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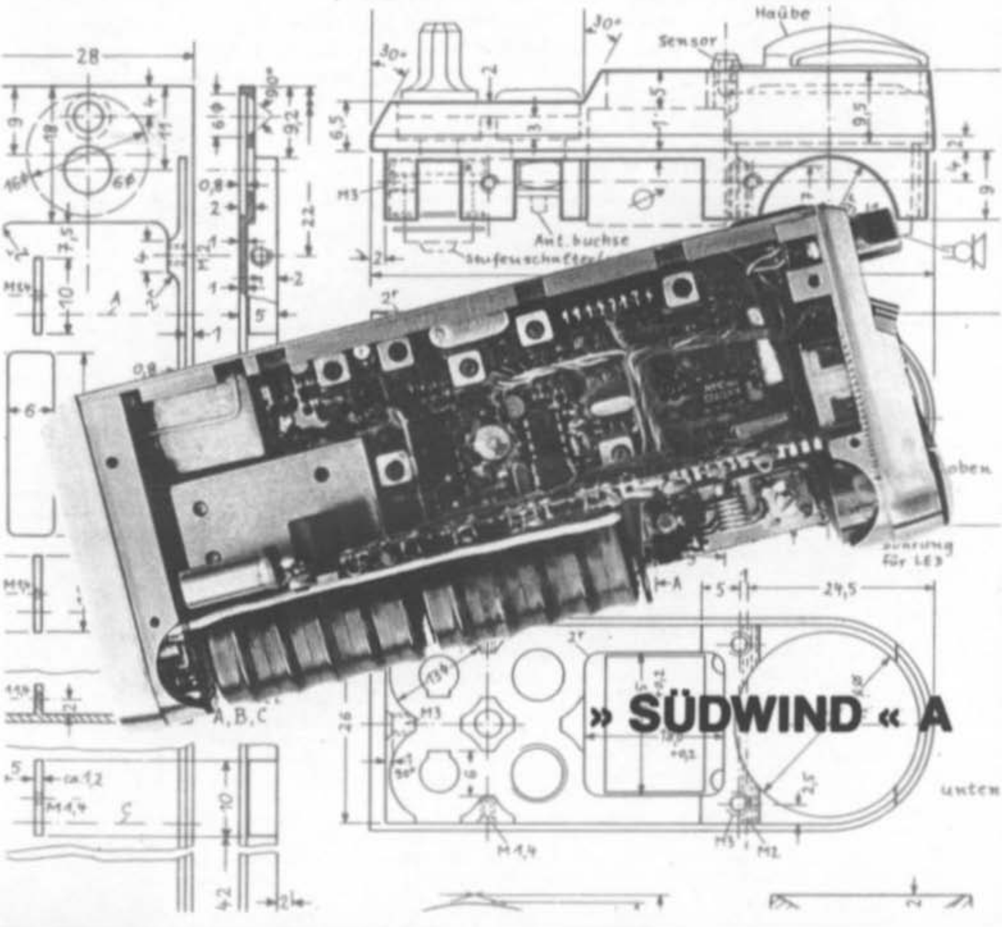

» SÜDWIND « A

VHF COMMUNICATIONS

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A PUBLICATION FOR THE RADIO AMATEUR
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VOLUME No. 10 WINTER EDITION 4/1978

This is the last edition of the tenth volume of VHF COMMUNICATIONS. The editors hope that they have been able to pay their contribution to the growth of VHF and UHF communications in these last ten years, as we intend to do in the next ten. We would be grateful if you were also able to forward our publication, since the number of subscribers has not increased very much since the commencement. We feel that there are many amateurs that do not know VHF COMMUNICATIONS, and it would be a pity for the magazine to stop due to lack of support.

We would like to remind you that we keep on reprinting the back volumes of VHF COMMUNICATIONS, and all back copies back to 1970 are still available. We can offer these volumes at the following prices:

Any three volumes	DM 36.—
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Our best wishes for 1979 to all of our subscribers.

»SUEDWIND« – A 2 m FM HAND-HELD TRANSCEIVER WITH 80 OR 396 CHANNEL SYNTHESIZER AND TOUCH-KEY OPERATION
Part 1: Circuit Description

by J. Becker, DJ 8 IL

1. CONCEPT

It is extremely difficult to maintain stable frequency control in handheld (Walky-Talky) transceivers due to the large variations in ambient temperature, and in their liability to mechanical shock. This means that crystal control must be used, especially in the case of miniature equipment. If more than a few frequencies are to be used, this would mean that it will be necessary for a synthesizer to be used to generate a large number of frequencies having a constant channel spacing from a single master crystal. The selection of the required frequency is made digitally. If operation is to be made over the whole frequency band and the required frequency should be selectable within a few seconds, it is necessary to use a synthesizer having a very fine channel spacing, in other words having a large number of channels.

The synthesizer of the »SUEDWIND« transceiver to be described in this and the next edition of the VHF COMMUNICATIONS is switchable in 5 kHz steps. Such a 400 channel transceiver for the FM mode is just as versatile as a FM transceiver equipped with a VFO. The bandwidth of a FM receiver ensures that signals that are 2 to 3 kHz from the center frequency can be demodulated correctly. In the case of the previously mentioned 5 kHz steps, the maximum frequency error to the partner station will be a maximum of 2.5 kHz.

The synthesizer is equipped with C-MOS integrated circuits, which ensures a low current drain. Another feature are the touch-keys for the most important functions such as transmit-receive switching, calling tone, frequency change up or down, channel indication and squelch. Furthermore, special attention has been paid to the modulation quality and the external appearance of this very handy transceiver. The S-Meter has a virtually dB-linear relationship in a range of 66 dB, which means that the receiver is also suitable for fox hunts (DF).

The following specifications show the results that were achieved with this concept in practice.

1.1. Specifications

Output power at 12 V DC:	0.95 W
Overall efficiency:	35 %
Current drain on transmit:	225 mA
Current drain on receive: (LED-readouts approx. 25 mA)	40 mA
Squelch sensitivity:	0.08 µV
Sensitivity for 10 dB (S+N)/N: (4 kHz deviation)	0.20 µV
Input voltage for 20 dB (S+N)/N: (4 kHz deviation)	0.45 µV
S-meter virtually logarithmic in the range of:	0.1 to 200 µV

Suppression of harmonics and spurious signals:

Adjacent channels and within band:	> 65 dB
In the range of 0 to 1.2 GHz:	≧ 46 dB
Dimensions in case A:	184 mm x 74 mm x 28 mm
Weight with case A and λ/4 telescopic antenna:	680 g
Dimensions in case C:	179 mm x 75 mm x 28 mm
Weight with case C:	600 g

The following can be connected:

External supply with 10.5 to 16 V; electronic voltage division and stabilization built in. Microphone for mobile operation, with remote control of PTT, calling tone and frequency selection.

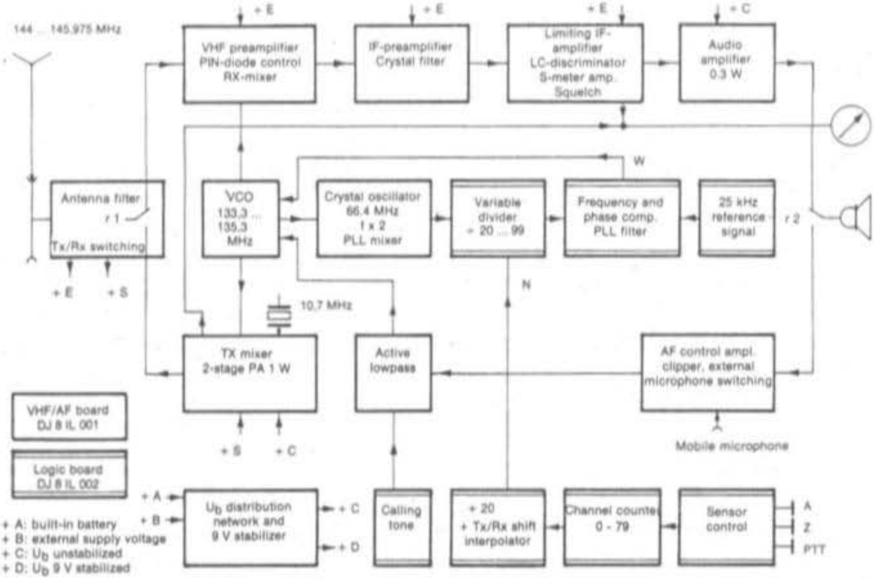


Fig. 1: Block diagram of the »SUEDWIND« FM transceiver above: receiver – center: PLL oscillator bottom: transmitter, modulator, auxiliary stages

2. CIRCUIT DESCRIPTION

2.1. Receiver

As can be seen in the block diagram given in **Figure 1**, the receiver is a single-conversion superhet with a 10.7 MHz IF and LC-discriminator.

This has advantages over a double-conversion superhet, having a better large-signal behaviour and lower number of circuits; however, it has the disadvantage in that the limiting IF-amplifier at present requires a greater current drain. This is, at least, valid for those integrated circuits that are able to provide a reliable S-meter signal that increases approximately logarithmicly – in other words dB-linear – together with the received field strength. The distortion of the single conversion superhet is very low, as only one single high-quality monolythic crystal filter is used for narrow-band selectivity.

In contrast to the professional technology, a crystal discriminator is not used, but a more simpler LC-demodulator. Measurements have shown that the signal-to-noise ratio of the demodulated audio signal is not better when using a more extensive crystal discriminator. In fact, when using a crystal in the discriminator circuit, it will no longer be possible to use a field strength dependent squelch and the S-meter characteristic of the integrated IF-amplifier.

The dynamic range of the RF-circuit was extended by 43 dB, using a four-stage control. A limitation of the signal previous to the crystal filter, which will cause intermodulation products only occurs at input voltages in excess of 0.25 V.

2.2. Transmitter

Two concepts were also discussed for the transmitter: direct generation of the required frequency using two separate oscillators (VCO) for transmitter and receiver, or conversion of the transmit frequency from the common VCO and the intermediate frequency. In the first method, the control signal for the transmit amplifier chain will not receive any unwanted conversion products that can appear with a spacing of $\pm$ n x f_{IF} of the required signal and their harmonics. However, the reason why the second method was still used in the »SUEDWIND« transceiver is because the harmonics generated in the transmit amplifier chain and their conversion products appear in the output spectrum of the transmitter at far higher levels, and because the filtering of these also allowed the spurious signals in the vicinity of the required signal to be also suppressed sufficiently. If a miniature construction is also required, the room required by the complex VCO module with respect to an integrated self-excited transmit mixer is an important consideration. The latter offers at least a power gain of ten times, which means that the VCO need not be designed for a high power output, but exclusively for a low residual frequency deviation (1).

2.3. Special Features

New developments in the »SUEDWIND« circuitry are the RF gain control in the receiver, the modulator having a low-distortion, dynamic control of 30 dB, and subsequent balanced limiting, as well as a digital interpolator in the frequency selection logic. This multiplies the number of channels of the synthesizer from 80 channels when using a 25 kHz spacing to 400 channels when using a frequency spacing of 5 kHz. The high comparator frequency of 25 kHz is maintained in this case which results in a correspondingly short transient time of the frequency selection on changing channels (repeater shift). A conventional 80-channel synthesizer can be obtained by deleting two integrated circuits.

2.4. Operating Values

The operating points and amplitudes are given extensively in the circuit diagrams (Figures 3 and 4). They do not only simplify the functional check and alignment during the construction, but also allow comparisons to be made when designing the different circuits. For this reason, the text is to give details regarding positions that are critical to design and adjust, as well as mentioning possible alternatives and variations of the individual modules together with measured values.

3. MECHANICAL CONCEPT

Before discussing the circuit in more detail, this short section is to give an impression of the mechanical construction of the transceiver.

3.1. PC-Boards

The electronic modules are distributed onto two PC-boards so that one accommodates all C-MOS integrated circuits with associated RC-circuits, as well as the 25 kHz reference source; all other stages are accommodated on the RF/AF board. This separation makes it easy to carry out modifications on the logic portion at a later date, for example, when new interesting LSI-circuits become available on the market. The actual RF-part of the circuit remains intact and is ready for experimentation using different types of logic. Two interconnection plans are provided to ease wiring.

3.2. Case

Three designs (A, B, C) have been made for the case. All of these use the same basic construction using a C-shaped chassis, onto which the operating controls are mounted, and a U-shaped cover. The RF/NF board is screwed to the chassis frame, which also accommodates the NC-battery. The logic board, on the other hand, is mounted on a threaded bar so that it can be folded out when ready for operation. This means that both boards are accessible from both sides.

The three different types of case differ in the method of making the chassis frame. They are given more as examples so that the reader takes the design of such a transceiver into consideration. In the case of models B and C, the same PC-boards are used. With version A (Fig. 2a), the antenna filter, RF-output rectifier, and DC-voltage circuit are accommodated separately on a third, small PC-board at the base of the transceiver. Furthermore, a different relay (Siemens Minipol) is used. The framework for cases A and B only be manufactured in a workshop (approximately two days work for type B.) The author is willing to provide a copy of the workshop drawing in exchange for an addressed envelope with IRC.

In the case of type C, which is to be described in part 2 of this description in detail, it is possible for the case to be home-made. A plastic case is not to be recommended, since the squelch sensitivity of the receiver will deteriorate.

3.3. Operating Controls and Connections

Most of the operating controls are accommodated on the upper side of the »SUEDWIND« transceiver: the microphone-loudspeaker (right), combined S-meter and RF-output meter, antenna connector, volume and on/off switch as well as the switches for repeater-shift and interpolation. The two mounting screws on both sides of the meter (Fig. 2b) are also touch-keys (sensors) for the frequency selection, calling tone, squelch, and channel indication. With the case types A and B, they are to be found to the right of the S-meter.

Fig. 2a:
Photograph of the author's prototype (Case type A)

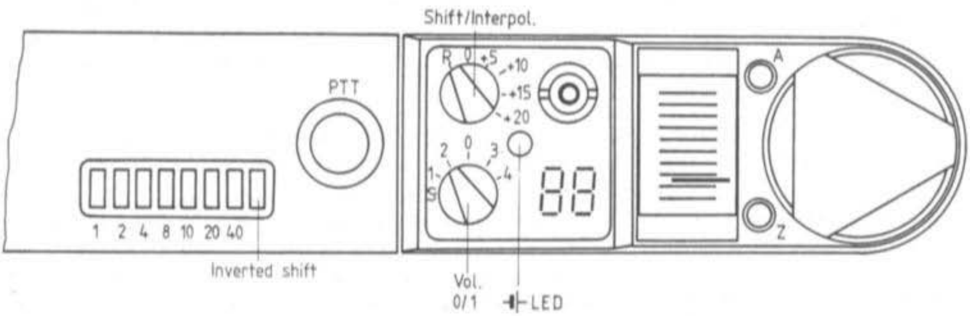


Fig. 2b: Diagram of the sensor keys and the shift-selector switch

The frequency is indicated as channel number from 00 to 79 using two 7-segment LED read-outs. They are located in the free area in front of the antenna connector and are provided with an optical filter to increase contrast.

A sensor key is also provided at the top of the left-hand side panel for the PTT-switch; an 8-position miniature set of switches is mounted 3 cm below this which allows the repeater shift to be selected.

A connector is provided on the rear panel of the unit for mobile and stationary operation. Connections are provided for a mobile antenna (50 Ω), or a power amplifier, for external power supply, charge current of the accumulator, AF-output, external microphone, and remote control of the three sensors.

4. CIRCUIT DETAILS (Fig. 3 and Fig. 4)

4.1. Frequency Selection and Operation of the Sensor Keys

The 80 main channels with a frequency spacing of 25 kHz cover the European band of 144.000 MHz to 145.975 MHz corresponding to channels 00 to 79, and can be selected in a slow or fast mode upwards or downwards. On switching on the transceiver, the priority channel, which is selected with seven bridges on the selector switch (2 x 4029, see Fig. 3) will be immediately available. The two 7-segment channel readouts will only light up (when the built-in accumulator is in operation) if one or both sensor keys adjacent to the S-meter are touched. They will go out again two seconds afterwards in order to save battery current.

If the rearmost sensor A is touched, the squelch will switch alternately off and on afterwards. In the case of the foremost sensor Z, the frequency selection process will be activated at a speed of two channels per second. This speed is too slow if a large part of the band is to be covered; for this reason, a fast speed of 10 channels/sec is provided when sensor A is touched at the same time.

If one has passed the channel, it is not necessary to retune over the whole band: it is only necessary to briefly release the Z-sensor and the channel selector will run in the opposite direction on reactivating.

The frequency selection is blocked in the transmit mode. Instead of this, a calling tone of 1.75 kHz will be radiated on depressing the Z-sensor key. The A-sensor retains its function of activating the channel readout.

For repeater operation, it is possible for the receive frequency to be shifted in an upward direction by n x 25 kHz. This is done by entering the required shift as n = 0 - 79 as a BCD-digit using the lower 7 of the 8 miniature switches on the side panel of the transceiver. If the upper switch is on, the shift will be added in the transmit mode, which means that the transmit and receive frequencies are inverted. For normal repeater operation, the eight switches are set to 00100100, which can be easily remembered, and corresponds to a frequency shift of 24 x 25 kHz = 600 kHz. If frequencies in excess of the upper band limit of 146 MHz result in error during the frequency selection, the logic circuit will shift these down by 2 MHz back into the amateur band. The LED-readouts will always indicate the actual transmit or receive frequency.

More to the operation of the shift and interpolation switch: In the »R« (Repeater) position, the shift will be operative as described; in all other positions, the transmit and receive frequency will be identical (simplex). In positions R and 0, operation will be made on the 80 main channels, where 99 % of the 2 m operation takes place. In the interpolation positions + 5 to + 20, the carrier frequency is shifted up by 5, 10, 15, or 20 kHz. However, the LED readout is hardly usable, since the digits of the two neighbouring channels are indicated one above the other. When using the described logic circuit, it is not possible to interpolate in excess of 145.975 MHz, corresponding to channel 79; for this reason, the transceiver does not have 400 channels, but only 396.

Usually, it only takes about ten minutes to get used to the operation of the shift/interpolation switch and the two sensors A and Z. After this, it is usually possible for any required frequency to be selected in less than 10 s, even in the dark.

