5.6 kΩ
5	R 105, R 109 R 111, R 113 R 114	Resistor 10 kΩ
1	R 125	Resistor 12 kΩ
1	R 108	Resistor 22 kΩ
1	R 112	Resistor 30 kΩ
9	R 115, R 116 R 118, R 120 R 121, R 126 R 129 – R 131	Resistor 39 kΩ
1	R 110	Resistor 82 kΩ
1	R 101	Resistor 100 kΩ
2	R 122, R 132	Resistor 470 kΩ
4	R 117, R 119 R 127, R 128	Resistor 1 MΩ
2	C 103, C 104	Capacitor 4.7 nF / FKC
2	C 108, C 113	Capacitor 10 nF / MKS
4	C 102, C 105 C 107, C 112	Capacitor 100 nF / MKS
10	C 117, C 119	Capacitor 100 nF / ceramic
4	C 101, C 109 C 114	Capacitor 220 nF / MKS
2	C 111, C 116	Capacitor 680 nF / MKS
1	C 106	Capacitor 1 µF / MKS
2	C 118	Capacitor 10 µF, 16 V / tantalum
2	C 110, C 115	Capacitor 47 µF, 16 V / tantalum
2	P 101, P 102	Potentiometer 10 kΩ (call 0613/300)
1	P 103	Potentiometer 4.7 kΩ (call 0613/300)
1	S 101	MFP 120 Knitter Switch
1	–	Connector strip 31-pin, DIN 41617
1	–	Front panel 1 inch Vero F1V1F
3	–	Telephone jacks insulated
–	–	Solder pins
–	–	Screws and nuts
2	–	Mica disk for TO 220
1	–	PC-board DLØHV 001, 100 x 160 mm, double-coated

**Table 3:
List of components
for the AF-module**

To be concluded in ed. 3/1984

Terry Bittan, DJ0BQ/G3JVQ

The GOES Series of Geostationary Weather Satellites

The characteristics and transmissions of the European geostationary weather satellite METEOSAT have been discussed several times in VHF COMMUNICATIONS. Since many of our readers are also or only able to receive transmissions from the US geostationary GOES-East or West, we consider that it is of interest to know more about the operation of this series of satellites, as well as the differences to the European METEOSAT-2.

1. HISTORY OF GEOSTATIONARY WEATHER SATELLITES AND WEFAX TRANSMISSIONS

The history of geostationary satellites and WEFAX transmissions is certainly of interest to our readers and is therefore to be mentioned briefly in this article.

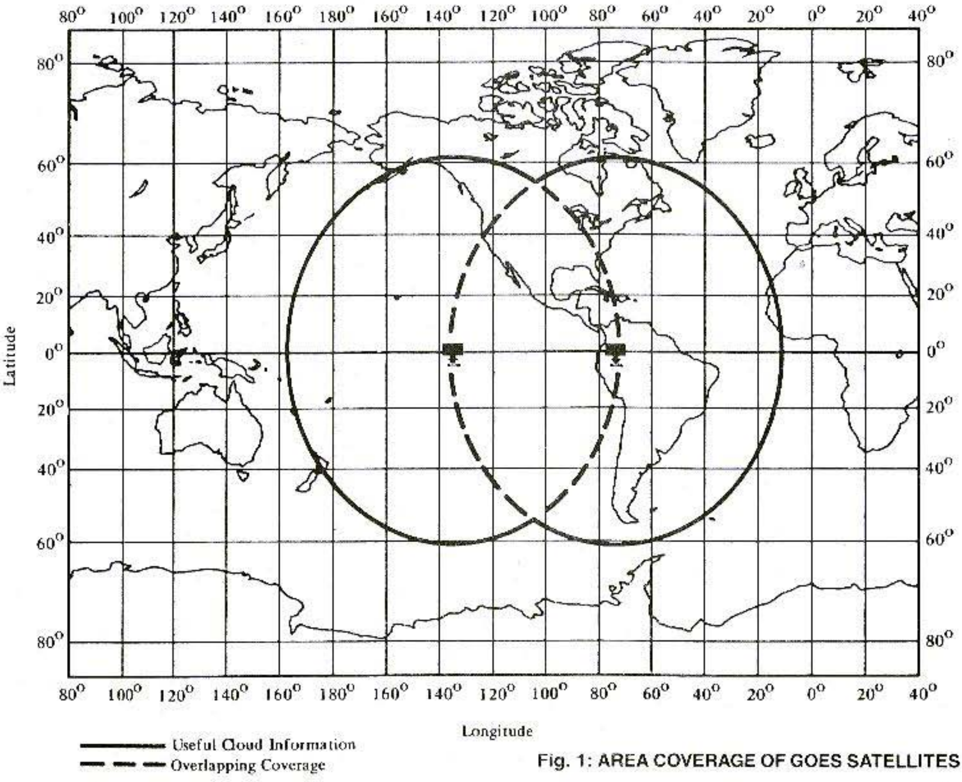


Fig. 1: AREA COVERAGE OF GOES SATELLITES

WEFAX was first introduced as a communications relay experiment on Applications Technology Satellites, (ATS) -1, later included on ATS-3, and subsequently refined for incorporation on the SMS/GOES. SMS/GOES satellites were developed by Philco-Ford Corporation and the National Aeronautics and Space Administration (NASA) for the National Oceanic and Atmospheric Administration (NOAA). This new generation of satellites carried a double nomenclature - SMS, for Synchronous Meteorological Satellite, (NASA's designation for the spacecraft during its years of development); and GOES - (NOAA's designation for NOAA-funded satellites). SMS-1, the first prototype of the GOES series, was launched May 17, 1974. NASA's SMS-2 was launched Feb. 6, 1975. The first NOAA-funded satellite, GOES-1, was launched Oct. 16, 1975. GOES-2 was launched June 16, 1977, and GOES-3 was launched June 16, 1978.

The WEFAX dissemination from the SMS/GOES satellites, which began in 1976, was and still is on a down-link S-band frequency of 1691.0 MHz. In the early days of SMS/GOES, it was assumed that many potential WEFAX users already had a Very High Frequency (VHF) APT receiving station operating in the 135 to 137 MHz region. These stations were used for receipt of ATS WEFAX and polar orbiting satellite APT. Technical standards for the S-band SMS/GOES WEFAX dissemination were developed by NESS, which allowed most of the existing APT ground receiving components to be used to receive SMS/GOES WEFAX. Although the transmission frequency from the spacecraft would be 1691.0 MHz, the 2400 Hz sub-carrier signal characteristics, the type of modulation (AM/FM), and image format were not changed. Therefore, the users' existing recording equipment could continue to be used. In addition, with a relatively simple VHF to S-band conversion kit, the existing receiver systems could also be used.

A brief explanation of the history and purpose of Coordination on Geostationary Meteorological Satellites (CGMS) is in order relative to the international aspects for positioning of the spacecraft in geostationary orbit. CGMS came into being in September 1972 when representatives of the European Space Research Organization, later to become the European Space Agency (ESA), the United States, Japan, and observers of the World Meteorological Organization (WMO) met in Washington, D.C., to discuss questions of compatibility among geostationary meteorological satellites. The meeting identified several areas, particularly for WEFAX dissemination, where both technical and operational coordination would be needed. The Soviet Union, having expressed its intention to develop and operate a geostationary satellite, joined the small group of satellite operators at the 2nd CGMS meeting in 1973. The WMO turned from its observer role at the first CGMS to full participation in the CGMS activities. Thus, since 1973, the WMO, ESA, Japan, the United States, and the USSR have cooperated in developing the technical elements and operational principles for the system of geostationary satellites. India, with planned geostationary satellites, joined the group at the 8th meeting of CGMS in 1978.

It was proposed at the second CGMS that five Geostationary Meteorological Satellites be placed in space. One important result in positioning the five proposed geostationary satellites was to achieve some overlap in the fields of view. This principle was sometimes contradictory to national or regional interests, and the geographical location of data reception stations and ranging stations also had an influence. The resulting positioning compromise was as follows:

METEOSAT (ESA)	0°
GOES-East (USA)	75°W
GOES-West (USA)	135°W
Geostationary Meteorological Satellite (JAPAN)	140°E
Geostationary Operational Meteorological Satellite (USSR)	70°E

In the early days of CGMS, the United States agreed to position GOES-East so as to make it visible from Lannion, France. Thus, in the METEOSAT WEFAX transmission schedule, provisions were made for receipt of WEFAX data from GOES-East to be relayed from Lannion, via METEOSAT, to users in the METEOSAT field of reception. For this reason, WEFAX transmissions from GOES-East have been of first priority. Another important factor for locating GOES-East in view of Lannion was the plan for Lannion to receive stored sounding data from the TIROS-N series satellites and to relay the data to the NOAA computer complex using the GOES-East Data Collection System (DCS). In 1975, the satellite operators agreed to provide WMO with WEFAX schedules. These are still published by WMO, and updating the schedules is a continuing activity.

Geostationary environmental satellites launched by these nations as part of the First GARP Global Experiment (FGGE) also provided WEFAX services. (GARP is an acronym for Global Atmospheric Research Programme). Compatibility exists among the WEFAX communications subsystems on all these spacecraft which permits ships or fixed ground stations the option of selecting WEFAX broadcasts from whichever spacecraft best meets their data requirements. Commonality in ground receiving equipment also exists, but schedules and products vary.

On Oct. 11, 1979, GOES-2 (after an earlier VISSR failure) was reactivated for WEFAX broadcasts and began operation as GOES-Central at 105°W. The position was later changed to 107°W to minimize the possibility of in-space collision of the spacecraft with space debris.

GOES-4 and GOES-5, the sixth and seventh spacecraft in this series, were launched on Sept. 9, 1980 and May 22, 1981, respectively. The GOES-4 and -5 spacecraft, developed by the Hughes Aircraft Corporation, are significantly different from the earlier SMS/GOES spacecraft. On-board is a visible and infrared **s**pin **s**can **r**adiometer (VISSR) **a**tmospheric **s**ounder (VAS). This instrument is a more sophisticated version of the VISSR on-board the earlier GOES spacecraft. The VAS has a new capability, atmospheric temperature sounding, for gathering infrared (IR) radiation data which can be used, with known atmospheric properties, to calculate atmospheric temperature profiles over a selected geographic area. With the positioning of GOES-4 at 135°W as the GOES-West operational satellite, and the positioning of GOES-5 at 75°W as the GOES-East operational satellite, the users receiving the WEFAX broadcasts from either of these satellites were required to change the polarity of their antennas. While the S-band antennas on SMS/GOES spacecraft through GOES-3 had linear horizontal polarization (polarization parallel to the plane of the Equator) the GOES-4 and -5 spacecraft have parabolic antennas which have linear vertical polarization (perpendicular to the plane of the Equator). This difference undoubtedly causes some inconvenience if a user wishes to switch reception of broadcasts from either the GOES-East or -West to the GOES-Central, since this requires a change in the polarity of the user's antenna.

2. THE GOES SERIES OF SATELLITES

As was briefly mentioned in Section 1, the GOES system comprises three geostationary satellites, two of which are used to actively observe meteorological phenomena. These two satellites GOES-East and GOES-West are located over the equator at 75°W and 135°W respectively, as can be seen in **Figure 1**.

The radiometer (VISSR – VAS) of each of these satellites scans the earth over a period of 20 minutes; this commences on the hour and half-hour in the case of GOES-East, and 15 minutes and 45 minutes after the hour for GOES-West. With certain exceptions, WEFAX data is transmitted in 10 min. time slots at half-hourly intervals, and this 24 hours a day. This means that GOES-East WEFAX images begin at approximately 20 min. and 50 min., and GOES-West at 5 min. and 35 min. past the hour. The transmission of WEFAX images is thus staggered between GOES-East and West. The advantage of this is that GOES East and West are not usually transmitting WEFAX at the same time, and it is possible for a single WEFAX station to switch the antenna to each of the satellites so that both can be received.

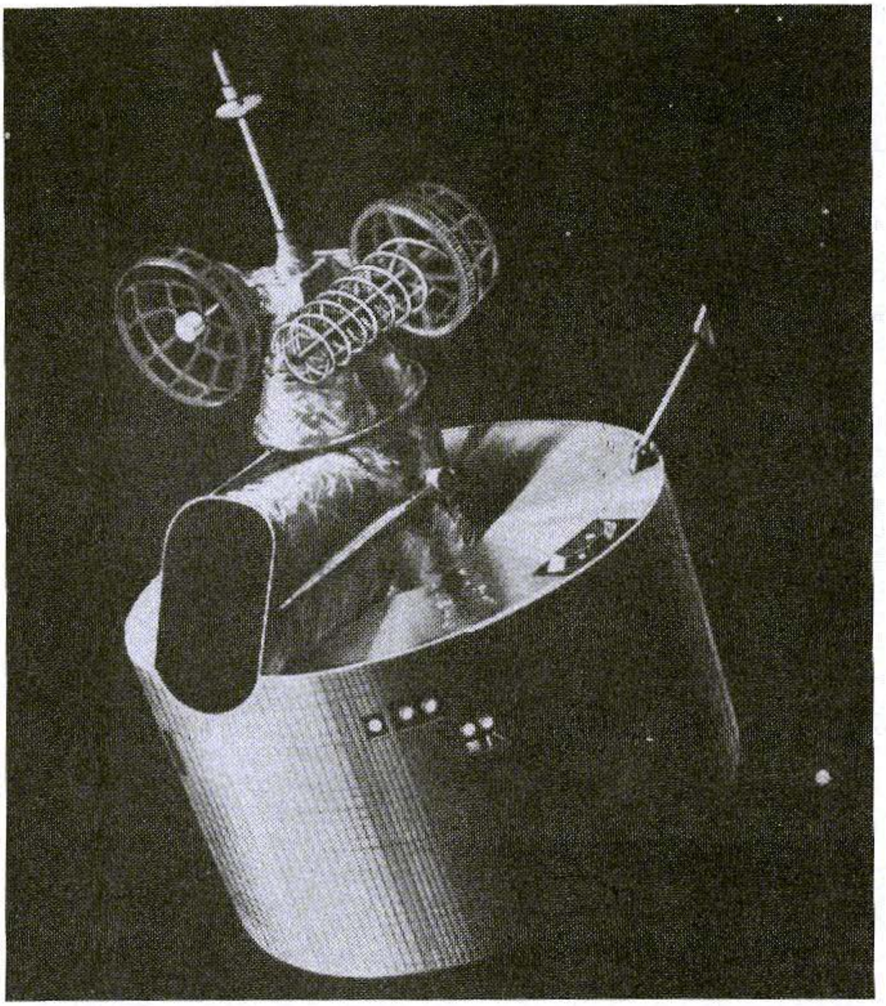


Fig. 2

Figure 2 shows a picture of a GOES geostationary satellite.

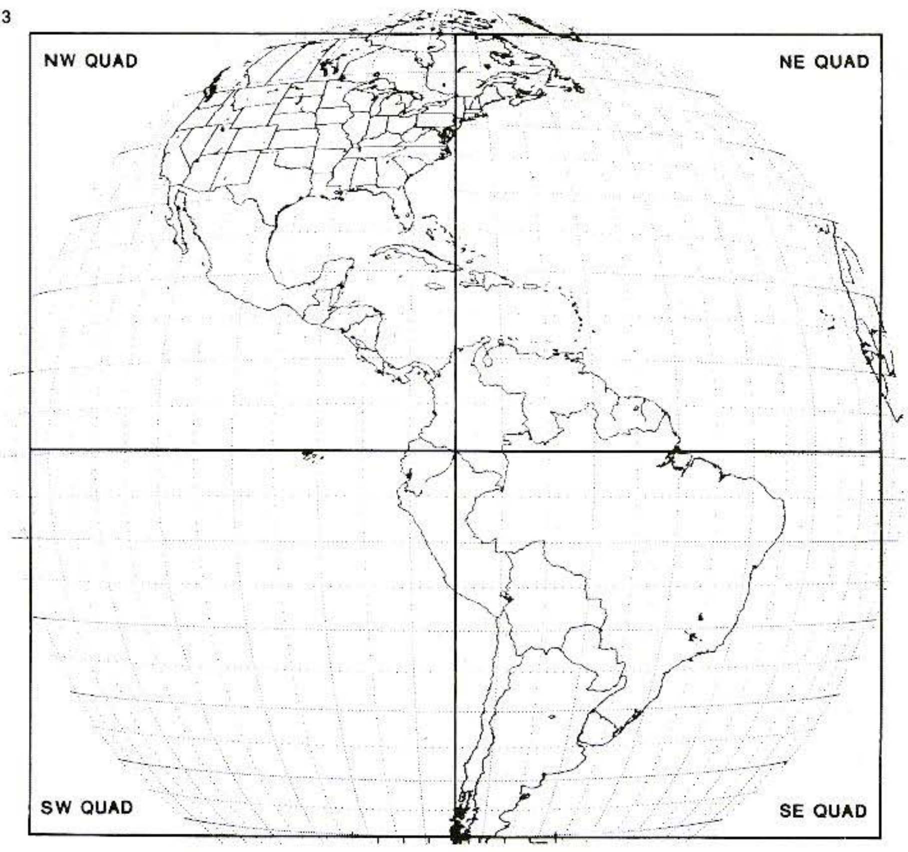
In contrast to this, GOES-central does not obtain any information from its VISSR-VAS system, which is defective, and is only used for relaying images received from GOES-East and West, and also for transmitting weather charts and NOAA imagery. GOES-central operates on the same frequency as the other two satellites but with horizontal instead of vertical linear polarization.

3. WEFAX PRODUCTS

3.1. VISSR – VAS Images

As has been briefly mentioned, the WEFAX products from the GOES series of satellites are quite different to those known from METEOSAT. Firstly, the global view of the satellite is not segmented into such small areas as with METEOSAT (9 main segments),

Fig. 3



but only into the four quads, as shown in **Figure 3**, and into two extra tropical segments as can be seen in **Figure 4** (GOES-E).

