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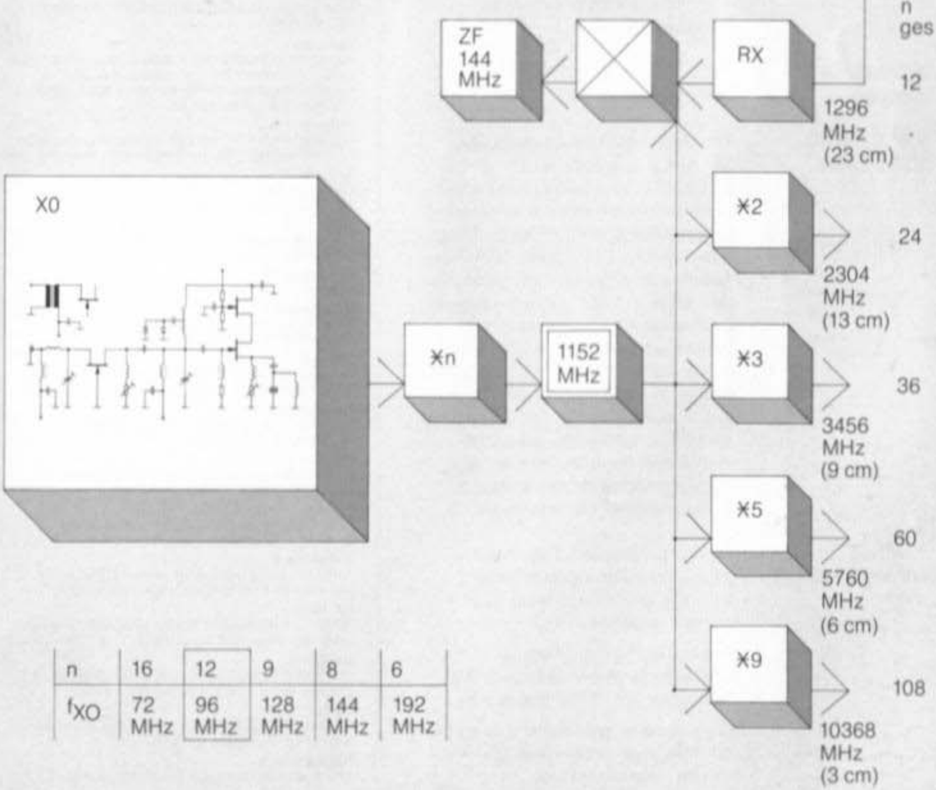
A Publication for the Radio-Amateur

Especially Covering VHF, UHF and Microwaves

VHF communications

Volume No.13 · Winter · 4/1981 · DM 6.00

An extremely low-noise crystal oscillator for microwave applications



n	16	12	9	8	6
f_{XO}	72 MHz	96 MHz	128 MHz	144 MHz	192 MHz

VHF communications

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Editorial

This edition completes the thirteenth year of VHF COMMUNICATIONS. We hope that our readers have enjoyed reading the magazine in 1981 and will not forget to subscribe to 1982.

We would appreciate you helping us by recommending VHF COMMUNICATIONS to your friends and other radio amateurs. It may seem strange, but the circulation of VHF COMMUNICATIONS has not increased much in the last ten years and has still not really reached the break-even point. This means that it is still being subsidized by the German magazine.

Since we only rarely have advertising income in the magazine, the subscriptions must pay for all the editorial work, the paper, drawings, photographs, printing, and binding. Not to forget the postage rates.

You will see that we have not increased the subscription rate for 1982, which remains at DM 20.00, even though the costs have risen considerably: For instance the postage rate by 38 % !

You can assist us, and ensure that we can continue to publish VHF COMMUNICATIONS by introducing the magazine to other amateurs so that the circulation can be increased to a more satisfactory level.

The publishers, editors, and staff would like to wish all our readers a very Merry Christmas, and a happy and prosperous New Year.

G 3 JVQ / DJ Ø BQ

An Extremely Low-Noise 96 MHz Crystal Oscillator for UHF/SHF Applications Final Part II

by B. Neubig, DK 1 AG

3. CONSIDERATIONS DURING THE DESIGN OF CRYSTAL OSCILLATORS WITH GOOD SHORT-TERM STABILITY

3.1. Selection of the most Favorable Output Frequency

The classical design of oscillators with a good short-term stability is based on a master oscillator at 5 or 10 MHz using precision crystals in their third or fifth overtone (6). This results in an extremely high frequency multiplication factor in the case of UHF and SHF frequencies. The noise component of the frequency multiplier stages adds a considerable amount to the total noise.

Due to technical advances in the production of crystals, it is now possible to use crystals having a higher resonance frequency. Fundamentally speaking, the Q of a crystal will increase on increasing the overtone (at the same frequency). It can be assumed that the product Q x f is a constant for a certain overtone. This means that the Q will deteriorate on increasing the frequency. The upper frequencylimitforcrystals onthe market is in the order of approximately 200 MHz. In professional technology, an output frequency range in the order of 100 MHz has been found to be most favorable. This means that our selection of 96 MHz represents the present state-of-the-art.

96 MHz crystals in fifth overtone are available readily in the larger HC-6/U or smaller HC-18/U case types. Comparison measurements madeby the author haveshown that the larger crystal has a more favorable Q. After carrying out a number of measurements, it was found that an HC-6/U crystal exhibited a mean Q of approximately 80 000, and 94 000 for HC-18/U. For comparison, crystals were also measured at their seventh overtone. This resulted in a mean Q of 106 000 in the case of HC-18/U crystals, which means approx. 13% higher Q than the fifth overtone. Precision crystals in a glass case exhibit a mean Q of 116 000 at the seventh overtone (see Section 4 for further details regarding the selection of a suitable type of crystal).

3.2. Selection of the Active Components

In the case of bipolar transistors, the noise is mainly determined by the base-emitter path. The noise of PNP-transistors is less than that of corresponding NPN-transistors. MOSFETs generate a high noise level, whereas the 1/f-noise dominates at low frequencies, and the thermal noise of the drain-source path dominates at higher frequencies. Junction field effect transistors generate less noise than bipolar transistors and MOSFETs (7). When field effect transistors are used, high-current types are preferable due to the larger, linear range and the low source impedance in a gate circuit. As can be seen in the measuring results given in section 4, transistor type P 8000 (now P 8002) is better than the approximately equivalent transistor BF 246 C.

3.3. Selection of the Most Suitable Circuit

The active stage of a crystal oscillator has two functions: The first is to provide a gain reserve for commencing and maintaining feedback conditions for oscillation, and the second is the limiting ofthemaximumpossible amplitude by reducing the gain at high amplitudes (saturation). In the case of an oscillator with good short-term stability, it is extremely important that both functions are separated from another and especially that the crystal will not »see« another component whose operating point will change during the commencement of oscillation, or after oscillation has been commenced. Otherwise, the fluctuating impedance in time with the RF-signal will cause a multiplicative conversion of noise sidebands.

Furthermore, a greatest possible feedback should be provided for the active stages, because a linear operating range, and a low-noise operation isonly possible inthismanner.

A decisive criterion for the selection of the oscillator circuit is the operating Q (see measuring results in section 4). A circuit that is able to fulfill these conditions was already described by the author in (6). The following information is to discuss the experience and measuring results encountered whilst obtaining a favorable design.

Figure 10 shows the principle of the circuit. As can be seen, it is a DC-coupled cascode circuit of two high-current field effect transistors. The basic oscillator circuit is a Collpits oscillator. If the source impedance of transistor T 1 is bridged capacitively, one will obtain a free-running LC-oscillator, where the feedback is generated using a capacitive tap on the resonant circuit L 1, C 1, C 2. Transistor T 1 operates in a common-source, and T 2 in a common-gate circuit.

The feedback isdesigned so that theoscillator will not oscillate on its own after removing the bypass capacitor at source 1. The resonant circuit comprising L 1, C 1, and C 2 is tuned to 96 MHz. If the crystal is now added into the source line, this will reduce the feedback of the source impedance due to the lower crystal impedance at the series-resonance frequencies (fundamental wave, third, fifth overtone etc.). The drain circuit ensures that oscillation is made at the required overtone. For this reason, it must have a sufficiently high Q. The ratio C 1 / C 2 determines the degree of feedback.

Both transistors operate in stable class A with a drain current of 40 mA. The limiting is made using two anti-phase Schottky diodes that are connected to a tap on the output circuit. The static capacitance of the crystal is compens-

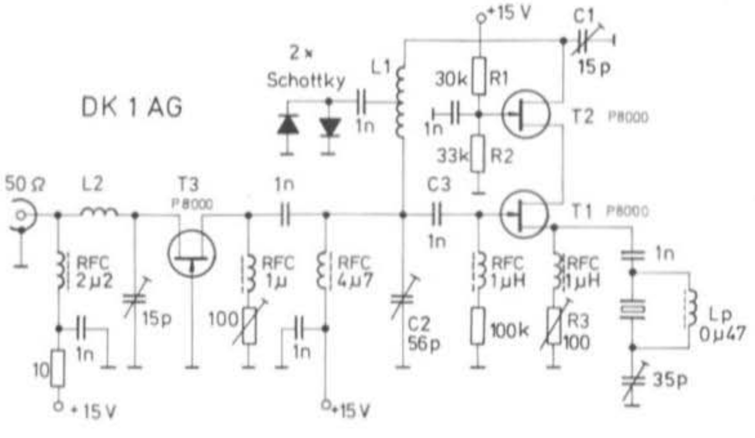

Fig. 10: Practical circuit of the 96 MHz oscillator

L 1: 6 turns of 1 mm dia. silver-plated copper wire wound on a 6 mm former, coil tap approx. at the center

L 2: as L 1, but without tap

ated for with the aid of L_p, in order to obtain a real crystal impedance, and a symmetrical phase run in the range of the series-resonance frequency of the crystal.

The cascode circuit equipped with the field effect transistors has characteristics similar to that of a tube pentode. The high internal impedance is characteristic, which means that the gain corresponds to the well-known tube formula

$$G = S \times (R_{in} \parallel R_{out}) \approx S \times R_{out} \quad (7)$$

The output impedance that can be realized in practice, is less than the internal impedance, which means in principle that a less than optimum matching is present. The same is valid for the drain of the first transistor. Its high-impedance output impedance is terminated with the low-input impedance of the source of the second transistor. In the opposite manner, source 2 will »see« the very high dynamic impedance of transistor 1, which is connected as constant-current source. This causes a very large feedback. A further feedback is provided by the source resistor R 3 in conjunction with the choke, which is in parallel with the series-resonance impedance of the crystal for RF.

The operating current of the cascode is adjusted with the aid of the source resistor; the voltage at the interconnection point of drain 1 and source 2 is adjusted to approximately half the operating voltage with the aid of the voltage divider R 1, R 2.

The two limiting diodes are removed for alignment, and the crystal is short-circuited. The free-running frequency is adjusted with the aid of the collector circuit to approximately 96 MHz. The compensating inductance L_p is aligned together with the crystal (after being removed) to 96 MHz with the aid of a dipmeter. The oscillator should operate in a stable manner after connecting the crystal and operating voltage. The oscillator frequency should only shift slightly on detuning the drain circuit, however, this should not be detuned too greatly in order to ensure that oscillation does not cease, or to avoid transcient problems on switching on. After adding the anti-phase diodes, the RF-amplitude should be limited to approximately half the value of the self-limiting oscillator by selecting the correct tap.

A subsequent buffer amplifier in a common-gate circuit (T 3) increases the output level to 18 dBm.

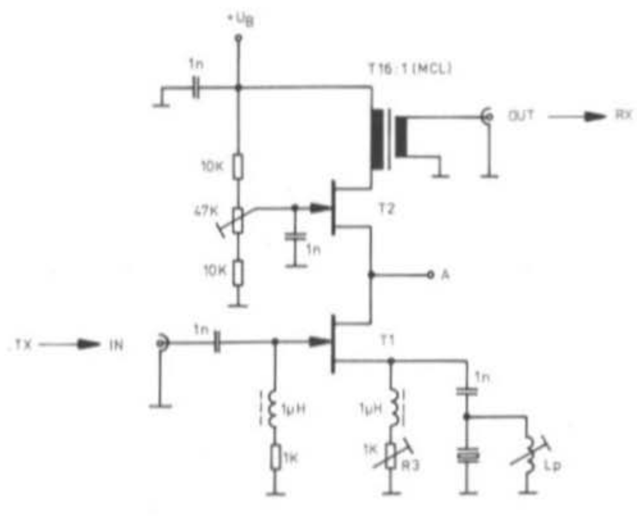

Fig. 11: Measuring circuit for determining the operating Q

4. MEASURING RESULTS

4.1. Operating Q as a Function of the Operating Point

The circuit was modified as shown in **Fig. 11** for measurement of the operating Q. The feedback line was cut, and a wideband transformer T 16-1 was used as output circuit. The 50 Ω signal from the signal generator of the spectrum analyzer was fed to gate 1, and the output signal fed to the 50 Ω input connector of the spectrum analyzer. If the crystal is bridged, the wideband gain can be read off from the spectrum analyzer.

On inserting the crystal, this gain will be fed back outside the resonances of the crystal, and one will obtain maximum gain at the series-resonance frequencies of the crystal, as well as at its spurious resonances. Inductance L_p can now be adjusted so that the resonance curve at 96 MHz is symmetrical.

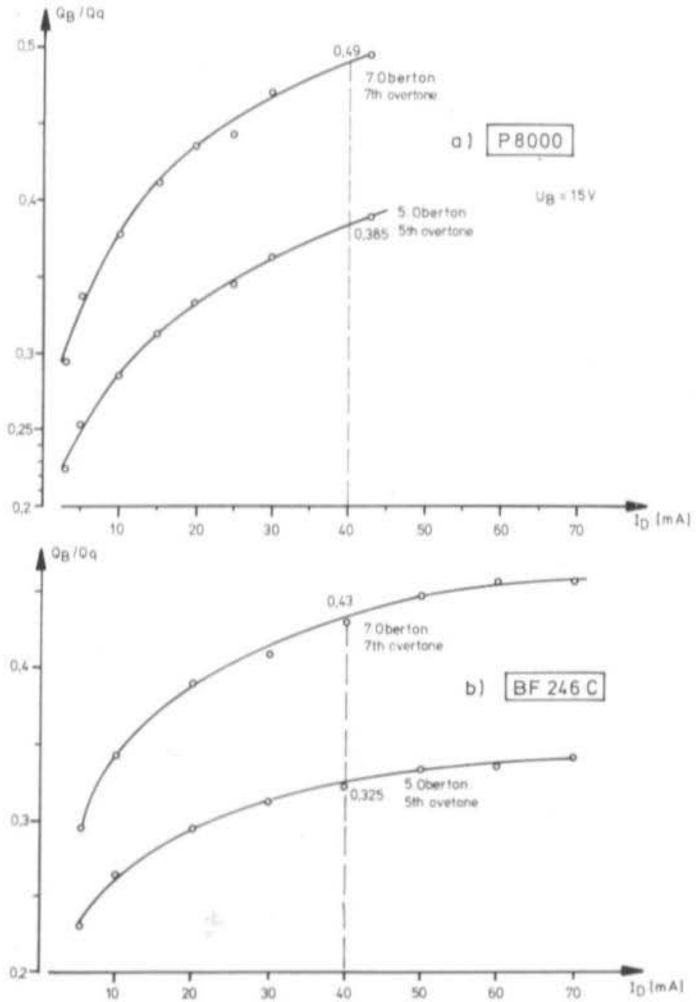

Fig. 12: Operating Q to crystal Q as a function of the drain current I_D

The voltage gain from gate 1 to drain 2 is now measured andthe bandwidth ofthe resonance curve is determined. This bandwidth is a measure of the operating Q.

$$Q_{op} = \frac{f_0}{\Delta f_{3dB}} \tag{8}$$

The operating Q as a function of the drain current was determined with a crystal operating in fifth overtone (Q = 89 300) and a crystal in seventh overtone in the glass case (Q = 127700) in conjunction with a transistor pair 2 x P 8000, and 2 x BF 246 C. The results are given in **Figures 12 and 13.**

Figure 12 shows the relationship between operating Q and crystal Q as a function of the drain current. The upper diagram is valid for transistor type P 8000, and the lower for transistor BF 246 C, and they show the results for both crystals. When using a BF 246 C, and an average fifth overtone crystal, it is possible to obtain a relationship of operating Q to crystal Q of between 23% and 33% according to the drain current. Using a transistor P 8000, one will obtain a value of Q_{op}: Q_{cr} = 37% when using the same crystal, and a drain current of 40 mA. The operating Q will be between 30 and 46% when using a seventh overtone crystal and a transistor BF 246 C, whereas a P 8000 will already reach the 50% mark at approximately 45 mA.

The operating Q is summarized as an absolute value as a function of I_D in Figure 13. One will notice the saturation effects in excess of 40 to 50 mA. For this reason, a drain current of 40 mA was used as operating point for the subsequent measurements. With this and with the inferior transistor and a fifth overtone crystal, an operating Q of approximately 28 000 was obtained, whereas approximately 62 000 (or more than twice) was obtained using a P 8000 and a seventh overtone crystal. Since the phase noise is mainly determined by the operating Q (see equation 1), it is easy to see that an improvement of more than factor 2 can be achieved by selection of the most favorable operating point and type of crystal.

Crystal Q is mainly dampened in the oscillator by two factors. The lowest component is the series connection of source-choke and R 3 that are parallel to the crystal, whereas the main component is the source-input impedance of transistor T 1. This is in series with the equivalent resistance of the crystal.