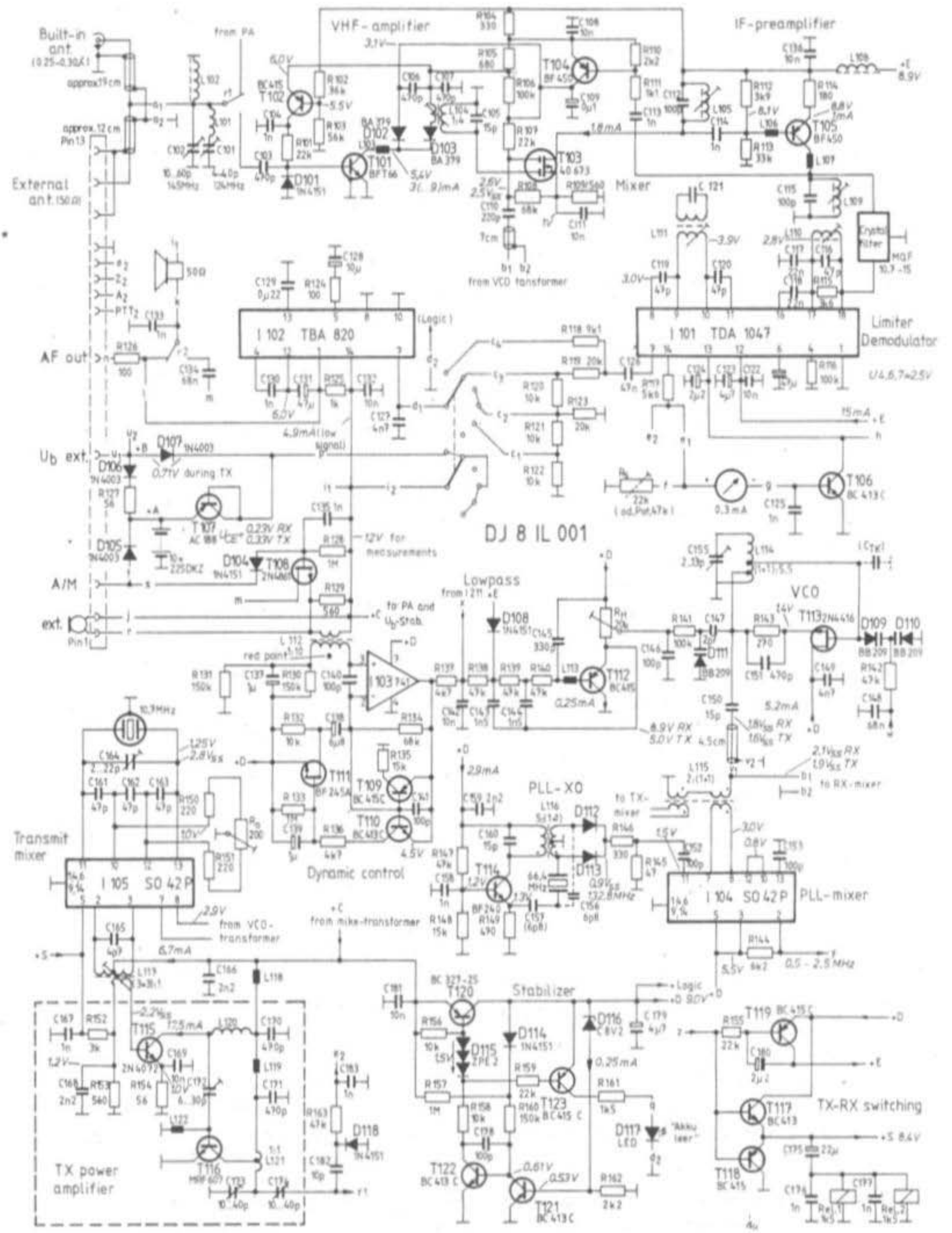


Fig. 3: Circuit diagram of the VHF and AF circuits DJ 8 IL 001

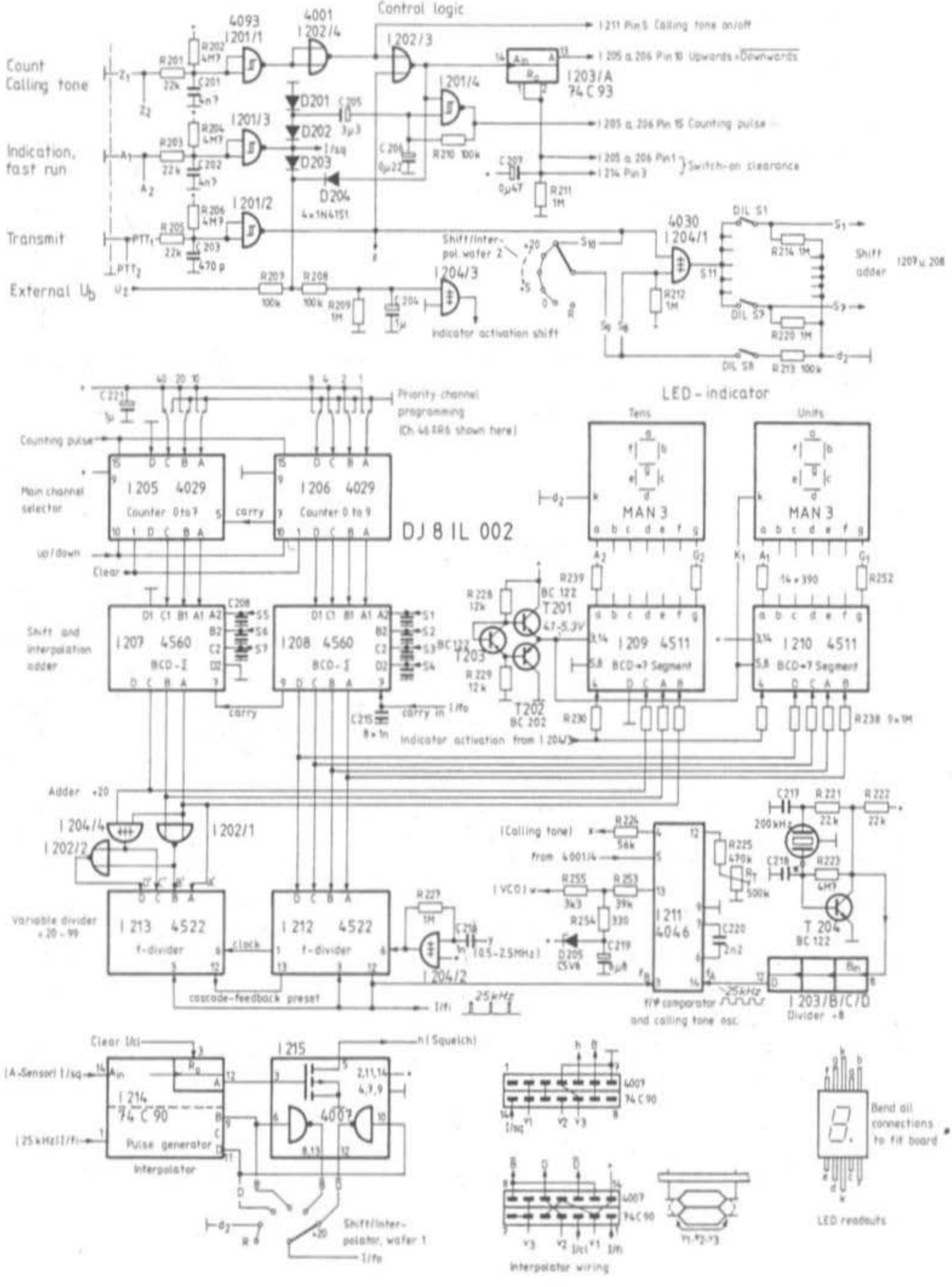


Fig. 4: Circuit diagram of the logic circuits DJ 8 IL 002

4.2. Frequency Plan

4.2.1. Phase Control Loop (PLL)

The VCO (see Fig. 1) operates in the range of f_{VCO} = 133.3 to 135.3 MHz, in other words 10.7 MHz below the 2 m band. This frequency is converted down in the PLL-mixer I 104 (SO 42 P, Figure 3) to f_{LI} = 0.5 to 2.5 MHz, which is the upper limit that can be reliably processed by C-MOS circuits. The heterodyne frequency of 132.8 MHz is generated in a crystal-controlled oscillator (T 114) with diode doubler. The intermediate frequency of the logic circuitry f_{LI} is amplified in an antivalence gate (I 204/2, Figure 4). It operates as an inverting operational amplifier as a result of the feedback via R 227; its operating point is approximately half the operating voltage.

The so-called variable divider of the PLL is a reverse counter (2 x 4522) that has been preset with the internal channel numbers N = 20-99. It will count from then on in time with f_{LI} until 0 is achieved, after which it will jump back to N. The counter generates an impulse which is fed to the frequency and phase comparator each time a count of 0 is obtained. The pulse frequency is therefore $f_B = f_{LI} \div N$; this is compared to a 25 kHz reference signal f_{ref}, which is obtained from a 200 kHz crystal oscillator (T 204, Figure 4) with subsequent divider (÷ 8). The fourth flip-flop of the divider-IC 74 C 93 can be used separately, and is used to control the count direction during frequency selection (see Figure 4).

The frequency and phase comparator (I 211: phase comparator II in 4046) is activated by the time sequence on the output slopes of signals f_{ref} and f_B. As long as $f_{ref} > f_B$, a high level (= + 9 V) will be present at the output (pin 13); if the opposite is the case: $f_B > f_{ref}$, the VCO frequency will be too high, and the output will remain at low level (= 0 V). When the two frequencies coincide, but f_{ref} is in advance of the phase of f_B, the phase comparator will provide positive impulses at pin 13 having a pulse duty factor proportional to the phase difference. The opposite will be the case when the phase of f_B is in advance of f_{ref}. The output voltage is filtered in the loop filter comprising R 253, R 254, and C 219 and controls the VCO thus closing the phase control loop. The phase comparator output will remain at high impedance when the two frequencies f_{ref} and f_B also coincide in phase. The selected tuning voltage is stored in C 219 and will not change.

The IC 4046 generates a control signal at pin 1 which will be always at low level when the phase comparator is generating correction pulses. Under locked conditions $f_{ref} = f_B$, $\varphi_{ref} = \varphi_B$, it will be at high level. It is thus possible for the transmitter to be blocked until the VCO locks in using a RC-filter link.

The zener diode D 205 ensures, that f_{VCO} does not drop to the image range of the logic IF when the voltage across C 219 is low: f_{VCO} < 132.8 MHz – 0.5 MHz (2). In this case, the PLL control would operate in the wrong direction: the VCO will run to the lower limit. When adjusting the VCO with the aid of C 155, it is necessary that channel 00 is not too near to the threshold of the zener diode, since the phase comparator must generate impulses of finite width as soon as zener current flows, in order to maintain the voltage across C 219. This causes a ripple on the tuning voltage, and thus leads to spurious lines in the spectrum that appear at ± n x 25 kHz spacing from the wanted signal **(Figure 5)**. These spurious lines should remain below - 60 dB.

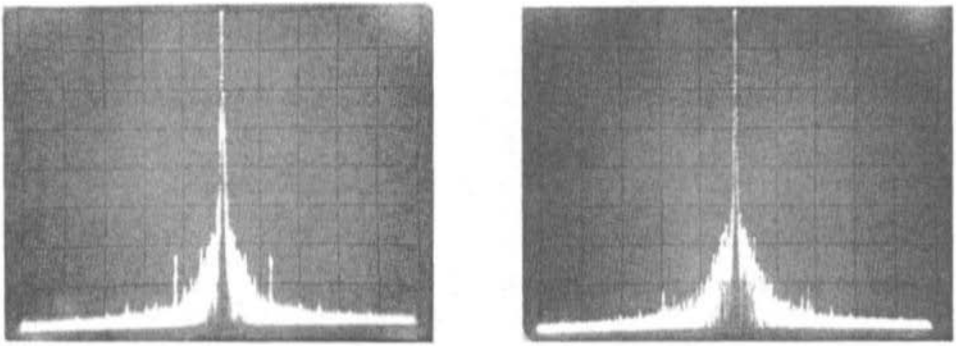


**Fig. 5: Transmit spectrum at channel 00 (left) and channel 78 (right).
Scale: Vertical: 10 dB/div. – Horizontal: 20 kHz/div. Measured with hp 141 T/8554 B/8852 B; Measuring bandwidth 300 Hz, trace speed 0.5 s/div. Video filter switched off.**

4.2.2. Operating Data of the PLL Stages

An antivalence gate (exclusive-or) has been selected as amplifier for f_{LI}, since it requires less current in this mode than an inverter, NOR and NAND circuit. A current of 0.3 to 0.8 mA has been measured in comparison to 1.1 to 1.3 mA at 0.25 to 2.5 MHz and 8 V. Furthermore, the output signals are steeper.

The 200 kHz transistorized oscillator (T 204) only requires 0.2 mA. A gate-oscillator would require 1 to 2 mA. Due to the insertion of R 221, a square-wave characteristic is obtained at the collector. Both series or parallel resonance crystals can be used. In the case of series resonance, C 217 - 270 pF, C 218 - 470 pF; bridge • on the PC-board is broken and an inductance L_S is provided. In the case of parallel resonance with a 30 pF load, C 217 = 47 pF, C 218 = 100 pF. It is not necessary to trim the frequency since the maximum effect will only be 1.7 % at the final frequency.

The frequency divider and comparator module equipped with ICs 4522 (2x), 4046 and 74 C 93 requires 0.7 to 3 mA at f_{LI} = 0.5 to 2.5 MHz and 9 V operating voltage. The variable divider (2 x 4522) operates reliably down to approximately 6 V, but not the phase comparator in the 4046. At the upper band limit, the phase comparator will receive impulses from the counter of only 0.2 µs duration. This means that not all ICs type 4046 operate reliably, not even at 9 V. It is advisable to use circuits manufactured by Fairchild or Motorola (3). A monoflop 4528 of max. 20 µs can be inserted between the divider and phase comparator (input f_B) to increase the length of the pulses.

The 66.4 MHz oscillator (T 114) is also a critical module. If the overtone crystal is really in series resonance at the nominal frequency, it is possible for its static capacitance C_0 to be compensated with C 156 at 6.8 pF; this results in a bridge circuit comprising C_0, C 156 and the two secondary windings of L 116. The advantage of this is that when L 116 is considerably detuned, the oscillator will not oscillate wildly (LC-oscillation), but will exhibit noticeable non-oscillation areas at both sides of the trimming range. If, however, the crystal is tuned too low, as is the case for oscillator circuits where the crystal replaces the base by-pass capacitor, it is necessary for a pulling capacitance C 157 to be provided. In this case, it may be favorable to delete C 156.

4.2.3. Frequency Selection Logic

The three sensor keys A, Z, and PTT are fed to Schmitt-triggers (I 201: 4093/1,2,3), that are activated at a touch resistance of approximately 2 MΩ. This value can be varied with the aid of R 202, R 204, and R 206. The fourth trigger gate operates as clock oscillator for the frequency scanning counter (I 205, I 206; 2 x 4029), in conjunction with R 210 and C 206. As long as the trigger I 201/3 has LOW level at the output (slow counting mode), C 205 will be recharged in parallel with C 206 via diodes D 201, and D 202. If sensor key A is activated, C 205 will be disconnected: fast counting mode.

In order to save current, the two 7-segment LED readouts and their decoders (4511) are connected in series. The intermediate level $U_S/2$ is generated in the voltage divider comprising T 201 - T 203, R 228, R 229. In this case, the current difference of the two readouts is accepted by T 201 or T 202 according to which of the positions is illuminating the largest number of segments. It would also be possible for an operational amplifier to be used instead of the three transistors that is able to supply the 25 mA (5 mA per segment). The resistances R 230 - R 238 operate as logic level converters from 0/9 V to 0/4.5 V, or 4.5/9 V together with the protective input diodes of ICs I 209, I 210.

The readout is activated via the OR-link comprising D 203, D 204, R 207 from sensor A or sensor Z (in receive mode), or by the external voltage supply. The circuit comprising R 209, and C 204 ensure a delay of 2 s before the readout is switched off. A non-inverting switching amplifier (1/4 I 204: 4030/3) ensures that the slowly falling voltage across C 204 does not switch off the units first and the tens afterwards.

The tens digit of the required channel number (with repeater shift) must now be increased by 2 before it can set the variable divider. This can be achieved using the three remaining gates of the control logic, since the range is limited to 0 to 7. The following can be taken from the coding plan:

$$\begin{aligned} A' &= A \\ B' &= \overline{B} \\ C' &= B \times \overline{C} \vee C \times \overline{B} = B \neq C \\ D' &= B \times C \text{ or } D' = \overline{B'} \times \overline{C'} = \overline{B' \vee C'} \end{aligned}$$

The available gates are sufficient if the second possibility is used to obtain D'.

Readout				Preselection				
Z	C	B	A	Z'	D'	C'	B'	A'
0	0	0	0	2	0	0	1	0
1	0	0	1	3	0	0	1	1
2	0	1	0	4	0	1	0	0
3	0	1	1	5	0	1	0	1
4	1	0	0	6	0	1	1	0
5	1	0	1	7	0	1	1	1
6	1	1	0	8	1	0	0	0
7	1	1	1	9	1	0	0	1

4.2.4. Digital Frequency Interpolator

The VCO will be tuned to a frequency between the two main channels if the variable divider in the PLL is provided with two different division ratios N_1 and N_2 at continuous alternation. If the alternation is fast enough, e.g. in time with the comparator frequency f_B, the loop filter will form a mean tuning voltage that corresponds to the changed frequency ratio $\tilde{N} = f_{LI} : f_{ref}$, which is between N_1 and N_2.

The carry-in input for the transfer from the lower-value position makes it possible to increase N by 1 in the shift-adder (4560). **Figure 6** shows the pulse plan for a division of N in 1/5 steps; for N = N + 0.4, for instance, twice N + 1 (carry in = HIGH) and three times N (carry in = LOW) are fed to the variable divider. In this case, the twice HIGH are split in order to obtain the rapid change and thus low residual ripple of the VCO control voltage. The pulse plan for N + 0.6 and N + 0.8 are the inverted form of N + 0.4 or N + 0.2; it is therefore sufficient for the latter two to be formed. The IC 74 C 90 (I 214) contains a ÷ 5-divider and the required time functions are present at its outputs B and D. The IC I 215 contains 3 p and 3 n-channel MOS transistors. Of these, two pairs are used as inverter. The third n-channel MOS is used as squelch-switch; it is driven from sensor A via the vacant fourth flipflop in the 74 C 90.

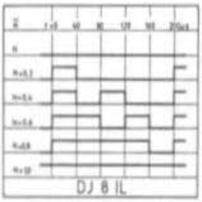


Fig. 6:
Carry-in pulse plan ($1/f_0$)
for digital frequency interpolation
at the shift-adder

The usability of this simple method is shown in studying the spectra given in **Figure 7**. As was to be expected, the same spectrum is obtained for N + 0.6 as for N + 0.4; the same is valid for N + 0.2 and N + 0.8. The interpolation spurious signals have a constant suppression of ≧ 65 dB down on the required signal throughout the whole 2 m band.

Fig. 7: Spurious waves in the transmit spectrum due to interpolation.
Left: channel 40 / + 15 kHz (N = 60.6); Right: channel 40 / + 20 kHz (N = 60.8);
otherwise measuring conditions as in Fig. 5

4.2.5. VCO

The FET oscillator (T 113) operates in a drain circuit with loose, inductive coupling to the resonant circuit. This arrangement combines a high Q together with a minimum of components driven with RF current. In the frequency range of 0.3 to 3 kHz, a residual frequency deviation of 30 Hz at 3 V and 20 Hz at 6 V and 9 V tuning voltage were measured using a RACAL 9009 modulation meter. This residual frequency deviation increased on connecting the modulator and the phase control loop, but remained less than 50 Hz at the output frequency. The output coupling capacitor C 150 can be aligned for maximum output power. However, this position is not critical, and it is more important to check for a reliable commencement of oscillation at channels 00 and 79.

4.3. Modulation and Dynamic Range of the PLL Control

4.3.1. Dynamic Control Amplifier

This circuit comprises the active components I 103, T 109 - T 111. The input signal from the microphone is provided with emphasis (4) and amplified to a constant peak-to-peak value of 1 V **(Figure 8)**. The module may seem complicated, but it offers a rapid, low-distortion control with virtually transient-free operation with defined adjustable values and good limiting balance. The maximum gain is given by the ratio of R 134 to the conductive resistance R_{DS} of T 111. The limit of the linear control range, and thus the limit threshold, is adjusted with the aid of R 132 ($V_{min} \approx$ R 134 ÷ R 132). Transistor T 110 is firstly used as control probe; afterwards, its base-emitter diode will limit the signal together with that of T 109. Resistor R 135 improves the balance, which remains intact even at high overload.

The range of the linear dynamic control should coincide to the range of voice voltages to be expected. Larger voltages are mostly short interference peaks, and these should be limited without altering the gain noticeably. The control speed can be slowed down by using larger values of C 139.

The noise level at the output of the amplifier is < 2 mV (peak-to-peak), which corresponds to a ratio of > 54 dB. The operational amplifier determines the upper frequency limit at V_{max}. Better results are provided when using more modern amplifiers such as the RC 4131, which is pin-compatible with type 741, however, the available signal-to-noise ratio is limited to approximately 40 to 45 dB due to the phase-locked oscillator.

This is then followed by an active lowpass filter as described in (5). It is designed to have 3 dB emphasis at 2.2 kHz and a - 3 dB cutoff frequency of 3.2 kHz. This means that it is matched to the frequency response of the microphone-loudspeaker capsule (see **Figures 9 and 10**). The calling tone is injected at the input of the filter (at x). It is taken from the VCO portion of the 4046; the frequency-determining impulses are C 220, R 225, and R_T. In the receive mode, the active filter is switched off with the aid of D 108.

Modulation and frequency control of the VCO are made using separate tuning diodes. This allows a constant frequency deviation to be obtained over the whole band: the calling tone deviation amounts to 4.1 kHz at channel 00 and 3.9 kHz at channel 79 (see **Figure 11**).

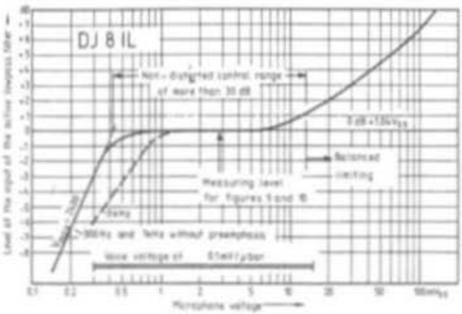


Fig. 8: Control characteristic of the dynamic compressor

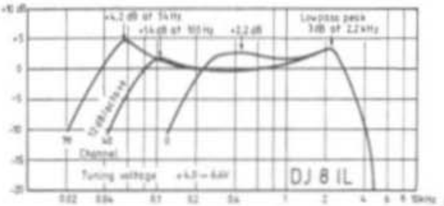


Fig. 9: Influence of the PLL on the modulation frequency response (without emphasis C 134, R 129)

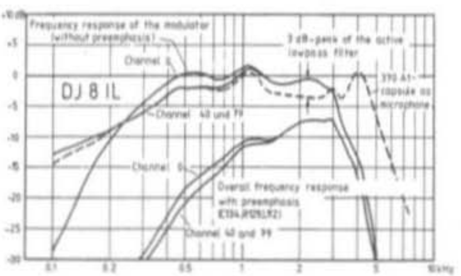


Fig. 10: Synthesis of the frequency response of the modulator

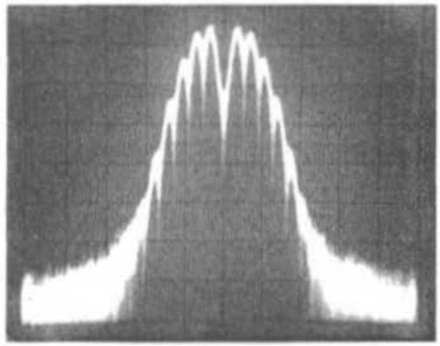


Fig. 11: Transmit signal modulated with calling tone. Channel 40; hor. 5 kHz/div. Other data as in Fig. 5

4.3.2. Design of the PLL Filter

The transient time of the VCO after changing frequency is determined, independent of the channel number N, by the time constant to be calculated from C 219 x (R 253 + R 254). However, it also causes a shift of the left branch of the modulation curves given in Figure 9 as a function of frequency. A shorter transient time will cause a higher lower modulation frequency limit.