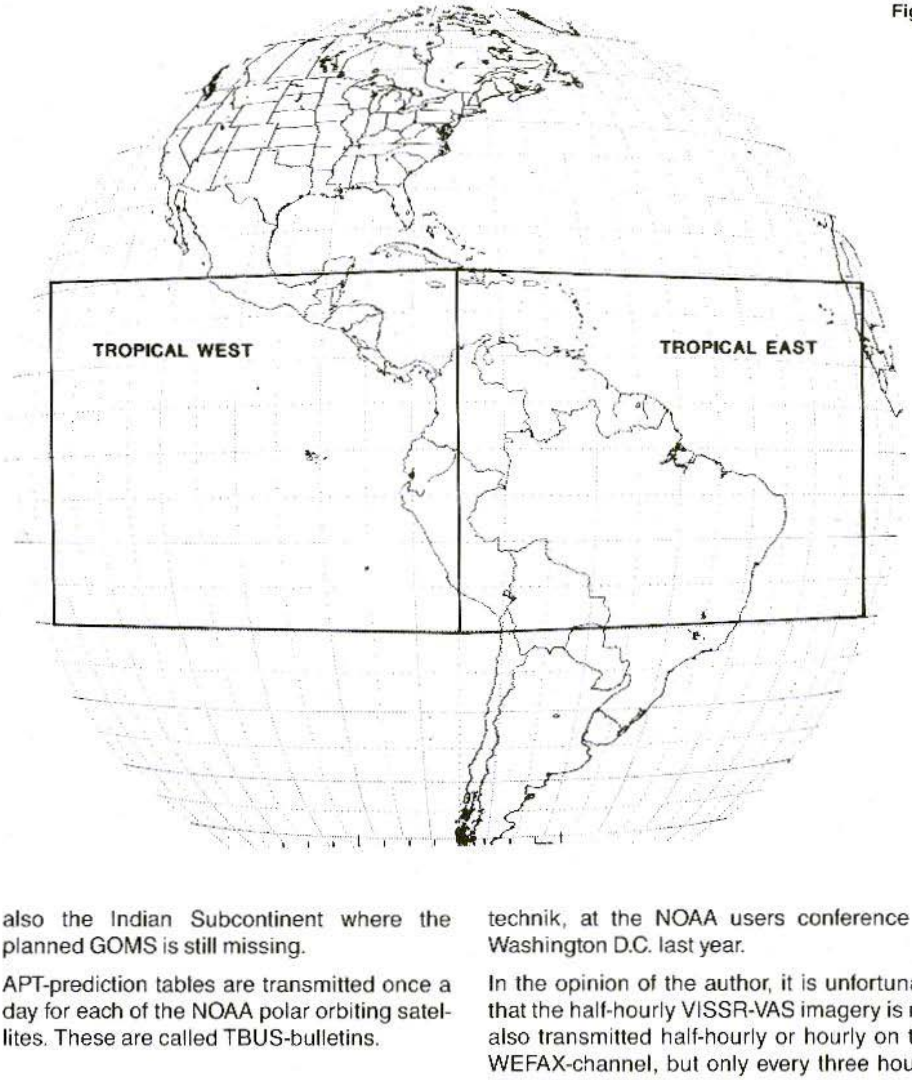
Similar segments are also obtained from GOES-W. However, most of the WEFAX transmission time is not taken up with GOES imagery (which would be much more advantageous in the opinion of the author), but in transmitting weather charts (which could be transmitted elsewhere) and NOAA images (which are very useful).

Although VISSR-VAS images are received every half-hour, most of them are not transmitted as WEFAX product but only about every three hours.

3.2. Polar Imagery

Approximately one third of the imagery transmitted on GOES-WEFAX are derived from the low-resolution visible and IR information recorded on-board the NOAA polar-orbiting satellites. They are reformatted for WEFAX-transmission. These products are of lower resolution than that of directly received APT-transmissions, and are in the order of 7 to 14 km. This imagery is compiled over a number of orbits and therefore does not always represent the actual situation. However, it does allow the weather services to study the worldwide situation, and also to relay information regarding areas not covered by geostationary satellites such as the Arctic, Antarctic, and

Fig. 4



also the Indian Subcontinent where the planned GOMS is still missing.

APT-prediction tables are transmitted once a day for each of the NOAA polar orbiting satellites. These are called TBUS-bulletins.

4. LIMITATIONS OF THE GOES—WEFAX SYSTEM

The author was able to monitor GOES WEFAX transmissions in the USA during several visits, including WEFAX imagery provided by UKW-technik, at the NOAA users conference in Washington D.C. last year.

In the opinion of the author, it is unfortunate that the half-hourly VISSR-VAS imagery is not also transmitted half-hourly or hourly on the WEFAX-channel, but only every three hours. The weather charts could be transmitted via normal telephone lines or satellite links, although this would mean that they would not be available readily to Central and South American users. The NOAA images are a worthwile addition to the WEFAX program, however, due to their limited resolution it is felt that they should be limited to those areas not covered well by geostationary satellites; the author would prefer an actual WEFAX image of a cer-

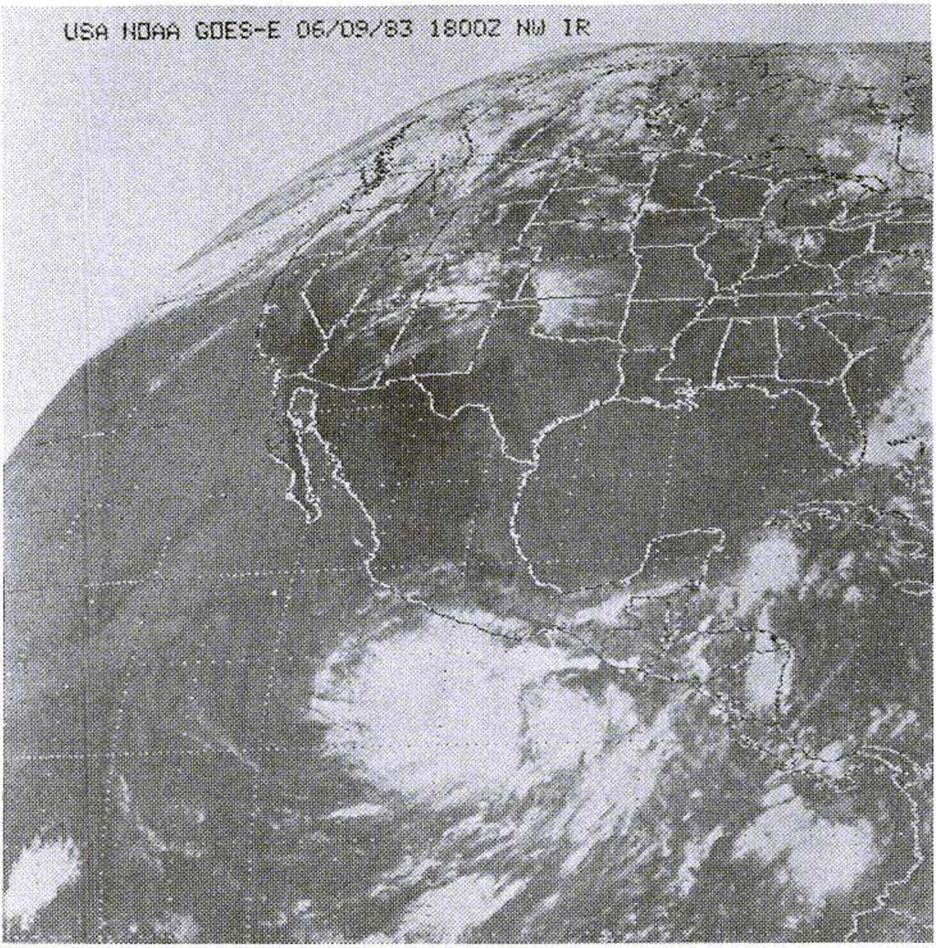


Fig. 5: GOES-East image of North America showing a hurricane to south of Mexico

tain area at full WEFAX resolution, rather than an old image at reduced resolution of the same area derived from a NOAA satellite.

The GOES series of satellites cannot transmit the stretched VISSR images (similar to METEOSAT PDUS transmissions) and WEFAX images at the same time (as is the case with METEOSAT). This means that WEFAX transmissions must be stopped during these periods.

However, there are a number of occasions when half-hourly VISSR-VAS transmissions are not sufficient and when more rapid scans are required, for instance, during severe thunderstorms, hurricanes, tornados, etc. Although WEFAX images are still transmitted, they are at reduced quality, since the WEFAX up-link signal is reduced from 500 W to 125 W, which is a drop of 6 dB. At this low level of WEFAX up-link power, the stretched VISSR-

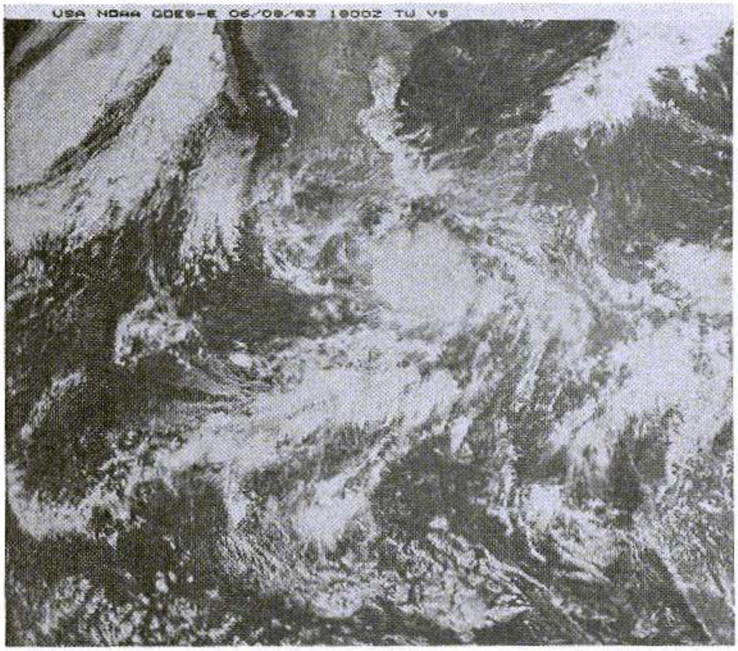


Fig. 6:
GOES-East visible image (Tropical West). The GOES vis. images do not have grids.

VAS will suppress the down-link power by about 12 dB. This type of interference is seen as white horizontal lines evenly spaced throughout the image.

Since the signal virtually disappears when using small antennas, this can cause synchronization problems with satellite reception systems using the 2.4 kHz subcarrier for synchronization, for instance, the YU3UMV-scan converter. However, it is possible to provide the 2.4 kHz accessory oscillator described in edition 1/1984 of VHF COMMUNICATIONS, which will solve problems of lost synchronization.

5. SUMMARY

European readers that are located to the west of the 0°-meridan, should be able to receive images from the GOES-East satellite if they have a good take-off to the south-west.

6. REFERENCES

(1) The GOES Users Guide
NESDIS-NOAA, June 1983

(2) The WEFAX Users Guide
November 1981

Further details on the GOES-program can be obtained from the following address:
Coordinator, Weather Facsimile Services
NESDIS-NOAA
E/SP21, WWB, Room 806
Washington, DC 20233 – USA

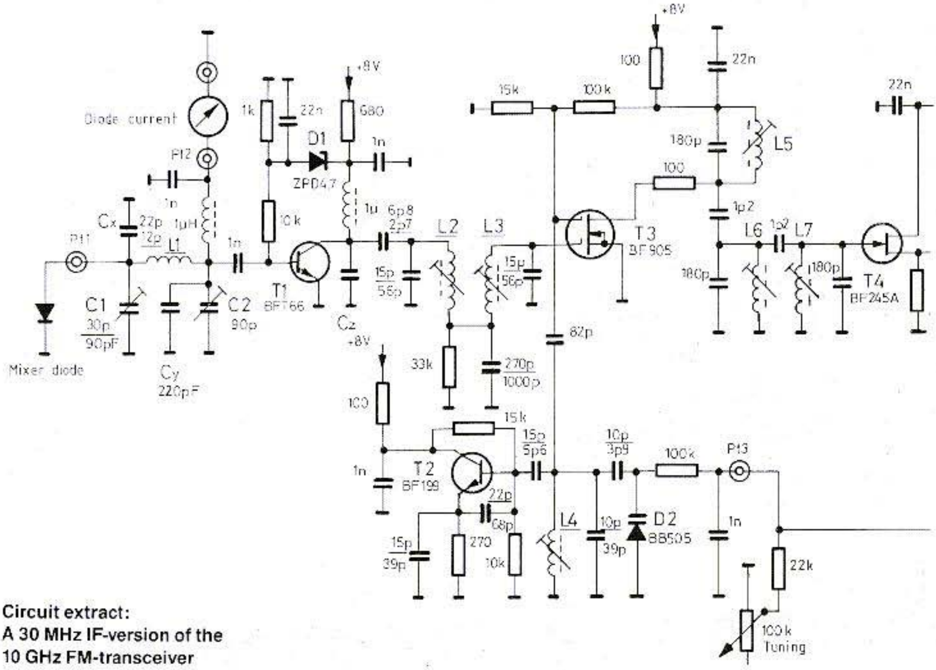
This article has only covered the WEFAX products of the GOES program. However, the program also provides high-resolution digital imagery, and relay facilities for digital telemetry data (DCP). Several excellent handbooks are available to interested users, and the author would like to thank the staff of NESDIS for many interesting discussions at the users conference in Washington DC in 1983.

Jochen Jirmann, DB 1 NV

A 10 GHz FM Transceiver with DSO

Another Version with a 30 MHz Intermediate Frequency.

Edition 1/1984 of VHF COMMUNICATIONS described a 10 GHz FM transceiver for an intermediate frequency of 105 MHz. This frequency is advisable due to the ease of filtering out the image frequency, and the sufficient spacing to the oscillator frequency (phase noise!). However, there are many areas where there is a large activity on 10 GHz using "Gunnplexers", and it may be adivsable to use the same IF of 30 MHz in order to allow duplex operation with such stations. The required modifications are given in the **circuit diagram**.



Circuit extract:
A 30 MHz IF-version of the 10 GHz FM-transceiver

The frequency-dependent components that must be changed have been marked and the values required for an IF = 30 MHz have been added. The four inductances should be wound as follows:

L 1: 9 turns of 0.5 mm dia. enamelled copper wire wound on a 5 mm former, self-supporting

L 2, L 3: 14 turns of 0.3 mm dia. enamelled copper wire wound in the coil set (10 x 10 x 15) supplied for 105 MHz. Core yellow or blue.

L 4: 7 turns of 0.3 mm dia. enamelled copper wire in special coil set 10 x 10 x 15, core blue or yellow.

We would like to take this opportunity of mentioning that the prototype shown on the title photograph of Edition 1/1984 is designed **for vertical polarization.** In the case of horizontal polarization, a 90°-polarization shift should be provided between the transceiver and antenna. If this is to be avoided, the microwave module can be rotated by 90° in the cabinet. Since the micrometer will then protrude to the side, the author decided to rotate the polarization between transceiver and antenna instead.

Günter Sattler, DJ4LB

A Control Circuit with four Time Steps for Transmit-Receive Switching

An UHF-transmitter can generate several millions of oscillations in the time that it takes for a coaxial relay to switch. This represents a great danger, mainly for the transistors in the output stage of the transmitter, and unnecessary wear and tear of the contacts of the coaxial relay itself. For this reason, it is not only useless but also dangerous to switch on a transmitter when it is still not connected to the antenna. If this is to be avoided, one must assure that the individual components of the transmit-receive system are switched on and off in the correct sequence and with a defined time spacing on switching from transmit to receive and vice versa.

A small, simple, and inexpensive module has been designed for this application that provides four switching steps, whose outputs can each be loaded with a current of 250 mA at 12 V. This means that it is possible for all available coaxial relays, as well as most preamplifier stages of receivers and transmitters to be directly connected. Higher currents, such as those of transistor power amplifiers, can be switched more favorably using higher power relays, which can be driven by the switching circuits.

The four switching steps are usually used as follows:

Step 1: The receive preamplifier
Step 2: The antenna relay
Step 3: The operating voltage of the PA
Step 4: The exciter.

This allows the expensive UHF-transistors of the transmit amplifiers to be especially protected, since it guarantees that they will receive their operating voltage, or RF-drive only after the coaxial relay has switched. This also ensures that the coaxial relay does not have to switch the RF-power. This will increase the life of the contacts considerably. The manufacturers of coaxial relays list the maximum power rating as a function of frequency, which is 2 to 5 times higher than the permissible switching power.

As can be seen in **Figure 1**, the transmit-receive button is followed by a circuit comprising four operational amplifiers (OP.AMPs). The first OP.AMP. shapes the negative pulse generated on pressing the button so that it has a square-wave characteristic; (the positive pulse generated on releasing the button is suppressed using diode D1). The subsequent differentiating link (4n7/47 k) feeds only the slopes of the square-wave impulse to diodes D2 and D3. The amplitude of the voltage fed back from pin 7/I1 via the 47 kΩ-resistor determines which of the two diodes conducts, and thus whether positive or negative slopes switch over the second OP.AMP., which is connected as a flipflop. It will maintain a clearly defined status after each connection of the operating voltage: Its output is connected to 0 V, which corresponds to the "receive" mode. The third OP.AMP. inverts the signal so that a voltage of 0 V is also connected to the output of the fourth OP.AMP. (an integrator) in the receive mode (in the transmit mode, approx. 10.5 V will be present).