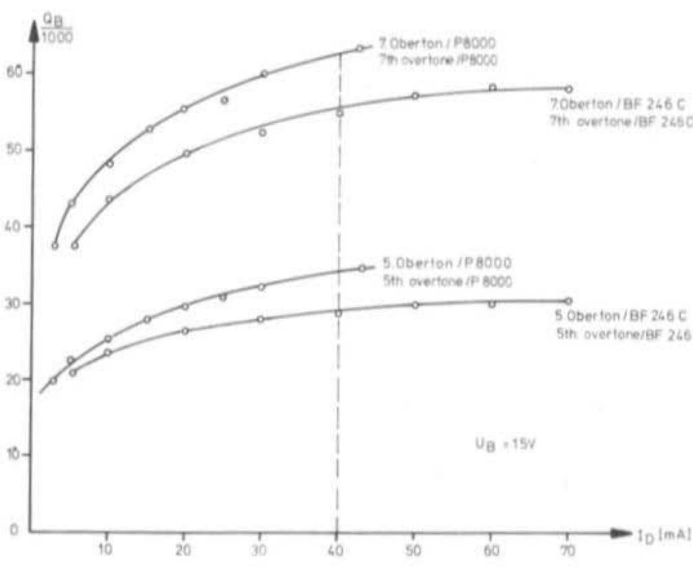

Fig. 13 Operating Q as a function of I_D

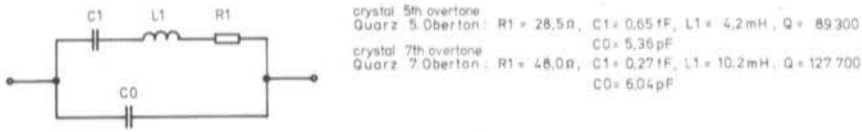

Fig. 14: Specifications of the measured crystals from Figures 12, 13, and 15

If one compares the equivalent data of both crystals (see **Figure 14**), one will see that the fifth overtone crystal possesses a lower Q. Furthermore, it exhibits a larger dynamic capacitance C 1 = 0.65 fF, which results in a series-resonance equivalent resistance of 28.5 Ω. The seventh overtone crystal exhibits a C 1 = 0.27 fF, which results in a resonance impedance of 48 Ω in conjunction with a Q of 127 700. The higher equivalent resistance of the seventh overtone crystal (in conjunction with a higher Q !) means that a certain source-impedance will deteriorate the overall Q less than when using a lower-impedance crystal. This means that the damping of the operating Q by the higher overtone crystal will be less than when using the fifth overtone.

Since the effective source impedance decreases on increasing the drain current, an increase of operating Q results on increasing the drain current. **Figure 15** shows this relationship with the aid of the effective source impedance determined with the aid of the above curves. This is in the order of 60 to 64 Ω for the BF 246 C at a drain current of 40 mA, and 45 to 50 Ω in the case of transistor P 8000.

Of course, a further improvement could be obtained when using a ninth overtone crystal, however, this would make the alignment of the oscillator more critical since the loop gain will be reduced considerably due to the higher source-feedback caused by the resonance impedance of the crystal.

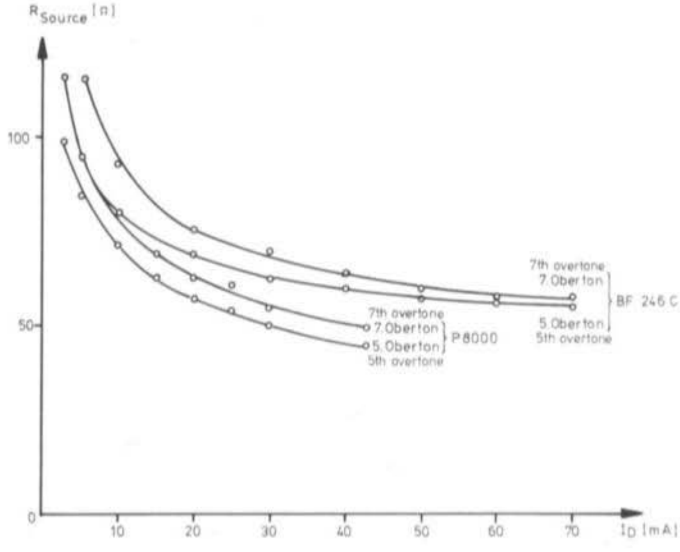

Fig. 15: Effective source impedance as a function of I_D

4.2. Most Favorable Output Impedance of the Oscillator Stage

A circuit as shown in Figure 10 was modified by removing the buffer stage, as well as interrupting the feedback by disconnecting C 3. The seventh overtone crystal is present in the source circuit. A low-level signal is injected to gate 1, and tapped off at drain 2 at high impedance (low-capacitance probe). The output level is measured at various terminating impedances.

Figure 16 shows the results:
The power gain (at I_D = 40 mA) reaches a saturation value of approximately 20 dB at $Z_T >$ 1 kΩ. The output power increases linearly up to Z_T = 800 Ω (dynamic slope = 14 mA/V) and falls off above this.

In order not to absorb too much power due to circuit losses, it is advisable to set the impedance to ≈ 1 kΩ. The output circuit comprising L 1, C 1, and C 2 is also used as π-link in order to transform Z_T to the 50 Ω input impedance of the buffer. The feedback ratio 1:20 ≙ 13 dB provides a sufficient gain reserve (g = 20 dB) to ensure a stable commencement of oscillation.

4.3. Measurement of the Noise Sidebands

Two oscillators were built up according to the circuit given in Figure 10, and one of these was aligned to a fixed frequency of approximately 96 MHz with the aid of a trimmer. In the case of the second oscillator, the trimmer in the vicinity of the crystal was replaced by a pair of varactor diodes (see **Figure 17**), and locked to the frequency of the first oscillator according to the measuring principle shown in Figure 7.

The measurement was made using a professional laboratory system (Wandel & Goltermann FSM-1095) in both the frequency and time domaines. It is possible in this manner to measure up to 15 Hz from the carrier. The result is given in **Figure 18** as the noise spacing 10 log £ (f) referred to the carrier, and recalculated for a measuring bandwidth of 1 Hz (»dB_c/Hz«) for a sideband. The following noise sideband values will be seen:

Spacing = 100 Hz:	113 dB_c
Spacing = 1 kHz:	140 dB_c
Spacing = 10 kHz:	147 dB_c

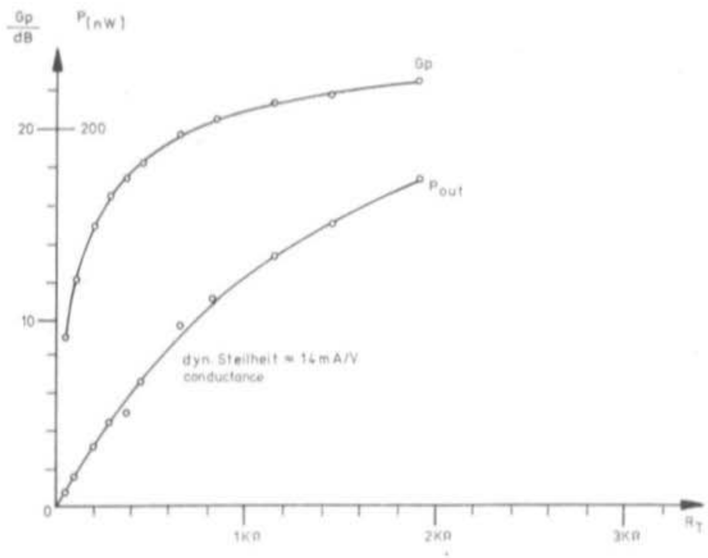

Fig. 16: Power gain (dB) and output voltage as a function of drain-terminating impedance Z_T of the cascode stage (using seventh overtone crystal in the source circuit)

The curve can be subdivided in straight lines of different slopes corresponding to the behaviour of flicker noise, white frequency noise, and white phase noise, etc.

The importance of these very good values is to be discussed with the aid of an example:

After a frequency multiplication of 36, required for the 3456 MHz band, a value of $-113\ dB_c$ means that an adjacent carrier spaced 3.6 kHz from the required frequency, will be superimposed with noise from the oscillator to the value of $-113\ dB_c$. If the carrier is S 9, corresponding to 50 µV or 34 $dB_{\mu V}$, the noise sideband will produce $-79\ dB_{\mu V}$ per Hz of bandwidth. At a bandwidth of 2.4 kHz (SSB), this corresponds to 10 log 2400 ≙ 33.8 dB more. This is equivalent to a noise floor of $-45.2\ dB_{\mu V}$ ≙ 5.9 nV, or 31 dB less than S 1. Of course, this is only valid when the unwanted carrier does not have any noise sidebands of its own.

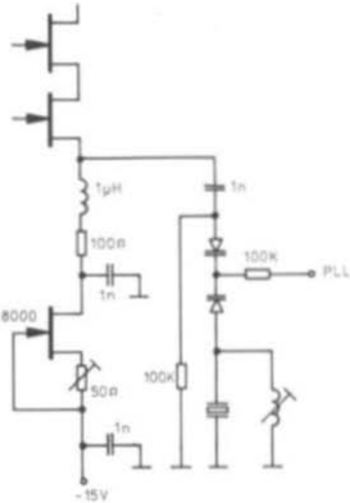

Fig. 17:
Improvement of the source circuit, and modification for pulling, or frequency modulation

5. PRACTICAL OPERATION AND IMPROVEMENTS

The operating current of 40 mA produces a considerable dissipation power, which is the price to be paid for obtaining a clean signal. The heat dissipation should be radiated effectively by a suitable mounting of both oscillator FETs. Furthermore, attention should be paid that it is not allowed to reach the crystal. It is therefore advisable for the crystal to be mounted outside of the oscillator case in order to ensure that no heating (frequency shift) occurs. For higher demands on the frequency stability, the crystal should be placed in a crystal oven, since even a crystal with a most favorable temperature response will drift by 1 to 2 ppm, equivalent to 100 to 200 Hz at 96 MHz for every 10°C.

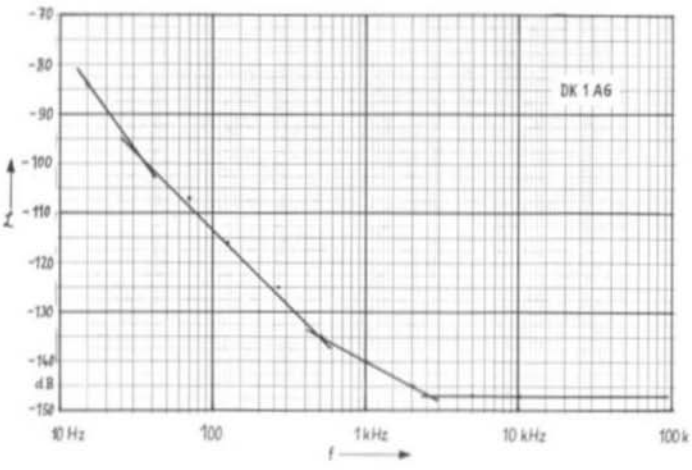

Fig. 18:
Noise rejection as a function of the frequency spacing from the carrier

If 40 mA is too high, it is possible to operate the circuit with 20 mA if a slight deterioration in stability is acceptable (see Figure 13).

The limiting diodes in the original circuit are Schottky types; if necessary, they can be replaced by noisier silicon diodes (1N 4151 or similar) or germanium diodes (AA 113 or similar).

The operating current of 40 mA is in the vicinity of I_D (0) in the case of some P 8000/P 8002 transistors, which means that the source leakage impedance, and thus the DC-feedback will be very low. This can be improved by providing a constant current source as shown in Figure 17, which can be fed from − 15 V. It may be sufficient when the source resistor of 390 Ω / 1 W is connected to − 15 V. Figure 17 also shows how a pair of varactor diodes can be inserted to allow frequency modulation. The pulling range that can be obtained is given in **Figure 19**. A linear range of ± 5 ppm (≙ ± 500 Hz) can be obtained with a pulling voltage of between 1 V and 7.5 V in conjunction with the seventh overtone crystal; approximately 2.5 times this is obtained in conjunction with a fifth overtone crystal.

The bias voltage for these varactor diodes must be taken from a low-noise voltage source, such as from the 10 V stabilizer REF-01, manufactured by Precision Monolithics (PMI) or a suitable, filtered battery.

A further increase of the phase purity can be achieved using a simple crystal filter as shown in **Figure 20**, which is connected to the output of the buffer. A cheap fifth overtone crystal is suitable, and one will obtain a 3 dB bandwidth of ± 3 kHz. When using a seventh overtone crystal, one will obtain a lower bandwidth of ± 1.4 kHz.

The alignment is made very simply by adjustment of C 2 and C 4 for maximum output level.

The obtained selectivity curve is given in **Figure 21**. The filter has no effect on the directly adjacent channel, but will attenuate the wideband noise spectrum by approximately 20 dB if constructed carefully.

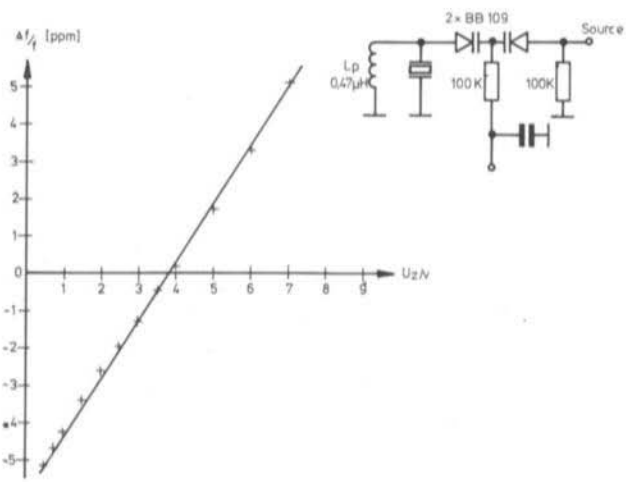

Fig. 19: Frequency pulling characteristics of a seventh overtone crystal

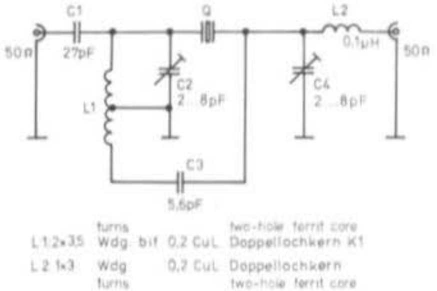

Fig. 20:
Crystal filter for improving the phase noise

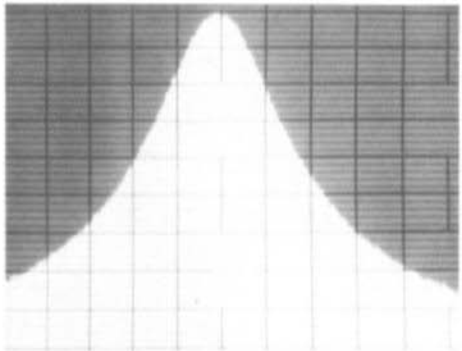

Fig. 21:
Selectivity curve of the crystal filter shown in Fig. 19 using a seventh overtone crystal
X = 1 kHz/div.; Y = linear

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A Linear 1 W Power Amplifier for 2400 MHz

by R. Heidemann, DC 3 QS

The following single-stage transistor linear amplifier possesses a gain of approximately 7 dB and an output power of 1 W in the 13 cm band (approx. 2400 MHz). It is thus suitable for use, for instance, together with the 13 cm transmit converter described in (1). This hybrid-power mixer equipped with a varactor diode type BXY 28 can generate between 200 and 250 mW. This means that the described power amplifier can provide a linear output power in the order of 1 W.

CIRCUIT DESCRIPTION

A transistor type F 1 E 6 manufactured by CTC is used. In the case of this transistor, the emitter is internally connected to the cooling bolt. This ensures that there are no problems with respect to the high-frequency grounding of the emitter. The circuit diagram of this amplifier is given in **Figure 1.** It shows no tendency to oscillation. The matching of the transistor input and output impedance to 50 Ω is

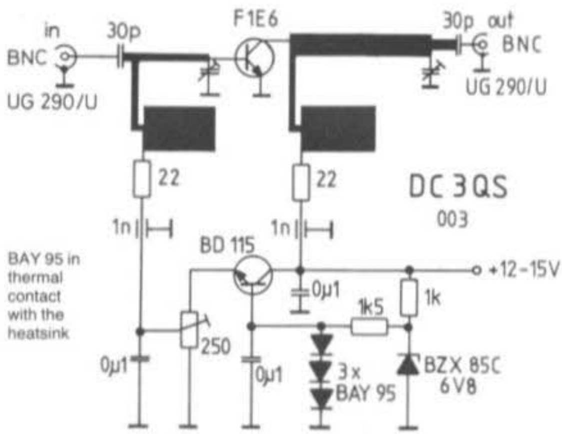

Fig. 1: This amplifier provides a gain of approx. 7 dB at 2400 MHz and an output power of 1 W

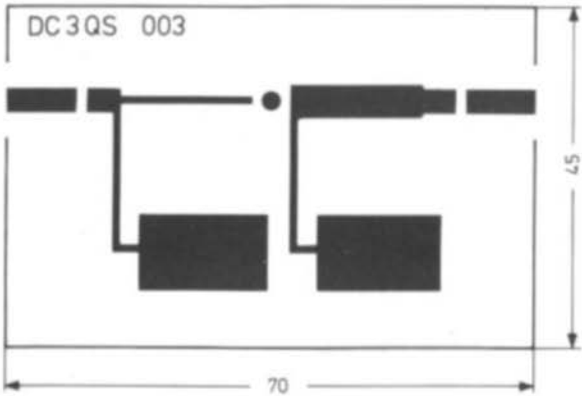

Fig. 2:
The PC-board consists of a double-coated glassfibre epoxy material

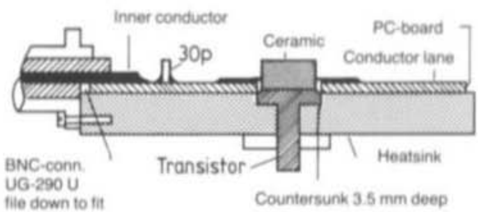

Fig. 3:
Mounting of the transistor and the two BNC connectors in a manner to avoid impedance jumps