With the aid of the time constant of C 219 x R 254, the phase control circuit is aligned for critical damping; this means that it is able to lock in as quickly as possible, but without overshoot. This is, however, only possible for one certain N; in addition to this, the amplitude and the position of the emphasis in the modulation frequency response in the vicinity of the lower frequency limit are also dependent on this: A greater (lower) value of resistor R 254 results in a greater emphasis in channel 00 [79]. The following transient times result when using the selected compromise:

Channels	00 ⇄ 79	46 ⇄ 70
Up	0.45 s	0.15 s
Down	0.18 s	0.07 s

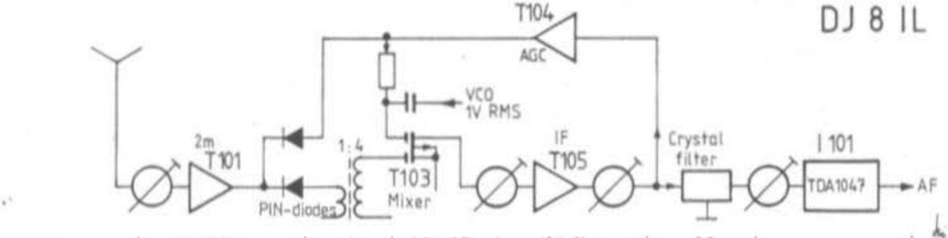
The additional filter comprising R 255 and C 148 is provided to suppress the pulse-type control commands from the phase comparator from the tuning voltage. One usually selects C 148 x R 255 ≈ (0.1 – 0.2) x C 219 x R 254 (6). If the filtering is too great, the residual frequency deviation will be increased.

4.4. Receiver Circuit

4.4.1. RF and IF Amplifier

Most of the RF selectivity is provided for transmitter and receiver using the common antenna filter which comprises a high Q resonant circuit with trap for the image frequency (L 101, C 101, C 102). This two-pole is easy to align and provides a higher image rejection and a lower noise figure than a two-stage bandpass filter of the same Q. The signal is processed at low impedance up to the mixer, in order to keep the danger of unwanted coupling as low as possible with the miniaturized construction used. For this reason, an antenna preamplifier circuit as described in (7) was used as RF input stage.

Figure 12 gives details as to the signal level present in the receiver. Larger variations are only to be expected with the dual-gate MOSFET mixer. The opinions differ greatly regarding the most favorable operating points. Measurements made in the author's own experiments using a MOSFET type 40673, showed that the best receiver sensitivity was obtained when using a positive bias voltage for gate 2 of approx. 1 to 2 V, and a source resistor of approx. 560 Ω, as described in (9). Capacitor C 110 was selected so that series resonance at the IF results in conjunction with the VCO network (L 115); this then results in maximum conversion gain (8). The overall gain is increased to such a level in the IF-preamplifier (T 105) that the noise level coincides with the limiting threshold of the TDA 1047.



Gain	9 (...0,9) + 19 (... – 1) dB	4 + 12 dB	2,5 (...0,7) + 8 (... –3) dB	6 (...3) + 15,5 (...10) dB	0,8 – 2 dB	ca. + 70 (...–15) dB

Noise (Δf = 15 kHz)	55 nV	0,7 μV	2,8 μV	7 μV	43 μV	35 μV	45 mV
Squelch threshold	95 nV	0,85 μV	3,5 μV	8,5 μV	50 μV	40 μV	High noise
10 dB $\frac{S+N}{N}$	200 nV	1,8 μV	7 μV	18 μV	105 μV	85 μV	70 mV
Large signal	200 mV	-	-		1,6 V	1,3 V	70 mV

Fig. 12: Level plan of the receiver

If the RF-voltage at the collector of T 105 attains 1 V, the control amplifier T 104 will commence operating as a peak-value rectifier. This will cause the voltage across C 109 to increase and will shift the operating point of the dual-gate MOSFET (gate 2) into a range having a more linear gain: The conversion slope will be reduced, and the large-signal capabilities will be improved.

If the voltage (C 109) approaches 6 V, diode D 102 will then change from the blocked to the flow range. D 102 and D 103 are PIN-diodes that have a behaviour of ohmic resistors at frequencies in excess of 1 MHz. The (real) RF-resistance depends on the direct current flowing through the diode. The following is valid for the BA 379:

I_f/mA	9	3	0.1	0 (at U = 0)
R_{RF}/Ω	5	12	220	2500

In the »SUEDWIND« transceiver, the PIN-diodes are connected as controllable current dividers: At low signal levels, the total RF-current from T 101 will flow via D 103 at the transformer L 104; at antenna voltages in excess of 5 mV, T 104 will provide direct current via D 102 and thus open a leakage path for the RF-current to ground via C 106 and C 107. The current through D 103 decreases at the same time nearly to zero since the voltage across R 105 + R 106 will return to U_{CEsat} (T 104).

The battery current required for the PIN-diode control is not lost, but increases the collector current of T 101 and thus improves the large-signal handling capabilities of the preamplifier stage (7). The value of I_C (T 101) controls T 102 via the base current so that the voltage at the emitter of T 102 remains constant at 6 V. The current I_C (T 101) then varies between 2.9 V ÷ (330 + 680) Ω and 2.9 V ÷ 330 Ω.

A fourth control function results from the increasing damping of the collector load resistor of T 105 when the base-emitter path of T 104 becomes more and more conductive. R 111 limits this control effect so that the AF-distortion factor does not become too great due to the mismatch of the crystal filter.

When using these measures, the control characteristic shown in **Figure 13** is obtained.

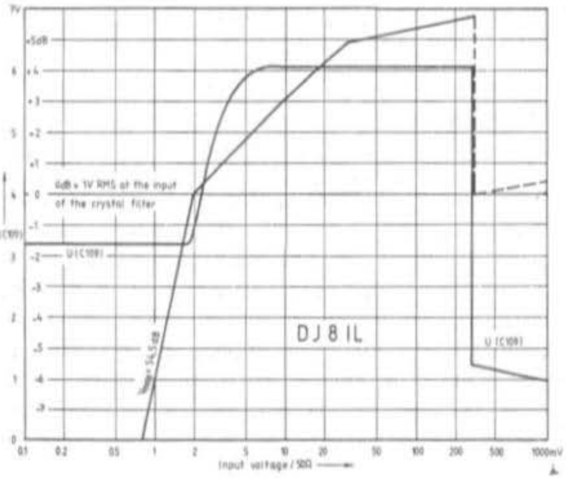
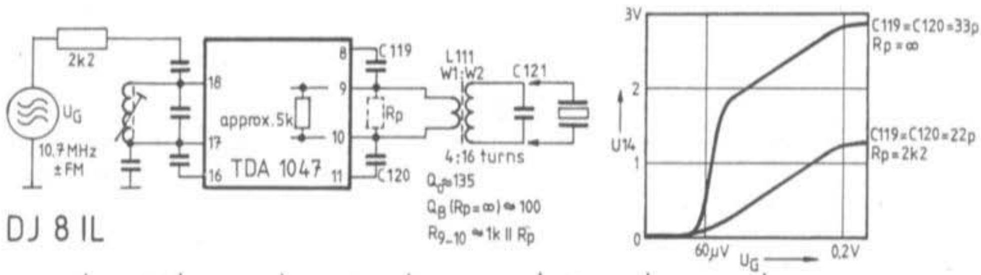


Fig. 13:
Control characteristic of the VHF/IF circuit

4.4.2. Discriminator

The limiting IF-amplifier and FM-demodulator are accommodated in the integrated circuit type TDA 1047 and generate a S-meter signal over approximately 3.5 decades of the input voltage in a similar manner to the well-known CA 3089 that requires 40 % less current. The S-meter characteristic and threshold of the integrated squelch are very dependent on the design of the discriminator circuit. Several typical results of measurements on demodulators with and without crystal are given in **Figure 14** and the associated table.



Rp/ Crystal	C 119, C 120 [pF]	AF at ± 4 kHz dev.	Limiting (-3 dB) at	Squelch via U_{15}	S-meter characteristic	Hump spacing	Noise deviation
2,2 kΩ	22	0,11 V_{ss}	-	> 1 mV	First	-	10 Hz
"	22	0,16 V_{ss}	[6,6 µV]	70 µV	linear	± 130 kHz	8 Hz
"	33	0,18 V_{ss}	35 µV [6 µV]	50 µV [14 µV]	segment	± 65 kHz	8 Hz
"	47	0,20 V_{ss}	[5,4 µV]	[8,5 µV]	too steep	± 120 kHz	[4 Hz]
Crystal	33	1,25 V_{ss}	75 µV	≈ 10 mV	→ f	± 22 kHz	10 Hz

Values in [] were measured with IF-preamplifier and squelch via T 106 (instead of U_{15}).

Fig. 14: Characteristics of the TDA 1047 using different discriminator circuits

For comparison, a KVG crystal discriminator with a hump spacing of ± 50 kHz supplied 0.11 V audio signal (peak-to-peak) at an IF-input voltage of 1 V and ± 4 kHz deviation, and a Japanese IF-circuit exhibited only 22 mV (with C 119 = C 120 = 33 pF).

In the section of the curve directly after the limiting threshold, the slope of the S-meter voltage (U_{14}) and the squelch-switching voltage (U_{15}) increase with C 119 = C 120 and R_{9-10} at resonance. Two examples of U_{14} are given in Figure 14. The demodulator slope U_{AF}/dev. will increase with the Q (L 111, C 121); the matching resonance impedance R_{9-10} is adjusted with the transformation ratio $w_1 : w_2$. The Q should not be too great so that the hump spacing is sufficient to compensate for the temperature response of the resonance frequency of L 111 and C 121. A sufficiently linearized S-meter characteristic with simultaneous high sensitivity of the squelch was obtained by adding transistor T 106; the emitter diode of T 106 is connected in series with the meter and compensates for the rapid increase of U_{14} at low S-values **(Figure 15)**. The greater the value of C 119 = C 120, the nearer the squelch switching threshold will be to the – 3 dB limiting threshold. The integrated circuit TDA 1047 also contains a tuning-dependent squelch, which, however, has tendency to instability in the narrow-band FM mode when using high Q discriminators. It is switched off with the aid of R 116 $\lesssim$ 100 kΩ.

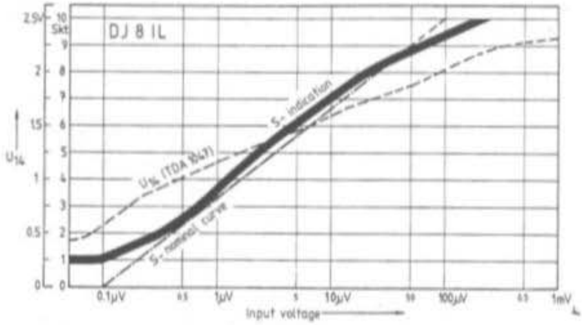


Fig. 15: Control voltage U_{14} for squelch and S-meter; scale division of the meter: 1 (not 0) to 10

4.5. Transmit Amplifier

The RF-circuits in the transmit amplifier are built up at a relatively high Q using a minimum of components in order to ensure that the required selectivity takes up as little room as possible and is easy to align. In contrast to normal practice, the driver is matched to the power amplifier stage using a single-stage resonance transformation together with the series capacitor C 172; the output stage of the transmitter also only uses one tapped inductance L 121.

The extract from the circuit diagram given in **Figure 16** possesses three additional ceramic disk capacitors over the circuit diagram given in Figure 2: C 186, C 185, and C 184. Capacitor C 170 has been supplemented with a larger capacitance C 186 having a lower Q, in order to increase the damping of the Pi-network C 171/L 119/C 170, which couples the driver to the power amplifier. Without these capacitors, a fluctuating, wideband base was noticed in the spectrum of some units, which indicated a tendency of the transmit amplifier to parasitic oscillation. Capacitors C 185 and C 184 form a Pi-network together with the inductance of the relay contact and the conductor lanes, which provides a further harmonic suppression in the order of 10 dB.

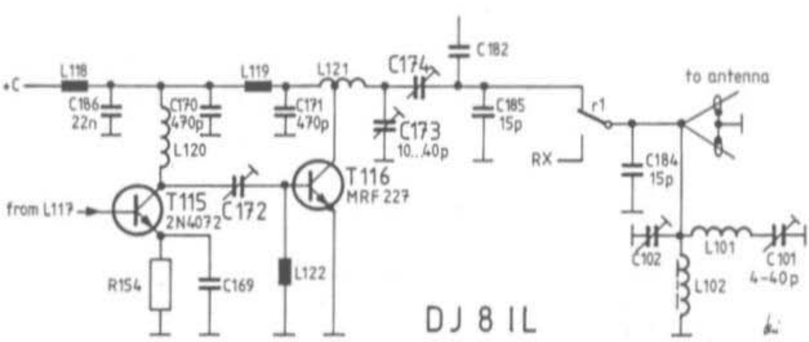


Fig. 16: Extract from the circuit diagram showing the new »SUEDWIND« transmit amplifier circuit

The higher the power gain of the final transistor, the lower will be the current drain at a certain output power P_0. For this reason, transistor type MRF 227 is used for T 116, because its power gain is 2 dB greater than type MRF 607 used previously. Furthermore, its metal case is connected to the emitter, which means that there are no heat dissipation problems. The output power, and thus the current drain can be adjusted by aligning trimmer capacitors C 174, C 173, and C 172.

4.6. Auxiliary Stages

With the exception of the power amplifier stages, the transceiver is fed with a stabilized operating voltage of 9 V. The stabilizer circuit designed for this is equipped with transistors T 120 - T 123, and has the following specifications: Impedance ≈ 0.07 Ω / Minimum voltage difference across T 120: $U_C - U_D \approx 0.5$ V / Intrinsic current requirements ≈ 0.6 mA. In the case of integrated »three-legged« stabilizers of the LM 78 L ... or LM 340 LAZ/... series, the corresponding specifications would be: $\leqq$ 0.6 Ω / 2 V / 3 mA.

The electronic distribution network comprising D 107 and T 107 ensures that the built-in NC-accumulator is always connected when the transceiver is operated from an external source. This ensures that a smooth transition takes place and that protection is provided against incorrect polarity and failure of the external power supply. The charge condition of the accumulator is monitored; a LED (D 117) will be activated as soon as the accumulator voltage falls below approximately 10.5 V. This LED is mounted between the loudspeaker switch and antenna connector.

If an external dynamic microphone is to be used, the built-in capsule will be switched off electronically. This is made with the aid of a FET (T 108) having a low conductive resistance. It blocks as soon as connection A/M is grounded. The acuumulator can be charged with a defined current via the same connection, independent of the external supply of the transceiver.

The transmit/receive switching is made electronically using the sensor-amplifier I 201/2 and T 117 - T 119. The antenna and microphone-loudspeaker are switched using polarized relais, which receive short current pulses via C 175.

The mechanical construction, wiring and alignment of the transceiver are to be described in the following edition of VHF COMMUNICATIONS.

A DF-RECEIVER FOR THE 2 m BAND EQUIPPED WITH INTEGRATED CIRCUITS, CRYSTAL FILTER AND S-METER

by M. Schmaußer, DL 2 DO

Many DF-hunt (fox hunt) enthusiasts think that a S-meter is not necessary on a DF-receiver, however, it is the opinion of the author that a minimum dip with the aid of a S-meter is far better than acoustically. A DF-receiver is to be described that allows a bearing to be taken even when only 1 m from the »fox« transmitter. Inspite of the good sensitivity, and large control range, the circuitry is not extensive and a relatively small PC-board is used. The most important components are only four transistors and three integrated circuits, as well as a cheap crystal filter. The photograph given in **Figure 1** shows the DF-receiver completely ready for operation including attached, flexible HB 9 CV antenna.

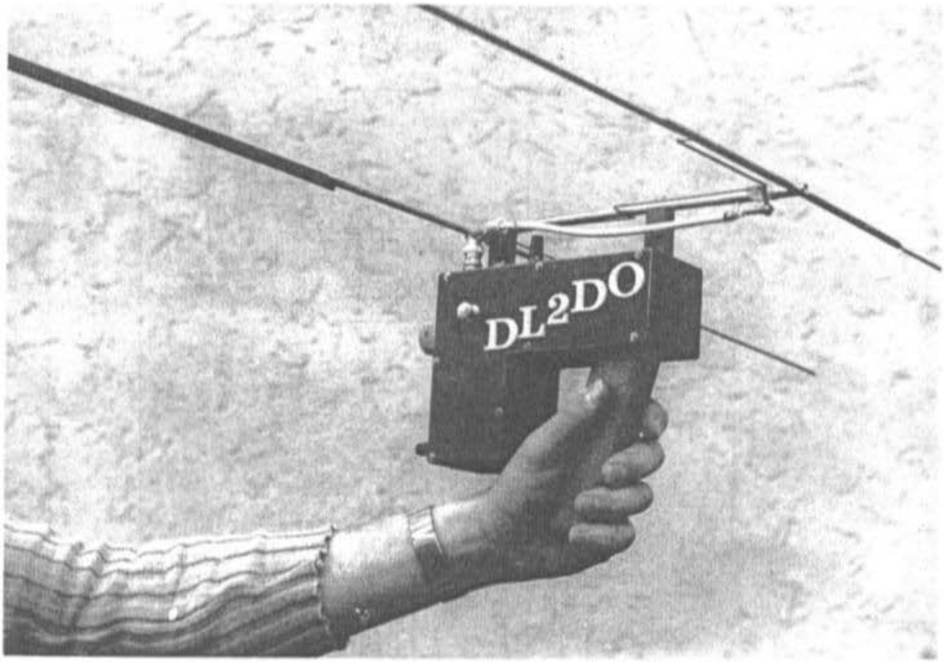


Fig. 1: A photograph of the author's prototype

1. CIRCUIT DESCRIPTION

The circuit diagram is given in **Figure 2**. It will be seen that a modern, high-gain, and low-noise dual-gate MOSFET type BF 900 is used in the VHF stage. It is possible to still DF-hunt under high signal level conditions in the direct vicinity of the transmitter, because gate 2 of the BF 900 can be biased with the aid of an extra battery to - 1.5 V. In addition to this, a potentiometer of approximately 500 Ω is provided at the antenna input, which allows the input signal to be attenuated until no desensitization occurs. Of course, these two methods are only effective when inductances L 1 to L 3 are screened so that the input transistor is not bridged and when a VHF-tight case is used.

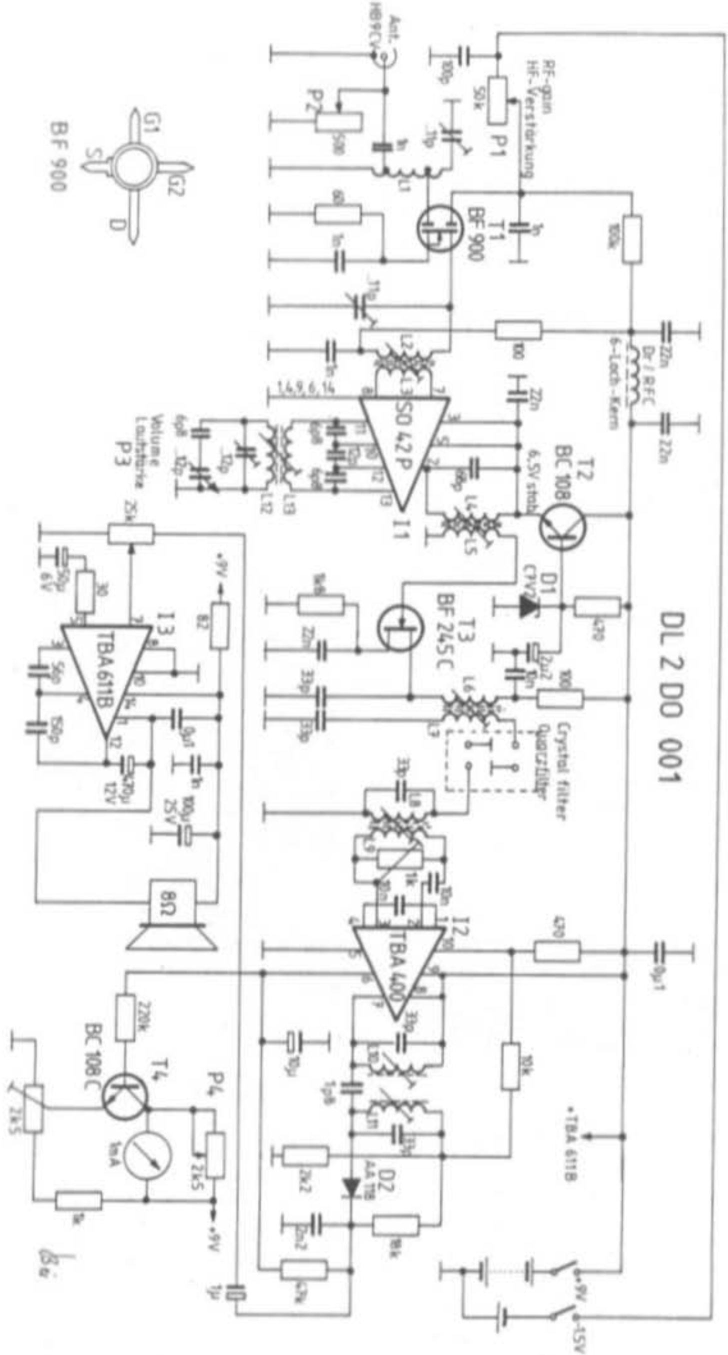


Fig. 2: Circuit diagram of the DF-hunt receiver for 2 m, equipped with ICs, crystal filter and S-meter

A push-pull mixer and variable oscillator are accommodated in the integrated circuit SO 42 P. These circuits are fed with a stabilized voltage via the circuit comprising T 2. Tuning is made using a miniature variable capacitor having 2 x 12 pF, of which only one set is used. A 3 : 1 reduction is provided on the variable capacitor; if an additional reduction of 6 : 1 is used, this will be fine enough for tuning.