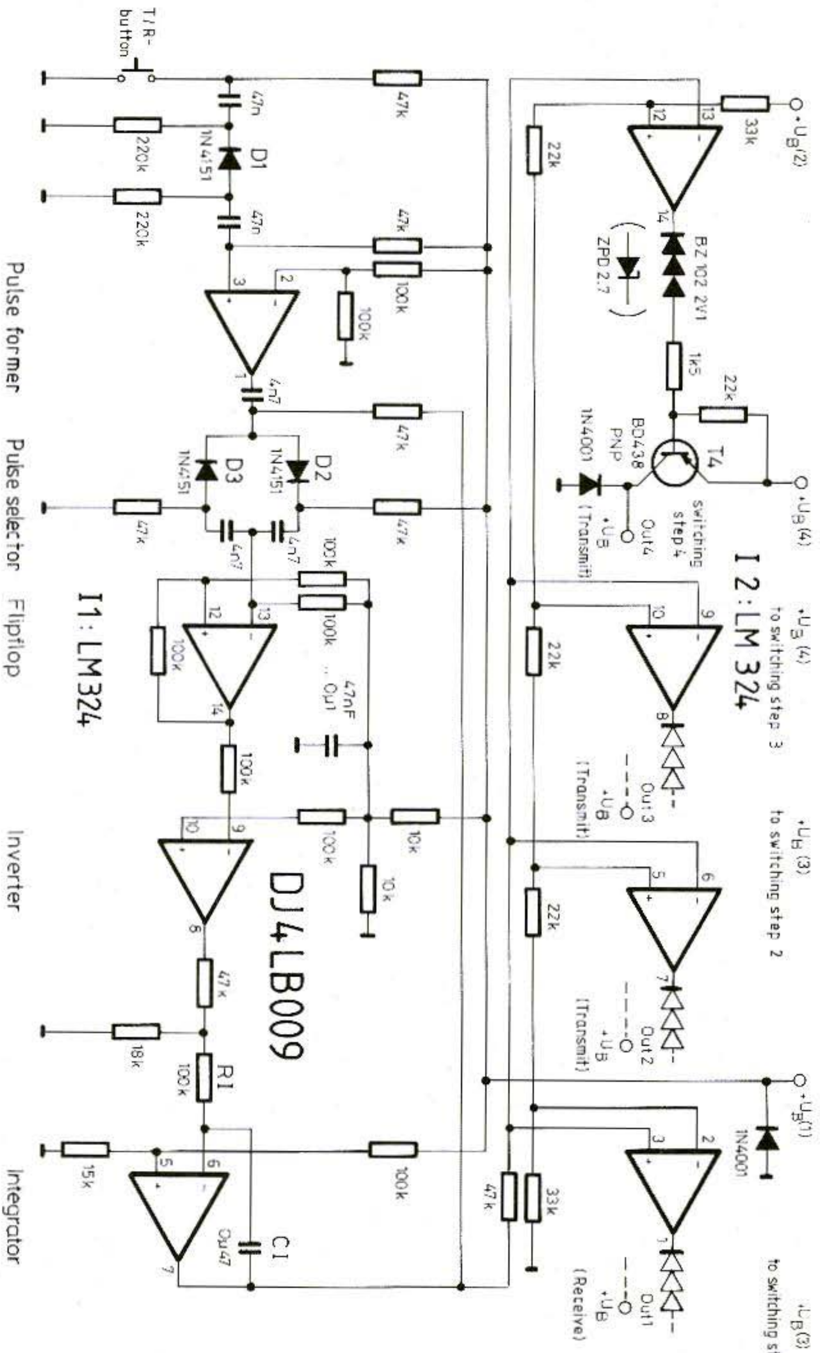


Fig. 1: Control circuit for transmit-receive switching with four time steps; the circuit of the identical switching stages 2, 3, and 4 is only indicated.

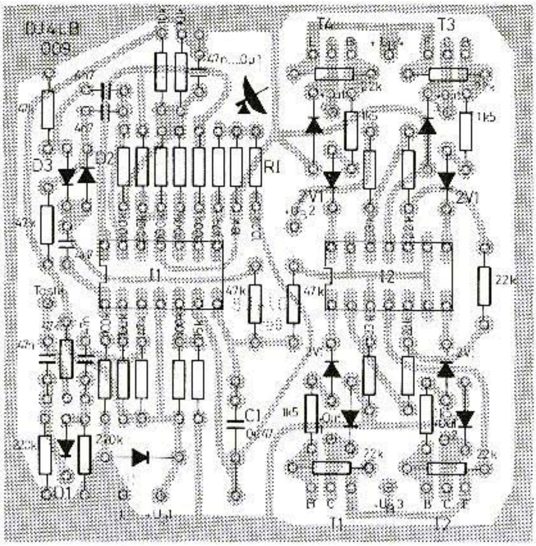


Fig. 2:
Switching transistors T1–T4 are screwed to the inside panels of the metal box.

The integrator drives four comparators (also OP.AMPs) which are adjusted to various threshold values using a number of resistors. Each of the comparators drives a switching transistor (only shown once). Due to the internal connections of the **LM 324**, the maximum positive output voltage is approx. 1.5 V lower than the operating voltage used. For this reason, multiple diodes, or zener diodes with a threshold of 2 to 3 V are required in order to block the switching transistors sufficiently. When switched on, the 1.5-kΩ bias resistors determine the collector-emitter saturation voltages of the switching transistors, which are less than 0.1 V up to collector currents of approx. 250 mA.

Due to the linear increase or decrease of the voltage at the output of the integrators after actuating the transmit-receive button, the four switching stages are switched on, or off, at the same time spacings.

Figure 2 shows the component locations on the 72 mm × 72 mm PC-board DJ 4 LB 009. It is single-coated and fits into a metal box with the dimensions 74 mm × 74 mm. The four switching transistors can be screwed to the side panels using mica insulation. The author's prototype is shown in **Figure 3**. One will see four LEDs, which are connected to the output via resistors of 1 kΩ, each, and are used to show the switching functions of the sequence.

Components

I 1, I 2: LM 324 N (var. manufacturers)
T 1 ... T 4: BD 438 (Siemens), PNP silicon transistors, 45 V/4 A, plastic case TO-126
D 1 ... D 3: 1N 4148, 1N 4151 or similar switching diodes

4 pcs. BZ 102-2V1 (TFK), or BZX 75/C2V1 (Philips) stabilizer-diodes. Instal so that the marker ring faces towards the OPAMP; if these are not available, one can use ZPD 2.7 (ITT), or similar zener diodes, which should be installed with the marker ring facing towards the 1.5-kΩ resistor.

5 pcs. 1N 4001 or similar 1 A rectifier diodes.

Alle resistors for 10 mm spacing.

All capacitors: 5 mm spacing, with the exception of the two 47 nF-capacitors at the switching input: 7.5 mm or 5 mm spacing; as well as the integrator capacitor C_I for which a maximum of 15 mm is available.

Special Notes:

The module is designed for an operating voltage of 12 V, but will operate between approx. 9 V and 30 V. When using 20 to 30 V, the resistors provided in front of the base connections of the switching transistors should be increased from 1.5 kΩ to 3.3 or 3.9 kΩ, in order to ensure that the LM 324 is not overdriven.

Any contacts can be used for transmit-receive switching, since the circuit is not sensitive to contact jump. It is also possible for several transmit-receive switches to be connected in parallel to the module. This means that the transmit-receive switching cannot only be operated from the transmitter itself, but also from the microphone, TV-camera, or teletype.

Any unwanted switching of the transmitter, e.g. when carrying out tests using receive preamplifiers, or during the warm-up period of a transmit tube, can be avoided by blocking the U_B1 voltage (only U_B1!).

The whole circuit is very insensitive to interference pulses and RF-interference due to the use of relatively slow OPAMPs, as well as switching transistors having a low cutoff frequency. On the other hand, it is advisable for the connections from the transmit-receive contact, at least, to be fed via a feedthrough capacitor (max. 5 nF) into the metal box.

Operation

The module is connected to the operating voltage of the transmitter or receiver.

On switching on, step 1 is in operation, which means that the station is operating in the receive mode.

On depressing the button, the station is

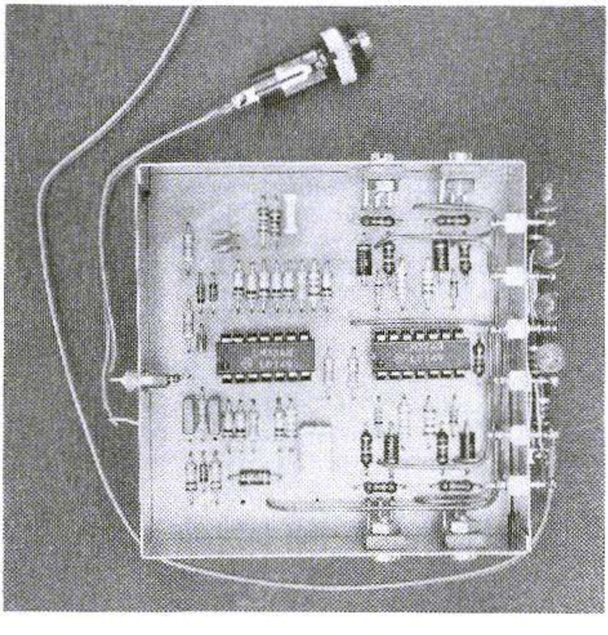


Fig. 3: The author's prototype has a LED connected to each output to demonstrate the switching state.

switched to "transmit" in the following sequence:

Firstly, step 1 is switched off, after which steps 2, 3, and 4 are actuated one after the other.

On depressing the button again, the station is switched to the "receive" mode:

Firstly, steps 4, 3, and 2 are switched sequentially, after which step 1 is switched on.

The time spacing between two switching steps can be varied within wide limits by changing the value of the integration link R_i, C_i, and can be calculated according to the following equation:

Time (from step to step):
T [ms] = 1.4 x R [kΩ] x C [μF]

Usually, coaxial antenna relays require approximately 10 to 20 ms between provision of the energizing voltage until closing the contact between transmitter and antenna connection. In some cases, this can be as long as 30 ms.

The switching times of small Reed relays are between 0.2 and 1 ms, and need not be taken into consideration here. Power relays usually have switching times of 5 to 10 ms.

With the values (100 kΩ/0.47 μF) given in Figure 1, a time spacing of approx. 66 ms results before switching to the next step, which is more than sufficient for most applications. It should be noted that the whole switching process from depressing the button until switching on the fourth step amounts to 5 × T, which corresponds to 1/3 s in the described case.

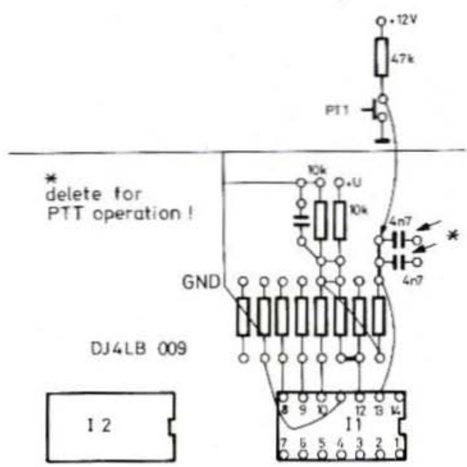
Modification for PTT-operation

In the original circuit, the switching from receive-to-transmit is made by keying the transmit-receive switch. The return to the receive mode is made with a second depression of the same switch.

If one wishes to operate in the PTT-mode (in other words to transmit as long as the button is depressed, and switching to receive on releasing the button), it is possible to wire the control circuit as follows:

Delete the 47 nF capacitor at the activating contact, and the two 4.7 nF capacitors at pin 13 of I1. Connect the actuating input to pin 13 of I1.

Connect a 47 kΩ resistor from the contact to the positive operating voltage in order to garantee the bi-stable operation of the flipflop. The rest of the passive components between the 47 nF and the two 4.7 nF capacitors can be deleted if the module is only to be used for PTT-operation.



Friedrich Krug, DJ3RV

A Receiver for the VLF Time and Frequency Standard Transmissions from DCF 77

This receiver can be used to synchronize a phase-locked crystal oscillator for use as frequency standard for amateur applications. The requirement of an accurate time and frequency standard for the realization of modern communication technology has been mentioned several times in VHF-COMMUNICATIONS. Especially in coherent telegraphy (CCW) or similar modern communications (such as spread spectrum technology), one requires an exact frequency and time base for the time and computer-controlled digital filters and evaluation circuits. The described receiver is designed for the reception of the VLF time and frequency standard DCF 77, which transmits at a frequency of 77.5 kHz. This transmitter can be received throughout Europe.

The time information is provided by the transmitter in the form of pulse-width modulated second pulses; and the extremely constant transmit frequency is used as reference frequency to synchronize a crystal oscillator with the aid of a phase-locked loop to form a frequency standard.

The applications of such a time and frequency standard can be made very versatile by adding additional modules. The most often used application is, of course, to evaluate the time information to form a very accurate clock. However, the extremely constant frequency is more important to us radio amateurs who can use it as frequency standard.

Examples for this are some modules realized by the author to improve the frequency stability of the VHF-oscillator with DAFC described by DJ7VY in Edition 2/1981 of VHF COMMUNICATIONS, in conjunction with the receiver concept described by the author in Edition 4/1981, as well as the 96 MHz crystal oscillator for UHF/SHF local oscillators described by DK1AG also in that edition.

The author has been using a 10 MHz frequency standard, which is locked to the DCF 77 receiver, for testing, and also as a clock for frequency counters.

The actual reason for developing this receiver was to obtain a time and frequency standard for a computer-controlled RTTY transceiver, in which all frequency and time signals are coupled together so that coherent communication will be possible. This requires a circuit that is designed for high reliability.

In the last ten years, a multitude of publications and construction articles have appeared regarding receivers for the time and frequency standard transmitter DCF 77. A certain number were constructed and operated with success by the author. However, the author was forced to carry out his own development since the operation of the available receivers was not possible in the vicinity of a TV-receiver, or computer with video display. It was found that in most cases an unlock condition of the phase-locked loop in the crystal oscillator circuit, or the time control circuit was caused by interference originating in the direct vicinity of the receiver that is mostly from one's own electrical equipment (man-made noise).

This led very quickly to a clean-up of the "electronic pollution", however, it was not possible to suppress all interference sources, which meant that a receive system was required that had a higher interference immunity.

1. RECEIVER CONCEPT

It is necessary to know the field strength to be expected and the interference field strength before conceiving the receiver. In a publication by Becker und Rohbeck (1) of the Federal Physical Laboratories (PTB) of Brunswick/ W. Germany, who generate the frequency and time signals for the DCF 77 transmitter, details are given about the usable field strength, which are based on measurements of the German PTT, as well as calculations of the propagation of the ground wave. According to this, an electrical field strength of 14.5 mV/m at a distance of 100 km, 2.3 mV/m at 400 km, and still 0.6 mV/m at 800 km is to be expected.

Of course, these values can be considerably less in the case of unfavourable locations, or where screening occurs. For instance, field strengths of between 0.6 mV/m and 8 mV/m were measured at the author's desk in a reinforced concrete building at a distance of approximately 150 km from the transmitter.

It was found that it was very much more difficult to determine the field strength of the interference. It was necessary to localize the interference sources, after which the measurement was not possible in most cases, or only with considerable expense of measuring equipment and time.

The interference sources can be split into three different groups.

A) Wideband Single-Pulse Interferences

These mainly originate from switching processes in electrical household equipment such as washing machines, electric cookers, or heaters. Most of this equipment is, unfortunately, still not switched during the zero pass of the power-line voltage using electronic components. Any radio amateur who has his antenna in the vicinity of an old lift system will know the problems encountered with this type of interference.

Thunderstorms also fall within this category of interference, as do discharge processes in non-grounded transmit antennas, and static discharge in the case of synthetic flooring and chairs. Even in the author's laboratory, a thermostat-controlled low-voltage soldering iron and the ignition processes of the fluorescent lamps caused strong spikes in the signal from the antenna.

Most of these interference sources cannot be suppressed. Any interference that is fed into the equipment by the power line, can be suppressed sufficiently well by using power-line filters in the form of lowpass filters, and by using a voltage stabilizer. Battery operation of the circuit is not possible due to the high current drain.

The interference spectrum received via the receive antenna must be suppressed as well as possible using narrow-band filters whose "ringing" is as low as possible, in order to ensure that the following circuits are not overdriven. Any residual phase jumps must be suppressed with the aid of the time constant of the phase-lock loop.

B) Low-Frequency Periodic Pulse Interferences

This is the largest category of interference sources. Frequency spectrums are generated by phase-control circuits of small electrical equipment, collector motors, vertical and horizontal deflection circuits of TV-receivers and monitors, switching power supplies, up to the data currents in computers; in the most favourable cases, this is in the form of a noise spectrum; mostly, however, they are in the form of descrete spectral lines of relatively high power density. The fifth harmonic of the line frequency of TV-receivers is extremely strong:

$$f = 5 \times 15625 \text{ Hz} = 78125 \text{ Hz},$$

which means that it is only 625 Hz above the transmit frequency of DCF 77. In the case of a portable TV-receiver owned by the author, the interference field strength of the fifth harmonic amounted to $\hat{E}$ = 300 mV/m! at a spacing of 3 m. Since the line frequency used for displaying the data of personal computers is only approximately 15625 Hz, the interference from such sources can be far nearer to the DCF-frequency.

C) Descrete Frequency Interference

This interference is usually caused in the shack itself, which means that it can very often be suppressed. Radiating shortwave feeders, and VHF-portable equipment should not be operated in the vicinity of the DCF-receive antenna.

As precautionary measure, one should provide a low-pass filter between antenna and antenna preamplifier.

Together with the knowledge of the interference effects and experience made by constructing the circuits published in (1), the concept was established for building a receive system for a simple frequency standard.

Of course, even a simple frequency standard should exhibit a good stability. For this reason, the long-term stability was determined over a period of 100 days (1), and was found to be approximately 2×10^{-13}, which was given by the signal of the DCF 77 transmitter.

The lower short-term stability of the DCF 77 signal caused by propagation and interference, requires a short-term stable crystal oscillator that is controlled via a PLL having a large time constant.