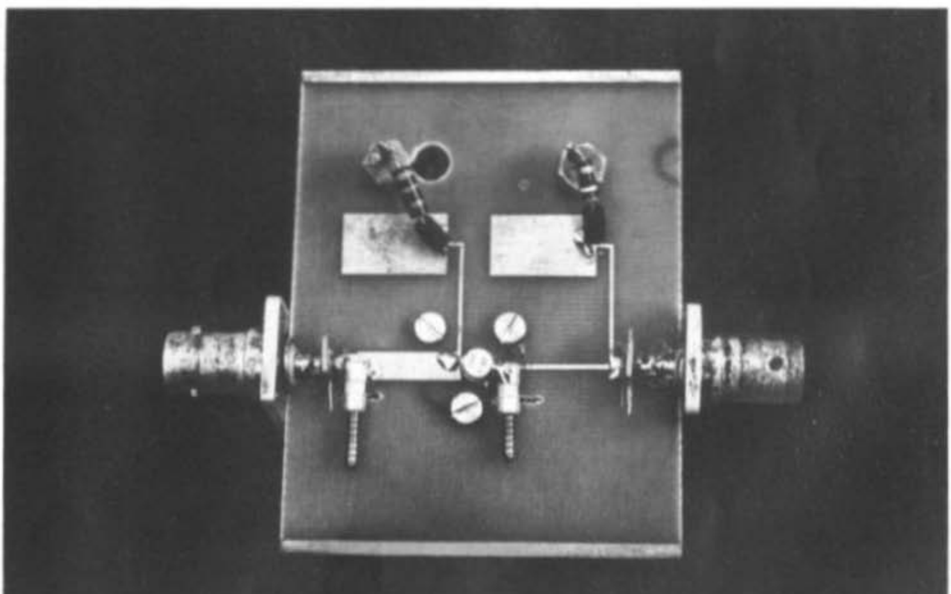

Fig. 4:
Photograph of the author's prototype using a piece of board that was available

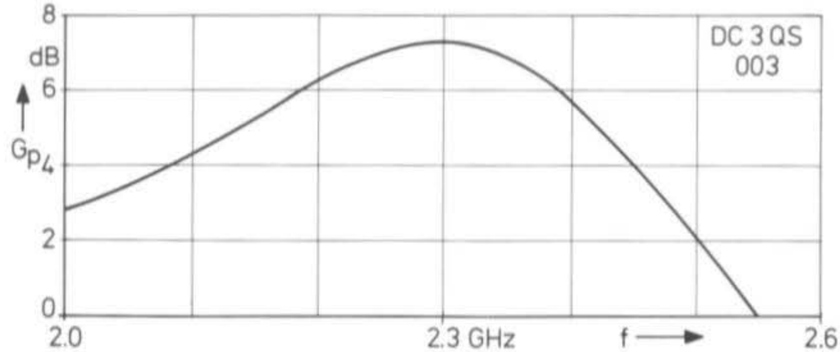

Fig. 5: The described amplifier can be used without tuning from 2.2 to 2.4 GHz

made using the standard method described in (2). The circuit for adjusting and stabilizing the DC-operating point is similar to the method described in (3).

CONSTRUCTION

Double-coated, epoxy glassfibre material is also suitable for applications in the 13 cm band. PC-board DC 3 QS 003 has dimensions of 70 mm x 45 mm, and is shown in **Figure 2**. Of course, it would have been possible to use the more expensive, non-readily available PTFE PC-board material, but this is not necessary for this application.

The cross-sectional drawing given in **Figure 3** shows how the transistor, and BNC-connectors for the input and output are mounted in such a manner that no impedance jumps occur.

Further details regarding construction are shown in the photograph given in **Figure 4**. One will notice the chip capacitors at input and output, whose values of approximately 30 pF are not critical, as well as the two miniature tubular trimmers (Philips, of approx. 3 pF), whose ground connections are placed through the slots in the board. The two ferrite beads on the connection wires of the 22 Ω resistors can also be seen.

Finally, **Figure 5** gives the low-signal frequency response of the described amplifier. The following measuring equipment was used for the measurements:

- Power meter HP 432 A
- Sweep generator HP 8690 A
- Network analyzer HP 8410 A
- Power signal generator AIL 124

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A Receive Converter for the Geostationary Weather Satellites METEOSAT – GOES – GMS

Part 1: The SHF-Module

by B. Rößle, DJ 1 JZ

The technology used in the 1.7 GHz converter to be described is not really new; it is based on the so-called interdigital filter converters (1) described in (2) for the 23 cm, 13 cm, and 9 cm bands. This converter is very suitable for use with the satellite reception system described by DC 3 NT (3). A large number of these converters have been constructed and have been in operation for several years now. The converter is to be described in detail to save the complex recalculation of the mixer and oscillator modules from their 23 cm or 13 cm counterparts.

Part 1 of this article is to describe the interdigital filter-mixer module and the IF-preamplifier; Part2 will describe the local oscillator module. The latter has two individual crystal oscillators for converting the two APT-channels of METEOSAT to the same intermediate frequency of 137.5 MHz. Further processing of the signal in a VHF-receiver can be made using modules DC 3 NT 003/004 (3) or modified modules DK 1 OF 034/035 (4) together with a home-made VFO. These are only two examples.

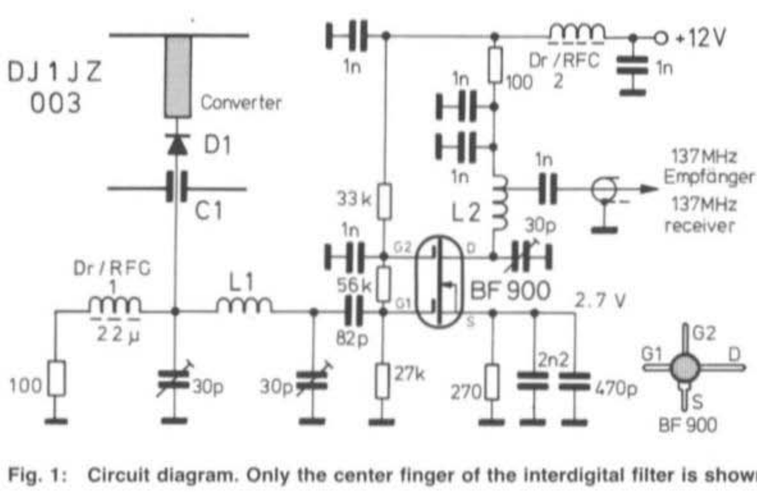

Fig. 1: Circuit diagram. Only the center finger of the interdigital filter is shown

CIRCUIT DESCRIPTION

The METEOSAT converter possesses a five-stage interdigital filter as mixer module in a similar manner to the converters described in (2). The input voltage from the antenna is fed to one end of the filter, and the local oscillator frequency injected to the other end. The mixer diode is connected to the center finger. The circuit diagram is limited to the mixer diode and the subsequent, low-noise IF-preamplifier **(Figure 1).**

For mechanical reasons, a Schottky diode in a small glass case (D 1) is used as mixer. The 2-GHz version HP 2817 provides a noise figure of between 7 and 9 dB. However, in order to achieve this, it will be necessary for the impedance transformation between the mixer diode and the preamplifier transistor to be optimized. For this reason, the circuit is provided with a Pi-link comprising two trimmer capacitors and inductance L 1.

Capacitor C 1 ensures a closed UHF-current circuit from the finger via the diode to the outside of the filter. This capacitor is home-made and should not have more than approximately 10 pF, since its capacitance is added to that of the first trimmer, and thus limits the adjustment range.

DC-current flows from the mixer diode via RFC 1 and the 100 Ω resistor to ground, and via the finger back to the diode. The voltage drop across the resistor allows the oscillator power to be checked. It should be between 1 and 10 mW and cause a direct current of approximately 1 to 10 mA. This means that one should be able to measure a voltage drop of between 100 mV and 1 V.

The intermediate frequency voltage is fed via the Pi-link to gate 1 of the dual-gate MOSFET BF 900. It operates in a standard circuit with approximately 10 mA drain current which allows transistor spreads to be virtually compensated for. The parallel-resonant circuit at the output comprises inductance L 2 and the 30 pF trimmer, as well as a 50 Ω tap, which means that the interconnection to the subsequent receiver can be made using virtually any length of coaxial cable. The gain of this stage is in the order of 20 dB when using transistor types BF 900 or BF 905, whereas approximately 25 dB can be obtained when using a transistor type BF 910 or BF 981. The »overall gain« of this converter will be obtained after deducting the actual conversion loss of approximately 8 dB.

ELECTRICAL COMPONENTS

D 1: HP 5082-2817 (Hewlett-Packard)

T 1: BF 900, BF 905, BF 910 (Texas Instr.) or BF 981 (Philips) or 3SK 88 (NEC)

L 1: 5 turns of 1 mm dia. silver-plated copper wire, wound on a 5 mm former, pulled out to fit the holes, self-supporting

L 2: 5 turns of 1 mm dia. silver-plated copper wire, wound on a 5 mm dia. coil former, without core. Coil tap 1 turn from the cold end.

RFC 1: 22 μH miniature fixed inductance or λ/4 choke for 137 MHz

RFC 2: 6-hole ferrite core choke (Philips)

3 plastic foil trimmers, 30 pF, 7.5 mm dia. (Philips: red)

Various ceramic disk capacitors, values uncritical, spacing 5 mm

Carbon resistors, spacing 10 mm

Coaxial sockets according to the type of construction and the case. Antenna input and oscillator input: N-connector or SMA; BNC for the IF.

MECHANICAL CONSTRUCTION

The principle of this receive mixer can be easily seen in **Figure 2.**

The base plate is 120 mm x 90 mm and is made from double-coated, epoxy glassfiber PC-board material. Whereas the lower side of the base plate is etched in the vicinity of the IF-preamplifier (see **Figure 3**), the upper sur-

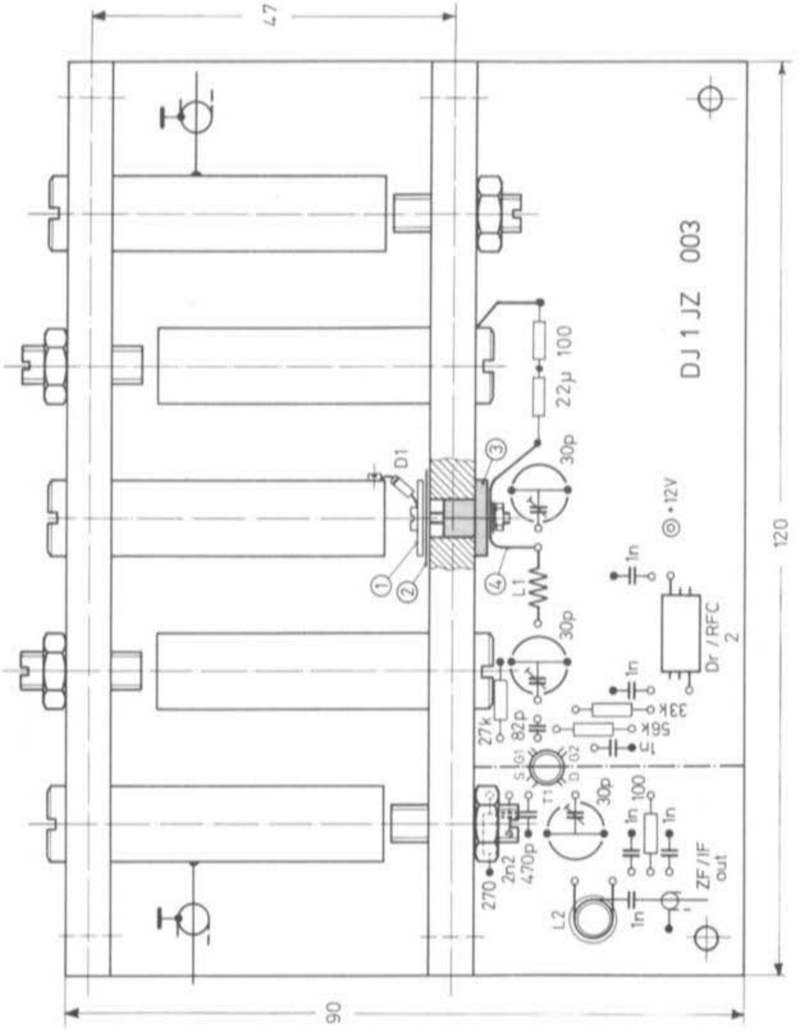

Fig. 2: Component locations and construction of the receive converter 1691/137.5 MHz

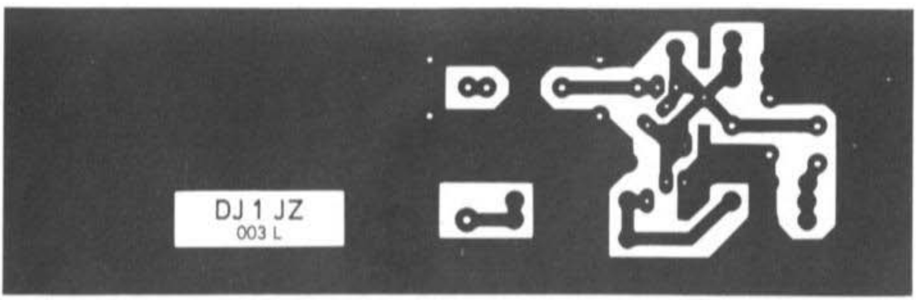

Fig. 3: Preamplifier part of the 120 mm x 90 mm PC-board

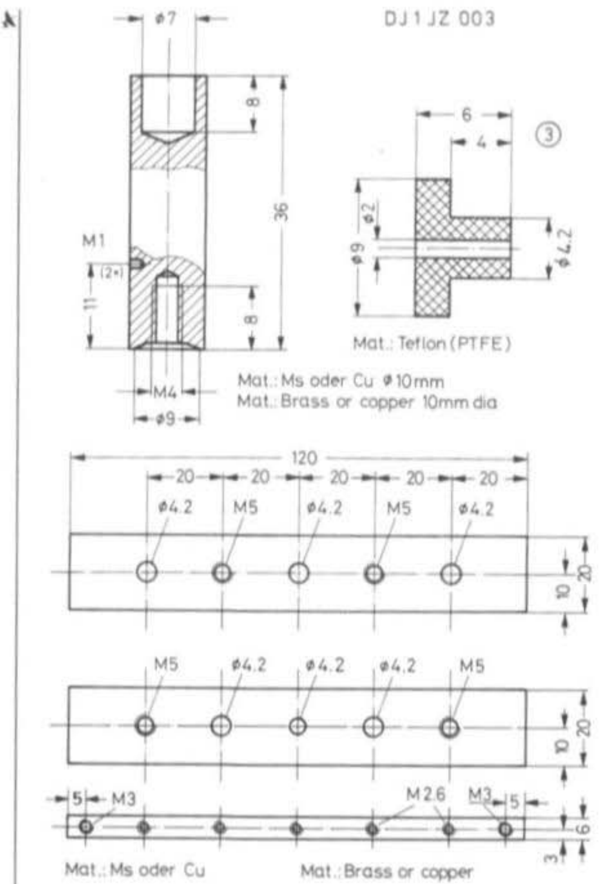

Fig. 4: Side pieces, resonators and PTFE-bushing for the filter

face remains as ground surface, and is only removed using a 3 mm diameter drill at those positions where component connections are passed through the PC-board to a conductor lane. The parts shown in **Figure 4** are constructed before mounting the components onto the board.

The interdigital filter comprises 2 side pieces constructed from 6 mm thick copper or brass plate. These are later screwed to the base plate with the aid of seven 3 mm screws, each. The five interdigital resonators are mounted alternately on these plates. The interdigital fingers are made, as can be seen in Figure 4, from 10 mm diameter copper or brass rods. The 4 mm diameter mounting hole is countersunk using a flat, 9 mm drill in order to ensure that the resonators have a good contact all around the edge with the side panels. If the resonators are to be made from tube, it will be necessary for M 4-nuts to be soldered into place, which should not coincide directly to the edge of the tube, but should be depressed somewhat so that the electric contact around the edge of the tube is also guaranteed.

Two of the resonators are provided with an additional tapped hole – if possible of 1 mm diameter – for connecting the inner conductors of short coaxial cables, preferrably semi-rigid cable. The center resonator is provided with a similar hole at one end for connecting the mixer diode.

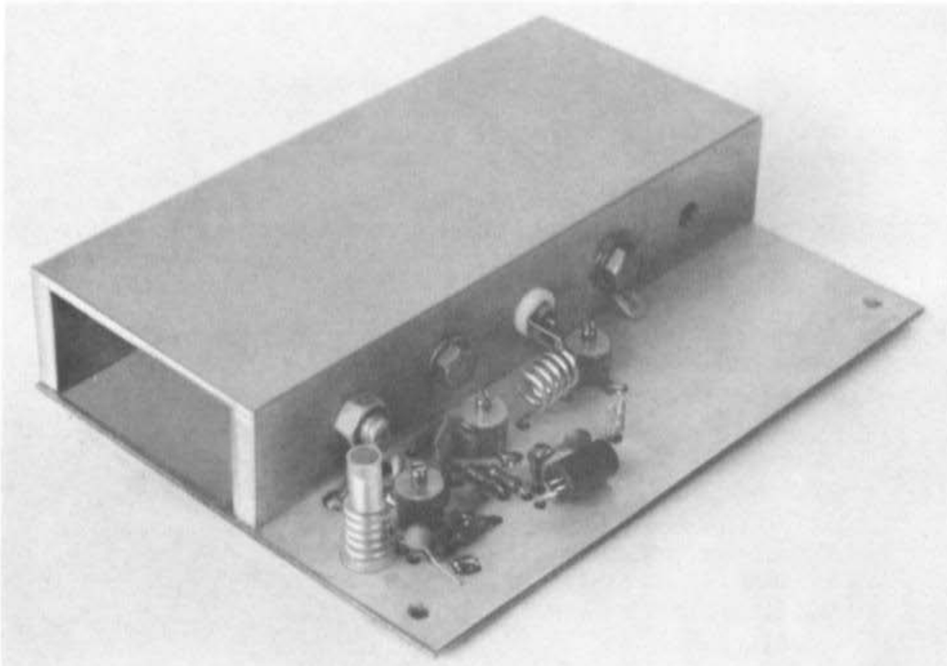

Fig. 5:
A voltage stabilizer or a remote-feeding filter can be mounted on the free-surface of the PC-board

It is now necessary for the UHF-bypass capacitor for the mixer diode to be described briefly. It consists of a metal disk ❶ and a PTFE foil ❷, whereas the other pole of the capacitor is formed by the side piece. This plate capacitor is held together by a M2 screw. The lathed PTFE piece ❸ shown in Figure 4 is required to ensure that no ground contact is made. A double solder tag ❹ is fitted to this as well as a disk and a nut. Choke RFC 1 and the 100 Ω resistor are soldered to one end of the solder tag, whereas the other end is bent toward inductance L 1.