The 10.7 MHz IF-signal is preamplified in the field-effect transistor T 3, matched to the crystal filter by L 6 /L 7, and finally fed to the wideband amplifier TBA 400 via L 8 / L 9. This three-stage amplifier has a balanced input and output, 75 dB gain (measured at 36 MHz) and can be controlled by 60 dB at pin 6. The rectified voltage of the demodulator diode D 2 is fed to this pin, as is the S-meter amplifier equipped with transistor T 4. The S-meter can be corrected using the trimmer potentiometer used as emitter resistor. A special feature of this receiver is the potentiometer in parallel with the S-meter. This allows the required reading to be adjusted according to the distance to the transmitter.

The integrated AF-amplifier type TBA 611 B finally provides an audio power of 1 W to a 8 Ω loudspeaker. If a switching connector is used, it is possible for an earphone to be also connected (not shown in the circuit diagram).

According to the operating period, either large or small batteries can be used for power supply. The author's prototype is equipped with six »baby« cells having a nominal voltage of 9 V, and one »mignon« cell (1.5 V) for the negative gate bias voltage of T 1.

2. SPECIAL COMPONENTS

T 1:	BF 900 (Texas Instruments)	
T 2, T 4:	BC 108 (C) or BC 413 (C) or similar AF transistor with high current gain	
T 3:	BF 245 C (TI)	
I 1:	SO 42 P (Siemens)	
I 2:	TBA 400 (Siemens)	
I 3:	TBA 611 (SGS)	
D 1:	C 7 V 2 zener diode	
D 2:	AA 118 or similar germanium diode	
Crystal filter:	FILTECH:	6 dB bandwidth: 13 kHz, 80 dB bandwidth: 35 kHz Ultimate selectivity > 100 dB; Insertion loss < 4 dB Input and output impedance: 2700 Ω Dimensions: 20 mm x 12 mm x 13 mm high
L 1:	6 turns of 0.8 mm dia. silver-plated wire wound on a 6 mm former, self-supporting, antenna tap: 1 turn from the cold end; G 1-tap: 0.75 turns from the hot end	
L 2:	5.5 turns of 0.8 mm dia. silver-plated copper wire wound on a 5 mm coil former with VHF core, in screening case 12 mm x 12 mm x 15 mm high	
L 3:	1 turn wound at the cold end of L 2	

Inductances L 4 to L 11 are wound with stranded coil wire of 10 x 0.05 mm in special coil sets available from the publishers.

L 4:	18 turns	L 5:	10 turns wound onto L 4
L 6:	18 turns	L 7:	20 turns wound onto L 6
L 8:	20 turns	L 9:	2 turns wound onto L 8

L 10, L 11: 18 turns

L 12: 3.5 turns of 0.8 mm dia. silver-plated wire on 5 mm former, VHF core

L 13: 2 turns of wire wound on the cold end of L 12

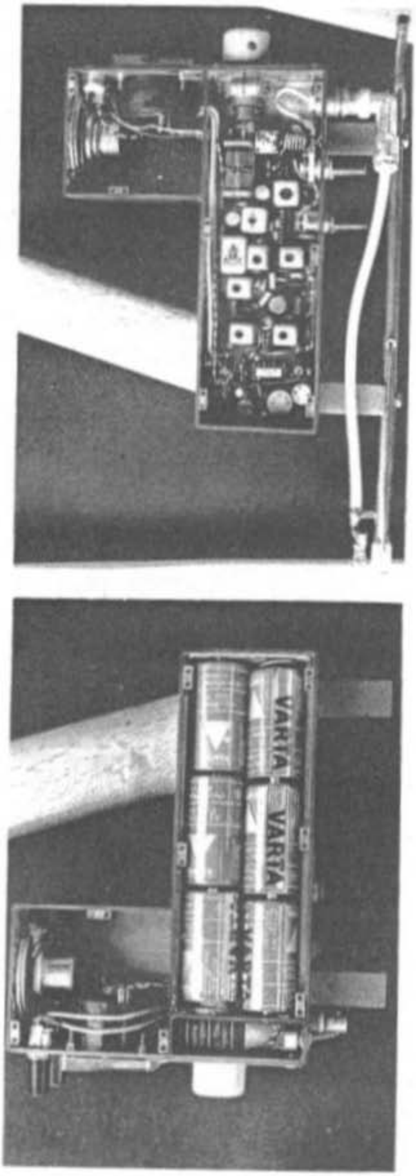


Fig. 4 + 5: Photographs of the author's prototype

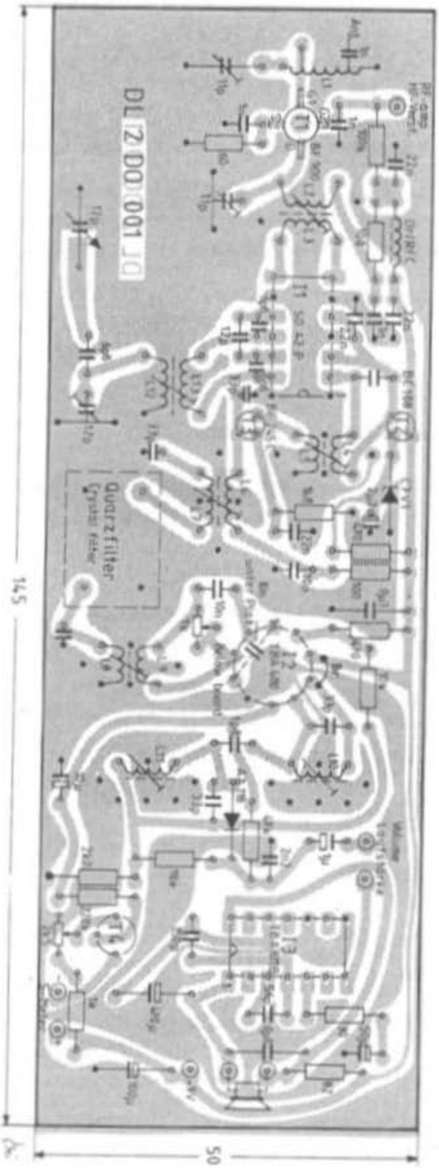


Fig. 3: PC-board DL 2 DO 001 of the receiver

3. CONSTRUCTION

The circuit can be constructed using the PC-board shown in **Figure 3**. The dimensions of this single-coated board are 150 x 50 mm. The TBA 400 is somewhat critical due to its enormous gain, and thus its tendency to oscillation. It is therefore necessary for all connections to be kept as short as possible in its vicinity, and the 10 nF capacitor between connections 1 and 4 is soldered directly across the IC (TO 5 metal case). As much copper surface as possible should remain on the PC-board as ground surface.

Potentiometers P 1 and P 4 are mounted on the front panel at suitable positions; the antenna input is in the form of a BNC connector for the HB 9 CV antenna, which can be mounted on top of the case. The outer elements of the antenna are made from flexible metal strips from a metal tape measure, and have been painted black so that other competitors cannot see to which direction one is hunting. Of course, there are sufficient possibilities to extend this basic design to any further requirements.

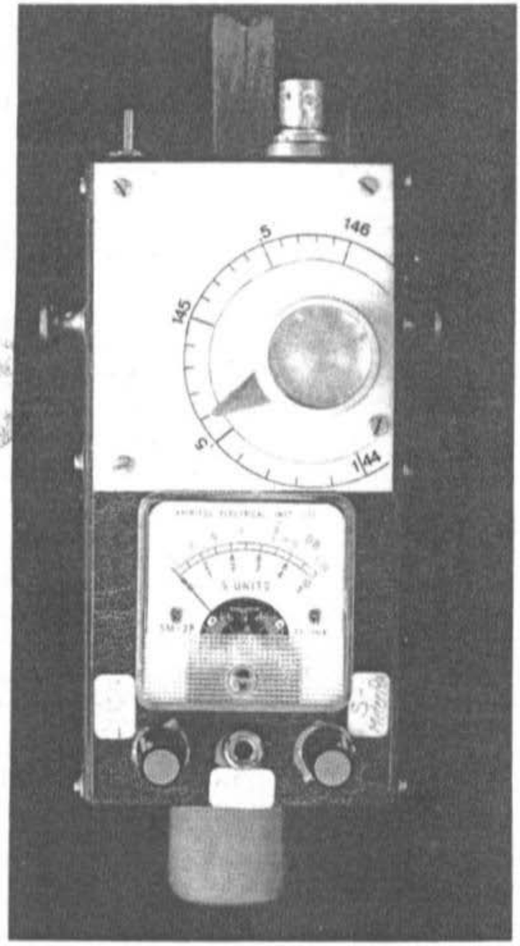


Fig. 6: Front panel of the receiver

A MODERN RECEIVE CONVERTER FOR 2 m RECEIVERS HAVING A LARGE DYNAMIC RANGE AND LOW INTERMODULATION DISTORTIONS

by M. Martin, DJ 7 VY

Although transistor technology has forced out tubes from receiver circuits, the large-signal handling capabilities of tubes have still to be obtained in professional equipment, although a number of circuits have been described that possess a sufficiently high performance. The following article is to describe a receiver input section (converter) that possesses very high large-signal capabilities, which is just as suitable for construction of 2 m converters, or for construction of single-conversion superhets for 144 MHz or shortwave. It is only necessary for corresponding circuits to be added or deleted from the system board. A further feature of this input circuit is an efficient noise blanker that does not distort under large-signal conditions, and is able to suppress pulse-type interferences such as from vehicles or radar.

1. GENERAL

Sensitivity, large-signal handling capabilities, and selectivity are measures of the quality of a receiver. Since multi-pole crystal filters can achieve any required selectivity easily, attention is to be paid to the first two characteristics. It has been found that it is especially difficult to develop input circuits that are simultaneously sensitive and have large-signal capabilities. The problems occuring in this respect and the measuring methods were described in detail in (1).

A new type of preamplifier with very low noise figure (NF) was described in (2) that allowed the excellent characteristics of high-current Schottky diode mixers to be utilized to the full in a receive converter. In this case, its third order intercept point (IP), which is used as measure for the large-signal capabilities, was only decreased to the value of the preamplification, since the mixer together with its matching amplifier determines the overall quality of the input circuit.

At first, it may seem possible to achieve an overall improvement by only increasing the intercept point of the mixer. It is true that active FET mixers are available with IP = 40 dBm, NF = 9 dB, G_p = 0 dB at P_{LO} = 27 dBm (P_{LO} = local oscillator power; G_p = power gain), however, it is virtually impossible to achieve the wideband ohmic termination with an intercept point of 40 dBm, or even 30 dBm, when using a low-reactive amplifier with NF $\leqq$ 4 dB using components that are now available on the market.

The maximum intercept point of the 50 Ω amplifier obtainable with moderate means is IP = 26 dBm. This means that an overall IP of maximum 32 dBm at the mixer input at NF = 10 dB is possible when using a passive mixer, which exhibits a signal loss of approximately 6 dB. When using the previously mentioned active mixer, an overall IP of only 26 dBm with NF = 13 dB at G_p = 0 dB would be possible. The intercept point of the crystal filter has not been taken into consideration; unfortunately, this is not infinite, but is in the order of 25 to 30 dBm according to the quality of the filter. An improvement over the Schottky mixer was only possible after splitting the mixer output signal using power dividers to eight subsequent amplifier chains followed by power combiners which resulted in an increase of the IP to 35 dBm. The cost of such a system stopped further developments in this direction.

A low-noise oscillator signal is absolutely necessary if the large-signal capabilities provided by the mixer are to be used to the full. This can only be achieved by careful selection of suitable components in the oscillator and multiplier circuits, and is just as valid for crystal control as it is for a variable oscillator.

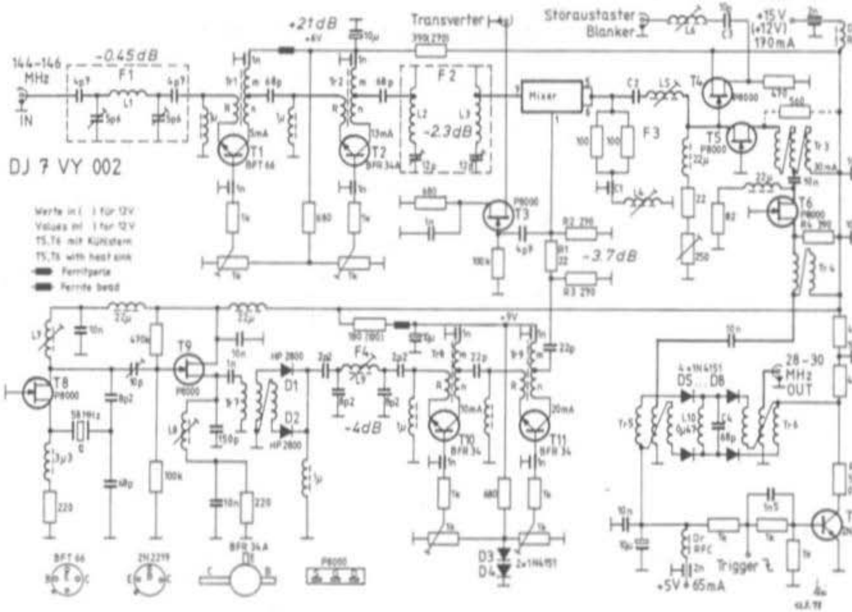


Fig. 1: A Schottky mixer (converter) or input circuit with large dynamic range

2. THE CIRCUIT

The circuit of the receive converter is given in **Figure 1**. It is very similar to that described in (1) and (2) and possesses the following specifications:

Noise figure NF = 1.74 ≙ 2.42 dB (IF-amplifier NF = 10 dB)
NF = 1.68 ≙ 2.27 dB (IF-amplifier NF = 6 dB)

Gain G_p = 20 dB

Third-order intercept point: with SRA-1 : IP = 1.75 dBm (measured: 0 dBm)
with SRA-1H: IP = 7.75 dBm

Input impedance R_{in}: 50 - 75 Ω (as aligned)

Output impedance R_{out}: 50 Ω

Image rejection (86-88 MHz) G_p (f_{im}): – 90 dB ! see **Figure 3**

Oscillator sideband noise: < – 160 dB/Hz at > 100 kHz spacing

Desensitization P_{in} max: – 12 dBm ≙ 56 mV into 50 Ω, see (4)
i.e. 560 mV at the output of the converter !

Blanking suppression: 88 dB

Blanking speed: 25 ns with TTL-low-signal

Oscillator coupling to the input: 116 MHz: – 108 dBm ≙ 0.9 µV
58 MHz: – 90 dBm ≙ 7 µV

Oscillator coupling to the output: 116 MHz: – 72 dBm ≙ 56 µV
58 MHz: – 79 dBm ≙ 25 µV

Required IP for IF-amplifier: 20 to 30 dBm !

Power requirements: 15 V (at least 12 V) stabilized, 170 mA
5 V, stabilized, 65 mA

Dimensions in mm: 152 x 77 x 45

The level values are given in the form of a block diagram in **Figure 2**.

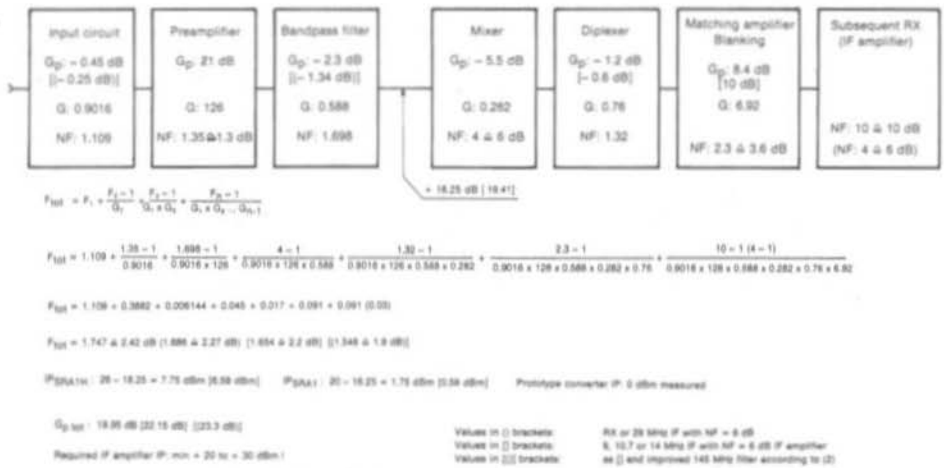


Fig. 2: Block diagram of the converter with level values

3. CIRCUIT DESCRIPTION

In the converter version shown in Figure 1, the input signal is fed via the filter F 1 having G_p = – 0.45 dB. This is followed by the two-stage preamplifier with G_p = 21 dB and a noise figure of 1.3 dB, as well as filter F 2 with G_p = – 2.3 dB. The frequency response of the RF-circuit up to this point is given in **Figure 3** with maximum drive, and, with 20 dB overload (upper curve).

The subsequent mixer will exhibit the following values when a SRA-1H is used:
G_p = – 5.5 dB, NF = 6 dB, IP = 26 dBm at an oscillator power P_{LO} = 17 dBm ≙ 50 mW. If the cheaper mixer SRA-1 or IE-500 is used instead of the SRA-1 H, it is necessary to reduce the oscillator power P_{LO} using the attenuator comprising R 1, R 2, and R 3. Measurements have shown that an IP of approximately 20 dBm can be achieved with these mixers at P_{LO} = 13 dBm ≙ 20 mW. On the other hand, it would be possible when using the high-power mixer RAY 3 to achieve an IP-value of approximately 29 dBm at P_{LO} = 23 dBm ≙ 200 mW. This corresponds to an input IP in the order of 10 dBm, even at the same noise figure !

The mixer is followed by the diplexer F 3 that ensures a wideband 50 Ω termination. This circuit was recommended by DK 1 OF in (3). It provides a better termination for the image frequency $f_{LO} + f_{in}$ = 116 MHz + 144 MHz = 261 MHz than would be provided by the subsequent high-power FET type P 8000 on its own, since the gate-source capacitance is present in parallel with the input impedance when used in a common gate circuit. The P 8000 is very similar to type CP 643 described in (1) and is used in many stages of this converter due to its favorable cost. If the IF-trap comprising L 4 / C 1 is constructed from conponents with connection leads, it will block the required frequencies, but will not sufficiently allow the image frequency of 261 MHz to pass due to the inavoidable series inductivity of the capacitors. This leads to an unfavorable high standing-wave ratio at this frequency and thus to a low overall IP. A high winding capacitance of L 5 would have the same effect. This problem can be solved by using a bypass capacitor without connection leads for C 1. The two 100 Ω resistors are soldered to the upper side with as short connections as possible; the connection length of L 4 is not critical.

The quality of diplexer F 3 can be seen in the oscillogram given in **Figure 4**: The return loss amounts to 30 dB, which corresponds to a VSWR = 1.07.

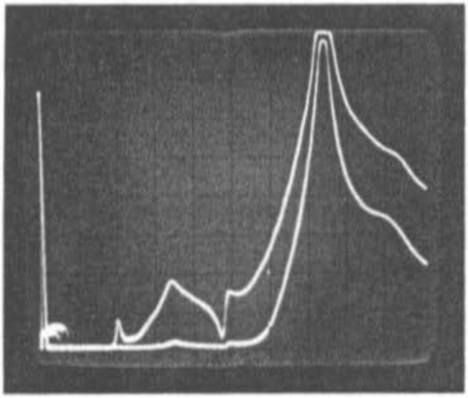


Fig. 3:
Frequency response of the HF-amplifier including filters F 1 and F 2
H.: 20 MHz/div., V.: 10 dB/div.
upper curve with 20 dB overload

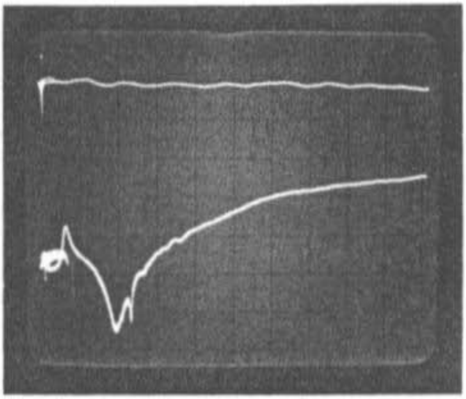


Fig. 4:
Input return loss of the diplexer with subsequent P 8000
H.: 50 MHz/div., V.: 10 dB/div.
upper curve: 0 dB line ≙ VSWR = ∞

The values for L 4, L 5, and C 2 must be calculated for the IF to be used, and according to the value of the capacitor C 1 (large spread).