A frequency of f_0 = 3.1 MHz was selected as oscillator frequency. This is the 40th harmonic of the transmit frequency of DCF 77. If the crystal frequency is divided by 31, a decadic frequency of 100 kHz will result, from which all further signals are derived.

Of course, other oscillator frequencies are possible, however, require an even multiple of the DCF-frequency. This results in several useful reference frequencies, according to whether a harmonic is required that can be divided decadically, hexadecimally, or binary.

Table 1 shows the frequencies divided into primary factors, for which the PC-board layout can be used so that it is possible to also employ a crystal having another frequency.

77.5 kHz	$= 2^2 \times 5^4 \times 31$ Hz	$= f_{DCF}$
3.1 MHz	$= 2^5 \times 5^5 \times 31$ Hz	$= 40 \times f_{DCF}$
2.79 MHz	$= 2^4 \times 3^2 \times 5^4 \times 31$ Hz	$= 36 \times f_{DCF}$
3.72 MHz	$= 2^6 \times 3 \times 5^4 \times 31$ Hz	$= 48 \times f_{DCF}$
4.96 MHz	$= 2^8 \times 5^4 \times 31$ Hz	$= 64 \times f_{DCF}$

Table 1: Divided multiples of the DCF-frequency

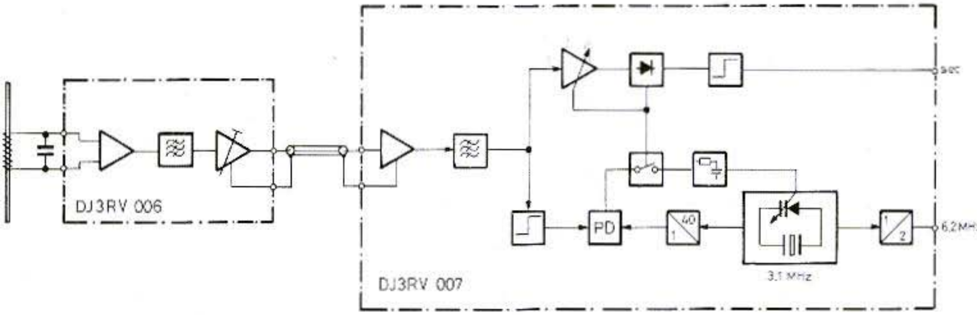


Fig. 1: Block diagram of the DCF 77 receiver

If the oscillator frequency is selected as clock for a computer – microprocessors such as the 8085 or Z 80 require clock frequencies between 2 MHz and 8 MHz – one will ensure that the signals and evaluation processes generated in the computer are directly coupled to the standard frequency.

Figure 1 shows the block diagram of the receiver concept. The signal is received using a ferrite or loop antenna. The signal is then fed to a preamplifier equipped with a narrow-band filter for suppressing the harmonics of the line frequency and is followed by an impedance converter. The antenna module DJ3RV 006 can be accommodated at a low-interference position and connected to receiver DJ3RV 007 with the aid of conventional coaxial cable of any required length (up to 100 m RG-58/U).

The receiver DJ3RV 007 possesses an impedance converter at the input which is followed by a further filter for improving the ultimate selectivity. After this, the signal is split up and fed firstly to a controlled amplifier with demodulator. This provides the time-information, and is used simultaneously for indicating a missing transmit signal. The second signal path of the 77.5 kHz signal is fed via a limiter to the phase discriminator PD. The phase comparison is made to the signal derived by dividing the oscillator frequency by 40. The phase deviation provides a DC-voltage that is passed via a low-pass filter having a large time constant, which is then used for synchronizing the oscillator frequency. If the transmitter is inoperative for any reason, it is possible for the control circuit to be disconnected.

In order to ensure a low-reactive output coupling, the oscillator signal is fed out via a frequency doubler as 6.2 MHz signal.

This concept was accommodated on two PC-boards and the author's prototype has been running for more than a year in virtually continuous operation.

2. THE ANTENNA

The first experiments with DCF-reception were made by the author approximately 10 years ago using a ferrite antenna from an old portable radio. The long-wave coil on this ferrite rod was resonated to 77.5 kHz by adding parallel capacitors, and used together with an antenna preamplifier as described by (1). This antenna had a measured bandwith of 2.6 kHz together with the relatively large parallel capacitance, which results in a Q of approximately 30. As later experiments with various antennas showed, this antenna represented a good compromise between receive energy and sensitivity to interference (purely by chance), and was only exceeded by using loop antennas.

The disadvantage of ferrite antennas is the non-linear behaviour of ferrite materials. This causes a phase modulation of the input signal in the case of strong magnetic field interferences from welding transformers, chokes in fluorescent lamps, deflection coils on TV-tubes, and the chokes in switching power supplies. The higher the Q of the antenna, the greater is the phase deviation. The interference compensates for itself statistically due to the long-term evaluation of the low-pass filter, however, it can cause a phase jump of the phase-control circuit in the case of additional interference pulses.

2.1. Calculation of the Antenna Voltage

For space reasons, only resonant loops or ferrite antennas can be used for reception of the VLF DCF 77 transmissions. The dimensions of these antennas are very small in comparison to the wavelength. Such antennas mainly utilize the magnetic field components of the signal to be received, and the directional characteristics of the antenna correspond to that of a short dipole, which is a "8" with a flat maximum and sharp minima. The output voltage of the antenna described in (3) amounts in the following, when optimumly aligned to the magnetic field component H:

$$U_{eff} = \frac{\pi \times f}{\sqrt{2}} \times A \times N \times Q \times \mu_{eff} \times \mu_o \times H$$

and to the following, due to the combination of the magnetic field component H with the electrical field component E with the aid of the field impedance characteristic Z_o of free space:

$$H = \frac{1}{Z_0} E = \sqrt{\frac{\varepsilon_0}{\mu_0}} \times E$$

The following is valid:

$$U_{eff} = \frac{\pi}{\sqrt{2}\,\lambda_0} \times A \times N \times Q \times \mu_{eff} \times E$$

where:

- λ_0 – Wavelength of the receive signal
- A – Loop surface, or cross section of the ferrite surface
- N – Number of turns
- Q – Operating Q of the antenna system
- μ_{eff} – Effective permeability number of the ferrite material

In the case of the loop antenna, $\mu_{eff} = 1$.

The output voltage of the antenna should be as high as possible in order to ensure good reception.

As can be seen in the equation, the voltage at the output of the antenna is directly proportional to the surface A, the number of turns N, and the Q, and to the effective permeability μ_{eff} in the case of a ferrite antenna, for any given field strength at the reception location. Unfortunately, these magnitudes cannot be increased infinitely.

The loop surface A is limited, if the antenna is to remain within handy dimensions, and is given by the cross section of the ferrite rod in the case of a ferrite antenna.

The number of turns N is limited by the stray capacity C_s of the coil. The intrinsic resonance of the coil should not be less than the receive frequency.

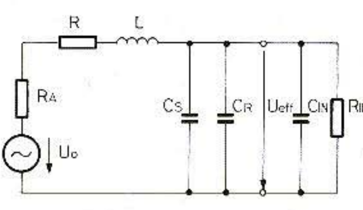


Fig. 2: Equivalent diagram of the receive antenna

The operating Q of the antenna is dependent on several magnitudes, that must be considered further in order to approximate their effects.

Figure 2 gives the equivalent diagram of the receive antenna; U_0 is the source voltage resulting from the field strength at the receive location. If the Q is not too small ($Q > 10$), the resonant frequency of the antenna is determined by the inductance L and the stray capacitance C_s of the coil, the input capacitance C_{in} of the antenna preamplifier, and the resonant circuit capacitance C_R.

$$f = \frac{1}{2\pi\sqrt{L \times (C_S + C_{in} + C_R)}}$$

The operating Q can be calculated as follows:

$$\frac{1}{Q} = \frac{2\pi f \times L}{R_{in}} + \frac{(R + R_A)}{2\pi f \times L} + \left(\frac{1}{Q_M} \times \frac{3\,l'}{4\,l}\right)$$

As can be seen in the equation, the input impedance R_{in} of the antenna amplifier connected in parallel to the resonant circuit, the loss impedance R of the wire in series with the coil, and the radiation resistance R_A determine the Q of the inductance.

In the case of ferrite antennas, one must also take the Q of the material into consideration,

$$Q_M = \frac{\mu_r''}{\mu_r'}$$

however, only when the inductance length l' is wound over the whole length of the ferrite rod. The term given in parenthesis is a coarse approximation that was found by the author experimentally.

The radiation impedance R_A of the antenna described in(3)

$$R_A = 20\,\Omega \times \left(\frac{2\pi}{\lambda_0}\right)^4 \times A^2 \times N^2$$

will become very low for large wavelengths and can usually be neglected with respect to the loss resistance R of the coil. This loss resistance results from the resistance of the wire itself, its increase due to the skin effect, as well as the eddy current losses in the wire. The skin effect and the eddy current losses can be reduced considerably by using stranded wire, which means the DC-resistance of the wire can be considered as the loss resistance R.

The inductance L is decisive for the Q of the antenna. It is dependent, however, on the surface area A and the number of turns N, and in the case of ferrite antennas, also on the effective permeability μ_{eff}.

The inductivity L can be calculated for a round loop antenna as short, wide inductance as shown in (4):

$$L = \mu_0 \times \frac{d}{2} \times N^2 \times (\ln \frac{4d}{l} - 0.5)$$

where l is the width of the inductance and d is the diameter of the coil.

In the case of the ferrite antenna in the form of a thin, long coil, the following is valid:

$$L = \mu_0 \frac{\pi \times d^2 \times N^2}{2\,(l+l')} \times \mu_{eff}$$

where l is the length of the rod, l' is the length of the coil, and d is the diameter of the ferrite rod. As can be seen in the diagram given in (5) shown in **Figure 3**, the effective permeability is dependent on the basic permeability μ_i of the material and the length-to-diameter l/d of the ferrite rod itself. Conventional ferrite rods for antennas having a l/d of 10 to 25, will have a μ_{eff} of between 50 and 200, which is virtually independent of the material used.

The stray capacity C_s and the input capacitance C_{in} can only be determined with the required accuracy by measuring the resonance frequency of the antenna. If the

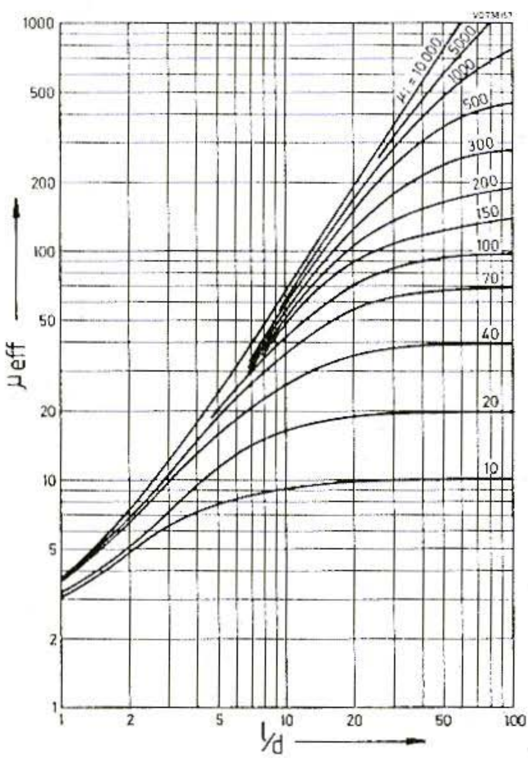


Fig. 3:
Effective rod permeability μ_{eff} as a function of the l/d ratio according to (5)

antenna possesses a resonance frequency f_1 without the resonant circuit capacitor C_R, and a resonance frequency f_2 results when using a certain capacitor such as C_R = 270 pF, the following is valid:

$$C_s + C_{in} = \frac{C_R}{\left(\frac{f_1}{f_2}\right)^2 - 1}$$

This measurement allows the actual inductivity to be determined:

$$L = \frac{1}{(2\pi\, f_1)^2 \times (C_S + C_{in})}$$

The practical use of the equations is now to be shown with the aid of a loop and a ferrite antenna.

2.1.1. Loop Antenna

Loop diameter d = 50 cm

Coil width l = 1.3 cm

Number of turns N = 50

Stranded wire of 45 x 0.05 mm dia. silk-covered enamelled copper wire is to be used which has a resistance of 0.2 Ω/m.

Input impedance R_{in} = 220 kΩ

Permeability constant $\mu_0 = 4\pi \times 10^{-9} \frac{V\,s}{A\,cm}$

$$L = \mu_0 \times \frac{d}{2} \times N^2 \times (\ln \frac{4d}{l} - 0.5)$$

$$= 4\pi \times 10^{-9} \times \frac{50}{2} \times 50^2 \times (\ln \frac{200}{1.5} - 0.5)\ \frac{Vs}{A}$$

$$= 3.45\ mH$$

$$R = \pi \times d \times N \times 0.2\ \Omega/m = 15.7\ \Omega$$

If R_A is neglected, the loop antenna will have a Q of

$$Q = \frac{1}{\frac{R}{2\pi f \times L} + \frac{2\pi f L}{R_{in}}} = 58.9$$

The following was measured on the completed loop antenna:

L = 3.48 mH

$C_S + C_{in}$ = 770 pF

Q = 46.5

A capacitance value of C_R = 442 pF is required for resonating the antenna to f = 77.5 kHz. Since the stray capacitance of the winding is very dependent on the winding itself, it is necessary for C_R to be determined individually for each antenna.

The antenna voltage U_{eff} is calculated as a function of the electrical field strength E with the following measured values:

$$U_{eff} = \frac{\pi}{\sqrt{2}\ \lambda_0} \times A \times N \times Q \times \mu_{eff} \times E$$

$$= \frac{\pi}{\sqrt{2} \times 3871\ m} \times \frac{\pi \times 0.5^2\ m^2}{4}$$

$$\times 50 \times 46.5 \times 1 \times E$$

$$\boxed{U_{eff} = 0.26\ m \times E}$$

2.1.2. The Ferrite Antenna

A ferrite rod of a length l = 200 mm and a diameter d = 10 mm is used. According to the databook, the ferrite rod has a material Q of Q_M = 180 at 77.5 kHz, and a basic permeability of μ_i = 750.

According to **Figure 3** the result is μ_{eff} = 160.

Number of turns N = 150 with a coil width l' = 30 mm, and using stranded wire of 45 x 0.05 mm dia. silk-covered enamelled wire having a resistance of R = 1.4 Ω.

If R_A is neglected, and R_{in} is assumed to be 220 kΩ, the following will result:

$$L = \mu_0 \times \frac{\pi \times d^2 \times N^2}{2\,(l + l')} \times \mu_{eff}$$

$$= 4\pi \times 10^{-9} \times \frac{\pi \times 1^2 \times 150^2}{2\,(20 + 3)} \times 160\ \frac{VS}{A}$$

$$= 3.09\ mH$$

$$\text{and } Q = \frac{1}{\frac{R}{2\pi f\ L} + \frac{2\pi f\ L}{R_{in}} + \frac{3l'}{4l \times Q_M}}$$

$$= 119$$

The following was measured on the completed ferrite antenna:

$L = 2.99$ mH

$C_S + C_{in} = 74.3$ pF

$Q = 110$

A capacitance value of $C_R = 1.33$ nF is required for resonance to f = 77.5 kHz.

The antenna voltage U_{eff} is calculated from the following measured values:

$$U_{eff} = \frac{\pi}{\sqrt{2} \times \lambda_0} \times A \times N \times Q \times \mu_{eff} \times E$$

$$= \frac{\pi}{\sqrt{2} \times 3871 \text{ m}} \times \frac{\pi \times (0.01)^2 \text{ m}^2}{4}$$

$$\times 150 \times 110 \times 160 \times E$$

$$\boxed{U_{eff} = 0.12 \text{ m} \times E}$$

It will be seen that the ferrite antenna only provides half the voltage as the loop antenna for the same field strength. This cannot be increased condiserably by increasing the number of turns N, since this will decrease the Q. When using the same ferrite rod, the following will result:

N = 300 turns; Q = 32 ; $U_{eff} = 0.14$ m x E
N = 500 turns; Q = 12.5; $U_{eff} = 0{,}145$ m x E

Since, on the one hand, the interference load of the input stage of the antenna amplifier increases with reducing Q, and on the other hand an interference-phase modulation occurs at higher Q, it was found that the optimum is obtained with Q = 50. This corresponds to a number of turns N = 250 in the case of a 200 mm long ferrite rod, and N = 300 in the case of a 140 mm ferrite rod, when using the same material and wires.

If the antenna is to be used over a wide temperature range, e.g. such as under the roof, a lower Q will be advisable due to the large temperature coefficient of the ferrite material.

3. ANTENNA PREAMPLIFIER DJ3RV 006

In order to be able to mount the antenna at a position of low interference, the same concept is used as was described in (2), where the ferrite or loop antenna was operated in conjunction with an amplifier and coupled out at low impedance using a coaxial cable.

The antenna preamplifier has thus two main tasks:
The provision of selectivity with the aid of a narrow-band filter, and the matching to the feeder cable to the receiver. The DC-voltage supply is made from the receiver via the feeder cable.