The cover of the filter system is not shown in the diagrams. It is 120 mm x 53 mm, and is made from 1 mm copper (or brass) plate and screwed to the side pieces with the aid of 14 screws. The author's prototype shown in **Figure 5** is somewhat different from this description in that the cover is hard-soldered to the side pieces.

Regarding the material used, it should be mentioned that brass should be galvanically silver-plated, whereas copper need not be silver-plated as long as it is polished. However, the best material and the best description will be of no use if the contact conditions between the resonators, side pieces, base plate, and cover, as well as from the outer sheet of the coaxial cables to the base plate are not good. Also the clamping of the short connection wires of the mixer diode (polarity is not important) must be made with extreme care.

ALIGNMENT OF THE MIXER MODULE

For alignment, one requires a multimeter for DC-voltage, a local oscillator module (will be described in the next edition of VHF COMMUNICATIONS), as well as a 1691 MHz source

TABLE OF MECHANICAL PARTS

Part-No.	Material, Dimensions, Notes	Number
1	Disk, 10 mm outer dia., 2 mm hole, 1 mm thick, copper or brass	1
2	0.13 mm thick PTFE-foil, reinforced with glass fibre, 12 x 12	1
3	PTFE part as shown in Figure 4	1
4	Double-solder tag, if possible for M 2	1
	Screw M 2 x 15, with washer and nut	1
	Side piece, outer, as shown in Fig. 4, center	1
	Side piece, inner, as shown in Fig.4, below	1
	Cover, 120 x 53, 1 mm copper or brass plate, with two rows of holes of 3 mm dia.	1
	Resonators, round copper or brass, 10 mm diameter, see Fig. 4	5
	Threaded bolt, M 6 x 20 mm with nuts, for tuning	4
	Screws M 4 x 12	5
	Screws M 3 x 5	28
	Screws M 1 x 3	3

(crystal oscillator, multiplier with HP 2800, and an interdigital filter), or a parabolic antenna of approximately 2 m diameter pointing to METEOSAT 2 or another geostationary weather satellite.

Firstly check the drain current at the source of the IF-preamplifier; it should be between 2.4 and 3 V across 270 Ω.

This is followed by aligning the oscillator side of the filter for maximum mixer diode current by measuring the voltage drop across the 100 Ω resistor. The outer resonator exhibits a wide resonance due to the loading of the 50 Ω cable, whereas the second resonator is very sharp. For this reason, the locking of the tuning screw must be correspondingly secure. Both crystal frequencies should be switched on alternately and the tuning of the resonator aligned so that the diode current is equal at both frequencies. A voltage drop of 100 mV is good, but more (up to 1 V) is better.

This is followed by connecting the subsequent receiver and aligning the drain circuit and the Pi-link of the IF-preamplifier for maximum noise. It is sufficient when this is done temporarily at first; the fine alignment can be made later when receiving the required signal.

It is now possible for the inner resonator on the antenna side of the interdigital filter to be changed to the oscillator frequency, which will be observed as a deep dip of the mixer diode current. From this position, the tuning screw is rotated out somewhat. If a signal source of at least 0.1 mW at the required frequency is available, it is possible for the two resonators on the antenna side to be aligned for maximum current of the mixer diode, of course, after switching off the oscillator.

Either a noise measuring system or an accurately orientated antenna [see Figure 14 in (5)] is required for the fine alignment.

Whereas the drain circuit comprising L 2 and the inner resonator can be tuned for maximum field strength, the alignment of the first resonator on the antenna side, and the Pi-link between mixer diode and preamplifier must be made for best signal-to-noise ratio. Both METEOSAT channels (1691.0 and 1694.5 MHz) should be switched on alternately during this alignment in order to obtain a suitable passband curve.

If the IF-preamplifier should show a tendency to oscillation, it is possible for a screening panel to be placed across the board at the position shown by the dashed line in Figure 2. It should be 20 mm high. Sometimes, a ceramic capacitor of several 100 pF can be connected with the shortest connections to gate 2 and the source of the BF 900.

The noise figure of this converter will be in the order of 8 dB after the alignment has been optimized. This means that METEOSAT APT-images will be received with a parabolic dish of approximately 2 m diameter. A preamplifier having a noise figure of approximately 2 dB and a gain of at least 16 dB (better 20 dB) will be required if a parabolic dish of 1 to 1.2 m is to be used. Attention should always be paid to keep the cable loss between antenna and preamplifier at a minimum. Sufficiently low-noise figures can be obtained, for instance, using the two-stage preamplifier for the 13 cm band described by DJ 6 PI in (6), which can be simply realigned to 1.7 GHz. This preamplifier is equipped with transistors type NE 64535 and NE 57835 on a PTFE-board.

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Antennas for Reception of Orbiting Weather Satellites in the 137 MHz Band

by Terry Bittan, G 3 JVQ / DJ Ø BQ

Orbiting satellites usually transmit clockwise-circular polarisation in order to minimize the fading effects encountered in this type of reception. However, the received polarisation is usually only truly clockwise-circular when the satellite is passing directly overhead. This means that the polarisation is not constant during the pass, but is mainly elliptical, and can vary beween virtually linear near the horizon, and truly circular-clockwise when overhead.

BASIC SYSTEM CONSIDERATIONS

Professional receiving stations nearly always use tracking systems with several circular-polarized antennas. These are usually either crossed-Yagi or helical antennas. Such systems are large, extremely complicated and expensive, also due to the required computers to control them. Our experiments have shown that they are not usually necessary, and only bring advantages when the satellites are less than 10° above the horizon. However, if we are to use simpler antennas without tracking, this will mean that we have less antenna gain, and thus must increase the overall sensitivity of the system as far as technologically possible, in order to compensate for this.

This was mainly achieved in our case by use of a low-noise preamplifier direct at the antenna, with good quality N-connectors throughout. Even at 137 MHz one can measure noticeable system noise figure differences between N and BNC connectors in a system, and one should not consider using PL-259/SO-239 connectors for anything above 28 MHz.

The noise figure of the preamplifier was 0.9 dB when measured through the connectors and bypass relays. It was equipped with a BF 981 Dual-Gate MOSFET. Of course, it would be possible to reduce the noise figure still further by using GaAs FET's, however, these devices are still non-protected, and are thus more critical during construction and operation. Since the system noise figure was sufficiently low using the BF 981, no attempt was made with GaAs-FETS.

TYPES OF ANTENNAS CONSIDERED

Crossed Dipole Turnstile

The first antenna to be considered was a crossed dipole or turnstile antenna that was phased for clockwise circular polarisation as shown in **Figure 1**. This antenna seemed ideal, since it receives clockwise-circular polarisation when the satellite is directly overhead, and represents an omni-directional horizontally polarized antenna for signals arriving parallel to the ground surface. Unfortunately, the results were not as good as expected due to a considerable amount of fading observed during the pass, which was found to be the result of multipath propagation caused by ground and other reflections.

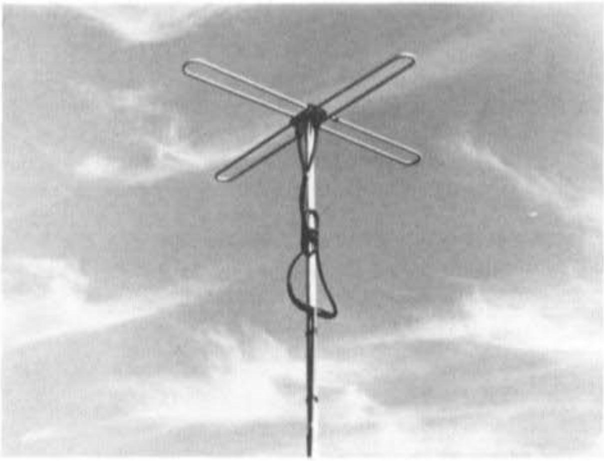


Fig. 1:
A crossed-dipole turnstile antenna

Discone Antenna

The discone **(Figure 2)** is a very useful wideband antenna that is mainly vertically polarized. In our case, it covered a frequency range of 80 MHz to 500 MHz. This antenna was selected for the second experiment, because it was not so strictly linear polarized as a conventional dipole. The results obtained with this antenna were better than the crossed-dipole at lower angles, and provided good signals throughout all passes with only some minor fading when the satellite is directly overhead. The disadvantage was not in the reception of the satellite signal, but in the reception of considerably more ignition and other electrical interference. Also, although the deep fading was less than with the crossed dipole, there were still certain angles where the signal disappeared into the noise due to multipath reception.

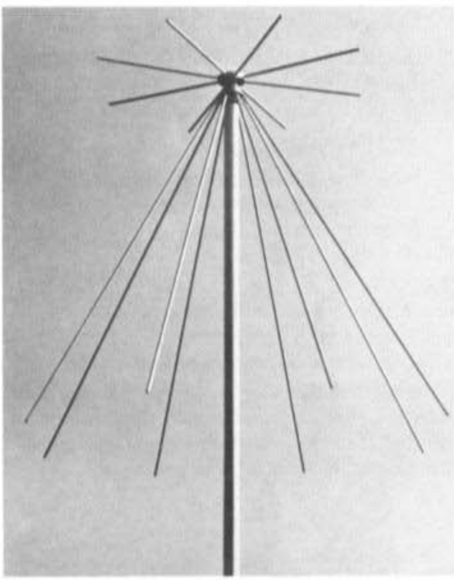


Fig. 2:
A DISCONE antenna for 80-500 MHz

Stacked Colinear Antenna

This antenna was mainly tried in order to increase signal strengths when the satellite is at or near the horizon. Of course, this antenna is completely unsuitable for overhead passes, and also received a considerable amount of ignition and electrical interference.

Fixed Helical Antenna

One of the easiest means of obtaining circular polarisation is to use a helical antenna, and this antenna was also considered. Unfortunately, truly circular polarisation is not obtained until using at least a five-turn helix. This means, however, that the antenna would have a beamwidth of approximately 50°, which is far too narrow for our application.

Big-Wheel Antenna

The big-wheel antenna was also tried at the same time as the discone antenna. It also provided good signals at the horizon and a good rejection of electrical and ignition interference.

However, considerable fading occurs at higher angles due to the signal reaching the three dipoles at different phase angles. This causes deep fading similar to the multipath propagation noticed with other antennas.

Fixed Two-Element Crossed Yagi

After some consideration, it was assumed that the deep fading encountered when using the crossed-dipole, big-wheel, and discone antennas was due to multipath propagation, and was mainly caused by ground reflections. These effects completely disappeared when extending the crossed-dipole to form a two-element crossed-Yagi by adding 2 reflectors at a spacing of $\lambda/4$ below the dipoles. Such an antenna is shown in **Figure 3**.

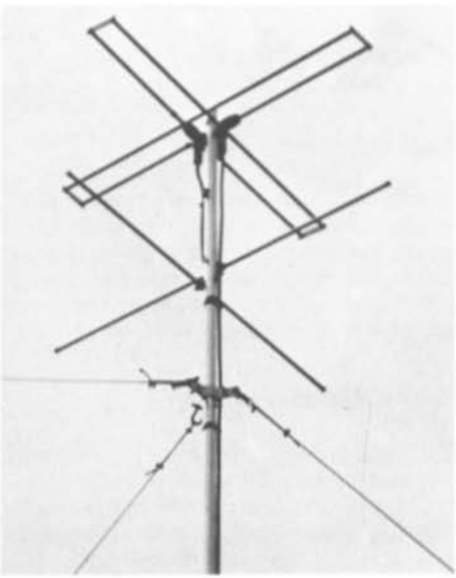


Fig. 3:
A fixed two-element crossed Yagi

The result was a constant signal throughout the pass with virtually no fading. The only disadvantage was that the reception at the horizon was not quite satisfactory due to the reduced beamwidth of the antenna.

At this point, the author would like to point out that a crossed-Yagi antenna phased for circular-polarisation will have a different beamwidth than a conventional 2-element Yagi antenna. Usually one can expect a beamwidth of approximately 75° in the E-plane, and about 140° in the H-plane. In the case of a crossed-Yagi where both dipoles are in use, the beamwidth is a mean value of H and E-planes, and is in the order of 100°. Readers should also remember that circular-polarized antennas will also provide 3 dB **more** gain than their linear counterparts when receiving circular polarisation !

FINAL ANTENNA CONCEPTION

The results of the above experiments showed that an antenna was required that was able to satisfy the following demands:

- Clockwise circular polarisation
- Suppression of ignition and electrical interference
- Directional characteristic to suppress multipath propagation
- Widest possible beamwidth in order to provide satisfactory reception at the horizon.

After determining that the 2-element crossed-Yagi fulfilled all these demands except for the requirement of a sufficient signal strength at the horizon, the author attempted to find ways of increasing the beamwidth of the antenna without losing the directional characteristics required for suppressing the ground and multipath reflections.

The solution was finally found by changing the dipole-reflector spacings from the original $\lambda/4$ required for maximum gain to a spacing of

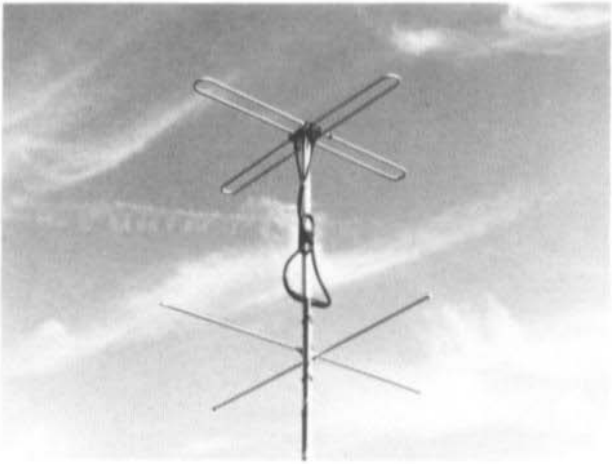


Fig. 4:
Final antenna conception with 3/8 λ spacing between radiator and reflector

3/8λ. **Figure 4** shows a photograph of the author's prototype. This caused the main lobe to start splitting into two lobes, which resulted in a flatter and wider main lobe. The resulting H-plane lobe is shown in **Figure 5** for a linear-polarized 2-element Yagi antenna. It is somewhat less for a circular-polarized crossed Yagi antenna due to the mean values between the H and E-planes. However, the increase in spacing causes the main beam to widen to the value of approximately 15%. The result is a noticeable improvement in the reception of satellite signals near the horizon.

PHASING

The same is valid for phasing this antenna for clockwise circular polarisation as has been already written in many articels in this magazine (1), (2). Firstly mark the hot end of each dipole: That is that end of the folded dipole to which the center conductor of the feeder cable is directly connected.

The feeders of the dipoles should be cut so that one feeder is an electrical halfwave, and

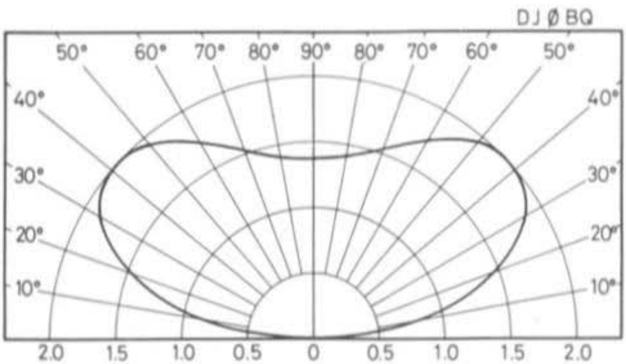


Fig. 5:
H-plane beamwidth of a two-element Yagi antenna with a spacing of 3/8 λ

the other is 3/4λ. When using either RG-213/U or RG-58/U, these lengths will be 72 cm and 107.9 cm respectively. Provide each of these feeders with a good quality N-connector. The two dipoles are now coupled together with the aid of a power divider as available from UKW-TECHNIK. Since these power dividers are low-loss wideband devices, it is possible to use the standard 144 MHz types.

The dipole with the shortest feeder is now mounted on the vertical boom. This is followed by mounting the second dipole so that the hot-end marking is 90° in a clockwise direction from the marking of the first dipole when viewed from the reflector. The antenna is now phased for clockwise-circular polarisation.

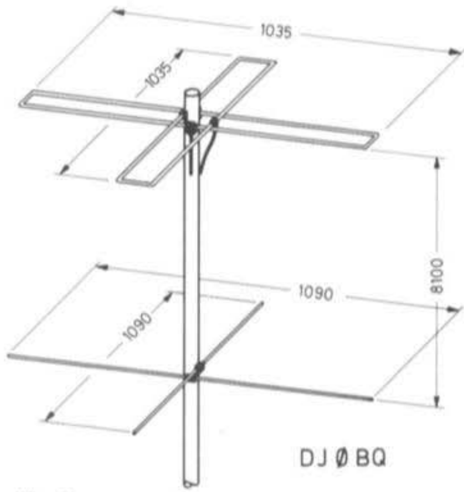


Fig. 6:
Construction details of the antenna

CONSTRUCTION

After taking the velocity factors into consideration, the dimensions of the antenna **(Fig. 6)** are as follows:

Dipole:	103.5 cm
Reflector:	109 cm
Spacing:	81 cm
Length of balun cable:	λ/2xVF = 143.8 cm

Forecasting for Reception Times of Orbiting Satellites

by Terry Bittan, DJ Ø BQ / G 3 JVQ

It is rather difficult for many of our readers to obtain the required equator crossing times and angles for the orbiting satellites in the 137 MHz band. This short article is to show where such information is available, and what one can do as an amateur to forecast one's own times.