Example: IF = 29 MHz; C 1 = 470 pF;

$$L\,4 = \frac{1}{\omega^2\ C\,1} = 64\ \text{nH}$$

With R 3 = 50 Ω the following will be valid according to (3):

$$B = \frac{1}{2\pi \times R \times C\,1} = 6.8\ \text{MHz}$$

This results in:

$$L\,5 = \frac{R}{2\pi \times B} = 1.2\ \mu\text{H} \quad \text{and} \quad C\,2 = \frac{B}{2\pi\, f^2 \times R} = 26\ \text{pF}$$

Table 1 gives the calculated component values for four different intermediate frequencies, together with the required oscillator frequency.

IF (MHz)	C 1 (pF)	L 4 (nH)	C 2 (pF)	L 5 (µH)	C 3 (pF)	L 6 (µH)	C 4 (pF)	L 10 (µH)	R 4 (Ω)	R 5 (Ω)	f_{LO} (MHz)
28-30	470	64	26	1.2	10	3	68	0.47	390	56	58 x 2
14-14.5	470	275	109	1.2	27	5	68	del.	270	82	65 x 2
10.7	470	470	188	1.2	39	6	100	del.	270	82	66.65 x 67.65
9	470	665	266	1.2	47	6	150	del.	270	82	67.50 - 68.50

The converted input signals are fed via the extremely linear matching amplifier (T 5) and a buffer, which is also equipped with a P 8000. The buffer is provided to ensure that no feed-back is made via the switching slopes via the interference channel receiver under noise blanking conditions.

This is then followed by the noise blanker switch comprising switching diodes D 5 - D 8 that exhibits a blanking of more than 80 dB within 25 ns. The diode switch is controlled by the switching transistor T 7, which is in turn controlled by a triggered Monoflop 74 LS 123. The full noise blanker circuit between the interference channel buffer T 4 which is connected in parallel to the matching amplifier T 5, and the trigger circuit, will be described in detail in a later publication and will be a further development of (5).

The two matching amplifiers with blanking switch exhibit the following specifications:

G_p = 8.4 dB, NF = 3.6 dB, R_{in} = 50 Ω, ± 0.5 %, input IP = 26 dBm, and output impedance $R_{out} \approx 50\ \Omega$.

In the oscillator circuit, a 58 MHz crystal is operated at its third overtone (series resonance). This is followed by a buffer which is in turn followed by a low-noise frequency doubler equipped with Schottky diodes D 1 and D 2. The required filtering is provided in the bandpass filter F 4 before the 116 MHz signal is amplified. This is achieved in the local oscillator amplifier comprising T 10 and T 11, which is very similar to the VHF-amplifier comprising T 1 and T 2, however, the low-noise BFT 66 has been replaced by a BFR 34 A in the first stage. In addition to this, the collector currents are aligned for higher values. A maximum of 18 dBm = 63 mW are available at the output of the amplifier to feed the mixer (R 1 = 0, R 2 and R 3 = ∞). This local oscillator provides an extremely low-noise oscillator signal with a sideband noise of less than – 160 dB/Hz at a spacing of more than 100 kHz from the carrier frequency !

In order to determine the overall noise of the converter, the values of the individual stages are inserted into the well-known sum equation:

$$F_{tot} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \times G_2} + \frac{F_n - 1}{G_1 \times G_2 \text{ to } G_{n-1}}$$

The result together with the noise components of the individual stages is given in the block diagram shown in **Figure 2**, where a receiver having NF = 10 dB was assumed as IF amplifier. The values given in parenthesis are valid for an IF amplifier with NF = 6 dB. This value can be obtained when the noise blanker switch is followed by a crystal filter having a G_p = – 3 dB and an IF-amplifier with NF = 3 dB.

The values given in the square brackets are valid when intermediate frequencies of 9 MHz, 10.7 MHz or 14 MHz are used with NF = 6 dB. The blanking transformers have less overall attenuation at these frequencies. L 10 can be deleted, and various R and C values must be changed (see Table 1).

It will be seen from the values given in **Figure 2** that none of the amateur radio receivers at present on the market possesses an input IP that is able to utilize the large-signal capabilities of this converter to the full. This would only be possible without limitation with a receiver according to (1). For this reason, it is advisable for the described circuit to be used as input circuit for a single-conversion superhet by connecting a low-noise variable oscillator to filter F 4. The VFO must provide an output power of approximately 1 mW (for the SRA-1H). When an intermediate frequency of 9 MHz is used, it is necessary for filter F 4 to be recalculated for a frequency of 136 MHz (4).

If, on the other hand, the circuit is to be used as RF-stages of a shortwave receiver, the preamplifier can be deleted and bandpass filters for the required reception range should be inserted instead of filter F 2. For shortwave reception using very short antennas, it may be advisable to use a single-stage preamplifier with approximately G_p = 12 dB, which can be switched in and out of circuit using two Reed relais.

The diplexer F 3 can be deleted in the shortwave mode, since the P 8000 matching amplifier (T 5) provides a very good 50 Ω termination on its own up to approximately 80 MHz, if it is operated with a drain current of 30 mA.

In order to reduce the lower limit frequency of the oscillator amplifier, and the preamplifier, if required, it is necessary for the emitter current chokes to be increased from 1 µH to 100 µH, and the coupling capacitors to 10 nF. If two-hole cores from material N 30 (Siemens B 62152-A 8-X 30) are used for the feedback coupling transformers, the lower limit frequency will amount to less than 1 MHz, which should be sufficient for any required oscillator and receive frequencies.

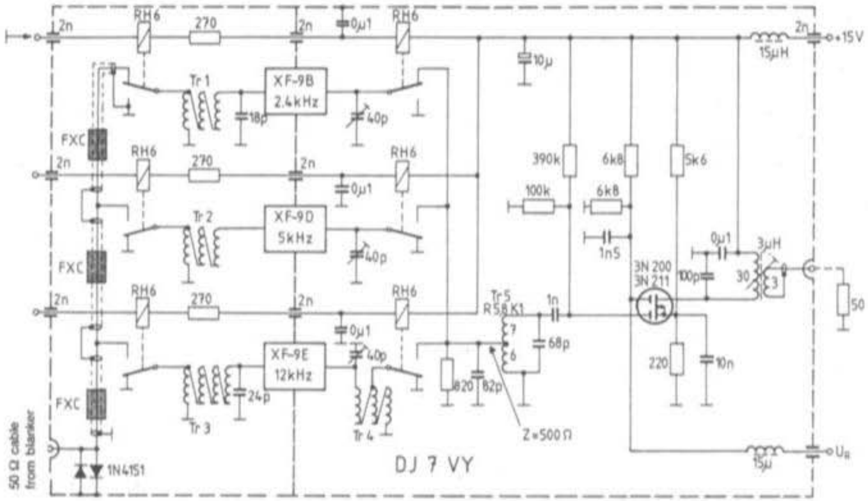


Fig. 5: Recommended 9 MHz IF-amplifier with various crystal filters suitable for use with the converter

Finally, **Figure 5** shows an example of a 9 MHz IF-amplifier including crystal filters for the various modulation modes which can be used in conjunction with the input circuit. This allows one to achieve the following at 9 MHz: G_p = 27 dB and NF = 6 dB. In the indicated position, the crystal filter XF-9 B is switched into circuit with a bandwidth of 2.4 kHz. For switching, two relais type RH-6 (National) are used per filter. The energizing coils are connected in series. The indicated screening and decoupling methods should be observed!

The three input transformers Tr 1 to Tr 3 are used for matching the crystal filters to the 50 Ω input: 500 Ω for crystal filter XF-9B and XF-9D, and to 1200 Ω for XF-9E. At the output, it is only necessary for the impedance of the XF-9E to be transformed down to the common value of 500 Ω (Tr 4), after which a resonant circuit with toroid coil transforms this to approximately 2500 Ω for the dual-gate MOSFET. **Table 2** gives the specifications for the toroid cores when used in the circuit given in Figure 5:

Tr	Dia. and Material	Siemens-No.	Turns	Wire
1	R 6, 3 N 30	B 64290 - A 37 - X 830	3 x 10	0.25 mm
2	as Tr 1			
3	as Tr 1 but		4 x 7	
4	as Tr 1			
5	R 5, 8, K 1	B 64290 - A 38 - X 1	6 + 7	0.25 mm

4. CONSTRUCTION

The converter can be accommodated on the double-coated PC-board DJ 7 VY 002 with through-contacts. The dimensions are 75 mm x 150 mm. In order to ease construction and save room, filters F 1 and F 2 (inductively coupled) are mounted on the board. Due to the smaller size in comparison to that described in (2), they possess a higher loss which in turn causes a somewhat higher noise figure; on the other hand, it increases the IP-value. In Fig. 2, the results that would be obtained using separate filters as described in (2), are also given. The image frequency rejection is virtually the same. If a VHF-transmitter in the range of 86 to 88 MHz is present with a signal strength of more than 90 dB over noise, it is possible for an absorption circuit similar to (2), **Figure 8**, to be used to increase the image frequency rejection to over 120 dB !

After winding all inductances and transformers according to the information given in (2), it is possible to commence mounting the components on the PC-board. It is important that the resistors of the 50 Ω termination (2 x 100 Ω) in F 3 are soldered into place with the shortest possible connections. If this diplexer is not required, the resistors should be removed, L 5 replaced by a wire bridge, and C 2 increased to 10 nF. When a noise blanker is not required, it is possible to delete T 4 with L 6 and C 3, and the 29 MHz output signal can be taken from the coupling capacitor between Tr 3 and T 6. In this case, T 6 can be deleted together with all subsequent circuitry, however, Tr 3 must be bridged using a 560 Ω resistor.

It is important to provide a consequent RF-tight screening of the board itself, as well as individual screening of the oscillator, mixer and input amplifier. Since the Schottky diodes in the mixer generate frequency components up to several GHz (!) when switching 50 mW oscillator power, special attention must be paid that this is not radiated. For this reason, all voltages are fed in via feedthrough capacitors, and 6-hole ferrite chokes.

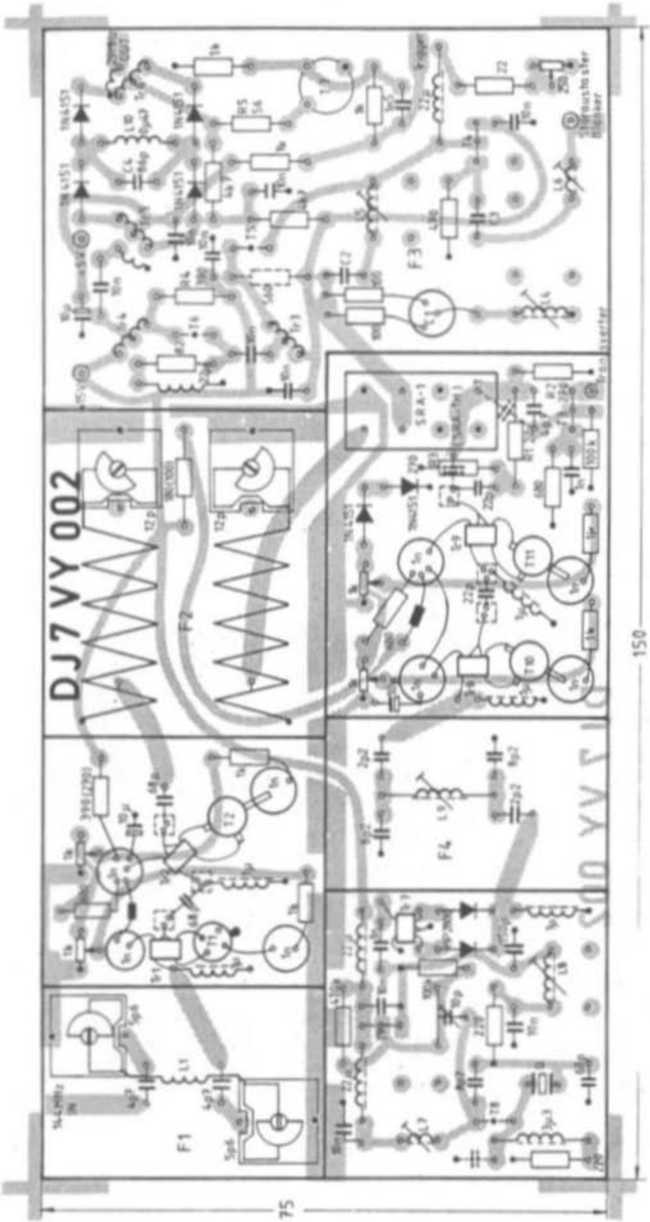


Fig. 7: Double-coated PC-board DJ 7 VY 002 with through contacts

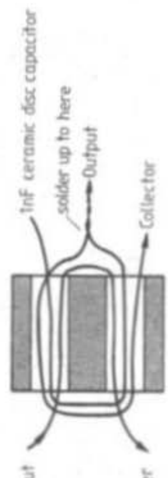


Fig. 6: Winding diagram for transformers Tr 1, Tr 2, Tr 8 and Tr 9

Intermediate panels of 30 mm in height made from 0.5 to 1 mm thick brass or tin plate have been found suitable. They should be soldered into place at the positions indicated in the component location plan. A 40 mm high metal panel is soldered into place around the edge of the board so that the upper edges of the panels are of the same height. On the lower side of the board, an approximately 8.5 mm high L-shaped screening panel is provided along the wide ground conductor with a bend at the corner in order to cater for pin 7 of the ring mixer.

If all upper and lower edges of the plates are not of the same height after the soldering has been completed, this can be obtained using a file or emery cloth.

It is favorable when the central screening panels on the upper side are soldered into place after the trimmer capacitors and resistors. After completing the board (CAUTION: R 1, R 2, and R 3 should not be soldered into place until the alignment is completed), the side panels should be soldered into place at the edge of the board; do not forget to drill the required alignment and power supply holes before mounting ! After this, the screening panel on the lower side of the board can be mounted. More cutouts must be provided in the screening panel for the lower side of the board at positions where conductor lanes are crossed. However, these should not be higher than 0.5 mm.

It is necessary for transistors T 5 and T 6 to be cooled and this is achieved using two brackets made from copper or aluminium plate and screwing these to the transistors. This can be seen in the photograph of the author's prototype given in **Figure 8**.

Tapped bushings are soldered into place at the corners for mounting the two covers. The case can be made RF-tight by glueing 0.1 mm thick copper foil into place, which is spring-loaded using 2 mm thick foam plastic. After screwing into place, this ensures that the copper foil is well grounded to all edges of the case.

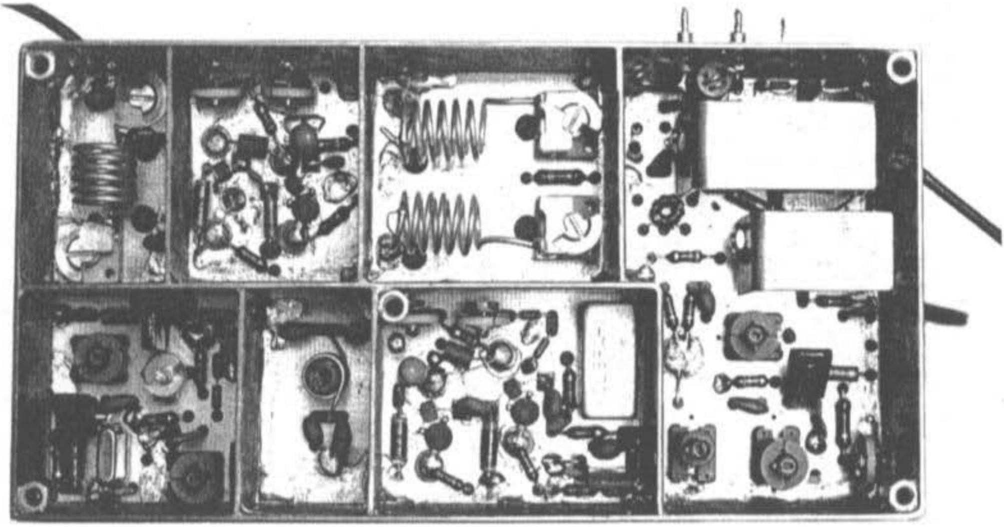


Fig. 8: A photograph of the author's prototype

4.1. Components

The given component values are valid for the 144/28 MHz version equipped with 58 MHz crystal.

T 1:	BFT 66 (Siemens)		
T 2:	BFR 34 A (Siemens)		
T 3 - T 6:	P 8000 (Texas Instruments)	T 8, T 9:	P 8000
T 7:	2 N 2219 A (various manufacturers)	T 10, T 11:	BFR 34 A

D 1, D 2:	HP 2800 or similar Schottky diode
D 3 - D 8:	1 N 4148 or similar silicon switching diode
Ring mixer:	SRA-1H or SRA-1 (MCL) or IE-500
Filter F 1:	
Trimmer:	5.6 pF tronser trimmer with 2 pins
L 1:	6 turns of 1.2 mm dia. silver-plated copper wire, wound on a 9 mm former, 10 mm long, between the stators of the trimmers (solder quickly !)
Filter F 2:	
Trimmer:	12 pF tronser trimmer with 2 pins
L 2, L 3:	6.5 turns of 1.2 mm dia. silver-plated copper wire, wound on a 8 mm former, 15 mm long, hot end soldered to the stator of the trimmer, tap 0.5 turns from the cold end; L 2 wound clockwise, L 3 wound anticlockwise !
Filter F 3:	
L 4:	2 turns of 0.3 mm dia. enamelled copper wire in special coil set (64 nH) with core (Fi 05 f 7) but without core cap and screening cover
C 1:	470 pF ceramic disk capacitor without leads
C 2:	see text, orientation value for 29 MHz: 26 pF (standard value 27 pF)
L 5:	see text; orientation value for 29 MHz: 1.2 μH. 10 turns of 0.3 mm dia. enamelled copper wire wound in a single layer in special coil set with core and cap, but without screening cover
L 6:	19 turns of 0.3 mm dia. enamelled copper wire, wound in two layers, in special coil set, without screening cover
L 7:	8 turns of 0.3 mm dia. enamelled copper wire, single layer, otherwise as L 6
L 8:	2 turns of 0.3 mm dia. enamelled copper wire, otherwise as L 6
Filter F 4:	

Two ceramic disk capacitors each of 2.2 pF and 8.2 pF

L 9:	6 turns of 0.8 mm dia. silver-plated copper wire wound on a 7 mm former; self-supporting, soldered into place in vertical position; a coil former of 5 mm dia. and 25 mm long with core Gw 3.5 x 13 of Fi 01 u 8 is placed into the coil
L 10:	0.47 μH; 16 turns of 0.25 mm dia. enamelled copper wire wound on a 100 kΩ resistor of 2.8 mm dia., coil length 3.5 mm
Tr 1:	R = 1, m = 4; n = 11 turns of 0.1 or 0.12 mm dia. enamelled copper wire, see Figure 6; R wound firstly with the ends to the left, followed by m and n, both with the ends to the right; connect one end of m to the start of n. Two-hole core Siemens B 62152 - A 8 - X 17
Tr 2:	R = 1, m = 3, n = 5 turns of 0.15 mm dia. enam. wire, otherwise as Tr 1
Tr 3:	3 x 9 turns of 0.25 mm enamelled copper wire, wound together on toroid core R 5.8 K 1 (Siemens B 64290 - A 56 - X 01)
Tr 4:	2 x 13 turns of 0.25 mm dia. enamelled copper wire, wound together on toroid core as for Tr 3
Tr 5, Tr 6:	as Tr 3
Tr 7:	4 + 2 x 4 turns of 0.12 mm dia. enamelled copper wire wound together on two-hole core as Tr 1; 4 turns below, then 2 x 4 turns wound together on top
Tr 8, Tr 9:	as Tr 2

2 ferrite beads

2 six-hole core chokes (designated as RFC in the circuit diagram for power supply line)

Other chokes: Miniature ferrite chokes (Delevan)
5 x 1 µH, 1 x 3.3 µH, 4 x 22 µH

Trimmer between T 8 and T 9: 2-12 pF plastic foil trimmer (yellow)

8 ceramic disk capacitors without leads, approx. 1 nF
1 ceramic disk capacitor without leads, 470 pF (C 1)
2 feedthrough capacitors of 1 nF, for solder mounting
3 tantalum drop-type capacitors 10 µV / 16 V
All other capacitors: ceramic disk capacitors for 2.5 or 5 mm spacing

All resistors: carbon, for 10 mm spacing
4 trimmer potentiometers 1 kΩ for vertical mounting with spacing 5/2.5 mm
1 crystal 58.000 MHz, HC-18/U or HC-25/U directly soldered to the board without socket
1 trimmer resistor 220 Ω for vertical mounting with 5/2.5 mm spacing

5. ALIGNMENT

The only equipment required for alignment are a 29 MHz receiver and a mA-V-meter with R ≧ 100 kΩ in the 5 V range (20 kΩ/V). The operating points of the preamplifier and oscillator amplifier are firstly adjusted to the nominal values. The mA-meter is then connected in series with the collector line. CAUTION: The adjustments affect another due to the temperature feedback circuit! The quiescent current of the matching transformer T 5 should be adjusted to 30 mA.