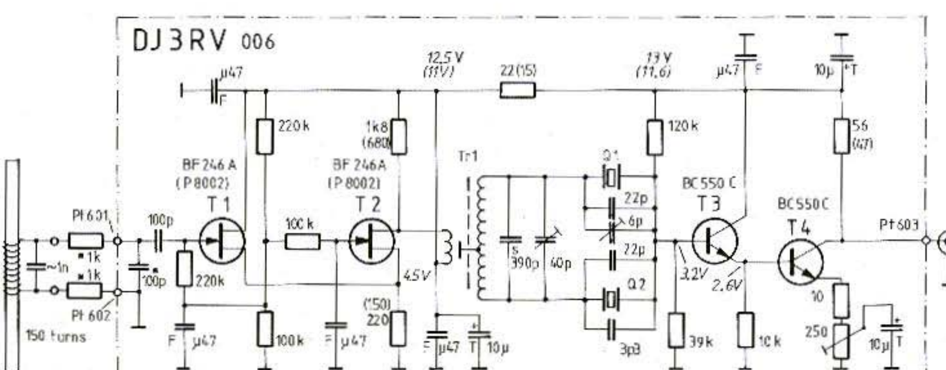


Fig. 4: Antenna preamplifier with two-pole crystal filter

3.1. Circuit Description

The circuit diagram of the module is given in **Figure 4.** In order to keep the interference load on the preamplifier as low as possible, a low and a high-pass filter are provided at the input. If the antenna is operated in the vicinity of the transmitter, it is advisable to provide the low-pass filter comprising 2 x 1 kΩ and 100 pF, , designated with * in the circuit diagram. This 100 pF capacitor is part of the resonant circuit capacitance C_R and transforms the two 1 kΩ resistances as additional loss resistances in the antenna circuit. The Q of the antenna will be reduced to Q ≈ 50 in the case of the ferrite antenna calculated in the previous section, having N = 150 turns and a circuit capacitance of approx. 1.3 nF, and a ferrite rod length of 200 mm (or approx. 1.5 nF with a length of 140 mm).

The symetrically wound transformer Tr 1 is decisive for the operation of the crystal filter. The two secondary windings must be wound at the same time and equally onto the whole core. The inductivity should amount to approx. 10 mH. Since the permeability of the cores fluctuates considerably, it is advisable to measure the coil. The number of turns can be between 2 x 27 turns, and 2 x 32 turns. When winding on or off, attention should be paid to the symmetrical characteristics of the windings. An exact alignment of the filter is otherwise no longer possible.

The high-pass filter is formed by the 100 pF coupling capacitor together with the input impedance of the amplifier, and is used for suppressing any low-frequency interference injected into the antenna.

The amplifier in front of the crystal filter comprises T 1 as source follower in order to achieve a high input impedance, and T 2 in a common gate circuit as low-reactive amplifier for decoupling the antenna and crystal filter. The 1.8 kΩ drain resistor of T 2 represents the required source impedance of the crystal filter transformed via Tr 1. The crystal filter has a bandwidth of ± 55 Hz and possesses two clearly defined attenuation poles, which allow interference carriers to be suppressed by more than 70 dB.

The terminating impedance of the filter is realized with the aid of the input impedance of the emitter follower T 3, and its base voltage divider. T 4 is used as amplifier and possesses a low collector impedance of 56 Ω, which is used as source impedance for the feeder cable to the receiver. By altering the emitter feedback at T 4, it is possible to adjust the overall gain of the module to between one and ten times to make it suitable for the various input levels.

As previously mentioned, the DC-power supply of the circuit is provided via the feeder cable. Transistors T 1, T 2, and T 3 are fed via the 56 Ω resistor connected to the drain of T 4. The filter capacitors in conjunction with the 22 Ω resistor neutralize the amplifier so that no self-oscillation can occur.

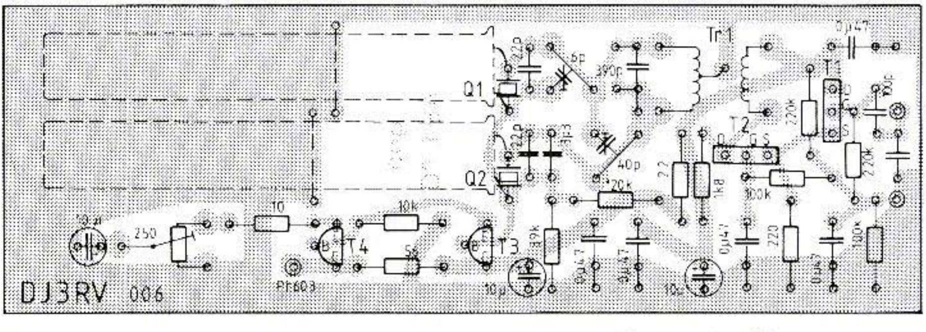


Fig. 5: Component location plan on the single-coated antenna preamplifier board DJ 3 RV 006

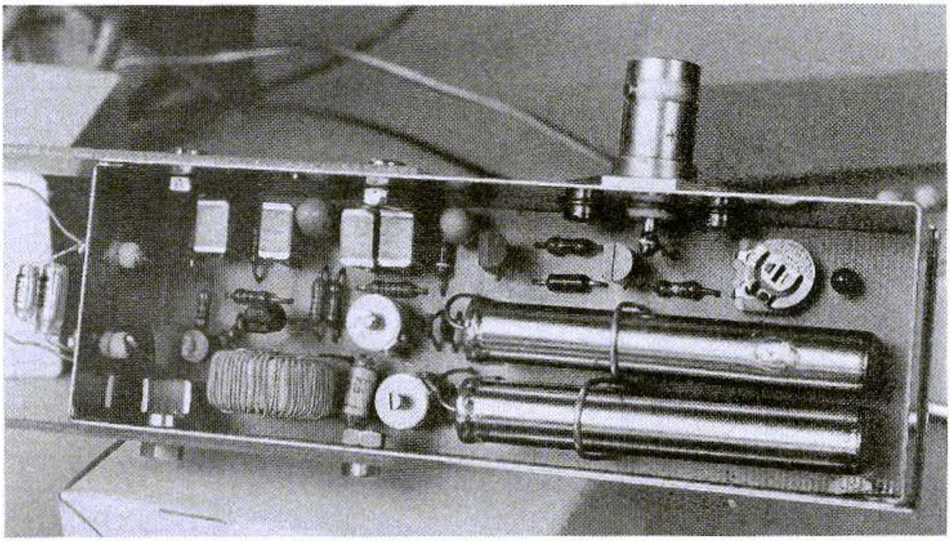


Fig. 6: Photograph of the author's prototype antenna preamplifier DJ3RV 006 complete with ferrite antenna

3.2. Construction and Alignment

The construction of the antenna preamplifier can be made without difficulties, and several DC-voltage values are given in the circuit diagram for checking the operating points. In order to ensure a balanced operating point of the input amplifier, transistors T1 and T2 should be selected to have the same U_{GS}.

The component location plan is given in **Figure 5**, and **Figure 6** shows a photograph of the author's prototype.

The following parts list contains all special components.

Parts list for DJ3RV 006

T1, T2: BF 246 A (T I)
Selected to have the same U_{GS} at I_D = 10 mA
Alternative: P 8002
Selected to have the same U_{GS} at I_D = 20 mA

T3, T4: BC 550 C (Siemens)

Tr 1: Toroid core, R 16 N 30 (Siemens)
Primary: 8 turns of stranded wire 45 x 0.05 mm dia. silk-covered enamelled copper wire
Secondary: 2 x 30 turns of stranded wire 45 x 0.05 mm dia. silk-covered enamelled copper wire

Q1: f = 77508 Hz, series resonance
L_1 = 53 H ± 20%
C_0 ≈ 10 pF

Q2: f = 77436 Hz, series resonance
L_1 = 53 H ± 20%
C_0 ≈ 10 pF

If a strong interference level is to be expected at the location of the antenna, e.g. in the vicinity of a TV-receiver or switching power supply, it is advisable to wind the antenna firstly, connect it to an oscilloscope, and align for resonance. In this case, it is possible to determine the interference level of the equipment in order to limit the number of surprises later.

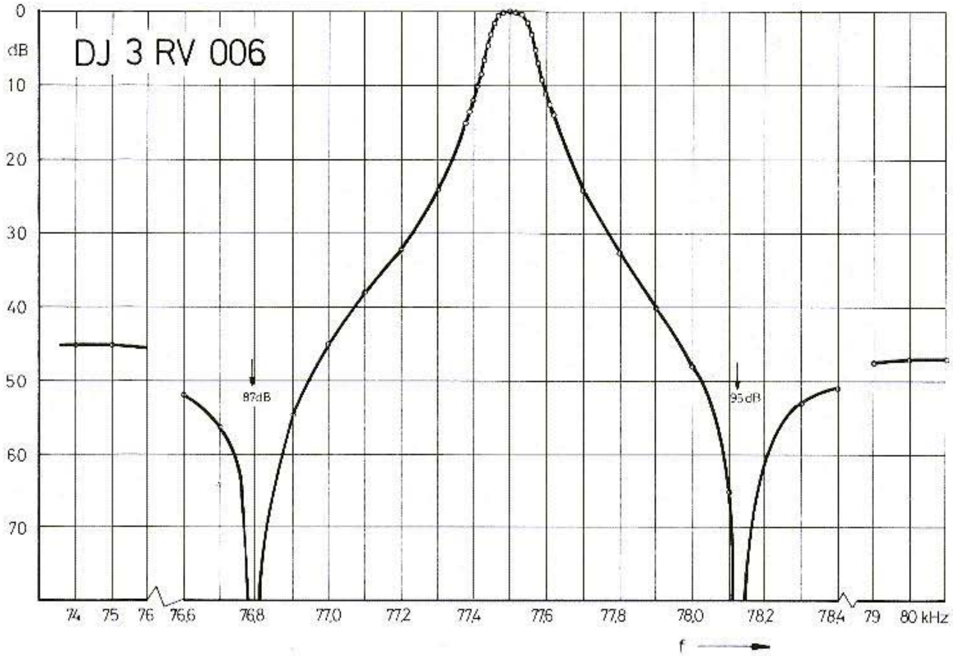


Fig. 7: The frequency response of the two-pole crystal filter used in the author's prototype

The amplifier previous to the crystal filter can be overloaded by interference pulses. At the drain of T2, limiting will take place at input voltage peaks of > 0.3 V at Pt601. If larger interference levels are to be expected, power FETs type P8002 can be used for T1 and T2. In this case, the values given in parenthesis in the circuit diagram are valid, and the primary winding of transformer Tr1 must be reduced to five turns. A limitation will then not occur until voltage peaks of > 0.8 V are exceeded.

The passband curve of the crystal filter given in **Figure 7** will result when using the crystals Q1 and Q2 listed in the parts list. The 40 pF trimmer is used to align the center frequency, and the 6 pF trimmer to adjust the position of the attenuation poles.

The alignment of the filter is made as follows:
- 6 pF trimmer to center position (approx. 4 pF)
- Align the 40 pF trimmer for maximum at 77500 Hz
- Align the 6 pF trimmer so that the attenuation pole is set to the fifth harmonic of the TV-line frequency f = 78125 Hz
- Align the 40 pF trimmer so that the attenuation at 75 kHz and 80 kHz are equal.

An alignment for the best symmetrical passband curve and ultimate selectivity can only be made with the aid of a stable generator. It is very difficult to sweep such narrow filters!

The antenna preamplifier can be used as a separate module if a coupling capacitor of at least 0.1 µF is connected to the output and the DC-voltage feed of + 12 V is fed to the collector of T3.

4. RECEIVER AND CRYSTAL OSCILLATOR MODULE DJ3RV 007

This module is designed to obtain the time

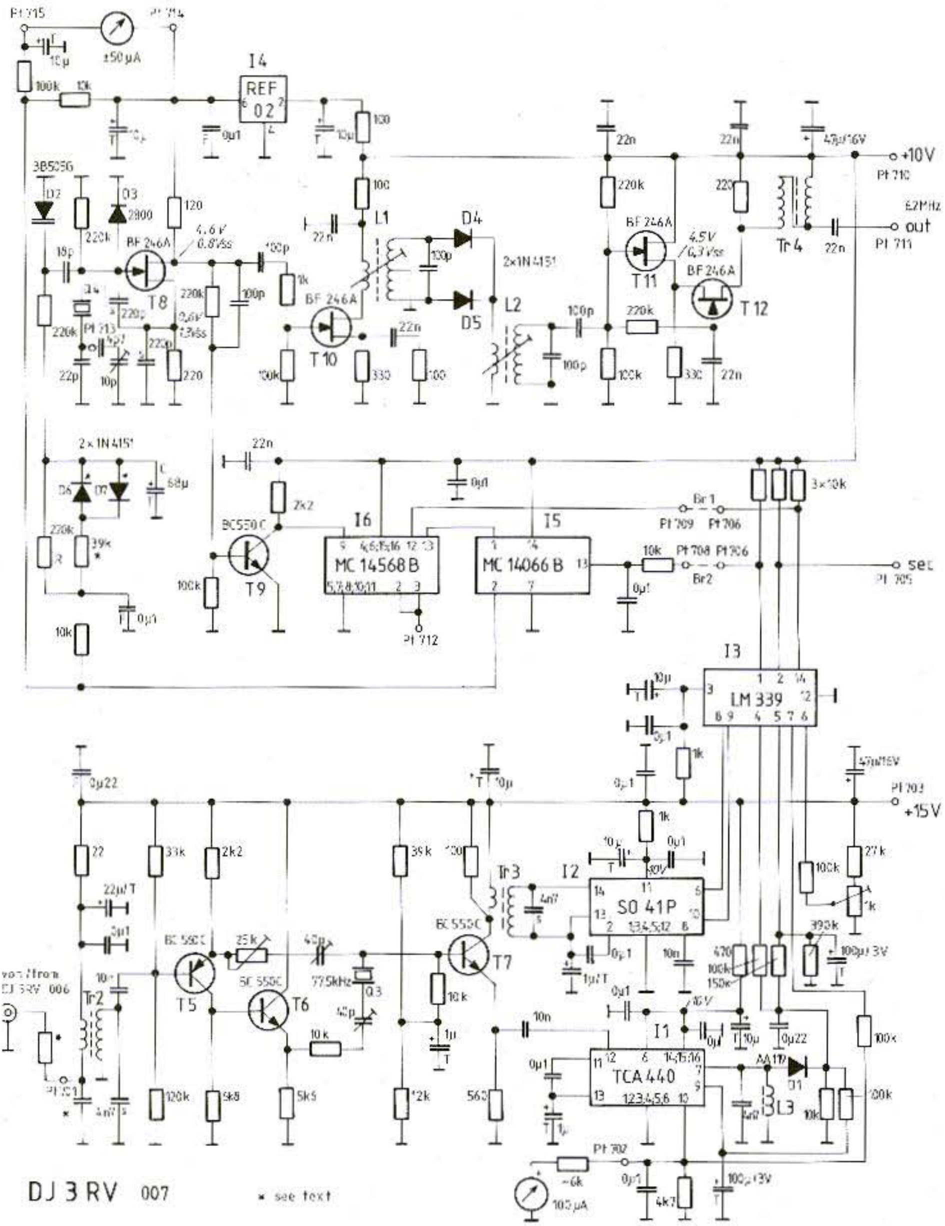


Fig. 8: Circuit diagram of module DJ3RV 007 with phaselocked crystal oscillator as frequency standard, and DCF 77 receiver with crystal filter, limiter for the reference frequency, and demodulator for the second pulses.

information and standard frequency from the DCF-77 signal provided by the antenna preamplifier. As can be seen in the block diagram given in **Figure 1**, the receiver and crystal oscillator are combined in a single module, which provides the second pulses and a 6.2 MHz signal with a level of – 6 dBm into 50 Ω.

The time information is decoded from the second pulses in a further module, which is to be described in a later edition of VHF COMMUNICATIONS. The 6.2 MHz signal is used as reference frequency for the generation of the auxiliary frequencies required for the RTTY transceiver circuits, which are also to be described later.

4.1. Circuit Description

The circuit diagram of the module is shown in **Figure 8.** The DCF 77 signal fed in from the antenna preamplifier via the coaxial cable, is fed at low impedance to Pt 701 of the toroid transformer Tr 2. At the same time, the DC-voltage supply of the antenna preamplifier is made via the primary winding. In order to improve the ultimate selectivity, the secondary winding is tuned to resonance with the aid of the 4.7 nF styroflex capacitor. This resonant circuit is very wideband due to the losses in the core material, and due to the loading of the input impedance of transistor T 5, which means that no alignment is required.

Transistors T 5 and T 6 generate the anti-phase signals for the crystal filter comprising Q 3. This part of the circuit is based on a publication of Hetzel and Rohbeck (2); the operating points and impedances were matched to the author's requirements.

Transistor T 7 isolates the filter from the subsequent stages. The limiting amplifier equipped with an SO 41 P is driven from the collector via Tr 3. The secondary winding of Tr 3 is also brought to resonance at 77.5 kHz with the aid of a 4.7 nF styroflex capacitor. The limited 77.5 kHz signal is fed via a comparator in I 3 (LM 339), which is used as level converter to CMOS-level at Pt 704. It is then passed via a bridge to Pt 709 where it is used as reference signal for phase comparison.

The DCF 77 signal is fed from the emitter of T 7 to the controlled IF-amplifier of the TCA 440. The other parts of the TCA 440 are not used, and are grounded, or connected to the operating voltage.

The amplitude-modulated second pulses are demodulated with the aid of diode D 1 (AA 119) from the amplified signal at the resonant circuit, comprising L 3 and the 4.7 nF styroflex capacitor. They are then passed via a second comparator in I 3 and converted to CMOS-level at Pt 705.

The DC-voltage from the diode is fed via a filter link comprising 100 kΩ/100 μF and used as control voltage at pin 9 of the TCA 440. Via an internal emitter follower, the control voltage is connected at low impedance to pin 10 (TCA 440), and to Pt 702. This signal is used for indicating when no transmission is being received, and is converted to CMOS-level with the aid of a third comparator in I 3 and fed to Pt 706.

A 100 μA-meter with an impedance of approx. 6 kΩ can be connected to Pt 702 for indicating the field strength.

The heart of the standard frequency generator is the low-noise crystal oscillator using a Colpitt circuit in conjunction with the junction FET T 8. The fundamental crystal Q 4 oscillates in parallel resonance and is coupled to the FET with the aid of the two 220 pF styroflex capacitors. The two capacitors should have approximately ten times the value of the alignment capacitance of the crystal. The author's prototype uses a crystal of 3.1 MHz with a parallel load of 20 pF. The frequency is aligned with the aid of the series capacitance of the crystal comprising 22 pF and 4.7 pF together with 10 pF trimmer at Pt 713. The frequency control via the PLL is made with the aid of varactor diode D 2 (BB 505 G) in series with the 18 pF capacitor at the low-impedance side of the crystal. This has several advantages. The circuit exhibits less noise, the AC-voltage amplitude at D 2 is always less than the DC-voltage amplitude, and the control slope only amounts to approx. 1 Hz/V.