AMERICAN SATELLITES OF THE NOAA SERIES

The equator crossing times and angles are available via theaeronauticalweather service. They are stored inthe data bank ofthe service in Brussels, and can be called off by teleprinter from any aeronautical weather office. This means that if you have good relations with your nearest weather office, you will be able to always obtain the required times.

During the winter months, the evening and night passes over the cold (low contrast) surface do not provide worthwhile infra-red images for us amateurs, and need not be considered. Even the morning passes of NOAA 6 are in the twilight zone and thus coldest part of the day, and will not be of great interest until the shortest days start to get longer. This means that it is mainly the afternoon passes of NOAA 7 that are of interest both in the infra-red and visible ranges. Of these, the overhead pass, and the previous and subsequent orbits to the east and west respectively, are of interest.

RUSSIAN SATELLITES OF THE METEOR SERIES

It is virtually impossible to obtain any information on the USSR weather satellites. No information is available in the European data bank, or through weather office sources. This means that one must rely on one's own observations. These times are not reliable for the following reasons:

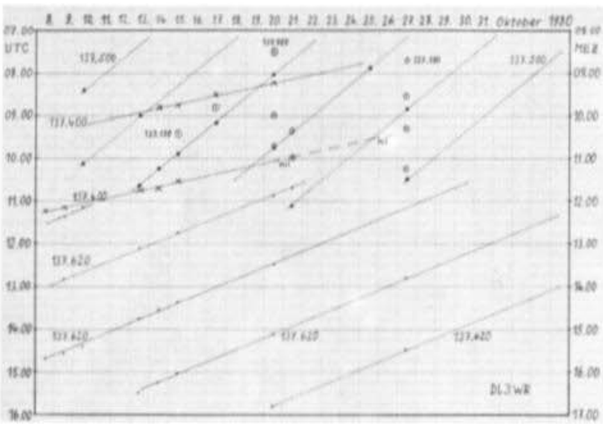
The Russians seem to have a large number of satellites in orbit, which they switch on and off according to the charge conditions of the batteries. At least this is assumed by the author, since there is no regularity to be found when observing these satellites over a period of time. This is just as true for the 120 line-per-minute satellites on 137.300 and 137.400 MHz, as for the 240 lines-per-minute satellites on 137.150 MHz.

Another problem is that the control stations switch off the satellite before it goes out of range of the USSR, presumably also to conserve battery energy. These satellites also only transmit during daylight hours, since they only provide visible, and no infra-red images.

FORECASTING THE TIMES OF RECEPTION FOR YOUR LOCATION

Assuming that you are not able to obtain the required information, it is possible to forecast the times of reception in a method used by DL 3 WR. After observing a frequency on two days, one can forecast the orbits for upto 30 days in advance. This is made in the following manner:

Firstly prepare a sheet of paper as shown in the following **diagram** with the time (either local or GMT) on the vertical scale, and the dates on the horizontal scale. The commencement of reception for each frequency is noted at the intersection of time and date. After two such points have been marked on different days it is possible to join these two points with a straight line, which can be extended to the end of the page. This should be made for each orbit and satellite (frequency). The times of reception for any subsequent days can then be found by following the date line until the orbit line is crossed and reading off the appropriate time. Since the orbital period is also known, it is possible to determine the previous andsubsequent orbit times.



Example for the reception times of several orbiting satellites of the NOAA and METEOR series

A Home-Made UHF/SHF Power Meter

by O. Frosinn, DF 7 QF

A power meter is to be described that operates according to the bolometer principle, and can be constructed with simple components and tools. The most important characteristics are:

- **The calibration canbemade with a DC-voltage**
- **The meter continues to use its linear scale**
- **Usable in the frequency range from DC to several GHz**
- **3 measuring ranges (full scale): 10 mW / 100 mW / 500 mW**
- **Accuracy: ± 0.5 dB**

The only disadvantages are that the measuring and cooling phase require several minutes, and that suitable attenuators will be required to measure higher power levels.

METHODS OF RF POWER MEASUREMENTS

The normal way of measuring RF-power is to rectify the RF-voltage across the consumer impedance, or a part of it, and to indicate the resulting DC-voltage. **Figures 1 and 2** show the principle of this in the form of diagrams: Home-construction is also possible, but the difficulties increase with the frequency. In the case of absolute-value indication, it is necessary for the power meters to be calibrated by comparing them to professional power meters.

Instead of measuring the DC-voltage after rectification, it is possible to measure the heating of a terminating resistor by the RF-power. Such circuits are shown in **Figures 3 and 4**, and the two interconnected circles represent a thermal coupling.

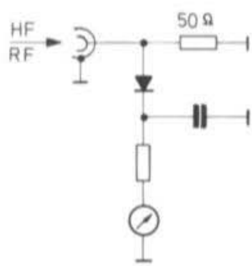


Fig. 1: Direct RF-voltage measurement at the load resistor

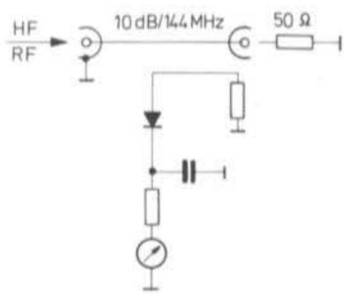


Fig. 2: RF-voltage measurement using a directional coupler

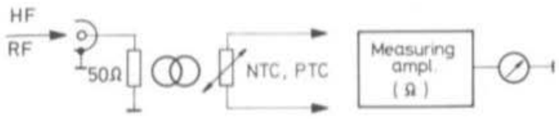


Fig. 3:
Heat measurement at the load resistor using a temperature-dependent resistor

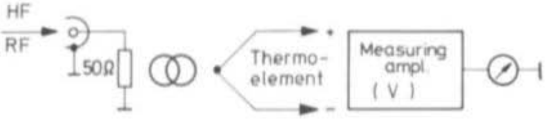


Fig. 4:
Heat measurement at the load resistor using a thermo-element

The heat generated in the 50 Ω resistor also heats a temperature-dependent resistor or thermo-element. This variation in the secondary circuit is evaluated, which means that the power effective in the primary circuit can be indicated. It is completely immaterial whether the electric energy used for heating is a DC-voltage, or alternating voltage, or whether it is a low-frequency or a high-frequency voltage. This allows the meter to be calibrated with the aid of a DC-power.

Measuring errors result due to the difference between the impedance of the terminating resistor and interconnection cable, in other words from too low a return loss, or too high a standing-wave-ratio. However, when the SWR remains below a value of 2, this error will be less than 0.5 dB, and usually negligible for amateur applications. These values can be taken from **Table 1**, which is given in more detail since it is not readily available to all radio amateurs.

In order to complete this section, it should be noted that all four methods of measuring power are used in commercially available power meters: Through-line and dummy-load wattmeters suchasthose from »Bird« operate according to Figures 1 and 2. Power meters manufactured by Siemens and HP usually operate according to the principle shown in Figures 3 and 4. For amateur applications, virtually only wattmeters operating according to principle 1 and 2 are available; their accuracy is very often unsatisfactory even on the specified frequency range. Outside of this range,itis always necessary for a recalibration to be made, which usually means that an extra scale must be made on the meter. Another problem with this type of measuring equipment is that the measuring diode is usually destroyed at high RF-power levels. If an absolutely identical diode is not available, it will then usually be necessary to make a new alignment, and redrawing of the scales.

THE BOLOMETER

The construction details of a reflectometer in (1) led the author to design a bolometer (heat meter) as the heart of a very-accurate power meter, which was usable from DC-voltage to several GHz. As can be seen in Figure 3, the power meter consists of three main modules:

- Terminating resistor (50 Ω)
- Temperature probe
- Measuring-amplifier with readout

The construction of these is now to be described.

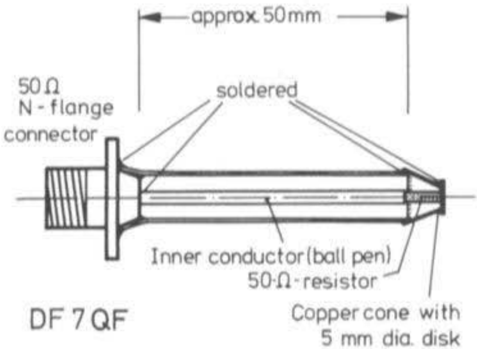


Fig. 5:
Coaxial terminating resistor onto which the temperature probe is mounted

THE TERMINATING RESISTOR

Since the temperature probe is in the form of a Wheatstone bridge, one requires two terminating resistors that should be as identical as possible. The construction is described in **Fig. 5** and should be made as carefully as possible, since the specifications are mainly dependent on it. After completion, the matching (VSWR or return loss) should be measured at the highest possible frequency. The best one is then used as measuring resistor, and the other as compensating resistor. The values obtained by the author are given in curve A of **Figure 6**; for comparison, curve B shows the VSWR-run of a 50 Ω terminating resistor manufactured by HP.

It is now necessary to obtain two suitable resistors. In order to ensure that low powers can also be measured, it is necessary to use a mechanically small version (providing sufficient heat at low power levels); of course, this also limits the rating towards higher power levels. The author used metal oxide resistors manufactured by Draloric, type MLAD 0309/ 50 Ω. These resistors do not possess helical windings and are rated at 62.5 mW. However, the measuring resistor has been loaded up to 500 mW in the described circuit for more than three years, and even overloaded up to several Watts for short periods. No deterioration has been noted. If other resistors are used, it may be necessary for the mechanical dimensions of the inner and outer conductors to be changed in order to compensate for the jumps in diameter.

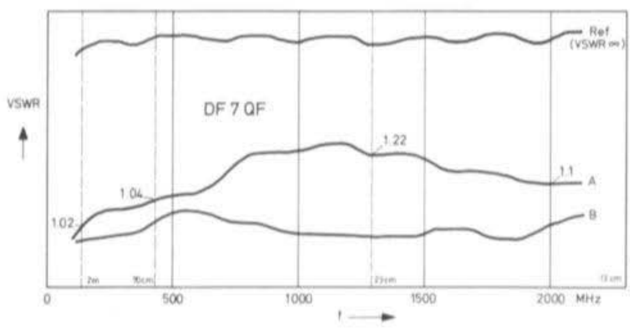


Fig. 6:
Matching of the terminating resistor shown in Fig. 5 between 100 and 2300 MHz

Return loss a_r/dB	Reflection factor r	Matching factor m	Voltage standing wave ratio s	Attenuation due to mismatch a/dB
0,1	0,988	0,006	173,368	16,42
0,2	0,977	0,011	86,835	13,46
0,4	0,957	0,022	45,436	10,74
0,6	0,933	0,034	28,971	8,89
0,8	0,912	0,046	21,725	7,74
1,0	0,891	0,058	17,349	6,86
1,5	0,841	0,086	11,578	5,35
2,0	0,794	0,115	8,709	4,32
2,5	0,750	0,143	6,997	3,59
3,0	0,708	0,171	5,842	3,02
3,5	0,668	0,199	5,024	2,57
4,0	0,631	0,226	4,442	2,20
4,5	0,595	0,254	3,933	1,90
5,0	0,562	0,280	3,571	1,65
5,5	0,531	0,306	3,263	1,45
6,0	0,501	0,332	3,012	1,25
6,5	0,473	0,358	2,793	1,10
7,0	0,447	0,382	2,618	0,97
7,5	0,422	0,406	2,463	0,85
8,0	0,398	0,431	2,320	0,75
8,5	0,373	0,453	2,203	0,66
9,0	0,355	0,476	2,101	0,58
9,5	0,335	0,498	2,008	0,51
10,0	0,316	0,520	1,923	0,45
10,5	0,298	0,541	1,848	0,40
11,0	0,282	0,561	1,780	0,36
11,5	0,266	0,579	1,726	0,32
12,0	0,252	0,598	1,671	0,28
12,5	0,237	0,618	1,618	0,25
13,0	0,224	0,634	1,578	0,22
13,5	0,211	0,650	1,538	0,20
14,0	0,199	0,668	1,497	0,17
14,5	0,183	0,684	1,462	0,15
15,0	0,178	0,699	1,480	0,13
15,5	0,165	0,716	1,396	
16,0	0,158	0,727	1,374	0,11
16,5	0,150	0,740	1,350	
17,0	0,141	0,752	1,329	0,09
17,5	0,133	0,766	1,304	
18,0	0,126	0,777	1,285	0,07
18,5	0,119	0,789	1,268	
19,0	0,112	0,799	1,251	0,05
19,5	0,106	0,809	1,235	
20,0	0,100	0,819	1,220	0,04
20,5	0,094	0,828	1,208	
21,0	0,089	0,887	1,193	
21,5	0,084	0,846	1,180	

Return loss a_r/dB	Reflection factor r	Matching factor m	Voltage standing wave ratio s	Attenuation due to mismatch a/dB
22,0	0,079	0,853	1,171	0,03
22,5	0,075	0,861	1,160	
23,0	0,071	0,868	1,151	
23,5	0,067	0,857	1,142	
24,0	0,063	0,882	1,133	0,02
24,5	0,060	0,888	1,124	
25,0	0,057	0,894	1,118	
25,5	0,053	0,900	1,111	
26,0	0,050	0,904	1,105	0,01
26,5	0,047	0,909	1,100	
27,0	0,045	0,914	1,094	
27,5	0,042	0,919	1,088	
28,0	0,040	0,924	1,082	
28,5	0,038	0,928	1,078	
29,0	0,035	0,932	1,073	
29,5	0,034	0,934	1,069	
30,0	0,032	0,938	1,064	
30,5	0,030	0,942	1,060	
31,0	0,028	0,945	1,056	
31,5	0,027	0,947	1,054	
32,0	0,025	0,951	1,051	
32,5	0,024	0,953	1,048	
33,0	0,022	0,956	1,045	
33,5	0,021	0,958	1,043	
34,0	0,020	0,961	1,040	
34,5	0,019	0,963	1,038	
35,0	0,018	0,965	1,036	
35,5	0,017	0,967	1,034	
36,0	0,016	0,969	1,032	
36,5	0,015	0,971	1,030	
37,0	0,014	0,972	1,028	
37,5	0,013	0,974	1,027	
38,0	0,013	0,975	1,025	
38,5	0,012	0,976	1,024	
39,0	0,011	0,973	1,022	
39,5	0,011	0,979	1,021	
40,0	0,010	0,980	1,020	

Table 1:
Four different ways to look at mismatch;
Last column: Attenuation caused by mismatch

CONSTRUCTION OF THE TERMINATING RESISTORS

As can be seen in Figure 5, the terminating resistor is constructed in conjunction with an N-flange connector. The inner conductor of this connector is lengthened using a metal tube (from a ball pen) which also has a diameter of 1/8 inch. The outer conductor is made from a thin brass plate by winding this around a 7.5 mm diameter drill, after which it is soldered. This arrangement results in an impedance of approximately 51 Ω ($Z = 138 \log D/d$).

The lacquer at the ends of the resistor is carefully removed after which the connection wires are soldered out, since they are only 1 mm from each other in the ceramic body of the resistor. It is now possible for the resistor to be soldered onto the tinned inner-conductor tube.

The outer conductor should be cut at one end, bent out in the form of a cone, and soldered to the flange of the connector as shown in Fig. 5.

The other end of the outer conductor is cut to the same length as the inner conductor so that the resistor now protrudes.

This is followed by making a suitable cone from 0.1 to 0.2 mm thick copper foil which is then soldered to the outer conductor. The narrow end is finally soldered quickly to the resistor and a small copper disk of 5 mm diameter. The temperature probe is glued to this disk afterwards with a quick-drying glue.

Both terminating resistors are now mounted onto a thick aluminium front panel so that the connectors are accessible from the front, with a spacing of several cm from another. A heat-insulating casing should be constructed, which is placed around both terminating resistors. This can be made from aluminium plates onto which styrene foam has been glued. This ensures that no short-term temperature fluctuations will have an effect on the measurement.

TEMPERATURE PROBE

The temperature probe circuit given in **Fig. 7** is in the form of a Wheatstone bridge. Nothing is connected to input »K«; this resistor with temperature probe maintains the bridge balance during slow variations of the ambient temperature. If, on the other hand, power is provided to connector »M«, the terminating resistor will heat the PTC-resistor, which unbalances the bridge. A voltage is generated between connections A and B, which is evaluated and indicated.

A silicon temperature sensor with a high positive temperature coefficient (PTC-resistor) has been found successful as temperature probe. Experiments with NTC-resistors and with diodes, on the other hand, were not satisfactory. Such a probe is found in the Siemens product range as type KTY 10, with four different tolerances of the nominal resistance of 2000 Ω: KTY 10 A: ± 1%; .....B: ± 2%; .....C: ± 5%; ..... D: ± 10%. All four classes are suitable; however, the alignment is easiest when using themore expensive,low-tolerance types A, or B. By the way, the external shape is in the form of a plastic case similar to TO 92.

Due to their low temperature coefficients, metal oxide resistors are used in the bridge circuit (2 x 2 kΩ, 1 x 100 Ω, and 1 x 20 Ω). The 500 Ω potentiometer is a ten-turn helical trimmer potentiometer, and the 100 Ω potentiometer is a carbon potentiometer with linear characteristic. This potentiometer can be operated on the front panel. The two 2 kΩ resistors are coupled together thermally using a strip of copper foil.

The construction of the bridge circuit is made according to Figure 7 on the rear of the heat insulating case of the two terminating resistors, and is not in the form of a PC-board. The voltage stabilizer 7806, however, must be mounted so that it does not heat this case and the bridge circuit.