The oscillator is now aligned to the required level of 17 dBm ≙ 50 mW ≙ 4.5 V (peak-to-peak) into 50 Ω. This is made by soldering two 100 Ω resistors with short leads to one end of a short piece of 50 Ω cable and connecting the voltmeter via a germanium point-contact diode (e.g. OA 182 or 1 N 21) with one connection to the inner conductor and the other to the ground shielding of the cable in parallel with the termination. A capacitor of 10 nF should be provided as integration capacitance for the rectifier circuit. The other end of the cable should be connected with short leads (less than 2 mm length) into position instead of R 3.

The oscillator coupling trimmer of 10 pF is firstly adjusted for minimum capacitance. The oscillator is aligned with the aid of inductance L 7 so that a secure commencement of oscillation is guaranteed, (approx. 0.25 turns of the core from maximum). Inductances L 8 and L 9 are aligned for maximum voltage on the voltmeter. The coupling trimmer should now be increased in capacitance until the meter indicates 2.15 V DC, and the detuned circuit should be realigned with the aid of inductance L 7.

The cable should now be removed from the board and resistors R 1, R 2 and R 3 soldered into place. The conductor lane from pin 1 of the mixer to the vicinity of R 3 should be removed, since it is only required when using a SRA-1H. If the SRA-1H is to be used, a coupling capacitor of 22 pF should be installed so that it is connected to pin 1 of the mixer with the aid of the direct stripline (shown as a dashed line in the component location plan). In this case, the line to R 1 / R 2 should be cut. The 4.7 pF capacitor to the buffer for transverter operation should be soldered directly to pin 1 of the mixer on the lower side of the board.

If the voltage at the oscillator output of the mixer is measured with the aid of a RF probe, it will be less than previously measured due to the limiting effect of the diodes. However, since it is the switching **current** that is required for the large-signal characteristics, this measurement is not important !

After completing the alignment of the oscillator, the input of the converter is connected to an antenna and a constant 2 m signal, e.g. from a beacon transmitter, is tuned in on the subsequent 10 m receiver. Filters F 1 and F 2, as well as L 4 and L 5 in filter F 3 are now aligned for maximum field strength. A transformation of the preamplifier input impedance to the impedance of the feeder cable can also be carried out when aligning F 1. The standing wave ratio (VSWR) should reach a minimum at this point since a power matching is used with this type of preamplifier. In the case of a 50 Ω feeder, both trimmers should have the same capacitance, whereas with 60 or 75 Ω cables the antenna trimmer will have slightly less capacitance. This is only valid, of course, when the antenna is perfectly matched to the feeder cable, e.g. with VSWR = 1.

Inductance L 6 is aligned together with the noise blanking circuit. The noise blanking switch is working correctly when a 80 dB 2 m-signal disappears on grounding the trigger input.

Finally, **Figure 9** shows how a transmit mixer can be connected to the oscillator output of the described receive converter.

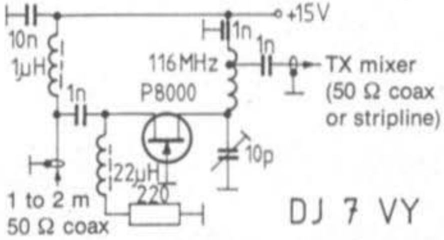


Fig. 9: Buffer for driving a transmit mixer

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MORE DETAILS ON RECEPTION OF THE EUROPEAN WEATHER SATELLITE »METEOSAT«

by R. Lentz, DL 3 WR

Due to the great interest in this satellite, more details are to be given regarding the transmitted images and about the levels of processing these images. Basic information was given in (1) and (2) in the last edition of VHF COMMUNICATIONS. As was mentioned in (1), the METEOSAT weather satellite is located above the intersection of the zero meridian and the equator at a height of approximately 36.000 km. Its image processing system always sees the same section of approximately 33% of the total surface of the earth as can be seen in **Fig. 1.**

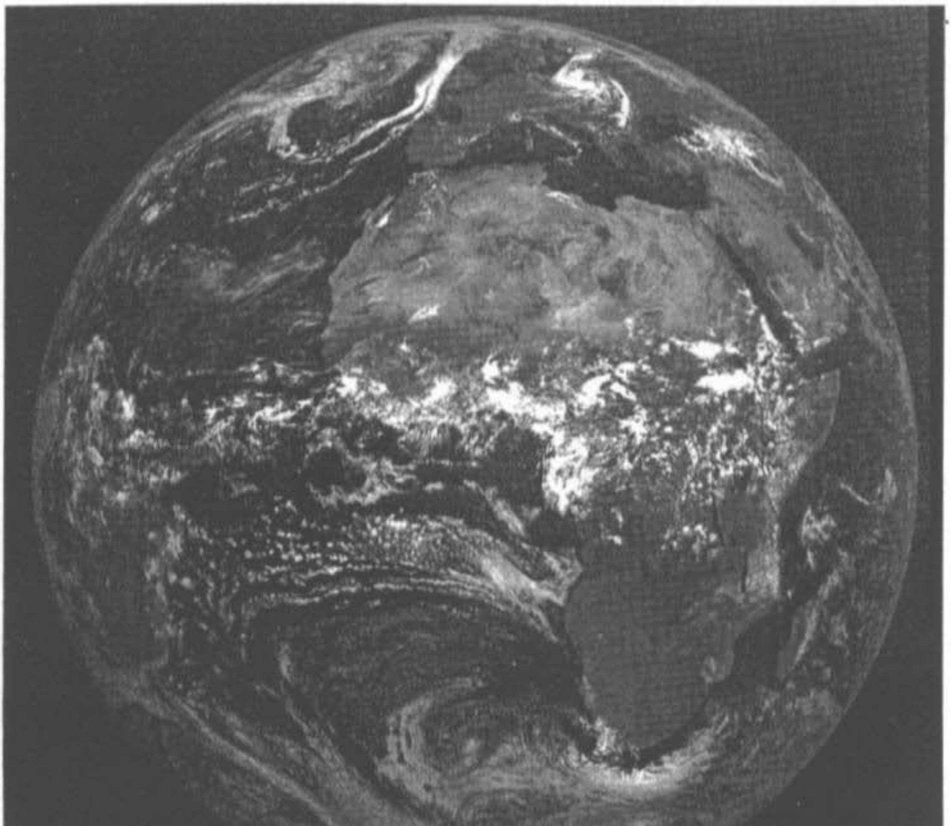

Fig. 1: Photograph of the earth as transmitted as a raw image by METEOSAT

This so-called raw image originates from a radiometer, a system that cannot really be classed as a camera. It is more like a mirror-telescope that scans the earth over a period of 25 minutes. A mirror system distributes the received radiation into three spectral ranges and reflects these to three detectors or sensors. One of these is sensitive to (visible) light (500 - 1000 nm), a second to infrared radiation at 10500 - 12500 nm (heat radiation), and the third to infrared radiation at 5700 to 7100 nm (hydrogen absorption band).

The signal from the selected detector is amplified, filtered, analog-digitally converted and stored. Since the earth only covers an angle of 18° at a distance of 36.000 km, the stored image will contain 19 parts space to one part of the earth surface. For this reason, the sector of interest is electronically expanded by 20 times so that a continuous transmission of digital image signals of the earth is made to the ground station.

HIGH-RESOLUTION DIGITAL TRANSMISSION AND PROCESSING OF THE RAW IMAGE

The main ground station of the METEOSAT satellite is located southeast of Frankfurt in West Germany. The received raw image is transmitted by a cable to the Operations Center of the European Space Agency in Darmstadt (ESOC). The images are then fed to one of the largest data processing systems in Europe, where they are geometrically corrected, calibrated and evaluated. In order to obtain an idea of the extent of this task, one should know that METEOSAT supplies one image every half hour in the visible and infrared range (water vapor images less frequently). The images in the visible spectral range consist of 5000 lines of 5000 points per line, in other words of 25 million image points. Each image point has one of 64 possible brightness steps. The infrared images are formed of 2500 lines of 2500 video points each, whereas the intensity information of each point is coded into 256 steps. The water-vapor images, finally, consist of 6.25 million points as the infrared images, but each point is »only« coded in 64 steps.

It should be mentioned at this point, that the previously mentioned number of image points would allow a theoretical resolution of 2.5 km in the visible range, and 5 km in the infrared range. These optimum values are valid for the area directly below the satellite (Gulf of Guinea).

METEOSAT AS TRANSPONDER

The raw images are processed and evaluated in a number of ways in the Darmstadt data processing system to make them more suitable for weather research and forcasting. METEOSAT is also used as transponder (relay transmitter) for distribution of selected images to meteorological centers, research institutes, and other interested parties throughout Europe. These selected images are transmitted to the satellite from the ground station near Darmstadt and retransmitted from the satellite at frequencies of 1691.00 MHz and 1694.50 MHz. It is these transmissions that are of interest to radio amateurs and amateur weather observatories.

DISSEMINATION – DIVISION OF THE RAW IMAGE

The images that can be received at the two given frequencies, are segments of the original image in the so-called APT format. The processed raw image is split into 24 square images at the ESOC in Darmstadt in the case of images in the visible range, or into 9 sectors in the case of infrared images. **Figure 2** shows the image segments for images in the visible spectral range, and **Figure 3** the larger areas for infrared images.

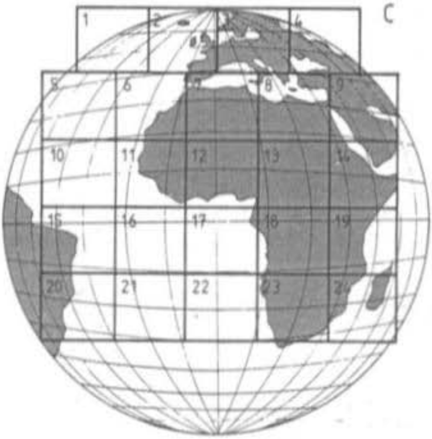

Fig. 2: Image segments in the visible range

Fig. 3: Image segments for IR-images

Each of these sectors is provided with a data line giving date, time (GMT), type of image (visible = VIS, infrared = IR), image segment (e.g. CO 3 = Central and Eastern Europe), as well as coded information with respect to image quality. When required, the computer system will also provide coordinate lines with a spacing of 10° and/or contours of continents or countries.

These images are now transmitted on the two frequencies of 1691.0 MHz and 1694.5 MHz according to a changing time plan. Normally, only IR-images (D and E) are transmitted during the night, images in the visible range (VIS) are added to them in the morning and evening hours (especially segments C02 and C03 that are important for Europe), whereas during the day time mainly VIS-images (C) are transmitted.

WHAT IS APT ?

APT stands for Automatic Picture Transmission and is a world-wide image transmission system used for optical/mechanical image production, in other words is used with a relatively slow transmission speed. In principle, it is similar to slow-scan TV as used by radio amateurs. However, the METEOSAT images are transmitted with a higher resolution – both with respect to the number of image points, and to the number of grey steps. The APT-format used for transmission of METEOSAT images is shown in **Figure 4**. The given times correspond to a speed of 240 lines per minute, with 840 image points per line (40 white points, and 800 usable image points). A total of 3 minutes and 33 seconds are required per image including the 300 Hz commencement tone, 5 s of white and 450 Hz completion tone. The transmission time plan provides a pause of 27 seconds between each image, which means that a new image is transmitted every 4 minutes.

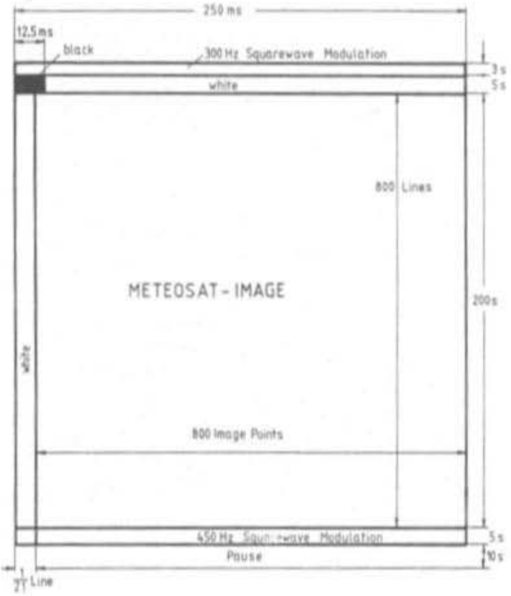

Fig. 4:
APT WEFAX format for METEOSAT transmissions

TECHNICAL REQUIREMENTS FOR RECEPTION OF METEOSAT IMAGES

Professional Reception

A professional receive station for the METEOSAT transponder frequencies 1691.0 and 1694.5 MHz is shown in form of a block diagram in **Figure 5**. Either phased array antennas with a side length of 1 m x 1 m (Rohde & Schwarz) or parabolic dishes of 2 to 2.5 m diameter are used. This is followed by a low-noise preamplifier equipped with several bipolar or GaAs-FET stages, a frequency converter to 137.5 MHz, FM receiver with subsequent AM demodulator and lowpass filter, and finally an image recorder. The most important specifications are given in this block diagram.

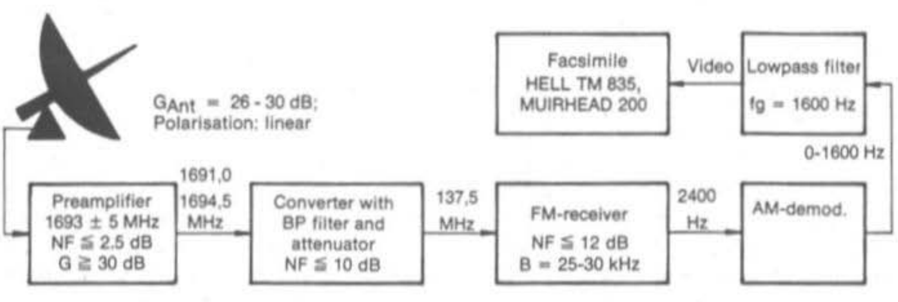

Fig. 5: Block diagram of a professional receive system

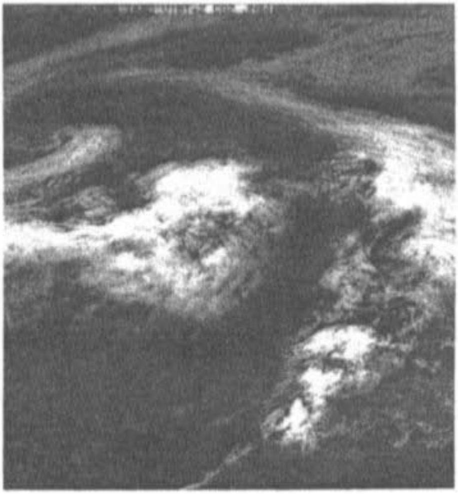

Fig. 6 + 7: Weather pictures taken over Europe on the 5.9.78 at 14.30 GMT

Such a system will not be realized with amateur means, mainly due to two reasons: A multi-stage, low-noise preamplifier and matching image recorder are most certainly out of the price-range of most amateurs, and are virtually unavailable on the surplus market. However, experience has shown that it is possible to receive METEOSAT images using a simpler system. For instance, the photographs given in **Figures 6 and 7** were received by Manfred Fütterer, DC 6 FM, in Bochum, West Germany, using a 2 m parabolic antenna without pre-amplifier, and a converter as described in (6).

Demodulation of METEOSAT's FM/AM Signal

The carrier frequency of 1691.0 or 1694.5 MHz are frequency-modulated with a sub-carrier of 2400 Hz. The maximum frequency deviation is 9 kHz, which means that a transmission band-width of a maximum of 26 kHz is required. This means that a (crystal) filter is required in the IF amplifier that exhibits a 3 dB bandwidth of 26 kHz. If a filter having a lower bandwidth is used, modulation distortion will occur in the same manner as encountered during FM operation. If, on the other hand, a filter having too large a bandwidth is used, this will deteriorate the signal-to-noise ratio. There are a number of surplus filters on the market which were used for the old FM channel spacing of 50 kHz. These exhibit a bandwidth of approximately 30 kHz.

In the FM mode, the AF signal-to-noise ratio will increase more quickly than the RF signal-to-noise ratio on increasing the signal as soon as the threshold, which is dependent on the modulation index, is exceeded. This is designated as demodulation gain. When used in conjunction with the transmit characteristics of METEOSAT, this means that at a RF signal-to-noise-ratio, measured between the filter and the demodulator in the IF-amplifier, of 15 dB, a demodulation gain of approximately 20 dB will be added. This means that the audio frequency video signal would have a signal-to-noise ratio of approximately 35 dB.

The 2400 Hz carrier recovered after FM demodulation, is amplitude-modulated with the video signal. The maximum degree of modulation amounts to 80 %, which represents the white level; zero percent modulation correspond to black level. Frequencies of 0 to 1600 Hz are present in the AF-video signal.

In order to achieve the signal-to-noise ratio mentioned, it is necessary for the video bandwidth to be limited to 1600 Hz using a lowpass filter. Such a lowpass filter is already given in the block diagram.

Finally, to close this section a practical measured value: M. Fütterer, DC 6 FM, was able to obtain an IF signal-to-noise ratio of 17 dB when using a 2 m parabolic dish and a converter as described in (1) and (6) with NF = 9.5 dB. This resulted in noise-free images.

Fig. 8: Reception of METEOSAT signals by DJ 1 JZ (not in picture) and DL 3 WR

Amateur Reception of METEOSAT Images

Nearly noise-free signals were received in July 1978 by the author using a system, which is probably the minimum possible **(Figure 8)**. A parabolic dish of only 1.2 m diameter was equipped with a tubular radiator as described in (5), recalculated for 1693 MHz, and the received signal was fed to an interdigital converter **(Figure 9)**. The converted signal was then fed to a home-made FM-receiver. The most important characteristics are given in **Figure 10**.

However, the most decisive contribution to amateur reception comes from Dr. K. Bauer of Stuttgart University, Institute of Physical Electronics: Replacement of the video recorder by an electronic reproduction system for satellite images which can be constructed by radio amateurs.

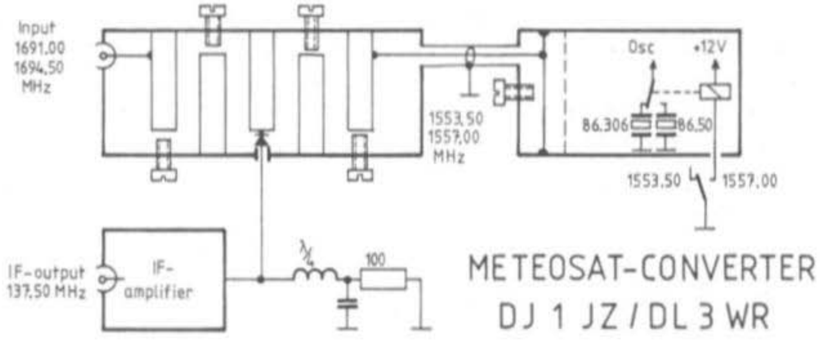

Fig. 9: The METEOSAT receive converter used

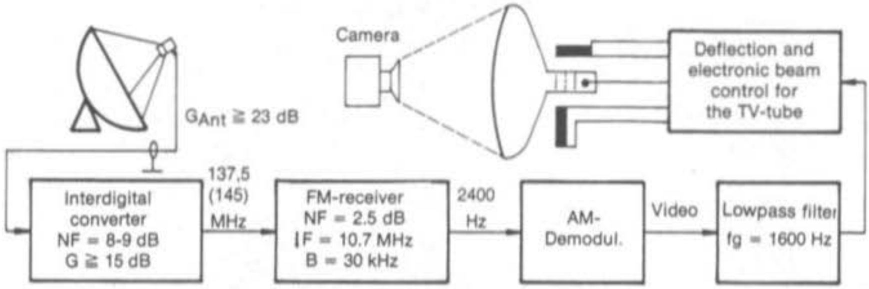

Fig. 10: An amateur receive station for METEOSAT transmissions

This unit can be switched to various standards, and is equipped with error detection and correction circuits. Since not all of this is required by radio amateurs for METEOSAT reception, a simplified version is to be developed, which will be described in VHF COMMUNICATIONS at a later date. For this reason, the following description is limited to the block diagram and a short description by Dr. Bauer.