If the receiver is to be often switched on and off, this control slope will be too low, and the transition time will be too long. The PC-board layout therefore provides the possibility of connecting the frequency control to Pt 713. It is necessary for the capacitance values to then be changed correspondingly.

The Schottky diode D 3 is used for amplitude limiting.

The oscillator signal is coupled out at low reaction from the drain of T 8, after which T 9 amplifies it to CMOS-level, and drives the frequency divider in I 6 (MC 14 568 B). I 6 possesses a phase comparator and two programma-

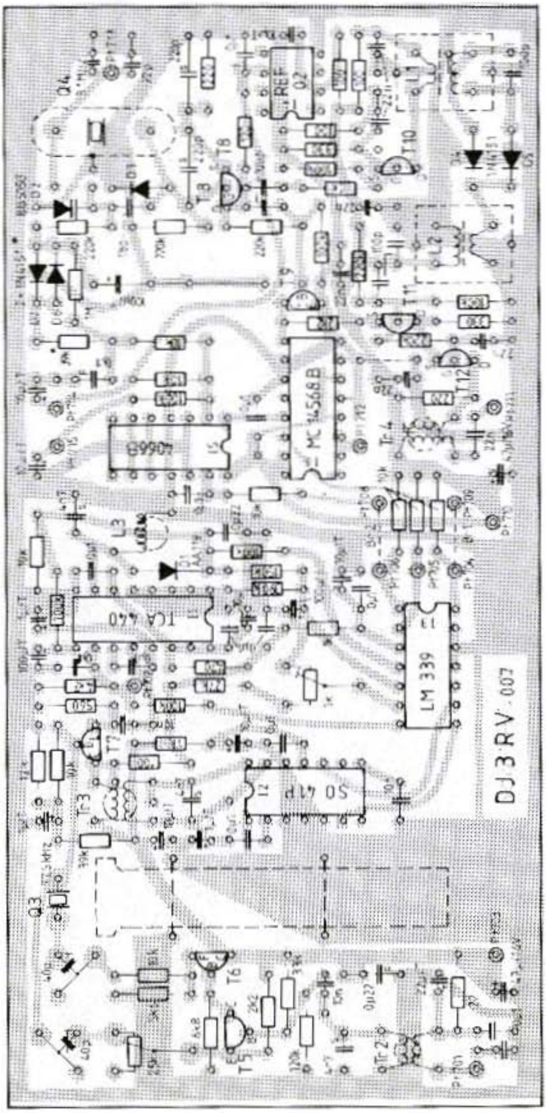


Fig. 9: Component locations on the single-coated PC-board DJ3RV 007

ble dividers. In order to obtain a frequency division factor of 40, it is necessary for both dividers to be connected in series: The first is programmed for a frequency division factor of 1:4, and the second for 1:10.

The output of the frequency divider chain is connected to Pt 712 for alignment, and internally to the phase comparator. The reference signal from the transmitter is fed via Pt 709 and pin 12 to the phase comparator. The phase differential signal is taken from the tristate output (pin 13) and controls the phase of the oscillator with the aid of varactor diode D 2 via CMOS switch I 5 (4066) and the low-pass filter comprising R = 220 kΩ and C = 68 μF.

If no signal is received from DCF 77, the control voltage will disappear at I 1, and switch I 5 will open the control circuit. Diode D 2 is then connected to the reference voltage of +5 V generated by I 4 (REF 02). The oscillator should therefore be aligned so that it has its nominal frequency with +5 V at D 2. A deviation from this value will be indicated by the meter connected between Pt 714 and Pt 715 when the control loop is closed.

The value of the low-pass filter time constant is very important. It should be as large as possible in order to reduce the effects of interference, however, be less than the instability caused by the components. This means that a large time constant requires a high-quality (preferably aged) crystal and the operation of the circuit at a virtually constant ambient temperature.

Too large a time constant will lead to an overshoot of the control circuit on switching on, or after a long period when the transmitter is not received. In the case of the described circuit, a time constant of τ = R x C = 100 s represents the upper limit. The author selected a value of τ = R x C = 220 kΩ x 68 μF = 15 s, which represents a useful compromise. Inspite of this, it will take approximately 3 minutes until the frequency is stable to ± 1 Hz after carrying out a 180° phase jump of the reference signal, which can be simulated by

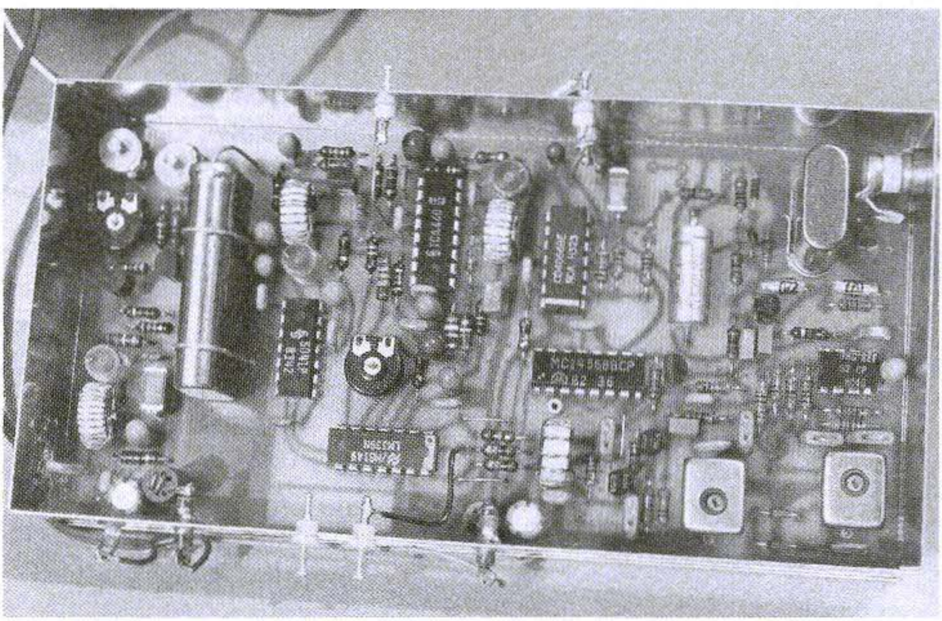


Fig. 10: Photograph of the author's prototype receiver module DJ3RV 007

rotating the antenna by 180°.

This time period can be reduced by connecting the series circuit of the anti-phase diodes D 6 and D 7 and the 39 kΩ resistor, designated with ✻ in the circuit diagram. These will limit large control deviations, and reduce the time constant. Since, however, any short-term phase deviations due to interference will have a large effect on the frequency stability, this possibility was not used in the author's prototype, although provided in the layout.

In order to couple out the oscillator signal with the lowest reaction possible, it is amplified with the aid of T 10 and doubled in frequency using diodes D 4 and D 5. L 1 is tuned to the crystal frequency of 3.1 MHz and L 2 to twice that frequency (6.2 MHz). The signal is amplified with the aid of T 11 and T 12 and transformed to a source impedance of 50 Ω with the aid of Tr 4.

4.2. Construction and Alignment

As can be seen in the component location plan given in **Figure 9** and the photograph of the author's prototype given in **Figure 10**, the receiver and oscillator circuit are accommodated on a common PC-board, which can be incorporated in a metal box in order to ensure that there is no danger of self-oscillation due to feedback to the antenna.

The PC-board can be completely equipped with the exception of the bridges Br 1 and Br 2. The MOS-components I 5 and I 6 should be inserted last; they can be inserted into sockets. I 1 and I 2 must be directly soldered to the board, since a tendency to self-oscillation can occur when using sockets. In order to simplify alignment of the circuit, it is advisable to measure the inductivity of the coils and the transformers before installation. The values can be taken from the following parts list:

Parts list for DJ3RV 007

T 6, T 7, T 9:	BC 550 C (Siemens, etc.)
T 5:	BC 560 C (Siemens, etc.)
T 8, T 10, T 11, T 12:	BF 246 A (TI, etc.) T 11 and T 12 selected to have the same U_{GS} at I_D = 10 mA
I 1:	TCA 440 (Siemens, etc.)
I 2:	SO 41 P (Siemens)
I 3:	LM 339 (National, etc.)
I 4:	REF 02 (PMI, Bourns)
I 5:	4066 B (RCA, etc.)
I 6:	MC 14568 B (Motorola)
D 1:	AA 119 (Siemens, etc.)
D 2:	BB 505 G (Siemens, etc.)
D 3:	2800 (HP, etc.)
D 4, D 5, D 6, D 7:	1 N 4151
Tr 2, Tr 3:	Toroid core R10 N30 (Siemens) prim.: 3.5 turns approx. 0.35 mm dia. enamelled copper wire sec.: 23.5 turns approx. 0.35 mm dia. enamelled copper wire, 0.9 mH
Tr 4:	Toroid core R10 N30 (Siemens) 2 x 12 turns, approx. 0.35 mm dia. enamelled copper wire
L 1:	Special coil kit D 41-2165, colour/orange 2 x 25 turns approx. 0.15 mm dia. enamelled copper wire, 26 μH 8 turns approx. 0.15 mm dia. enamelled copper wire
L 2:	Special coil kit D 41-2165, colour/orange 25 turns, approx. 0.15 mm dia. enamelled copper wire 7 μH 6 turns, approx. 0.15 mm dia. enamelled copper wire

L 3:	Toroid core R10 N30 (Siemens) 23.5 turns approx. 0.25 mm dia. enamelled copper wire, 0,9 mH
Q 3:	77.5 kHz, parallel 40 pF
Q 4:	3.1 MHz, parallel 20 pF
Trimmer:	40 pF plastic foil trimmer 7.5 mm dia. (Philips: violet) 10 pF air-spaced trimmer, Johanson type 5200

All other components not specifically indicated in the circuit diagram, are standard components of the given values.

Resistors: Composite carbon, spacing 10 mm, possibly metal-layer resistors in oscillator and phase-locked loop

Capacitors: Ceramic capacitors with spacings of 5 mm, some for 2.5 mm; in the oscillator circuit, use capacitors with a TC = NPO

The designations of the other capacitors are as follows:
S = Styroflex (polysterene film dielectric)
F = Plastic foil
T = Tantalum

A hermetically sealed type should be used for the low-pass filter capacitor C = 68 μF.

A signal generator, frequency counter, oscilloscope, and a multimeter are very useful for setting up and aligning the module. The following procedure is recommended by the author:

- Connect the operating voltage of + 15 V to Pt 703. The current drain should amount to approx. 25 mA without antenna preamplifier. Check the operating voltage at I 1 and I 2.
- Alignment of the crystal filter.
  Connect the signal generator via a **coupling capcitor** to Pt 701 (Attention: DC-voltage!), and connect the oscilloscope to the emitter of T 7. Align the signal generator to 77.5 kHz and select an input level of approx. 3 mV.

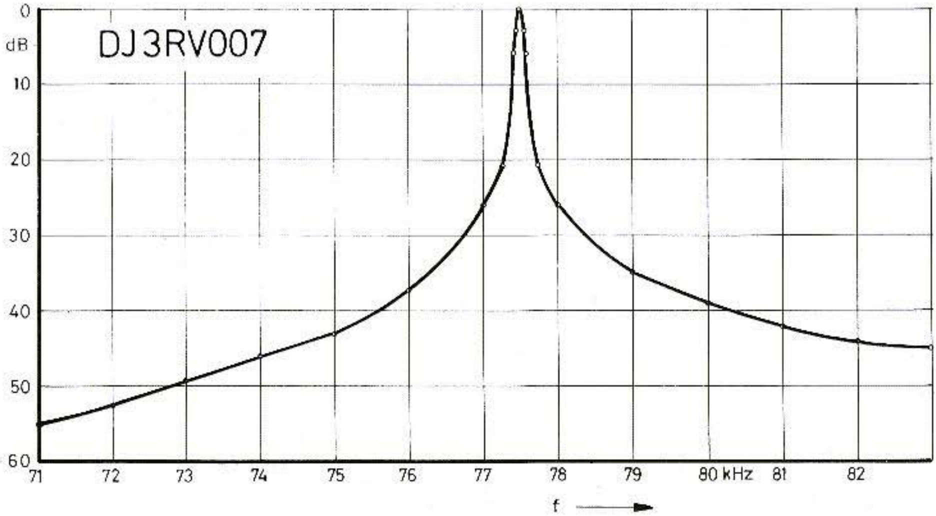


Fig. 11: Frequency response of the crystal filter in the receiver module

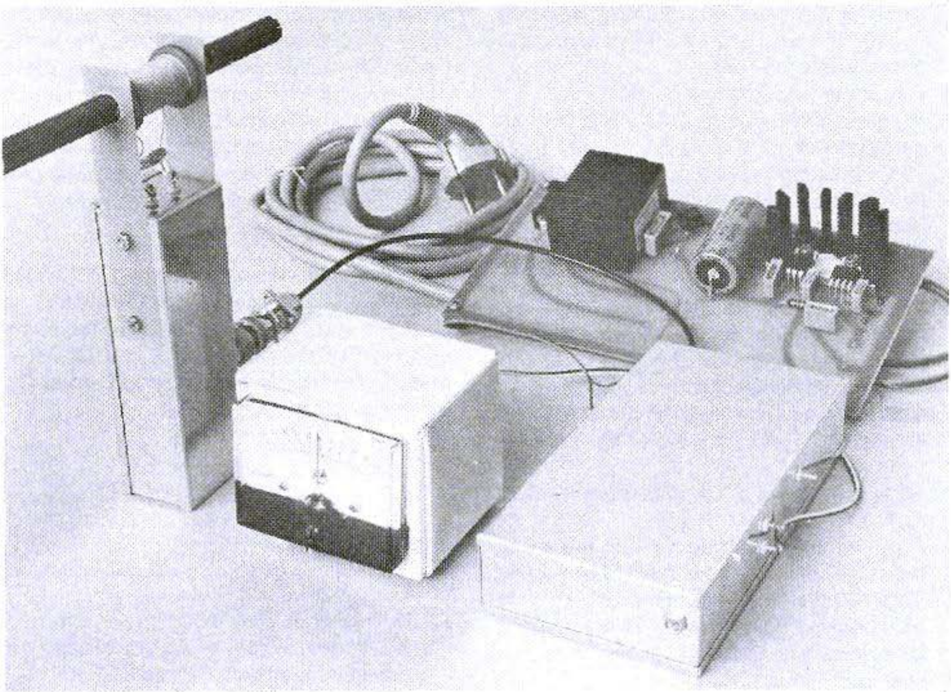


Fig. 12: The complete DCF 77 receiver ready for experimental operation

Align the trimmer in series with Q 3 for maximum level on the oscilloscope. Align the signal generator to 75 kHz, or 80 kHz with an input level of approximately 100 mV, and align the crystal filter with the bridge elements of 25 kΩ and 40 pF for best symmetrical characteristics of the ultimate selectivity. The measured frequency response of the crystal filter including the selectivity of Tr 2 is shown in **Figure 11** for the author's prototype.

- Check the receive level.

  Connect the antenna preamplifier module DJ3RV 006 to Pt 701, and point the antenna to the DCF 77 transmitter. Pay attention to any interference from the oscilloscope!

  The current drain at Pt 703 should be between 50 mA and 75 mA.

A DCF 77 signal of at least 1 mV and maximum 200 mV should be present at the emitter of T 7. A value within this range must be adjustable with the aid of the level control in the antenna amplifier. Too high a value will overdrive I 1 and I 2 and cause phase modulation of the second pulses. Too low a value will reduce the interference suppression. If the receive location is too near to the transmitter, and cannot be reduced sufficiently, it is possible for a resistor to be connected in parallel with the antenna for attenuation. It is also possible to reduce the value of the drain resistor of T 2 and simultaneously to reduce the number of turns of the primary winding of Tr 1, which will ensure a level reduction and increased interference suppression.

- Connect the operating voltage of + 10 V to Pt 710. The current drain should amount to approximately 30 mA.
  Check the following test points:
  A squarewave signal of 77.5 kHz and an amplitude of 10 V must be present at Pt 704, the second pulses at Pt 705, + 10 V at Pt 706, and + 5 V at Pt 714. The signal at Pt 706 indicates that no transmission is being received. By rotating the antenna by 90° (minimum signal), it is possible to simulate no-signal conditions. The 1 kΩ potentiometer should be adjusted so that the voltage at Pt 706 drops to approx. 0 V.
- Alignment of the oscillator circuit
  Check the voltage level at the drain and source of T 8 with the aid of the oscilloscope.
  Align L 1 and L 2 for maximum level at Pt 711.
  Align the crystal frequency with the aid of the 10 pF trimmer capacitor so that exactly 6200000 Hz can be measured at Pt 711. For checking the frequency divider in I 6, it is necessary for 77500 Hz pulses to be present at Pt 712 at CMOS-level.

  If the crystal cannot be pulled to the nominal frequency, it may be necessary for the 22 pF capacitor in series with the crystal to be increased or reduced.

- Finally, insert bridges Br 1 between Pt 704 and Pt 709, and Br 2 between Pt 706 and Pt 708. It is, of course, necessary to switch off the operating voltage and to ground the circuit and soldering iron during this process.

  After switching on the module, the PLL should synchronize the crystal oscillator to the DCF 77 signal within several minutes. This can be seen on the meter connected between Pt 714 and Pt 715 when the needle no longer jumps, but indicates a steady value.

  It may be necessary to align the frequency now and again, until the preliminary aging of the crystal and the circuit have been achieved. Any deviation from the mean value will be seen on the meter.