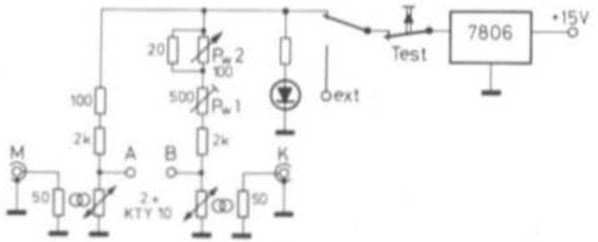


Fig. 7: A Wheatstone-bridge as temperature-probe circuit

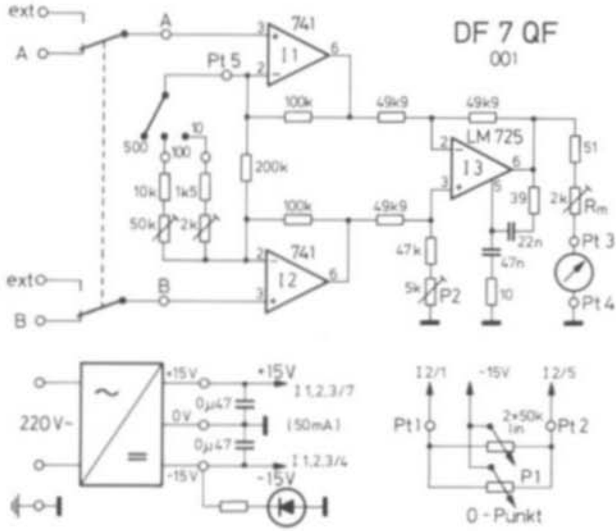


Fig. 8: The evaluation circuit of the bridge voltage, including power supply

MEASURING AMPLIFIER

As can be seen in **Figure 8**, three integrated amplifiers are connected together in such a manner that the common-mode suppression at P 2 is brought to values of over 100 dB. The gain is determined by only one resistor, and both inputs have approximately the same impedance.

If type 741 is used as integrated amplifier, Pin 5 of I 3 may not be connected. The given frequency response compensation need only be provided when a LM 725 is used for I 3.

A 80 mm x 70 mm PC-board was designed for accommodating the amplifier circuit as shown in Figure 8 (without small power supply), which is shown in **Figure 9**. Sockets are provided for the operational amplifiers. The PC-board is soldered into an RF-tight case, and all connections are made with the aid of feed-through capacitors.

All lines are screened, and the power supply is also screened in an RF-tight manner, since it was found that the amplifiers could be effected by spurious RFvoltages. An abrupt deflection of the meter during a measurement will show that an unwanted RF-voltage is being rectified somewhere and indicated. If this is noticed, the screening should be checked and the leads bypassed in a more efficient manner.

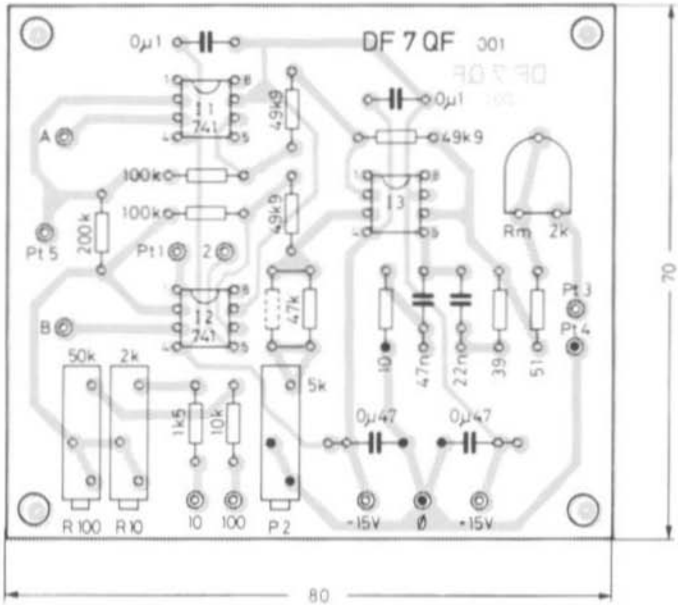


Fig. 9: Component locations on the single-coated PC-board for the evaluation circuit

ALIGNMENT

Resistor R_m is set to its maximum value, potentiomters P 1 and P 2 set to a center position, and the range switch set to position 500 mW. Connections A and B are connected together and to ground. The »Test«-switch is placed in the »open« position, after which the power supply of ± 15 V (approximately 50 mA) is switched on.

The needle of the meter is now set to zero with the aid of P 1. Close the test switch and align the indication once again to zero with the aid of P 2. This optimization of the common-mode suppression is repeated and improved in the 100 mW and 10 mW ranges.

The short-circuit and the ground connection of points A and B are now removed, after which they are connected to A and B of the bridge circuit. The test switch is closed in the 500 mW range, and the meter-reading set to zero with the aid of P_W 1. Potentiometer P_W 2 is in its center position, and is only used for fine alignment during practical operation . This bridge alignment is also made in the 100 mW and 10 mW ranges, whereby the amplifier is set to zero with P 1, with the »Test«-switch open.

The range switch now is set to 500 mW, and 5 V DC-voltage connected to the measuring input M. The meter needle starts to move slowly. When the power across the 50 Ω resistor amounts to 500 mW according to Ohm's Law, this value should also be indicated. After a period of several minutes when the needle stops moving, it should be adjusted to read 500 with the aid of R_m. The large time constant is the disadvantage of this measuring principle.

After allowing several minutes for cooling, the 100 mW full scale can be adjusted with R 100 (spindle trimmer), and the 10 mW full scale with the aid of R 10. The required voltage U can be calculated according to Ohm's Law:

$$U = \sqrt{P \times R}$$

This results in the following values:

100 mW: U = 2.236 V
10 mW: U = 0.71 V

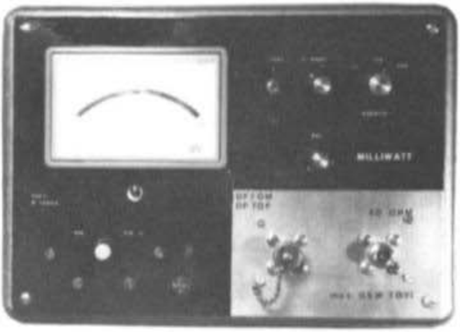


Fig. 10:
The author's prototype

Fig. 11:
Inside are 3 screened modules and a meter

It is advisable for the whole alignment to be repeated several hours later. If everything is stable and working correctly, it is now possible for the power meter to be installed in a case.

FINAL NOTES

Before every measurement, the wattmeters should be allowed to »warm up«. After this, the zero alignment should be made with the aid of P 1 with the »Test«-switch open, which is then followed by balancing the bridge with the switch closed with the aid of P_w2.

As can be seen, the measuring accuracy is mainly dependent on the accuracy of the DC-voltage used for calibration, on the accuracy of the 50 Ω resistor up to the highest possible frequencies, on the linearity of the PTC-resistor, and on the accuracy of the meter. However, this should not stop you constructing this power meter even if you do not have access to high-precision laboratory measuring equipment. A good meter, and a good (borrowed) multimeter – or better a digital multimeter – for calibration will allow you to construct a valuable power meter that can be used also for microwave measurements !

The author is also experimenting with a measuring and compensating resistor that can be connected to the »external« input. This is even more simple to construct, and seems to allow measurements even in the 3 cm band.

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Hewlett Packard, August 1977
Application Note 64-1

An Easy to Build TV Pattern Generator

by L. Damrow, DC 7 EP

Pattern generators are very useful for checking and aligning TV-transmitters and receivers. Whereas three boards full of TLL-integrated circuits were required some years ago (1), it is now possible for this to be achieved using a single integrated circuit manufactured by National. This integrated circuit is designated MM5322N; it is a pattern generator in MOS-technology, whose power requirements are only 35 mA at 15 V.

This integrated circuit operates according to the American standard NTSC with 60 Hz/525 lines. However, the author has been operating this pattern generator in conjunction with CCIR receivers and has found that the output signal is within the lock-in range of most CCIR receivers – the line length only differs by 0.6 µs.

This means that the described circuit can be used both for those countries using the US-standard, and for those countries where the CCIR-standard is used.

The integrated circuit can generate 16 different patterns, which can be selected together with a frequency divider that can be switched in BCD-code. The IC also contains three rainbow patterns, which can be used in conjunction with an accessory circuit to form a NTSC-colour composite video signal.

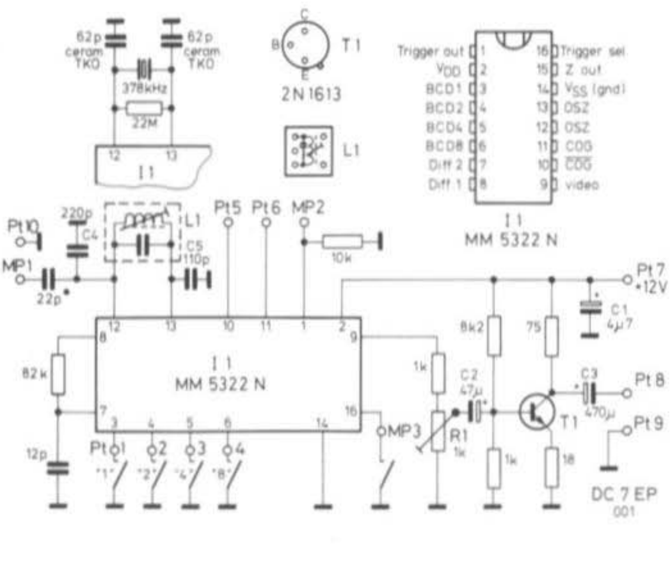

Fig. 1:
The output coupling capacitor at MP 1 should be as small as possible so that the connected frequency counter is just able to indicate.
The switch at MP 3 is for selecting the vertical or horizontal trigger pulses at MP 2. Value and TC of the ceramic capacitors adjacent to the crystal depend on the actual characteristics of the crystal itself.

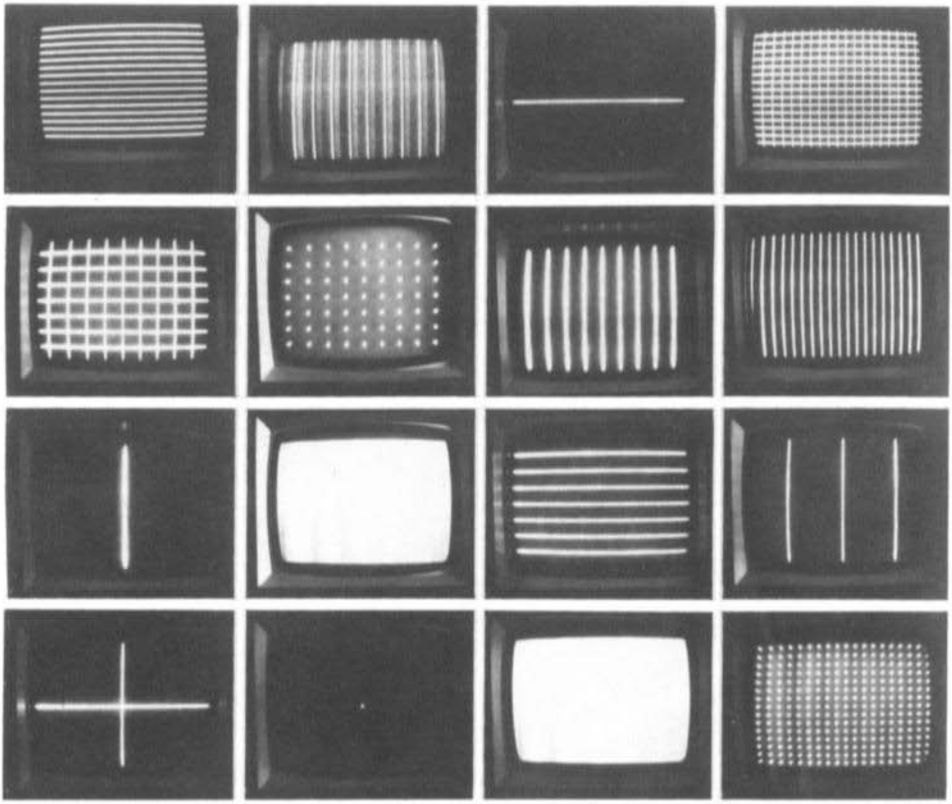

Fig. 2:
These 16 patterns can be selected using the four switches connected to Pt 1 - Pt 4.
All images were taken in conjunction with an ATV-transmitter as described by DJ 4 LB.

Since the majority of our readers are not in NTSC-areas, the following article is to be limited to the monochrome (black/white) applications of this circuit.

CIRCUIT DESCRIPTION

Figure 1 shows the complete circuit of the small pattern generator including the external components, and also includes an alternative oscillator circuit. The integrated circuit is provided with the control signal for the pulse processing, which usually uses a 378 kHz crystal. The author's experiments have shown, however, that it is sufficient for amateur radio applications to use a 455 kHz resonant circuit instead of the crystal. In this case, the resonant circuit is tuned to 378 kHz and will provide sufficient stability.

A composite video signal will be available at pin 9 of the MM 5322 N, which is inverted and amplified in the subsequent wideband amplifier. With the aid of trimmer R 1, it is possible to select the standard video level of 1 V peak-to-peak at 75 Ω at the output connections Pt 8/Pt 9.

The signal for controlling the colour burst can be taken from pins 10 and 11 (Pt 5 and Pt 6).

The required test pattern can be selected via pins 3-6 (Pt 1 - Pt 4). The easiest manner is to use four toggle switches, with which one can set the following valencies according to BCD-code:

Switch to Pt 1:	Valency 1
Switch to Pt 2:	Valency 2
Switch to Pt 3:	Valency 4
Switch to Pt 4:	Valency 8

It is now only necessary to study **Table 1** in order to find out which pattern can be selected with which BCD-number:

1	2	4	8	Pattern
L	L	L	L	15 horizontal lines
H	L	L	L	Rainbow
H	H	L	L	1 horizontal line
L	L	H	L	Fine grid
H	L	H	L	Wide grid
L	H	H	L	63 dots
H	H	H	L	9 vertical lines
L	L	L	H	19 vertical lines
H	L	L	H	1 vertical line
L	H	L	H	Rainbow
H	H	L	H	7 horizontal lines
L	L	H	H	3 horizontal lines
H	L	H	H	Cross
L	H	H	H	Dot in the centre
H	H	H	H	Rainbow
L	H	L	L	285 dots

Table 1:
H = Switch open
L = Switch closed

Figure 2 shows these 16 patterns in the given order.

CONSTRUCTION

The complete circuit is accommodated on a single-coated PC-board of only 73 mm x 73 mm. The component locations are given in **Figure 3**. This board has been designated DC 7 EP 001.

A 16-pin DIL-socket is soldered into the board for accommodating the integrated circuit, which should not be inserted until all other components have been mounted. All unused connections of the resonant circuit for the clock oscillator should be removed before mounting.

The PC-board should be mounted in a screened case after completion. The dimensions of this case are 75 mm x 75 mm x 30 mm. The operating voltage and the connections for the four switches are made via feed-through capacitors of approximately 2 nF soldered to the side panel of the metal case. The composite video signal is taken via a thin coaxial cable such as RG-174/U. **Figure 4** shows a photograph of the author's prototype using a Vero-board.

SPECIAL COMPONENTS

I 1:	MM 5322 N (National)
T 1:	2N 1613, 2N 2219 A or similar
L 1:	455 kHz-resonant circuit (Japan) white, yellow, or black with built-in capacitor
C 1, C 2, C 3:	Aluminium electrolytic for 5 mm spacing, or tantalum electrolytic
C 4:	220 pF to 270 pF, styroflex
C 5:	110 pF to 150 pF, styroflex

All other capacitors:
ceramic tubular or disk type

All resistors for 10 mm spacing.

ALIGNMENT

The alignment is also very simple. After connecting a stabilized operating voltage of 12 V, it is possible to measure a frequency of approximately 400 kHz at testpoint MP 1 with the aid of an oscilloscope or frequency counter. This can be brought to an exact frequency of 378 kHz with the aid of L 1.

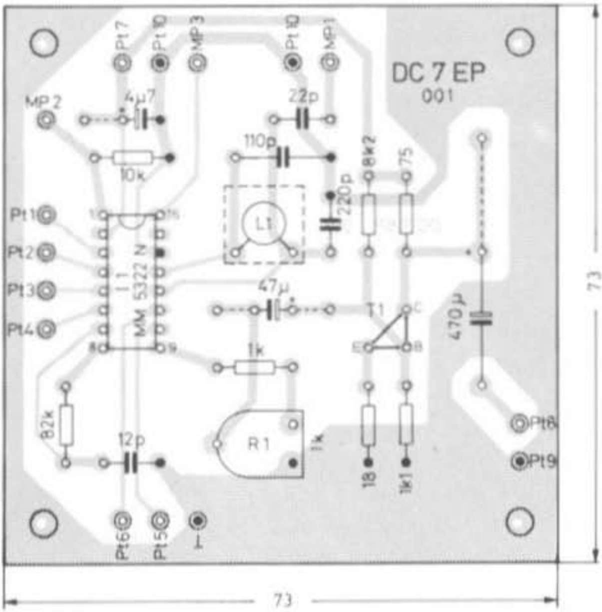

Fig. 3:
The single-coated PC-board for the pattern generator. Dimensions are 73 mm x 73 mm.