AN ELECTRONIC DISPLAY SYSTEM FOR SATELLITE IMAGES

The display system, whose block diagram is given in **Figure 11**, is suitable for all video signals operating according to the APT-system. This means that it is suitable for all weather satellite transmissions such as in the NOAA and METEOSAT programmes. The images are produced in real time – e.g. within 3.6 minutes in the case of METEOSAT – and are written on the screen of a 31 cm TV-tube. In order to store the image, a camera is mounted on a tube in front of the screen so that ambient light is not present, and the image photographed in a similar manner to oscilloscope traces. The camera shutter remains open for the whole period of 3.6 minutes. The use of polaroid cameras is very advantageous, since the image is immediately available; for higher-quality images, 35 mm cameras or similar can be used.

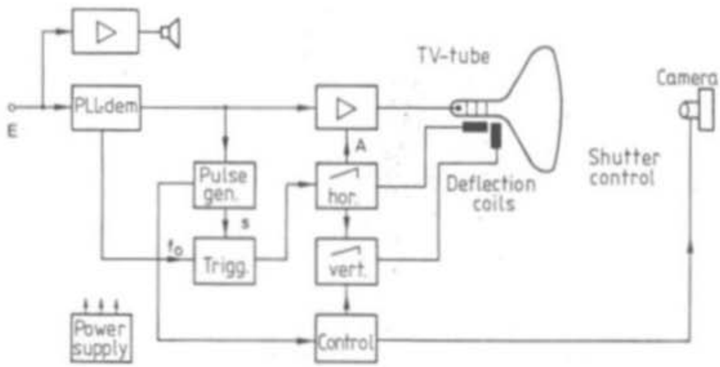

Fig. 11: Block diagram of the electronic APT system (Dr. Bauer)

The audio-frequency video signal and the synchronizing signals are modulated on a 2.4 kHz subcarrier in the APT-system. Demodulation is made using a PLL circuit which simultaneously generates the fundamental frequency f_0 for phase-locked triggering of the electronic beam with the synchronizing signals.

The video signal is fed to the cathode of the TV-tube where it modulates the amplitude of the electron beam, e.g. brightness. The pulse processing and trigger circuit feed the deflection system for horizontal and vertical deflection, and also supply the start and stop pulses for actuating the shutter of the camera. An audible signal is actuated at the end of the image. In the case of motorized cameras, it is possible for the film to be transported automatically. The complete video display is about the size of a small portable TV-receiver of 32 cm x 32 cm x 32 cm and can be operated from 12 V DC, or 220 V AC.

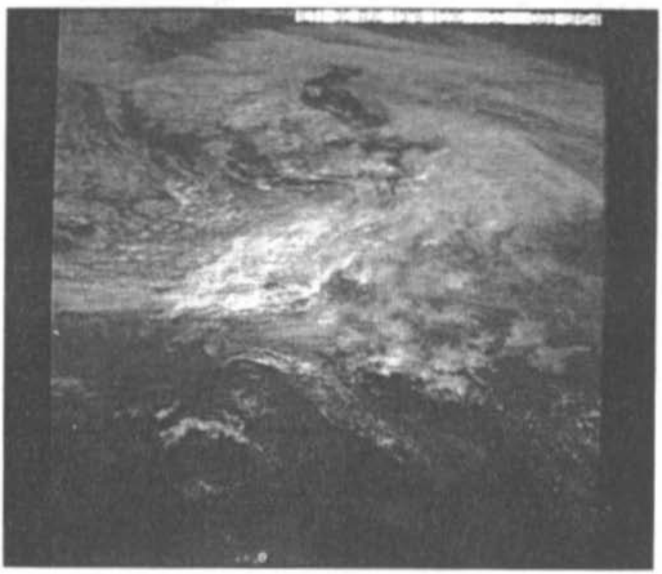

Fig. 12: Southern and central Europe on 30.8.1978 at 10.30 Z, received on the above system

Figures 12 and 13 are two examples of the quality of the METEOSAT images that can be photographed using this system. Figure 12 was taken with a polaroid camera, whereas for Figure 13 a plate camera was used. The test pattern which was also taken via the METEOSAT satellite, shows a few small geometric errors, that are caused by the curvature of the picture tube.

A much simpler but basically similar system is described in (9).

Fig. 13: ESA test pattern showing the quality of the receive system

RECEIVING TRANSMISSIONS FROM THE NEIGHBOURING GEOSTATIONARY SATELLITES

As was shown in Figure 1 of (1), neighbouring satellites of the same series are to be found 70° west and 70° east of METEOSAT (at present both American satellites). M. Fütterer, DC 6 FM, calculated that if the ground station has a good clear location, satellites with a synchronized orbit will be visible at ± 75° from that geographical location. This means that geostationary satellites up to 68° west and 82° east will be visible from Bochum (7° east), which has been proved in practice, or in the case of a line Augsburg-Nürnberg (11° east) from 64° west to 86° east. According to the geographical position, and exact position of the satellite, it should always be possible to receive at least one of the neighbouring satellites. UK ground stations should be able to receive both the GOES-SMC satellite at 70° W and the American satellite at position 70° east in addition to METEOSAT.

It should be mentioned that the American GOES satellites do not use the second frequency 1694.5 MHz, but use 1691.0 MHz exclusively for transponder operation.

ELEVATION AND AZIMUTH POSITION OF THE ANTENNA

A method of calculating this was given in (2). After publishing this information, DC 6 FM provided the author with reference (8) which contained a diagram for this. It is simple to use and is to be explained with the aid of an example.

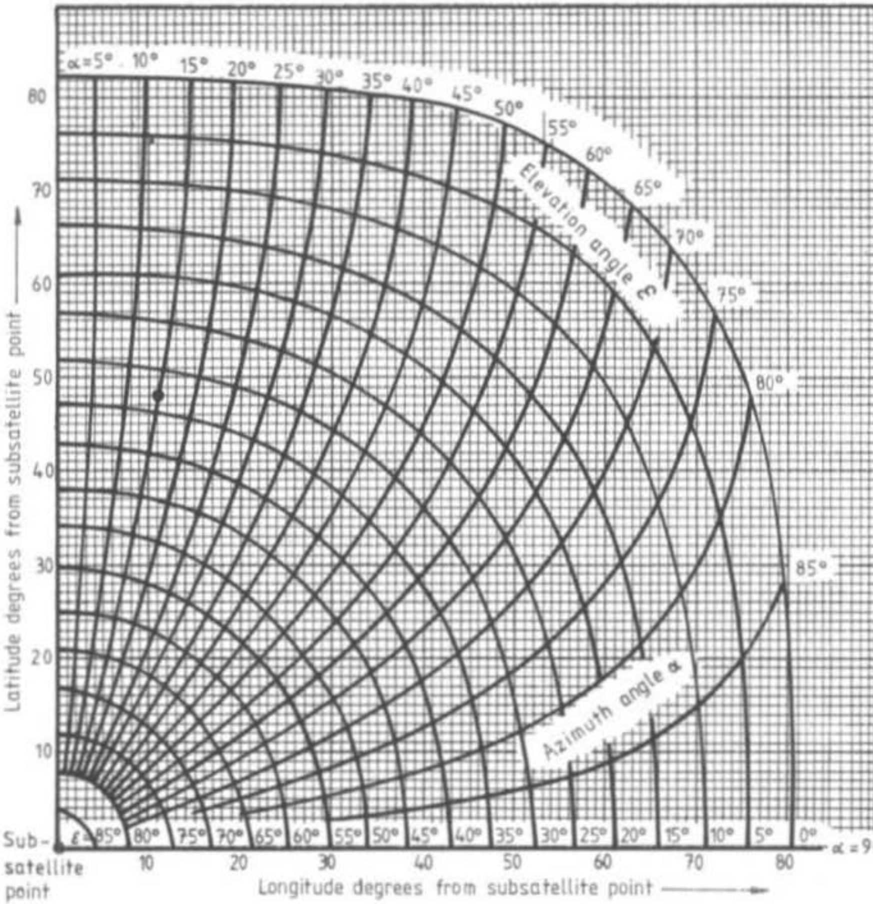

Fig. 14: Azimuth and elevation angles can be taken from this diagram

Figure 14 can be used for any geostationary satellite. However, it is very easy to use in conjunction with METEOSAT, since the sub-satellite point as origin of the coordinates corresponds to zero degrees of longitude, and latitude. It is only necessary to go to the right by the amount of longitude of the antenna location, and to go up to the value of latitude; the azimuth and elevation angles corresponding to this point can be read off directly. As an example, let us select the antenna coordinates for METEOSAT reception in Augsburg/West Germany. Augsburg is approximately 11° east and 48.4° north. It is therefore necesssary for us to go 11° to the right in Figure 14, and then approximately 48° upwards. This point is marked in the diagram and provides us with the following angles for the antenna: Elevation angle ≈ 34°, azimuth angle ≈ 15°. This means that the antenna must be pointed to a point in the sky approximately 34° over the horizon and 15° to the west of true south.

As an example for stations located to the west of the Greenwich meridian, let us determine the values for Liverpool, England: Coordinates for Liverpool: 53° 20.5' N and 2° 52.8' W. When entered into the diagram (Figure 14), this results in the values:

$\alpha \approx$ 4° to the east

$\varepsilon \approx$ 29° elevation

Stations located south of the equator should use the diagram accordingly.

PHOTOGRAPHS:

Fig. 1:	Copyright ESA
Fig. 6 + 7:	M. Fütterer, DC 6 FM, Bochum Observatory
Fig. 8:	Author
Fig. 12 + 13:	Dr. Bauer, Stuttgart University, Institute of Physical Electronics

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A 1268 MHz LOCAL OSCILLATOR MODULE FOR DF 8 QK 001

by U. Beckmann, DF 8 QK

This local oscillator module for 1268 MHz was mentioned briefly in edition 2/1978 of VHF COMMUNICATIONS, and is now to be described in more detail complete with PC-board and alignment details. Since modifications have been made to the design in the area of T 5 and inductances L 8 and L 9, the modified circuit diagram, and inductance values are to be given once again.

1. CIRCUIT DESCRIPTION

The circuit of the first four stages is very similar to that of module DJ 4 LB 003. The crystal oscillator frequency of 70.444 MHz will have been multiplied six times up to this point, as can be seen in **Figure 1**. The main feature of the described circuit is the tripler circuit using a transistor (T 5). The efficiency of this circuit, which is equipped with a BFR 34 A, is considerably higher than when using a tripler equipped with a varactor diode. The matching at 422 MHz is made with the aid of a Pi-filter comprising inductance L 6, so that transistor T 5 can be driven in the most favorable manner with the 25 mW that is available. An output power of approximately 15 to 20 mW is available at 1268 MHz at the output of the two-stage λ/4 bandpass filter.

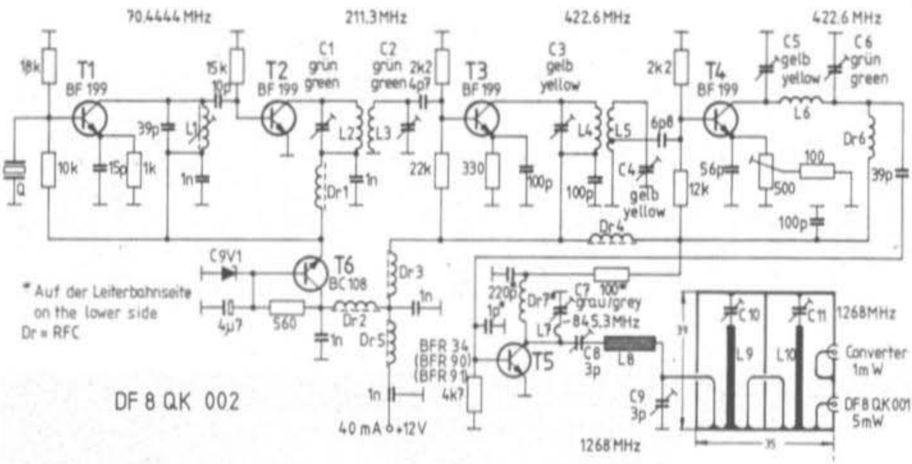

Fig. 1: Local oscillator module with transistor multiplier

Components C 8, L 8, and C 9 are used to match the collector of T 5 to the bandpass filter. Since frequencies of 422 MHz and 844 MHz are present at the collector in addition to the required frequency, it is necessary for a termination to be provided for these if maximum efficiency is to be obtained. An idler circuit comprising inductance L 7 and trimmer capacitor C 7 is provided for 844 MHz; for 422 MHz, no idler circuit is provided so that the alignment does not become too critical. However, the inductance of the collector choke Ch 7 is sufficiently low to ensure that the 422 MHz frequency is practically short-circuited.

A clean output signal is extremely important when a hybrid ring mixer is to be used. For this reason, the bandpass filter is built up coaxially using PC-board material on the ground (component) side of the PC-board. The output power can be matched to the mixer by aligning the input trimmer of T 4, and by varying the spacing of the output coupling link.

2. CONSTRUCTION

The local oscillator module is accommodated on a double-coated PC-board of 170 mm x 50 mm. The components are mounted on the ground side of the board. The position of the bandpass filter box on the PC-board is shown in the component location plan, given in **Fig.2**. This box can be made from PC-board material or from metal plate. Its height amounts to approximately 20 mm. The wire of L 8 is directly soldered to the coating of the spindle trimmers C 8 and C 9. The photograph given in **Figure 3** shows further details regarding the construction. The four components indicated with a star in the circuit diagram that are in the vicinity of transistors T 5, and T 5 itself are soldered into place on the conductor side of the board. Attention must be paid during this that the small 1 pF capacitor at the base of T 5 is kept as short as possible. The connection for the emitter of transistor T 5 is made using an approximately 2 mm wide copper foil strip. Choke Ch 7 is by-passed using a disk capacitor (chip) without leads. The ground surface around the hole for trimmer C 8 is removed for approximately 1 mm so that no short-circuit occurs. All grounded components are soldered to the top and bottom of the board.

After mounting the components, the PC-board should be provided with approximately 40 mm high screening panels made from PC-board material or metal plate.

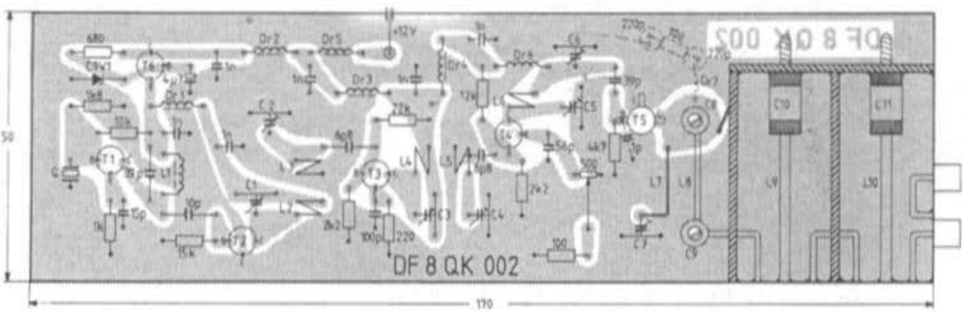

Fig. 2: Component locations on PC-board DF 8 QK 002

Fig. 3: Photograph of the author's prototype

2.1. Component Details

Silver-plated copper wire of 1 mm diameter is used for the inductances L 1 to L 7.

L 1:	4.75 turns wound on a 6 mm coil former with core (red)
L 2, L 3:	1.75 turns wound on a 5 mm former, self-supporting, spaced 2 to 3 mm from the board
L 4, L 5:	1.75 mm turns wound on a 4 mm former, self-supporting, spaced 1 to 2 mm Tap on L 5: 0.5 to 0.75 turns from the cold end.
L 6:	As L 4, self-supporting, spaced 2 mm from the board
L 7:	25 mm long, shaped as shown in the component plan, spaced 1 to 2 mm from the board
L 8:	17 mm long, from 2 mm dia. silver-plated copper wire
L 9, L 10:	32 mm long, from 2 mm dia. silver-plated copper wire
Ch 1 - Ch 4:	4.5 turns through ferrite bead
Ch 5:	6-hole score, ferrite choke (Philips)
Ch 6:	3 turns of enamelled copper wire wound on a 3 mm former, self-supporting
Ch 7:	2 turns of a 0.5 mm dia. silver-plated copper wire, wound on a 4 mm former, self-supporting, mounted on the conductor side of the board
Q:	Series resonant crystal, 3rd overtone, HC 18 or HC 25, 70.444 MHz
C 1 - C 7:	Plastic foil trimmer, 7 mm dia., 22 pF (green), 12 pF (yellow), 6 pF (grey)
C 8 - C 11:	Ceramic miniature spindle trimmer of 3 pF

3. ALIGNMENT

No further details need to be given for the alignment of stages T 1 to T 4. The emitter trimmer of T 4 should be placed in its fully anti-clockwise position, C 5 set to half capacitance, and C 6 and C 7 set to approximately 1/4, C 8 to 2/3, and finally C 9 virtually at full capacitance.

An absorption wavemeter is now placed into the box of the bandpass filter and tuned to the required frequency of 1268 MHz. Trimmers C 7 to C 10 should then be aligned to maximum. The output frequency of 1268 MHz should be available at the output after aligning C 11. In order to achieve maximum output power, it will be necessary to align trimmers C 5 to C 11 several times.

THE 10 GHz AMATEUR BAND
Considerations of Present and Future Technologies

Based on a Lecture at the 1977 VHF Convention in Weinheim

by D. Vollhardt, DL 3 NQ

Activity on the 3 cm amateur band has increased phantastically since 1977, as can be seen in the number of interest in contests. This favorable trend is partly due to recent publications (Ref. 1-4), but also due to the fact that components are available at reasonable prices. The availability of ready-to-operate modules (Gunn-oscillators, receive mixers, horn antennas etc.), has also played its part in increasing activity of this amateur band. Now it is probably the time to discuss the possibilities and limits of the different systems and modes by comparing them and to underline the necessity of coordination to interested radio amateurs.

1. SIGNAL GENERATION IN THE 3 cm BAND

The components that can be used for RF frequency generation in the 3 cm band require various different circuit technologies, and partially very different operating techniques. Without going into details, the various possibilities are to be discussed in the following sections.

1.1. Gunn Oscillators

Gunn elements have probably played the most important part in the rapid expansion of activity on the 3 cm amateur band. Gunn diodes allow transmit and receive oscillators to be constructed simply and at relatively low cost. Signal processing using integrated circuits ensures that radio equipment equipped with Gunn oscillators is very compact, easy to construct, and if required can operate for many hours from an accumulator.

Such equipment is very suitable for portable operation.

1.1.1. Simple Oscillators and Self-Excited Mixers

The frequency stability of Gunn oscillators is not too good, and if construction is not favorable, they can even be extremely unstable. For this reason, wideband FM-technology is prevalent. The IF bandwidths are in the order of 200 to 300 kHz similar to that used in VHF-FM broadcasting. For this reason, a VHF-car radio tuned to 104 MHz ± 4 MHz is often used as IF-amplifier (3), (5).

The Gunn oscillator can also be »misused« as self-excited mixer and thus be used simultaneously in the transmit and receive mode. Since the receive frequency will be shifted to the value of the IF-frequency used, it is necessary for the partner station to use the same intermediate frequency and for their Gunn oscillator to be tuned to the receive frequency of the first station. This is sometimes known as the »IF shift method«, and was also described in (5). However, the Gunn-element is a very insensitive mixer, which can be 20 dB down on a »true« mixer diode operating correctly.

This loss can be avoided when a real mixer is used and driven with the oscillator signal via a directional coupler. Such a construction was described in (6); however, two antennas will be required. A more elegant solution using only one antenna can be constructed in a transceiver equipped with a circulator. Unfortunately, such a circulator still costs approx. DM 200,—.

One handicap remains with all these methods:

Station A can communicate with station B and also station C, however no communication will be possible between B and C, unless one of the two detunes their oscillator to the value of the IF (100 or 30 MHz), or if both by half this amount (in the correct direction !). With the type of construction usually used, this is only possible by changing the effective dimensions of the cavity resonator mechanically, in other words by rotating either poorly or non-calibrated screws and plungers. What this can mean when sitting on the top of a tower in bad weather with gusty winds and drizzle, is best illustrated by a commentary heard regarding the first GHz-contest in 1977: »At first, station X knew roughly at which frequency he was operating; however, after the first few QSOs everything was completely misaligned. While they were shifting all over the band, other stations were chasing after the »new«! carrier which was rather weak, but was running around all over the place .....« Of course, all new technologies are difficult at first; however, this state of affairs cannot continue indefinitely.

1.1.2. Gunn Oscillators with Varactor Tuning

The first technical improvements were provided by modules that were described in (3), which used a tuning varactor in the resonator of the Gunn diode. This type of electrical tuning is possible within a range of at least 60 MHz. Also the mixer diode is integrated in a simple circulator in these modules, which means that a single antenna can be used without switching. Such an arrangement possesses two main advantages:

- Tuning can be made with the aid of the varactor voltage which allows a simple but accurate calibration of the frequency.
- The microwave circuits can be mounted separately from the operating portions and remote-controlled. This is not only suitable for occasional portable operation (especially in bad weather !), but also for fixed-station operation.