When using a long coaxial cable between the antenna preamplifier and receiver, it is possible for RF-signals to be injected into the cable from HF and VHF transmitters. This can be suppressed using an RC-link comprising 47 Ω and 2.2 nF connected to Pt 701. This is especially necessary when using an antenna preamplifier as described by (2).

If FETs P 8002 are used in the DJ3RV 006 module, it is necessary for the RC-link to be changed to 22 Ω and 2.2 nF, and the impedance of Tr 2 reduced from 22 to 10 Ω. Otherwise, the voltage drop of the operating voltage will be too great.

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Hans Ulrich Schmidt, DJ6TA

Introduction into Spread Spectrum Technology

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

The spread spectrum technology is practically unknown to most radio amateurs although the theory has been known for fourty years. Spread spectrum technology is used in space technology and increasingly for military applications.

1. INTRODUCTION

Spread spectrum technology is a method of increasing the bandwidth of a radio frequency carrier beyond the bandwidth usually required for it to be transmitted. This may seem stupid for us radio amateurs who are always trying to keep the bandwidth as low as possible for reasons of sensitivity (CW, SSB) or limited frequency spectrum (narrow-band FM). Actually, the opposite method, the spreading of a modulated RF-carrier over the largest frequency range, really does possess a number of considerable advantages:

- Interference signal suppression (increased independence of interference of a two-way communication with respect to wanted or unwanted interference from other sources)
- Reduction of the spectrum energy density, thus less likely to be detected, and reduction of interference from other communication networks
- Secrecy of communications
- Accurate distance measurements.

Of course, these tasks that can be solved using spectrum technology, have already been solved using conventional methods. Interference freedom is usually encountered by increasing the output power; the identification of one's signal, or interference to third parties can be reduced by dropping the output power; the secrecy of the transmission is made with the aid of coding systems; interference and jamming transmitters can be avoided by using a number of previously arranged frequencies and/or previously arranged time intervals. All these methods are used by spread spectrum technology automatically.

The fundamentals of this technology were made in the USA before the Second World War by Shannon (2) as part of his studies regarding "Communications in the Presence of Noise" and was brought before the IRE (Institute of Radio Engineers) in 1940. The first publication of this work was delayed due to the war until 1947 (2). These ideas were not realized rapidly at that time since fast digital circuits, very wideband modulators, and modulation methods, as well as very wideband transmitters and receivers were required. In the fifties, one still had problems obtaining the required useful-modulation bandwidths and one was not able to imagine increasing these bandwidths by a power of ten. In the meantime, the US-military authorities, and several US-companies were working on research prototypes that covered the whole radio frequency range.
The real chance of realizing the spread spectrum technology was firstly to be seen in

the introduction of fast switching transistors from approximately 1960, and the use of integrated circuit technology from 1965. The military authorities in the USA recognized very quickly the value of this method for the realization of interference-free, and secure radio communications, and developed the first operational system.

The actual breakthrough for the spread spectrum technology was, however, not made in the military field, but in space and satellite technology. The Jet Propulsion Laboratory (JPL) of the NASA in Pasadena/California especially studied this method in the sixties for the design of communication methods for interplanetary missions that did not only provide a far higher data security of the information, but also allowed exact distance measurements to be made up to and in excess of the solar system. Nowadays, the various spread spectrum methods are "state-of-the-art" in space technology and are used in the military field for point-to-point telecommunications. This technology is still not used for military multichannel systems, or at least has not gone beyond prototype status.

Radio amateurs had their first contact with this method of communication relatively early in the research program. Costas (2) mentioned "Poisson, Shannon, and the Radio Amateur" in his fundamental IRE-publication of 1959. In this publication, he stated that an excessively full frequency band, such as the 40 m amateur band, could be arranged so that a considerably more effective channellization system can be provided with far more usable telecommunication channels with the aid of spread spectrum technology than could be achieved by using continuously decreasing bandwidths (the SSB-technology was just about to become popular in the shortwave range). However, the technical possibilities of this new technology were not available to radio amateurs at that time.

At the present time, experiments have already been made with the spread spectrum technology by radio amateurs. P.L. Rinaldo, W4RI, mentioned in a QST-article that the AMRAD (Amateur Radio Research and Development Corporation) intended to build up several experimental groups, apply for FCC authorizations, and carry out tests (4). In May 1981 (5) a limited authorization was given to AMRAD on several shortwave frequencies and for a larger number of experiments in the 70 cm band. Recommendations were then made to the FCC in December 1981 (6) to allow all extra class and advanced class licences to carry out spread spectrum experiments on the VHF and UHF bands.

As far as we know, no active experiments have been made by radio amateurs in the spread spectrum technology. However, quite a number of amateurs will suffer passively from such signals, since the AWACS early warning system will be using spred spectrum signals on the L-band (0.9–1.3 GHz), which will increase the overall noise level in the 23 cm amateur band by several dB.

2. FUNDAMENTALS OF THE SPREAD SPECTRUM TECHNOLOGY

The idea of increasing the security of a radio link using an artificially increased bandwidth is a direct result of the examinations made by Shannon (2) regarding the behaviour and the capacity of a telecommunication channel under interference conditions. Using several mathematical simplifications, it is possible for the relationships to be shown as follows:

The following basic magnitudes are used:

Power of the wanted transmitter	S [W]
Power of the interfering transmitter	J [W]
Bandwidth of the telecommunication channel	W [Hz]
Bit-rate of the information	R [Hz]

(for simplicity, binary signals are to be assumed, but the above considerations are just as valid for other analog signals)

The above allows the following magnitudes to be defined:

Power density of the interfering signal	$N_o = \frac{J}{W} \left[\frac{W}{Hz}\right]$
Received energy per bit of the required signal	$E_b = \frac{S}{R}$ [Ws]

The quotient E_b/N_o is without dimension and describes the signal-to-interference ratio and, according to Shannon, is inverse proportional to the bit error rate:

$$\frac{E_b}{N_o} = \frac{S \times W}{J \times R}$$

By rearranging this equation, it is possible for the "jamming margin" to be given; this is the ratio of interference power to signal power by which a certain bit error rate will be maintained:

$$\frac{J}{S} = \frac{\frac{W}{R}}{\frac{E_b}{N_o}} \quad \text{"Jamming margin"}$$

With conventional systems, the bandwidth W of the transmission channel is equal to the bit error rate of the information (e. g. for SSB), or a multiple of this for technical reasons (e. g. AM, DSB). This means that W/R is a constant and thus that the above equation is trivial; the jamming margin is determined by the signal-to-interference ratio.

The new idea of spread spectrum technology is that W/R should no longer be seen as a constant. It is permissible for the bandwidth W of the transmission channel to be artificially wide with respect to the minimum bandwidth determined by the bit error rate R of the information. As can be seen in the above equation, the jamming margin will increase linearly with the ratio W/R with a constant bit error rate.

$$G_p = \frac{W}{R} \qquad \text{"Processing gain"}$$

The "processing gain" is very often given in logarithmic magnitude:

$$G_p' = 10 \log \frac{W}{R} \text{ [dB]}$$

For judging complete telecommunication systems, it is usually necessary to take the internal losses V_{sys}, as well as the signal-to-interference ratio V_D at the demodulator into consideration, by which a certain bit error rate is not exceeded. In this case, the jamming margin is calculated as follows:

$$\frac{J}{S} \text{ (dB)} = G_p' \text{ (dB)} - V_D' \text{ (dB)} - V_{sys}' \text{ (dB)}$$

An example is now to show which increase of the jamming margin is possible:

Information bandwidth:	5 kHz
HF-bandwidth (spread):	50 MHz
Processing gain: $W/R = 10^4$; $G_p' = 40$ dB	
Required S/N at demodulator:	
(e.g. PLL-RTTY demodulator)	+ 1 dB
System losses:	3 dB
Jamming reserve of the system:	36 dB

Due to the spreading of a narrow-band signal (5 kHz) to 50 MHz, the interference signal must be 36 dB stronger (approx. 5000 times) than the required signal (at the receiver input, in the passband range of the spread spectrum receiver), before the demodulation of the required signal is affected.

With the aid of this example, one can see which high spreading rates are usually required in order to obtain satisfactory processing gain values and why this was virtually impossible to obtain in the days of tube amplifiers.

3. SPREAD SPECTRUM PROCESS

The main requirement for realizing the spread spectrum technology are methods of spreading the RF-carrier over a sufficiently large bandwidth. The simplest form of a spread spectrum signal could be a wideband FM-signal with which the interfering signal suppression increases rapidly on increasing the frequency deviation. Such a FM-system should have a processing gain of $G_p \approx 3 \times M^2$ (M = modulation index). However, actual spread spectrum methods are only

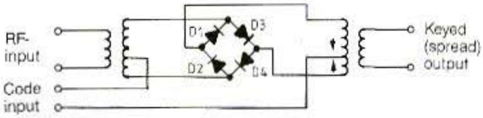


Fig 1: Ring mixer as 0°/180° phase shifter

those by which the RF-carrier is not only modulated by the telecommunication signal, but with a further signal that is not used for communications but only for spreading.

Three main methods are used for this:

3.1. The direct sequence method (DS)

With images of the direct sequence method, the RF-carrier which may have been already modulated, can be increased in width by using a binary phase-shift keying with a certain, suitable binary code **(see Fig. 1)**. A balance mixer is used for the phase-shift keying, such as a four-diode ring mixer, where the IF-drive is made with a constant positive or negative voltage so that it is used as a 0°/180° phase-shifter. **Figure 2** shows the spectra of the individual signals more accurately.

The carrier to be spread is not modulated in our case, and therefore is shown in the frequency spectrum as a single spectral line. The impulse sequence used for spreading the signal has a line-type frequency spectrum whose envelope has the function of $(\frac{\sin \omega}{\omega})^2$. The zero positions correspond to multiples of the clock frequency.

The spectrum of the phase-shift carrier also has a line-spectrum with an envelope of $(\frac{\sin \omega}{\omega})^2$, which is located symmetrically around the original carrier frequency. In order to process this signal, it is usually sufficient for the frequency-band between the first zero points to the left and right of the center frequency to be used so that the minimum required bandwidth is identical with twice the clock frequency of the code sequence.

With sufficiently large spread bandwidths and suitable code sequences, (see Section 4) which have a virtually random character, the RF-power of the original carrier will be distributed virtually constantly over a wide frequency range, and the power density (W/Hz) will have a very low value. When received on a narrow-band (non-authorized) receiver, this signal will be heard as noise only, which will hardly change when tuning over a wider frequency range, and which can be so weak that it disappears into the interference and noise level. As can be seen, such transmissions are very secure and difficult to discover, and they will hardly ever cause interference to other telecommunication links using the same frequency range, due to their very low spectral power density. The prerequisite is, of course, a sufficiently high carrier suppression of the mixer, and good linearity of the amplifier so that the carrier is not regenerated due to intermodulation.

At the receive end, the phase-shift keying of the required carrier is demodulated using the same method (see **Figure 3**).

Of course, very many characteristics of the transmitted spread spectrum signal must be already known at the receive end (authorized receiver):

- The code sequence used for spreading
- The exact clock frequency of the spreading code
- The exact starting time of the code (phase position)

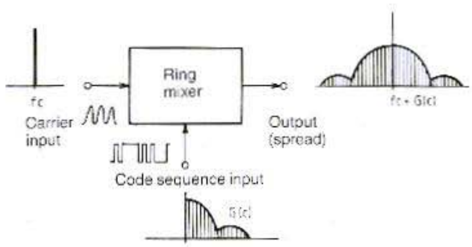


Fig. 2: Direct sequence method (DS); phase keying of a carrier using a PN-code

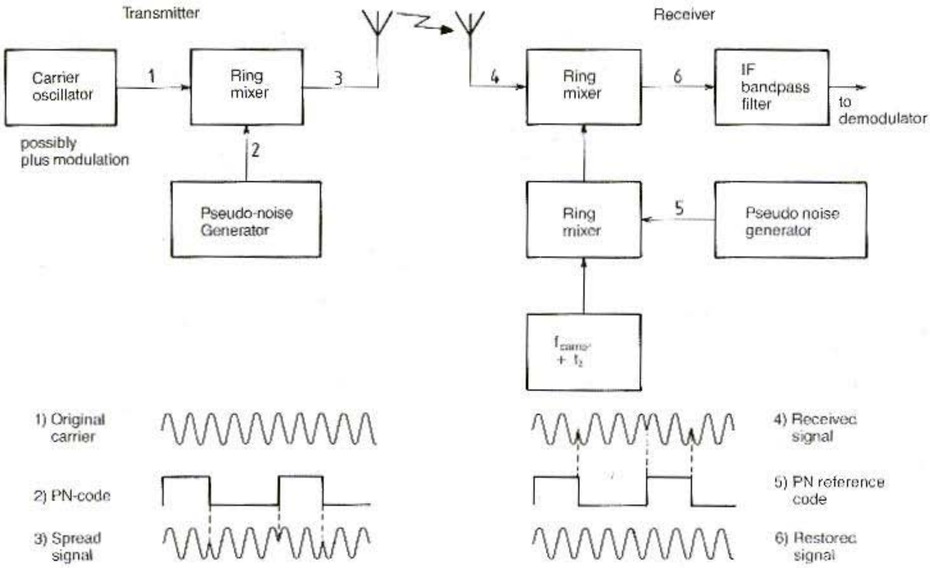


Fig. 3: Construction and operation of a direct sequence system

- The exact carrier frequency
  (It is advisable for carrier and clock frequency to be derived from a standard frequency oscillator at both ends using division or frequency synthesis).

The original, required modulated signal appears now at the output of the receiver-side phase-keyer (often called correlator), after which it can be fed to a conventional, narrow-band receiver for further processing and demodulation. Usually, the ring-mixer used as correlator also serves as first mixer by using an already spread RF-carrier that has been shifted to the value of the IF. The operation is not changed by this.

The behaviour of DS-spread spectrum communications with respect to narrow-band and wide-band interference, and the interference suppression that is achieved in this way is shown clearly in **Figure 4**.

The whole power contained in the spread spectrum signal is compressed in the correlator to a signal bandwidth that is suitable for processing in the subsequent receiver and demodulator. A narrow-band interference signal (such as a continuous carrier) will, on the other hand, be spread by the correlator in the same manner as the required signal in the transmitter, and only a fraction reduced by the value of the spreading factor (processing gain) will fall into the subsequent passband range. In addition to this, the interference signal appears at the output of the correlator as a quasi-noise signal, which means that the further processing in the IF-circuit will only cause a reduction of the signal-to-noise ratio and will not cause a correlated interference. Other wideband signals that are not correlated with the spreading code, will be further spread in the receiver correlator. This means that a far lower power density is present in the processing bandwidth than would be the case with narrow-band interference.

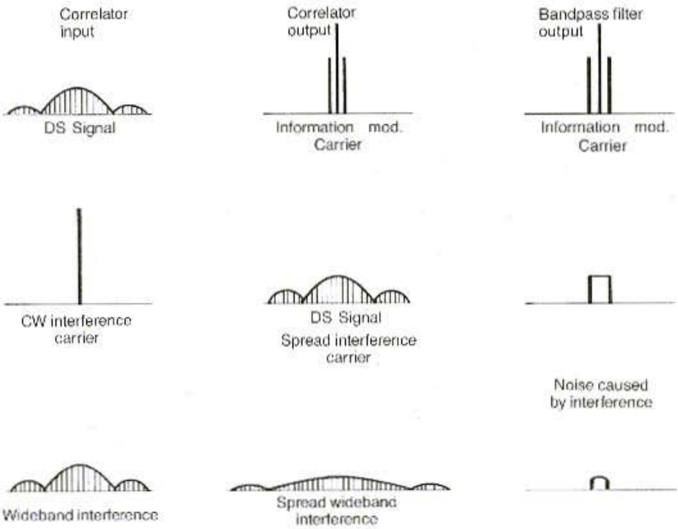


Fig. 4: Behaviour of a DS-system in the case of interference; narrow-band interference (cw), wideband interference (e. g. DS)

Such wideband interference can, for instance, be a second DS-spread spectrum signal having a different code, and this shows that a large number of DS-signals with a differing spread code can be accommodated in the same frequency spectrum. This method allows even a higher number of channels per frequency band than would be possible using conventional, narrowband systems since steep filter slopes, adjacent channel interference, multi-channel intermodulation, and the required safety spacings need not be taken into consideration in DS-systems.

A further advantage of such a channel distribution is that a DS-system is far less affected by overload conditions than a narrow-band channel system. In the case of the DS-system, only the signal-to-noise ratio of all channels will be reduced proportionally to the overload, whereas a narrow-band system will very soon become unusable under overload conditions.

3.2. The frequency hopping method (FH)

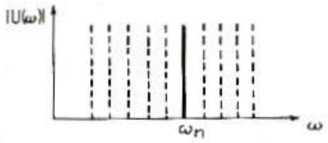
The second method of spreading the frequency spectrum of a transmit signal is to allow the carrier frequency to jump from one frequency to another by selecting one of a large number of available channels. The sequential selection of the frequencies is made according to a PN-code sequence, and will therefore appear to be random in a non-autorized receiver **(Fig. 5)**.

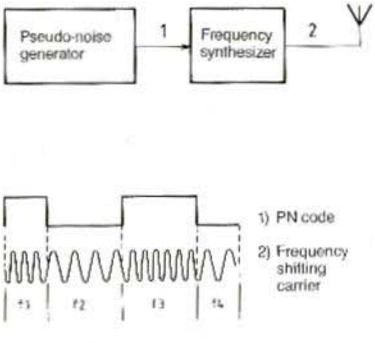
The autorized receiver controls its receive oscillator with the same PN-code sequence and will therefore follow the frequency hopping of the transmit frequency with the correct clock and phase.

This method can easily be realized with the present state-of-the art, and is naturally only possible since fast frequency synthesizers and the required programmable dividers, PLL-circuits and VCOs are available.