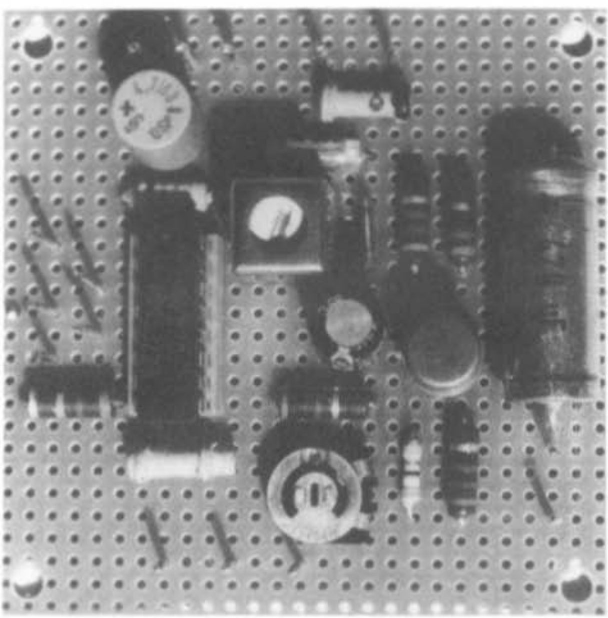

Fig. 4:
The uncritical circuit can also be constructed on a Vero-board

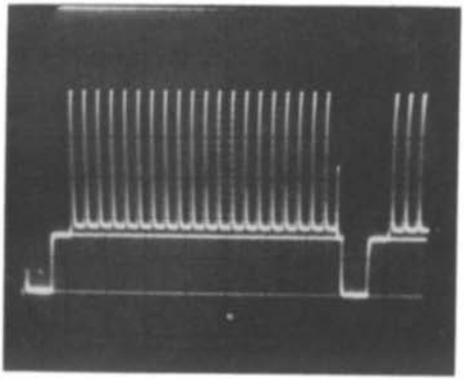

Fig. 5:
One line of composite video signal on an oscilloscope

Output Pt 8/Pt 9 is then terminated with a 75 Ω resistor and connected to an oscilloscope. The pattern generator is now switched for »19 vertical lines« (LLLH), and the oscilloscope set to 10 μs/cm and the amplifier to 0.2 V/cm. The oscillogram given in **Figure 5** should now result on the screen of the oscilloscope. The standard level of 1 V_{pp} is set with the aid of R 1. This completes the alignment of the pattern generator. The 75 Ω resistor should be removed after alignment.

MEASURED VALUES

Operating voltage:
12 V, stabilized (11-16 V stabilized is possible)

Operating current at Pt 7:
35 mA at 12 V; 60 mA at 16 V

Output impedance: 75 Ω

Output voltage (peak-to-peak): 1 V at 12 V
1.1 V at 16 V

Frequency stability of the LC-oscillator during operation:
Variation of the frequency on changing the operating voltage from 12V to 16V: – 470 Hz

Drift during the first 30 minutes:
less than 500 Hz

Drift after 30 minutes operation:
less than 20 Hz/h

The following experiment was made to determine the permissible tolerance of the oscillator:

The pattern generator DC 7 EP 001 was connected to an ATV-transmitter as described by DJ 4 LB. The transmit and receive antenna were both HB 9 CV types. On the receive side, the author used a Microwave Modules ATV-converter inconjunction with aJapanese TV-receiver.

No difficulties were observed on altering the operating voltage of the pattern generator between 11 and 16 V.

Synchronization was no longer possible after exceeding the following frequencies: 330 kHz, and 393 kHz respectively.

Since a frequency variation of 63 kHz cannot be expected even under the most unfavorable conditions, it is felt that the LC-oscillator provides a sufficiently good solution for amateur radio applications.

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A Wavemeter for the Frequency Range 23.5 to 24.5 GHz

by E. Schaefer, DL 3 ER

The wavemeter for the 1.5 cm amateur band to be described in this article is based on an older, American publication (1). This wavemeter exhibits a high resolution and excellent reproducibility of the calibration. Due to its construction, the wavemeter is insensitive to temperature fluctuations. However, the construction requires some mechanical skill, and access to a lathe with threading capabilities.

This article is firstly to describe the electrical fundamentals; the mechanical details regarding construction are limited mainly to the resonance chamber with tuning pin, and critical dimensions are to be underlined. No calibration scale is described, since there are many methods of doing this. It is possible, for instance, to use a micrometer screw or another form of drive – the author uses a home-made micrometer screw. The usual, decadic division was marked on this screw with the aid of a semicircle, and the coarse division is marked on the spindle case. The mechanical stability is not perfect due to the short threaded guide (play).

TECHNICAL DESCRIPTION

Most X and K-band wavemeters described for home construction usually work according to the self-calibrating principle (2), (3). The waveguide wavelength λ_H can be calibrated from the inner dimensions of the waveguide and the (hopefully) known and present wave mode. This is then measured and recalculated to the free-space wavelength λ_0, and finally to the frequency to be measured after considering the waveguide values.

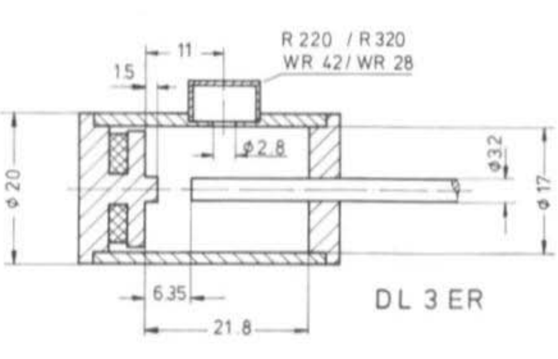

Fig. 1: Principle of operation, and critical dimensions of a wavemeter

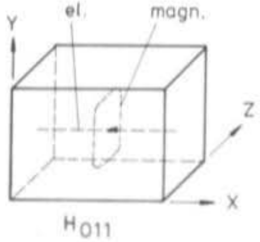

Fig. 2:
Characteristics of the electrical and magnetic lines of field of a H_{011}-mode

The disadvantage of this simple wavemeter – immaterial whether it is coaxial or waveguide – is its unfavorable tuning ratio $\Delta f/\Delta l$. This means a small variation of length Δl will cause a relatively large frequency change Δf. This results in a noticeable temperature dependence, and errors during the measuring process.

The construction to be described virtually avoids these disadvantages, since the tuning length amounts to approximately 8 mm for a frequency variation of 23.5 to 24.5 GHz.

However, an accurate frequency source must be available for calibration.

The wavemeter operates with wave mode TE_{011} (also called H_{011}) in a cylindrical cavity resonator. As can be seen in **Figure 1**, the frequency to be measured is coupled in from the waveguide via an iris. The resonance frequency is increased on inserting the tuning pin. It only causes a slight field distortion, which results in a small $\Delta f/\Delta l$. In addition to this, the Q is very high, and is in the order of approximately 5000 to 6000 in the case of the loosely coupled resonator, if all components are silver-plated.

The system is designated as a hybrid due to its partial, coaxial construction. Since radial currents are present in the terminating panels with the TE_{011} wave mode for both coaxial and cavity resonators (see **Figure 2**), it is possible for the tuning pin to be inserted through a central cutout without terminating choke or contact finger, and with a gap spacing of approximately 0.1 to 0.15 mm. Opposite to the tuning pin the cavity resonator is terminated using a choke-system, that is filled with lossy material. This is provided to avoid higher, unwanted wave modes. The protruding end of the choke opposite to the tuning pin is in the form of a small cylinder (4 mm dia., 1.5 mm long), as shown in Figure 7. This linearizes the tuning curve at the highest frequency (fully inserted).

CONSTRUCTION

The dimensions given in Figure 1 are the critical dimensions, and must be maintained as accurately as possible. They are valid for all 3 types of waveguide that can be used (4).

Figure 3 shows the construction and gives all dimensions. The individual parts are given in the following drawings.

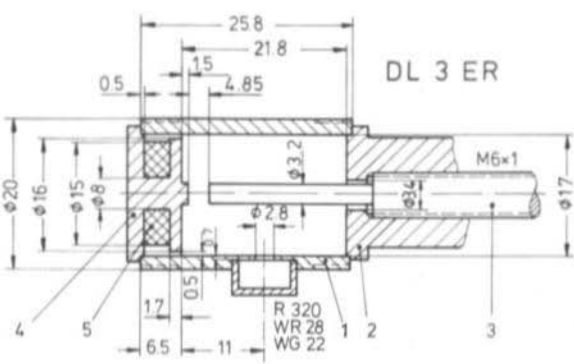

Fig. 3:
A micrometer satisfying the requirements is fitted to Part 2 and Part 3

The cavity resonator (Part 1, **Figure 4**) is made from a brass tube of 20 mm x 1.5 mm. This is cut exactly to the required length, and provided with a thread of M 20x1. This is the same threading as used for the well-known »spinner plug« 3.5/9.5. The choke holder (Part 4) and the cover with threaded bushing (Part 2) are screwed together using such nuts. The cover **(Figure 5)** and also Part 4 are provided with a crimp and are fitted without play into the resonator chamber (Part 1). A slight play can be compensated for if necessary by using a galvanic, silver-plating.

The tuning pin (Part 3, **Figure 6**) is held firmly in the very long thread in Part 2. A temperature-stable grease can be provided to ensure a smooth run, however, attention must be paid that no grease can penetrate into the resonator chamber.

A material (Part 5) is glued into the choke holder (Part 4, **Figure 7**) that exhibits a large loss at microwave frequencies. Suitable materials are pertinax, bakelite, plastic materials containing graphite, or a special microwave-absorption foil.

The width of the milled slot in Part 1 is dependent on the waveguide to be used. The author used waveguide R 320/WR 28. The position and diameter of the iris cutout are independent on the type of waveguide. A dip of approximately 40% results when using the described iris.

A wide side of the waveguide is removed; the size is dependent on the flat surface of the resonator, which means that the resonator replaces the walling of the waveguide. Waveguide and resonator are soft-soldered together at the end of construction. Residual solder and flux should be carefully removed afterwards.

The calibration scale depends on the possibilities and requirements. In the case of the author's prototype, Part 3 is provided with a home-made micrometer screw head in which a division of 0 to 90 in steps of ten, with a subdivision in »half steps of ten« has been marked. Part 2 is provided with the corresponding, linear millimeter division.

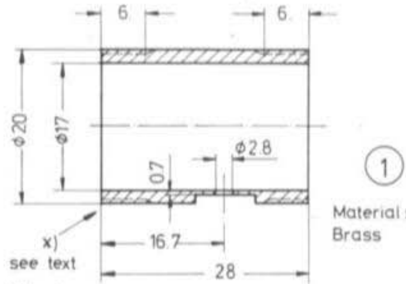

Fig. 4:
The two threads x must fit the required nuts; the milling must suit the waveguide to be used

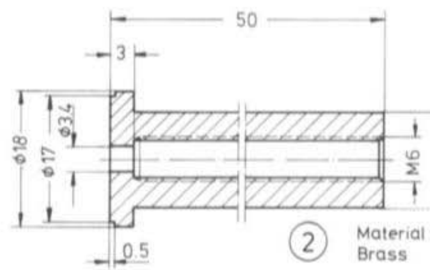

Fig. 5:
Part 3 should be held virtually without play by the long M 6-thread

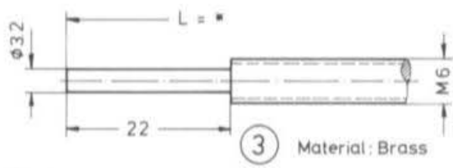

Fig. 6:
The extension to the right is dependent on the micrometer used

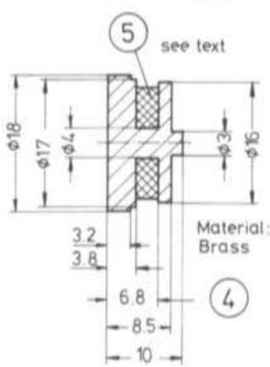

Fig. 7:
The choke attenuates unwanted wave modes using a lossy material (Part 5)

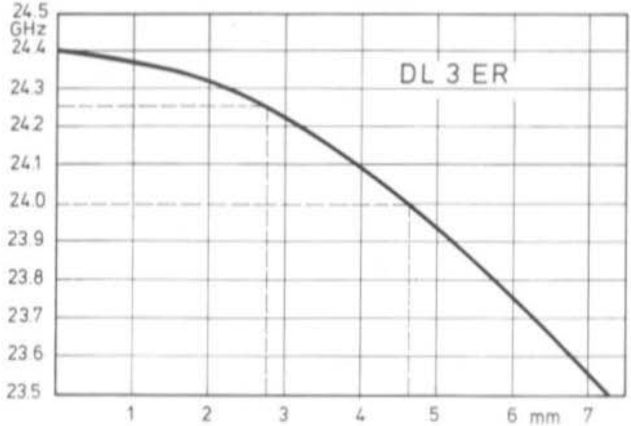

Fig. 8: Tuning range of the author's prototype wavemeter

Part 2 can also be modified for accepting an available depth-micrometer, which, of course, means that modifications must be made to Part 3.

However, the 24 GHz amateur band is in the range of 24.0 to 24.250 GHz, which is still covered in a virtually linear scale with a tuning range of approximately 2 mm.

CALIBRATION OF THE WAVEMETER

As has been previously mentioned, an accurate-calibration oscillator, or an oscillator together with another, exactly calibrated wavemeter is required for calibrating the homemade wavemeter. The resonance frequency dip is now sought with the aid of a connected, detector probe. This is repeated for the whole frequency range, and the result is a calibration curve similar to that of the author's prototype shown in **Figure 8**.

However, the author's prototype is somewhat too long due to tolerances in the resonator chamber. Furthermore, the cylinder on Part 4, which should be 1.5 mm long, is also missing. For this reason, the calibration curve is somewhat bent towards the high-frequency end.

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(3) Heubusch/Hock/Knauf: A Transceiver for 10 GHz, Part 3 VHF COMMUNICATIONS 9, Edition 4/1977, pages 247-255

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Line-of-Sight Microwave Communications

by H. Schlager, OE 3 HSC

This article requires some knowledge of trigonometry of an oblique-angled triangle so that the flat surface profile can be corrected for by the curvature. The required terms »4/3 curvature of the earth« and »Fresnel zone« are to be described.

The triangle M, H 1, H 2 is determined by the three values $r + h_1$, $r + h_2$, and α (see **Fig. 1**). The angle α results from the arc F_1F_2 (= b), and the distance (QRB) of the two base points F_1 and F_2.

$$\alpha = \frac{QRB \times 360^\circ}{2 \times \pi \times r}$$

where: r = radius of the earth, to be explained later.

The straight line H_1 and H_2 is thus calculated according to cosine law as follows:

$$A = H_1H_2 = \sqrt{(r+h_1)^2 + (r+h_2)^2 - 2(r+h_1)(r+h_2)\cos\alpha}$$

It is now possible for the sum of the side lengths (circumference) of the triangle to be calculated, and this will be called circumference = 2 s:

$$s = \frac{(r + h_1) + (r + h_2) + A}{2}$$

Now, φ_1 is determined as the next calculation, using the chord formula:

$$\cos\frac{\varphi_1}{2} = \frac{s\,[s - (r + h_1)]}{A \times (r + h_2)}$$

All elements of the triangle are now determined that are required for further calculations. A general point (base point) F can be calculated that is spaced b from base point F_1, as well as the height h of the radio beam, and other things.

From the value b' = F_1F the following will result:

$$\alpha' = \frac{b' \times \alpha}{b}$$

From the rule »the sum of angles = 180°«, the following will result:

$$\varphi = 180^\circ - (\alpha' + \varphi_2)$$

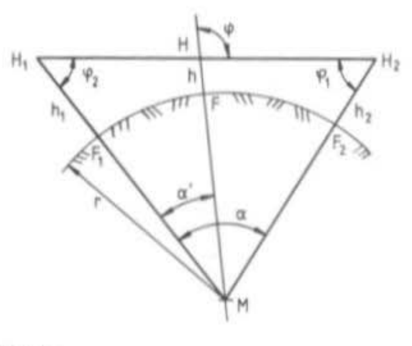

Fig. 1:
Trigonometry of the oblique-angled triangle H_1H_2 = A; F_1F_2 = b; F_1F = b'

With this and sine law it is now possible for the unknown side of the small triangle H_1MH to be calculated:

$$\frac{h + r}{\sin \varphi_2} = \frac{h_1 + r}{\sin \varphi}$$

$$h = \frac{(h_1 + r) \sin \varphi_2}{\sin \varphi} - r$$

THE TERM 4/3 THE EARTH CURVATURE

The propagation of electromagnetic waves in space and in a homogene atmosphere is a straight line.

The value of the relative dielectric constant varies with increasing height in the air layers near to the surface. One will find a gradient of the reflective index (ΔN) that is dependent on the temperature, and even more on the humidity.

The anomalities of this refractive index are responsible for the tropospheric over-the-horizon communication, which is so useful on the 2 m band.

The value of ΔN is (usually) negative, which means that it is normally more humid near the surface and decreases with height.

This gradient is found in the equation for the effective diameter of the earth (1):

$$r_{eff} = r_{geom.} \times k, \quad \text{where}$$

$$k = \frac{1}{1 + 6.37 \times N \times 10^{-3}}$$

According to CCIR, ΔN does not differ greatly from − 40, which means that k is 1.34, which corresponds to 4/3, and r_{eff} = 8488 km.

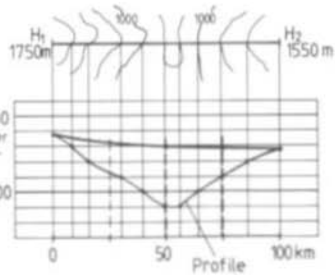

Fig. 2:
The points of intersection are now transferred from the lines of height on the map with the aid of additional lines onto the mm-paper

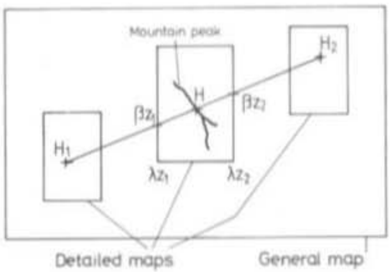

Fig. 3:
Detailed maps should be obtained for the critical points of the communication path

THE PROFILE OF A COMMUNICATION PATH

This is commenced by obtaining a map with a scale of approximately 1 : 500 000. There are some maps available which are only provided with lines of height, and lakes. A line is now drawn between the two required points (projection in the sectional plane). A page of mm-paper is now fixed approximately 1 to 2 cm from this line.

The line-of-sight orientated to the curvature of the earth is now inserted into the straight profile as a dot approximately every 25 km (**Figure 2**).

All points should be checked at which the radio beam is in the vicinity of the profile. This requires the study of more detailed maps (approx. 1 : 50 000) from which the exact position of H_1, H, and H_2 can be determined. These maps provide lines of height every 20 m.

It is best to provide a great circle through H_1 and H_2 and to study the exact path of the radio beam on the more detailed maps. It is thus possible for the actual height above ground to be determined.