By the way, the built-in varactor cannot only be used in conjunction with a good automatic frequency control (AFÇ), but also used in conjunction with a simple circuit to allow automatic frequency scanning. Furthermore, the frequency modulation need not be forced upon the Gunn element, but can be made using the varactor voltage in a sensible manner, right up to bandwidths suitable for TV transmission.

An electrical tuning range of only approx. 60 MHz requires a lower IF than 100 MHz if the previously mentioned problems of a mechanical tuning are to be avoided. Such a concept using a 30 MHz IF, AFC, and wideband FM, is described in (7). Due to the handiness of this system, hundreds of these have been sold in Germany and Switzerland in the meantime, and several dozens are in operation during contests.

1.1.3. Frequency-stabilized Gunn Oscillators

Varactor tuning is also required for the next step in the direction of frequency stability and reliable communications. This offers not only the possibility of locking one's own Gunn oscillator to the receive signal using a frequency discriminator (AFC), but also allows the transmit signal to be locked to a stable reference signal in the transmit mode. If a phase discriminator is used, this will then represent a PLL-system.

In the case of the »Gunnplexer« briefly described in the previous section, which transmits and receives simultaneously, it is basically only necessary for a pilot signal to be fed to the antenna that has a sufficient frequency spacing from the receive signal. This signal is filtered out in a separate IF-amplifier and provides a reference signal that can be processed relatively simply.

An other method is to extract a small portion of the Gunn oscillator energy and to feed this to a separate mixer. This method is also possible for systems where transmission and reception are not made simultaneously. Something similar is mentioned in (3), and various different versions would be possible.

One may ask whether such extensive measures are worthwile. The answer is that the disadvantages of the more complex construction are no longer so difficult due to the advances made in integrated technology, and are more than offset due to the advantages of the reproducibility of the frequency. This is especially true when the limits of possible antenna gain for the radio path have been achieved and when the beamwidths of the antennas are down to a few degrees.

Since the typical frequency drift of the Gunn oscillator is deleted when using PLL stabilization, it is possible for the bandwidth of the IF amplifier to be reduced and for defined channels, or pairs of channels to be designated. Such PLL-systems are also being developed, and there is no doubt that they will be utilized on the 3 cm band as they are at lower frequencies. However, FM-systems will probably always be used since the required frequency stability and low-noise required for the true narrow-band modulation modes of CW and SSB will not be realized in this way.

Practical operation of such equipment requires, however, that a frequency plan is agreed for the 10 GHz band. Only then will compatability be guaranteed. A preliminary band plan is given in appendix II.

1.2. Impatt Oscillators

Virtually the same is valid for IMPATT diode oscillators as was given in section 1.1. for Gunn oscillators. They are able to generate power levels in the Watt-range, however, require operating voltages between 50 and 150 V. Their efficiency of approximately 10 % is higher than that of Gunn diodes, as is the price. This component will probably not have a great interest for amateur applications in the 10 GHz band.

1.3. Frequency Multiplication using Varactor Diodes

Signal generation using frequency multiplication will become just as important on the 3 cm band as on the lower microwave bands. This will not be just a few Microwatt for PLL-systems or several Milliwatt for the first mixer of narrow-band receivers, but it will be also possible for higher transmit powers to be generated than would be economical when using Gunn diodes.

It will be seen later that real long-distance communications are only possible when using narrow-band techniques, e.g. in CW and later also in SSB. The interest in this form of 3 cm communications will most certainly increase in the next few years, however, due to the technical complicity, it will firstly be realized for fixed-station operation. The frequency range of 10368 to 10370 MHz has been set aside for the narrow-band modes.

For instance, if one possesses a crystal-controlled 1296 MHz transmitter for the 23 cm band with say 10 W output power, it is possible for a frequency of 10368 MHz to be achieved with an output power of approximately 1 W by frequency multiplication by eight in several steps. A similar result can be achieved when a 1152 MHz signal is available, which is often used as local oscillator for the lower microwave bands. This signal should be amplified up to an output of 10 W, and multiplied by nine (2 x 3 times). The main problem in the realization of such transmitters is the high price of the power-varactor diodes, however, Gunn diodes and IMPATT diodes having a similar output power are also expensive !

A SSB signal can also be processed in a similar manner. The method described in (10) where a varactor diode is used as power up-converter, offers interesting possibilities.

1.4. Transistors for Transmitter and Receiver Applications

The development of the microwave semiconductors of the gallium arsenide field effect transistors (GaAs-FETs) has been very rapid within the last three years. At first, attention was mainly paid to low-noise preamplifiers. There are some types available on the market that offer a noise figure of approximately 4 dB and a gain of at least 6 dB at a frequency of 10 GHz (Hewlett-Packard HFET 1101). Unfortunately the price is still high (DM 370,— in March 1978), and their use is still somewhat problematic. However, it is assumed that such transistors will be produced in large quantities for the 11 GHz TV system, which means that they will also be available at a reasonable price to radio amateurs.

Recently, great advances have been made in the production of power GaAs-FETs, however, the high price will limit their use for amateur applications for some time.

1.5. Beam Amplifier Tubes

This includes reflex klystrons, amplifier klystrons and travelling wave tubes. The smaller reflex klystrons such as type 2 K 25 can only be »squeezed up« into the amateur band with great difficulty and overloading them mechanically. The advances made in Gunn technology have replaced them completely. In addition to this, there are only a few types that are able to operate in the frequency range of 10250 to 10500 MHz – for example type X-13 that is able to generate some 100 mW. However, the stability is inferior to that obtained using good Gunn oscillators, and an extensive power supply is required.

Amplifier klystrons are relatively narrow-band and do not usually operate in the amateur band. In contrast to this, travelling wave tubes are extremely wideband, which means that several types operate well in the amateur band. They would be able to solve all the problems of a 3 cm amateur radio station with their gain values of 25 dB and output powers in the Watt-range – if one is lucky enough to have one ! It is necessary to also have the matching focalizer, and a quite complicated power supply. At present, travelling wave amplifiers are virtually non-existent in amateur radio applications, and for this reason they are not to be discussed in detail here.

2. PROPAGATION

2.1. Line-of-Sight

Up to now, 10 GHz radio communication has mainly be made over line-of-sight paths. Several mW of output power and relatively small horn antennas are sufficient for distances of up to approximately 100 km. Suitable locations can be found easily.

If communication is to be made over larger distances, it is necessary to study the path using a map. The topographic features in Europe limit line-of-sight communication to approximately 300 to 350 km; and it is only a few of the highest alpine peaks that offer more. Further details regarding this can be taken from **Figure 2**, which gives the line-of-sight horizon for heights up to 3200 m.

The height values above mean sea level (MSL) are given as ordinate in the center of the graph. Parabolicly shaped curves run from this point to the horizontal reference line corresponding to zero m. These curves represent the curvature of the earth for the selected scale and cross the reference line at the values + D and – D, which give the line-of-sight distances. For example, one can see 112 km from an altitude of 1000 m; or on the other hand, two points having an altitude of 1000 m would have a line-of-sight distance of 224 km to another. Examples regarding this are given in appendix I.

When using this method under practical conditions, it will be seen that it is not simple to find two locations of sufficient height having a suitable distance from another, and also offering the line-of-sight path without obstacles. If such a path was to be found, it would exhibit the following path loss at a frequency of 10 GHz:

$$a = 113 \times 20 \log 2\,D$$

2 D = distance between the antennas in km; attenuation a in dB

This basic path loss will increase by approximately 0.1 dB/km in the case of thick fog (30 m visibility). In the case of heavy rain (12 mm/h), a loss of approximately 0.3 dB/km is exhibited, which can increase by 4 to 10 times during heavy rain showers.

It should be considered, however, that although thick fog may cover the whole communication path, this will hardly be the case with heavy rain showers.

The maximum line-of-sight path from a height h in meters above mean sea level can be approximated in this direction using the following formula:

$$D = 3.54 \times \sqrt{h}$$

Line-of-sight can be classed only as the lower-limit. Tropospheric over-the-horizon communications due to inversions or ducting also take place in the 3 cm band. This method of communication can be classed as the other extreme that allows communication to be made at low power levels over very great distances, far in advance of actual line-of-sight. Between these two extremes, are the possibilities of quasi line-of-sight propagation using reflection, refraction, and troposcatter, which are used extensively by VHF and UHF amateurs. These types of propagation are to be discussed briefly in the following sections.

2.2. Quasi Line-of-Sight

In a calm troposphere having a normal temperature distribution, humidity and density, electromagnetic waves are bent to earth slightly in the less-dense upper air layers. This means that such waves do not follow a completely straight line but follow the curvature of the earth slightly (11). This effect can be approximated mathematically by using a somewhat greater radius of the earth in the calculation formula. This results in larger »line-of-sight distances«. This correction factor for the earth radius is designated »K«.

Extensive measurements made in densely populated areas in the northern hemisphere have shown that the statistic mean K-value in the order of 1.3 is exhibited. The value 4/3 is standardized and has been accepted internationally as basic value for the calculation of the radio horizon. Since this propagation behaviour cannot be achieved optically, the term quasi line-of-sight is to be used. The maximum line-of-sight from a height of h (in m over MSL in that direction) is:

$$D = 4.12 \sqrt{h}, \text{ in other words approx. 16 \% greater than } K = 1$$

Especially when planning line-of-sight paths that are marginal, it is necessary to remember the statistic character of K. It is not only possible for the value K to be considerably greater than 1, but also for it to obtain values of less than 1 ! Generally speaking, the value K tends to reduce with increasing latitude and in winter. Further details regarding this are to be found in (12) where a different type of signal path analysis is made.

2.3. Tropospheric Inversion

In a normal atmosphere, temperature drops with increasing altitude. If there is a reversal (inversion) of the temperature or humidity characteristic with height, it is possible for very large values of K to be obtained. Although, principally speaking, these are virtually the same refraction processes as were described in section 2.2., one usually speaks of over-the-horizon communication when these are in excess of K = 2.

In extreme cases, K cannot only be infinite (this would be the case with a flat earth surface), but can also obtain negative values, which would correspond to a concave curvature of the earth. In other words, strong inversions will return the electromagnetic waves back to the earth surface.

Double or multiple layers having alternate density variation, are especially of interest since they are able to transmit electromagnetic waves at low loss in a similar manner to a waveguide. This is called tropospheric ducting.

These effects are, of course, not specific to the 3 cm band and have been described in detail in (11). However, the following points can be important for long-distance communications at 10 GHz:

The main characteristic for the refraction effect of a layer is its density with respect to the wavelength of the wave front. This means that thin layers of several meters in thickness can have a considerable effect on 3 cm signals, whereas they are not able to provide any noticeable refraction for 2 m or 70 cm signals. However, it is not possible at this time to know if such thin layers are sufficiently still or continuous to allow communication.

In addition to the problem of having the correct height of the station with respect to the layer in question (which is frequency-independent), considerable differences can occur during 3 cm communication, whether one is able to radiate into the layer at a narrow beamwidth under a »favorable« angle or not.

Fixed parabolic antennas with large dimensions with respect to the wavelength (larger than 60 cm diameter) should therefore be able to be tilted vertically. Users of 25 dB EME antennas have observed such effects even in the 70 cm band !

In any case, tropospheric over-the-horizon communications can allow long-distance communications even in the 3 cm band, and are therefore even more interesting for fixed station operation than occasional portable operation. Unfortunately, multiple inversions of the required height able to duct the signal are rather rare. The path loss of this type of communication can be several orders of magnitude less than the normal free-space loss, which means that communication over many hundreds of kilometers can be made with just a few mW and moderate antennas.

Short-term inversions are observed relatively often 1 to 3 hours after sunrise or sunset. It is possible with the aid of these to also cover greater distances, however, a considerable power reserve is often necessary. This is especially the case when the layers are irregular or disturbed by turbulent air. In this case, only forward scatter will take place. A high signal reserve is also necessary when the wave is to be refracted on a strong single-inversion by which the wave is reflected once or more times from the earth surface, which has considerably more loss than water surfaces.

Unfortunately, the necessary high-pressure areas allowing such interesting propagation modes are rare and only take place every few years.

2.4. Reflection

The fact that wave fronts are reflected on mountains, quarries, buildings, etc. is not new. The author has used this effect even on 2 m in the early 1950's to make several »impossible« contacts. With increasing frequency, more and more structures such as walls, towers, and even factory chimneys can be used as passive reflectors, a mode that is used by many SHF amateurs daily.

The most important characteristics for the reflection capabilities of obstructions is their size with respect to the wavelength, shape, surface smoothness as a function of the wavelength and the conductivity of the surface. The size is virtually no criterion any more for 3 cm communications, since even small lattice masts represent effective reflection surfaces, especially due to their good conductivity. The opposite is the case with surface smoothness; for this reason, all wooded, or overgrown surfaces are unusable due to their high absorption. It is only possible for rock surfaces, quarries or similar to be used as natural reflectors. Communication via reflection is very interesting at 3 cm, since several buildings, church towers, masts, factory chimneys, television towers, castles, etc. can be seen form virtually every location.

For those amateurs that have been using passive reflection points for 23 cm communication, it may be of interest to know that the same reflection surface should provide an approximately 10 dB better reflection at 3 cm if the surface is not unfavorable. A slight vertical tilting of the 10 GHz antenna can also be necessary for this mode of propagation.

If one is able to estimate the size of the reflection surface and the distance, it is then possible to calculate the additional attenuation a_a for a metallic flat surface as follows:

$$a_a = 8.7 \times \ln \frac{d}{A} - 30.7$$

where
d = distance to the reflector in m;
A = effective area in m^2;
a_a in dB

Example: A gasometer is visible at a distance of 1 km having a metallic, but convex surface. For this reason, only 5 % of the visible surface of approximately 50 m^2 are placed into the equation as effective reflection surface.

$$a_a = 8.7 \times \ln \frac{1000}{2.5} - 30.7 = 8.7 \times 5.99 - 30.7 = 21.4 \text{ dB}$$

This surprisingly low additional loss should be added to the path loss that would be present under line-of-sight conditions.

Even when approximately 20 dB reflection loss is assumed for concrete towers or only an effective surface of 4 % for lattice towers, the additional losses of »only« 30 to 50 dB are exhibited, which can be compensated for on the equipment side as is to be described in section 3. Resonance effects on metallic structures such as neon lights, fences, lightning conductors etc. can also considerably improve the reflection characteristics.

2.5. Refraction

At 10 GHz, the refraction on obstacles can be used far less than at lower frequencies. The signal loss in the »shadowed« area behind the obstacle is more than one order of magnitude higher than at 70 cm. In addition to this, most natural obstacles are overgrown or wooded, which means that considerably higher absorption loss is exhibited. Exceptions to this are towers or buildings mounted on mountain peaks having direct line-of-sight to the transmit and receive antennas. In such cases, however, it is virtually impossible to differentiate between reflection and refraction effects.

2.6. Troposcatter

The forward scatter of signals on turbulent conditions that are always present in the lower atmosphere is often used comercially at UHF and SHF for over-the-horizon communications. Examples of this are the link between West Germany and West Berlin and several island-to-island paths. Since effective radiated power levels in the order of 10 kW and more are necessary for this form of communication, this mode is limited, at present, to the VHF-UHF bands, although it does become possible for 3 cm communications if approximately 5 W and a parabolic antenna of 2 m diameter are available.

MATERIAL PRICE LIST OF EQUIPMENT
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DL 2 DO 001	**DF-RECEIVER**		**Ed.4/1978**
PC-board	DL 2 DO 001	....	DM 11,—
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Semiconductors	DJ 7 VY 002	11 transistors, 8 diodes	DM 84,—
Ring mixer	IE-500	....	DM 38.—
Minikit 1	DJ 7 VY 002	4 airspaced, 1 foil trimmer, 3 feedthru, 3 tantalum, 9 ceramic without conn., 29 ceramic capacitors	DM 39.—
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Crystal	58.000 MHz	HC-18/U	DM 26,—
Kit	**DJ 7 VY 002**	....	**DM 288,—**
DF 8 QK 002	**1268 MHz LOCAL OSCILLATOR**		**Ed.4/1978**
PC-board	DF 8 QK 002	double-coated, with plan	DM 18,—
Semiconductors	DF 8 QK 002	6 transistors, 1 diode	DM 18,—
Minikit 1	DF 8 QK 002	11 trimmers, 1 feedthru, 9 ceramic, 1 tantalum caps., 1 coilformer with core, 1 choke, 4 ferrite beads, 1 potentiometer	DM 28,—
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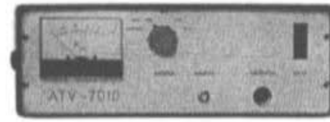

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SPECIFICATIONS

Transmitter	Maximum drive:	0.5 W
	Output power:	0.7 W typ.
	Carrier and image rej.:	20 dB min.
Receiver	Noise fig. (single-sideband):	4 dB min.
	Overall gain:	30 dB
	IF-bandwidth	2 MHz
	Operating voltage:	12.5 V ± 1.5 V
	Dimensions:	255 x 75 x 200 mm

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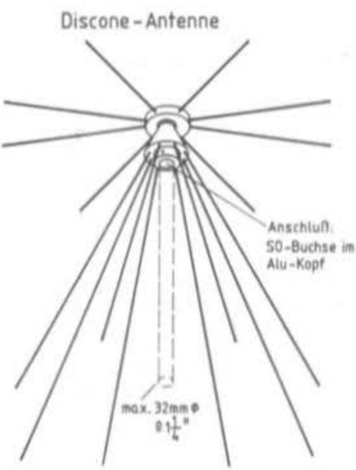

- Frequency range: 80 - 480 MHz
- Gain: 3.4 dB / λ/4
- Impedance: 50 Ω
- Power rating: 500 W
- Polarisation: Vertical
- Connection: SO 239 socket in the head
- VSWR: < 1.5 : 1
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- Dimensions: Height: 1.00 m / Diameter: 1.30 m
- Material: Aluminium
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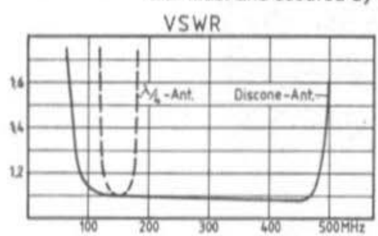

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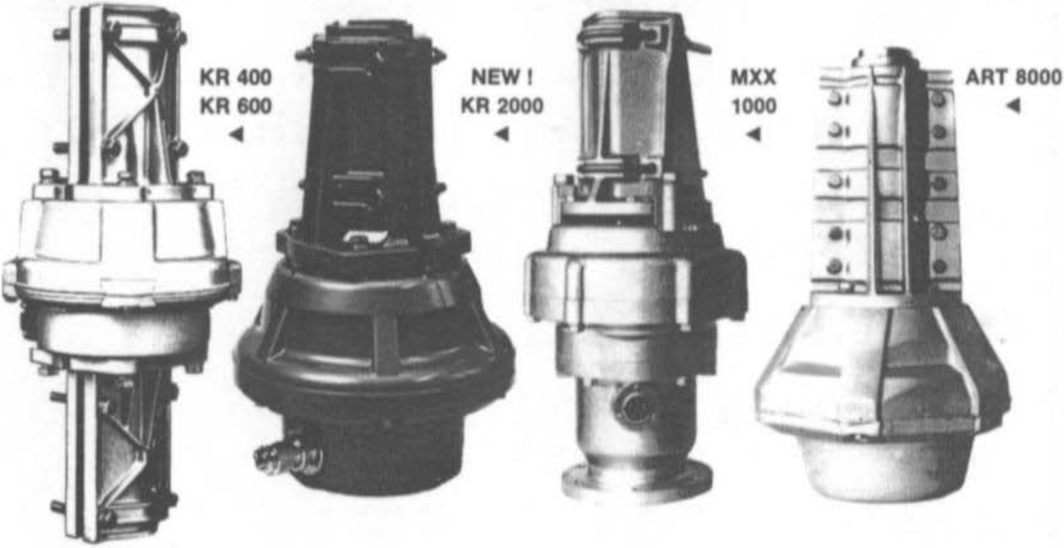

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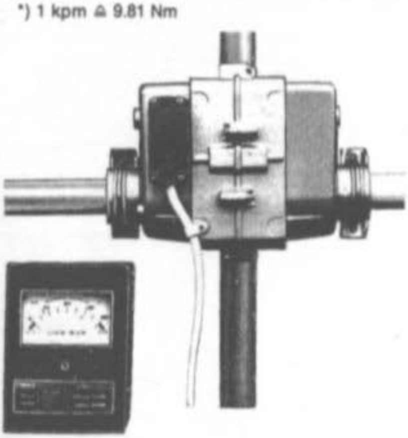
Pending torque	800	1000	1600	1650	2450	Nm *)
Brake torque	200	400	1000	1200	1400	Nm *)
Rotation torque	40	60	150	180	250	Nm *)
Mast diameter	38 - 63	38 - 63	43 - 63	38 - 62	48 - 78	mm
Speed (1 rev.)	60	60	80	60	60	s
Rotation angle	370°	370°	370°	370°	370°	
Control cable	6	6	8	7	8	wires
Dimensions	270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

*) 1 kpm ≙ 9.81 Nm



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