The frequency hopping (FH)-method also provides a processing gain in the same manner as with the DS-method. The theoretical gain $G_p = W/R$ was given here as the number N of the available channels:

$$G_p = \frac{W}{R} \rightarrow G_p\,(FH) = N$$





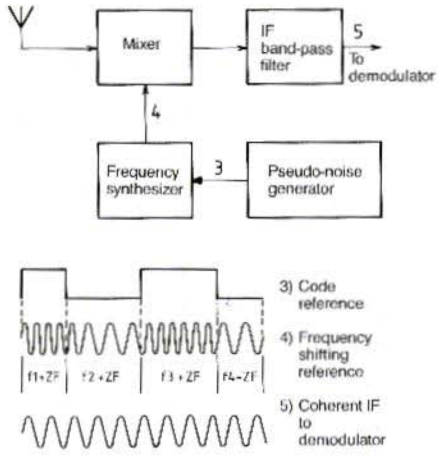


Fig. 5:
Frequency-hopping method
a) Frequency spectrum of a FH-system
b) Construction and operation of a FH-system

In the case of N = 1000 channels between which one jumps back and forth, the processing gain will be $G_p = 10^3$ or $G_p' = 30$ dB.

This can be seen easily in the following description, and digital modulation is again to be assumed for simplicity.

One channel is to be interfered with during these considerations, where the interference power should be greater than that of the required power. In a conventional narrow-band system, the error rate in this channel would then be = 1, or 100%. In the case of a FH-system, the error rate amounts to the following, if one data bit is transmitted per frequency jump:

$$p = \frac{\text{number of interfered channels}}{\text{number of available channels}} = \frac{J}{N}$$

A narrow-band interfering station with J = 1 would result in an error rate of $p = 1 \times 10^{-3}$ in the case of a 1000-channel FH-system with N = 1000. This is already a good value for voice communication; however, it is not sufficient for data transmission. For this reason, one has started to use more than one frequency jump per data bit (e.g. 3 per data bit), and to make a majority decision at the receive end (e.g. 2 from 3).

The possibility of errors can be considerably reduced in this manner. If 'c' represents the number of channels per bit, 'r' the required number of interfered channels in order to interfere with one bit, and 'p' the possibility of error for a normal FH-system, the new possibility of error P is as follows:

$$P = \sum_{x=r}^{c} \binom{c}{r} \times p^x \times (1-p)^{c-x}$$

In the above example with J = 1, N = 1000, $p = 1 \times 10^{-3}$, the possibility of error can be reduced to the following with c = 3 channels per bit and a decision of two from three:

$$P = \binom{3}{2} \times p^2 \times (1-p)^{3-2} = p^2 \times (3+p) \approx 3 \times 10^{-6}$$

3.3. The Chirp Method

The chirp method (pulse-FM) was mainly used in radar technology and not for telecommunications. However, it is also classed as one of the spread spectrum methods since the transmit spectrum is considerably spread using a special method. In order to complete our considerations, this is to be mentioned briefly in this article (**Figure 6**).

In the case of this method, the transmit frequency of a (radar) transmitter is continuously shifted during a keying time Δt by a frequency f_1 to a frequency f_2 (e.g. linearly with a sawtooth signal). The pulse transmit power is spread over the largest possible time range and frequency spectrum, and the mean power density will be reduced considerably. At the receive end, this frequency modulation is restored using a so-called dispersive filter with which the delay is dependent on the frequency. The power components of the different frequencies transmitted at differing times arrive simultaneously at the output; this means that a pulse compression occurs in the dispersive filter. This is a very useful effect for increasing the distance resolution of radar equipment.

In the case of a "compression ratio" $D = \Delta t \times \Delta f$, the signal-to-interference ratio will increase by $\sqrt{D}$, and the time resolution (= distance resolution) will be improved to $\Delta T = \Delta t / D$.

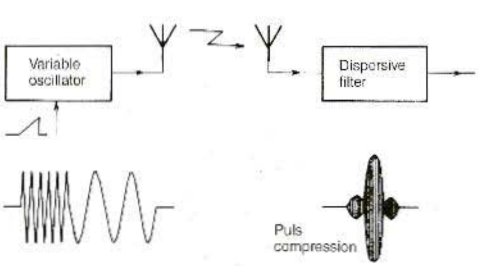


Fig. 6: Chirp method (radar technology)

3.4. The Time-Hopping Method

In publications, the so-called time-hopping method is often classed as belonging to the spread spectrum methods, although, when used on its own, it does not offer hardly any advantages. The principle is that the telecommunication transmitter does not transmit continuously, but at certain time intervals that are determined by a PN-code sequence. This method also considerably widens the spectrum of a transmit signal. It is sometimes used together with a DS or FH-method and can improve the characteristics of these systems.

3.5. Combination of the Various Spread Spectrum Methods

Very often, several spread spectrum methods are used together in order to improve the system characteristics further, or to satisfy certain technical demands on a system with more simple means.

In the case of a combined DS/FH-system, the direct sequence and the frequency-hopping method are used simultaneously. The processing gain is equal to the product of the gains of the individual methods:

$$G_p^{tot.} = G_p^{(DS)} \times G_p^{(FH)} = \frac{W}{R} \times N$$

or logarithmically:

$$G_p^{'tot.} = G_p^{'(DS)} + G_p^{'(FH)} \text{ [dB]}$$

This combination is of interest when a certain processing gain cannot be obtained using a single method on its own, or only with difficulties.

If, for instance, a gain of 47 dB is required at a data rate of R = 10 kHz for interference suppression, it will be necessary to spread the signal over a bandwidth of W = 500 MHz in the case of a DS-system, which will require a PN-clock frequency of 250 MBit/s. In the case of a FH-system, it will be necessary to switch between 50000 channels. Both possibilities are at the limits of what is technically possible. If, on the other hand, the direct sequence and frequency-hopping method are combined, it is possible using a DS-code of 10 MBit/s and a

50-channel frequency-hopping system to obtain the required processing gain of 47 dB, and this is possible with normal means.

In spread spectrum communication networks with a multiple access, the DS or FH-method is often used together with the time-hopping method (TH). In this case, the individual stations only transmit at several time intervals, which are determined by a common PN-code. This allows one to achieve the condition that only one station is transmitting at any particular time. Even though spread spectrum signals can be operated quite well in the same frequency band, it is possible for local transmitters and receivers to be affected by unwanted desensitization.

4. CODING AND SYNCHRONIZATION

4.1. Code Characteristics

The demands made on a spread spectrum system can only be achieved when the spreading code possesses certain characteristics:

One-Zero Balance (High-Low Balance)

The number of 1 and 0 bits in a code should, if possible, be identical. This allows the DC-current component of the pulse sequence to be zero. The DC-component of a DS-system will cause the ring mixer to be switched through for a certain percentage of time which means that the original carrier will appear in the output spectrum in a noticeable manner (inferior carrier suppression). In the case of a FH-system, this would cause one channel to be used in preference to all others, and be accentuated in the output spectrum. In both cases, this will mean that the spectrum is no longer "noise-like", which means that the carrier can be recognized, can be interfered with, or can even interfere with itself.

One-Zero Distribution
(High-Low Distribution)

The distribution of the 1 and 0-bits within a code also has a large influence on whether the output spectrum appears "noise-like". The 1 – 0 distribution should therefore also be as statistic as possible, although it is, of course, generated using a predetermined method. This is often called "pseudo noise" (PN).

Low Code Repetition Rate

The frequency with which a "pseudo statistic" sequence of 1 and 0 bits repeats itself should not fall into the required information (modulation) frequency range, since this could cause interfering effects. Normally, it is placed below the lowest modulation frequency. Since, on the other hand, the clock frequency of the code must be very high, this results in very long codes.

Good Auto-Correlation Behaviour

The code must be arranged so that it can be identified well by the authorized receiver (auto-correlation). Attention must be paid that a similar code cannot cause a positive identification of the decoder circuit (even when they will be weaker).

Good Cross Correlation Behaviour

When fed into an identification circuit that compares it to another code (cross correlation), the received code should not produce an output signal. The auto and cross correlation behaviour of the codes will be discussed further.

4.2. Code Generation

The given demands are fulfilled by a number of so-called "maximum" codes. All other codes are called "non-maximum codes". The generation of such codes is made most easily using shift registers where the input of such a shift register is provided with the binary sum from the output signal, and the signal itself.

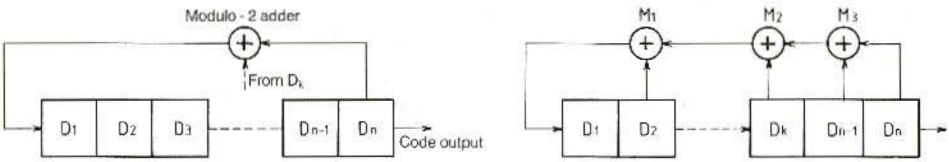


Fig. 7: PN-code generators using tapped registers

Due to the selection of the tapping point, various different codes can be generated (**Figure 7**).

Very often, several adders are used together with several tapping points, or combinations of signals from several shift-register generators. With the aid of microprocessor technology, it is possible for other methods and principles of generator circuits to be used for any required codes.

4.3. Auto-Correlation and Cross Correlation

In order to understand the receiver-end identification of the transmitted spread code, it is necessary to describe the terms "auto-correlation" and "cross correlation" in more detail. These correlation functions are a measure of the similarity of functions. In the case of the auto-correlation function

$$\varphi_A(\tau) = \int_{-\infty}^{+\infty} f(t) \times f(t-\tau)\, dt$$

a function f(t) (e.g. a time-dependent function such as sin ωt) is compared with itself, however, shifted by a magnitude τ (e.g. shifted in the time plane). This is achieved by multiplying all functional values with the associated values that have been shifted with the value of τ, and for all these products to be added (integrated) for all values of t. This function, of course, has its maximum at $\tau = 0$, which means that a function is most similar to itself without phase shifting. In the case of periodic functions, further maxima appear when τ is a multiple of the period.

The behaviour of the correlation function at other values than $\tau = 0$ determines how good the original function f(t) can be found again by variation of τ.

It is also possible using the same method to compare various functions f(t) and g(t) with the aid of the correlation function:

$$\varphi_K(\tau) = \int_{-\infty}^{+\infty} f(t) \times g(t-\tau)\, dt$$

This function is called cross correlation function. Since the functions to be compared are different, $\varphi_K(\tau)$ may never achieve the maximum value of $\varphi_A(\tau)$. The non-exceeding of a certain thresold is a sign that the functions are different.

In the case of the correlation of binary code sequences, the product in the above equations will obtain the value + 1, if the functional values coincide, and it is agreed that the value − 1 should be obtained (actually = 0), if they do not coincide. The integration then forms a summing of all bits of the code. The correlation value for a certain phase-shift can therefore be simply calculated by placing the bits over another and comparing then bit by bit. In this case, the correlation value is the sum of the coincidence or non-coincidence.

This is easily explained with the aid of an example. The maximum code sequence 1110010 is to be compared with itself in the seven possible phase shifts (**Figure 8**). The auto-correlation value is always − 1, except for the case of coincidence, where it is a maximum (number of elements). The amplitude of the maximum increases therefore with the length of the code, which improves the discrimination with respect to other codes (cross correlation).

Shift	Sequence	Agreements	Disagreements	A – D
1	0111001	3	4	– 1
2	1011100	3	4	– 1
3	0101110	3	4	– 1
4	0010111	3	4	– 1
5	1001011	3	4	– 1
6	1100101	3	4	– 1
0	1110010	7	0	7

Fig. 8:
Auto-correlation functions A–D for a maximum 7 Bit code

In the case of non-maximum codes, side maxima appear in the auto-correlation function (see **Figure 9**) and it is then important that a sufficiently large spacing exists between the main maximum and the side maxima at all times. Inspite of these disadvantages, non-maximum codes are often used, in order to exploit their advantages, as for instance their easier synchronization.

The time slope of the auto-correlation function is of special importance. The coincidence can be determined with an accuracy better than ± 0.5 Bit. This allows one to determine the delay times of spread spectrum signals using the phase-shift, and thus to determine distances extremely accurately. At a code clock rate

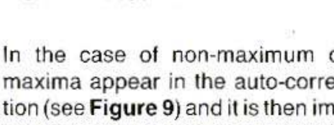


Fig. 9a:
Auto-correlation function for a maximum code

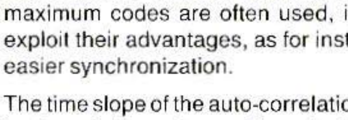


Fig. 9b: for a non-maximum code

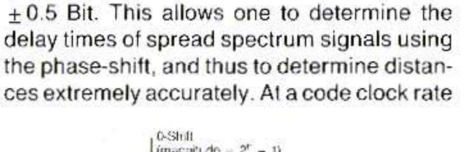


Fig. 9c: Auto and cross correlation for a 31 Bit code

of 100 MBit/s, it is possible to determine a distance resolution of better than 3 m!

The range of clear identification of the distance measurement is determined by the time-length of the code sequence. If this length is great enough, i.e. is always greater than, for instance, the duration of a space mission, this will mean that the distance measurement is always clearly defined. Which lengths are required for this and how many shift-register cells are necessary to generate it, can be seen in **Figure 10**.

One of the first large applications of spread spectrum technology was made due to this exact measuring possibility in the "Deep-Space Mission Program" of NASA.

The principle of the distance measurement is based on the measurement of the phase-shift of the comparison code on the receive end that is necessary in order to coincide with the originally transmitted code, and to result in maximum auto-correlation.

In many cases, it is not at all necessary for the receive end (e.g. space vehicle, satellite) to have its own spread spectrum installed. The whole spread spectrum signal can be converted to another frequency in a transponder and retransmitted to the transmit (ground) station. The evaluation can be carried out here much easier, since the spread code is already available, and no synchronization problems will appear.

The same method is used for determining the distance of METEOSAT, as well as the distance of the new amateur radio satellite OSCAR 10 from the command station on the ground (7).

Register Length n	Sequence Length	Sequence Period
7	127	1.27×10^{-4} sec
8	255	2.55×10^{-4} sec
9	511	5.11×10^{-4} sec
10	1023	1.023×10^{-3} sec
11	2047	2.047×10^{-3} sec
12	4095	4.095×10^{-3} sec
13	8191	8.191×10^{-3} sec
17	131071	1.31×10^{-1} sec
19	524287	5.24×10^{-1} sec
23	8388607	8.388 sec
27	134217727	13.421 sec
31	2147483647	35.8 min
43	879609302207	101.8 days
61	2305843009213693951	7.3×10^{4} yr
89	618970019642690137449562111	1.95×10^{9} yr

Fig. 10: Code length and period duration at a bit rate of 1 MHz

To be concluded in edition 3/1984 of VHF COMMUNICATIONS.

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 2/1984 of VHF COMMUNICATIONS

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 connectors (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,—

DJ4LB	**Control circuit with 4 time steps for transmit-receive switching**			**Ed. 2/1984**
PC-board	DJ4LB 009	72 mm × 72 mm, single-coated, silver-plated, undrilled	6838	DM 16,—
Components	DJ4LB 009	2 lin. ICs, 4 power-switching transistors, 3 switching, 4 zener and 5 rectifier diodes, 1 feedthru, 6 ceramic and 1 foil cap., 35 resistors, 5 PTFE-feedthru, 1 tinned-metal case 74 × 74 × 30	6839	DM 53,—
Kit	**DJ4LB 009**	**complete with above parts**	**6840**	**DM 67,—**

A System for Reception and Display of Weather-Satellite Images from METEOSAT/GOES, NOAA/METEOR

A) A complete system as kits

Description	Edition	Kit designation	Art. No.	Price DM
Parabolic antenna, 1.1 m diameter, 12 segments to be screwed or riveted together, 3 plastic supports for radiator, mast-mounting parts with elevation mechanism	3/1979	**Set of 12 segments**	0098	**180.00**
		Riveting machine + rivets	0105	**93.00**
		1.7 GHz Cavity radiator kit	0091	**90.00**
		3 radiator supports	0106	**29.00**
		Mast-mounting parts	0107	**85.00**
Low-noise amplifier for 1.7 GHz (Originally described for use at 2.4 GHz, this unit is tuned to 1.7 GHz)	1/1980	**DJ6PI 010**	6565	**225.00**
METEOSAT Converter, consisting of two modules – Output first IF = 137.5 MHz)	4/1981	**DJ1JZ 003**	6705	**189.00**
	1/1982	**DJ1JZ 004**	6714	**185.00**
VHF Receiver, frequency range: 136 – 138 MHz, Output: 2400 Hz sub-carrier	4/1979	**DC3NT 003**	6141	**225.00**
	1/1980	**DC3NT 004**	6145	**80.00**
Digital scan converter (256 × 256 × 6 Bit)	4/1982 1/1983	**YU3UMV 001** **YU3UMV 002**	6736	**675.00**
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,–**
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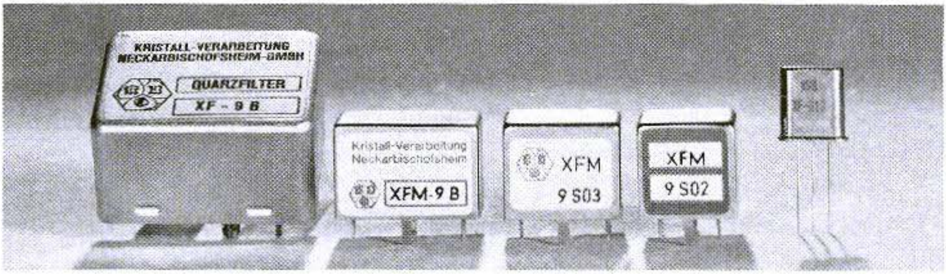

Case: 1 15 14 13 17

DISCRETE CRYSTAL FILTER	Application	MONOLITHIC EQUIVALENT with impedance transformation: Type	Termination	Case	without impedance transformation: 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

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