GREAT CIRCLE

The equation for determining the width β_z of a point on the great circle from H_1 (λ_1, β_1) to H_2 (λ_2, β_2) with the point of intersection on the meridian λ_z is:

$$\beta_z =$$

$$\text{arc tg} \frac{\text{tg}\, \beta_2 \times \sin(\lambda_z - \lambda_1) - \text{tg}\, \beta_1 \times \sin(\lambda_z - \lambda_2)}{\sin(\lambda_2 - \lambda_1)}$$

An example is given in **Figure 3**.

The height above ground for H can be read off from the map.

FRESNEL ZONES

Fading of the signal due to multipath propagation is well-known in mobile communication. Similar effects are also observed when a signal from a fixed station at location H_1 is not only received on the direct path to location H_2, but also, partially, via a reflection. This will cause a complete cancellation of the signal when the signals are of equal amplitude and opposite phase.

This means that it is not only necessary for direct line-of-sight conditions to be present between transmitter and receiver, but also to ensure that there are no reflections present with a path difference of $\lambda/2$ (or odd multiples of this).

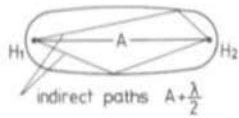

Fig. 4:
The Fresnel zone is an ellipse around the radio beam from H_1 to H_2

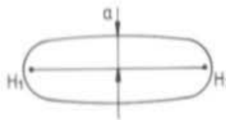

Fig. 5:
The critical spacing a is calculated according to equation 2

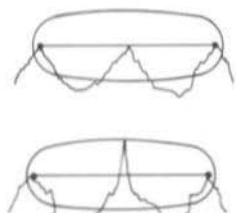

Fig. 6:
Anything protruding into the Fresnel zone will cause additional losses

The geometrical location (the assumed joining) of all these points is an ellipse around the radio beam, whose focal points are H_1 and H_2. This zone is called Fresnel zone (see **Fig. 4**).

The spacing »a« of the ellipse around the radio beam can be approximately given for any point of the path by the following (see **Figure 5**):

$$a = 548 \times \frac{b' \times (b - b')}{[b' + (b - b')] \times f}$$

with a in m, b in km, f in MHz.

Anything protruding into this ellipse will increase the path loss by 6 dB on approaching the first Fresnel zone (see **Figure 6a**) up to the radio beam, and by 16 dB when cutting the radio beam **(Figure 6b)**.

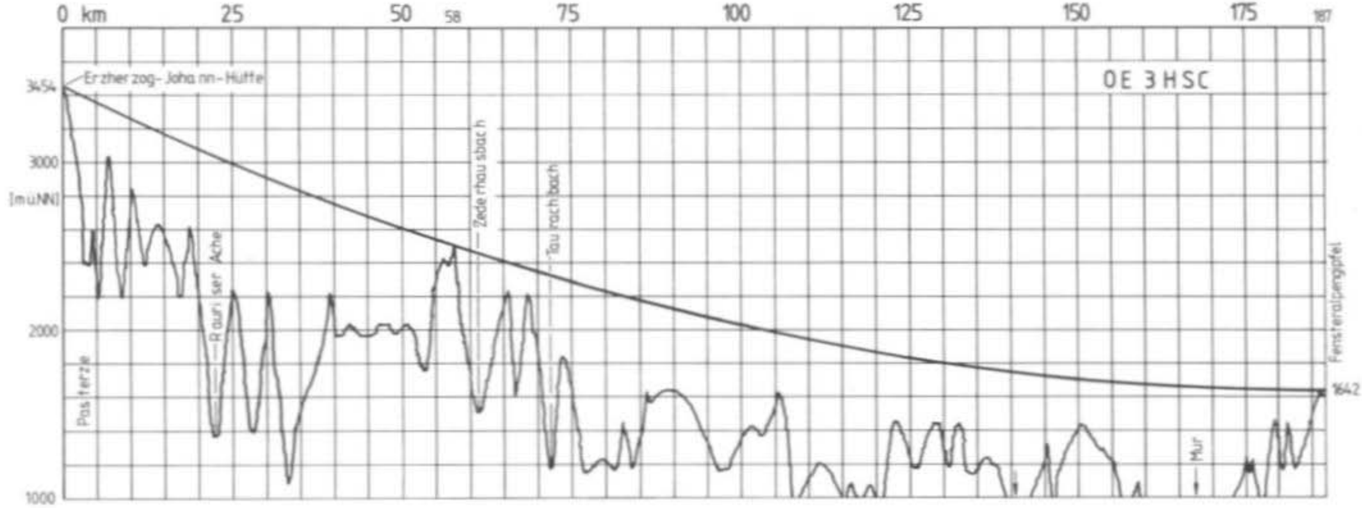

Fig. 7: Profile of a radio path from the Grossglockner to Fensteralpen-Gipfel in Austria

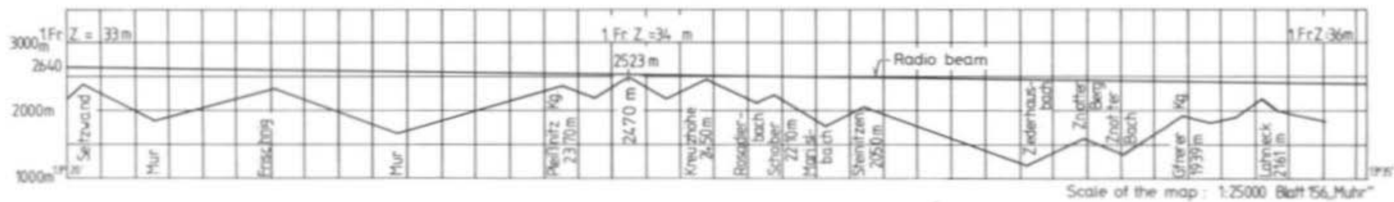

Fig. 8: At 58 km, the height between the radio beam and the mountain amounts to 53 m

These values are additional losses that must be taken into consideration on calculating the path loss along the communication path.

FINAL NOTES

The given calculations can be made easiest using a programmable calculator. Programmed cards for the HP 41 c can be provided at cost. Finally, the author would like to show this calculation in the form of a practical example: Profile and Fresnel zone consideration of a communication path in Austria for 10 GHz communication over 187 km: This is from the Erzherzog-Johann-Hütte on the Grossglockner to Fensteralpengipfel in OE 6. Both points are readily accessible using a cable railway, or road.

One will see in **Figure 7** that one part of the path is in the direct vicinity of the radio beam between 13°20' and 13°35' East.

With the aid of the equation given in the section »Great Circle«, the points of intersection of the great circle (determined by H_1 and H_2) were determined for the meridians 13°20' and 13°35'. These were 47°07'55" and 47°09'15". These coordinates have been marked in the detailed map **(Figure 8)** and joined with a pencil line. This allows the fine profile of this section of the path to be drawn, and indicates that the first Fresnel zone is not affected. The height of the radio beam amounts to 2523 m, and the height of the mountain is 2470 m according to the map. This means that there is a spacing of 53 m between the radio beam and the mountain peak.

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A Versatile IF-Module Suitable for 2 m Receivers, or as an IF-Module for the SHF Bands

by F. Krug, DJ 3 RV

Although modern receivers are able to satisfy the average need on the shortwave, 2 m, and 70 cm bands, they do still not satisfy a number of demands such as large-signal handling capabilities.

Virtually no ready-to-operate equipment is available for the amateur bands in excess of 1 GHz. This means that there is still room for experiments and home construction. Most of the constructional articles for the microwave bands, however, only describe converters that require a shortwave or 2 m receiver as IF-module, see (1) - (4). Unfortunately, many of the amateur receivers are not suitable for use for such applications. For instance, they are usually not screened well enough to avoid breakthrough of 144 MHz or 28 MHz signals.

A receiver or IF-module is required that can be used for all operating modes, and which is able to satisfy the high demands on intermodulation, rejection of signals at IF-level, and rejection of spurious reception points. Remember, no converter is better than the subsequent receiver or IF-module. A suitable IF-module can be home-constructed, and the following series is to describe such a high-quality IF-module.

A receiver suitable for use as an IF-module should be able to demodulate all operating modes encountered in amateur radio, should have low-noise characteristics, and exhibit a high overload and interference rejection. The following demands should be fulfilled:

a) Demodulation of all operating modes used for amateur radio applications
b) The bandwidth should be variable to suit all operating modes
c) The gain control should be suited for each operating mode
d) Good suppression ofinterference pulses and spurious signals
e) Variable sensitivity
f) Low noise
g) Freedom of spurious reception points
h) Good overload characteristics with respect to signals within and outside of the passband.

Of course, one will not be able to afford a receiver that satisfies every demand optimumly. The concept is designed for normal amateur radio communication, but can be extended as required by adding accessories. Inputs and outputs are provided in the module for this.

Demands a) to c) determine the extent of the IF-module, as well as the number of demodulators and filters, whereas demands d) to h) determine the characteristics of the input circuit of the receiver. In the case of the latter demands, a number of circuits have been described,and the input circuit can be realized with an input module that has already been described in this magazine. The subsequent IF-module is now to be described in detail.

1. RECEIVER CONCEPT

As previously mentioned, amateur radio receivers are usually used as IF-modules in conjunction with a microwave converter. In order to ensure sufficient image rejection, usually 144 MHz receivers are used. This has the advantage that a crystal-controlled oscillator can be provided in the receive converter. The required local-oscillator signals required for conversion are generated in a crystal oscillator with subsequent multiplier such as described by DK 1 AG in (5).

The disadvantage of this is in the direct breakthrough of strong 2 m signals. The receiver shown in **Figure 1** is suitable for such IF-applications. A further alternative is to use a subsequent receiver using a fixed frequency of, for instance, 80 MHz as shown in **Fig. 2**. Inexpensive, high-quality crystal filters are now available which allow the advantages of this concept to be used to the full.

The only difference between both concepts is in the input stage since they both use the same IF-module.

This article is now to describe the concept shown in Figure 1, and the second concept will be described as extension to this later.

1.1. Characteristics of the Receiver

Let us examine the required receiver characteristics in the order of the previously mentioned demands.

a) Operating Modes

Both amplitude and frequency modulation is permissible for amateur radio communications. The required bandwidth for the various operating modes is summarized in **Table 1**.

Operating Mode		Bandwidth
Television	A 5 C	4 MHz
Voice	F 3	15 kHz
Voice	A 3	5 kHz
FSK	F 2/F 4	5 kHz
FSK	A 2/A 4	5 kHz
FSK	F 1	2.5 kHz
Voice (SSB)	A 3 J	2.5 kHz
Telegraphy	A 1	0.5 kHz

Table 1: Bandwidth Requirements

If television is deleted due to the large bandwidth, the following modulation modes must be demodulated in the receiver: A 1, A 2, A 3, A 3 J, and F 3.

Special demodulators, and signal processing are required for F 1, F 2, F 4, and A 4 signals. This means it is better to provide a special IF-output for them.

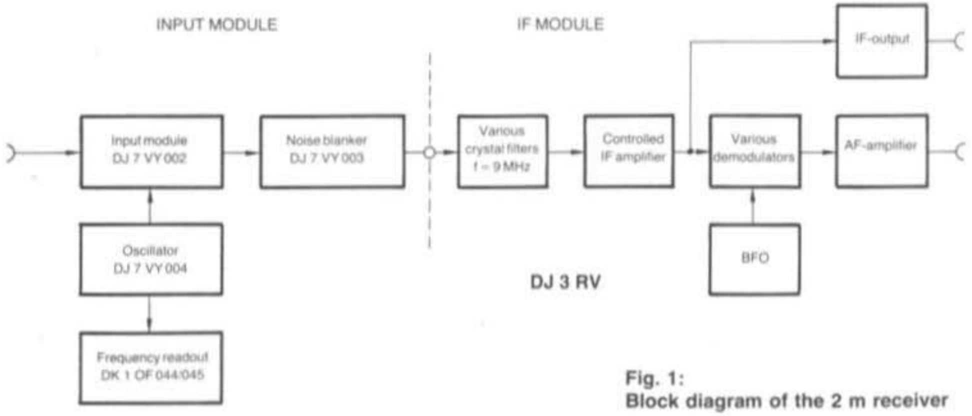

Fig. 1:
Block diagram of the 2 m receiver

b) Bandwidth

As can be seen in Table 1, the required filter bandwidths are 15 kHz, 5 kHz, 2.5 kHz, and 0.5 kHz. The selectivity characteristics are determined by the quality of the filters.

c) Gain Control

The gain control of the IF-amplifier must have a variable time-constant, as well as operating as limiter with a high AM-suppression in the FM-mode. A variable squelch for the AF-amplifier is also required.

d) Interference Suppression

A good interference suppression requires an effective noise blanker before the IF-filters, as well as a Notch-filter in the IF-amplifier. A further interference suppression at AF-level after demodulation is not advisable.

e) and f) Selectivity, and Low Noise

The IF-module should provide usable signals at an input voltage in excess of 0.1 µV, and possess a dynamic range of 100 dB. The gain of the input module should only be great enough to compensate for the noise figure of the IF-module. In order to ensure that the IF-module can also be used as a receiver, it must be sufficiently sensitive and have an input impedance of 50 Ω. The input impedance should be sufficient so that an input signal of 0.5 µV will provide a satisfactory signal after demodulation. An input dynamic range of approximately 100 dB can be obtained technically, and will mean that the maximum input voltage for the IF-module will be in the order of 50 mV. Greater signal levels should be attenuated using a variable attenuator in front of the input circuit. This ensures that the IF-module can be used as complete receiver when the gain of the input circuit is in the order of 15 dB.

g) Rejection of Spurious Reception Points

This is achieved by use of suitable circuit technology for suppressing the image frequency, and selection of the correct frequency plan. The image frequency suppression should be greater than the gain control range of the IF-amplifier so that no image frequency signals can interfere under full-gain conditions of the IF-amplifier. The frequency plan is determined by the selected intermediate frequency, which was 9 MHz for our application. Monolythic crystal filters are available inexpensively for the various bandwidths.

In the case of the first concept, the image frequency suppression is determined by the selectivity of the input circuit using a band-pass filter. In the case of the second concept, it will be determined by the ultimate selectivity of the 80 MHz crystal filter. One will notice the disadvantages of the first concept: The image frequency suppression should be in the order of 100 dB if the dynamic range of the IF-amplifier is also 100 dB. However, this cannot be achieved with simple means with an input module that is tuneable over a range of 2 MHz. When using a fixed frequency in the second concept, such values can be achieved using relatively narrow-band filters.

The image rejection of the converter must be taken into consideration when operating the receiver as IF-module. The high input frequency of 144/146 MHz (or 80 MHz) guarantees that a good suppression of the image can be made in the converter.

The rejection of spurious reception points is not only dependent on the circuit, but also on the construction and screening of the modules. The mechanical construction of such a module is considerably important for operation, and good workmanship is required.

h) Overload Rejection Characteristics

A low-noise input circuit having good overload characteristics requires a large, linear drive range. Suitable solutions were described by DJ 7 VY in (6).

1.2. Description of the Block Diagram

The block diagram of the receiver shown in Figure 1 is mainly equipped with modules previously described in VHF COMMUNICATIONS! The input circuit DJ 7 VY 002 was described in (6). It is used without crystal-controlled oscillator and noise-blanking gate. The local oscillator signal is provided by the variable oscillator DJ 7 VY 004 described in (7). The frequency readout is made with the frequency counter DK 1 OF 044/045 described in (8). The noise blanker module DJ 7 VY 003 described in (9) is directly connected to the mixer output of the input circuit. It is not absolutely necessary for the operation of the unit, and can be provided afterwards, if required. The output of the mixer and the input of the IF-module both have an impedance of 50 Ω and can be directly connected. It is necessary for all modules with RF-inputs and outputs to have a real impedance of 50 Ω, so that interconnections can be made with conventional coaxial connectors and cables. This has the advantage that no unwanted coupling can take place between the individual modules by radiating lines, and that the length of the cables will not have an adverse effect on the operation of the circuits. A further advantage is the ease of checking the individual modules, since most RF-measuring equipment possesses an input impedance of 50 Ω.

Due to the defined input impedance, the subsequent IF-module requires a matching amplifier to match the input circuit to the crystal filter. The switchable crystal filters determine the frequency response. The author plans to use monolythic crystal filters with bandwidths of 15 kHz, 5 kHz, and 2.4 kHz, as well as a discretely constructed CW-filter with a bandwidth of 500 Hz. However, other filters having a suitable size can be used if other bandwidths are required.

The subsequent, controlled IF-amplifier is equipped with dual-gate MOSFETs and is also provided with a switchable Notch filter. A separate demodulator is provided in the IF-amplifier for generation of the control voltage, which then also drives the S-meter. This ensures that a mean level is always present at the output of the amplifier independent of the modulation mode. This IF-signal feeds the demodulators and a buffer amplifier for outputting the signal. The output signal can then be further processed with the aid of accessories.

Demodulators are planned for FM, AM, and SSB/CW. The FM-demodulator is fed via a limiter amplifier, and is equipped with a crystal discriminator.

A synchronous demodulator is used for AM. This exhibits a better interference rejection than when using an envelope demodulator. The demodulation of SSB/CW signals is made in a push-pull mixer. A BFO supplies the required heterodyne signal. It is equipped with two sideband crystals, and a CW-crystal.

The subsequent AF-amplifier provides outputs for headphones and a loudspeaker.

1.3. Extension Possibilities

The construction of the module is made so that a number of extensions can be made.

Further demodulators can be connected to the IF-output, such as an FM-demodulator for FSK-signals having a lower shift, and this will be described later. It is also possible to connect a monitor for checking modulation. The outputs of the BFO- and local oscillator signal can be used to extend the receiver to form a transceiver. A suitable module for this is being developed by the author at present.

A further extension of the receiver is to use an input circuit as shown in **Figure 2**. This extends the 9 MHz IF-module by a 80 MHz input circuit. As can be seen in the block diagram, a variable attenuator is provided at the input to decrease too high an input level. This is followed by a low-noise, amplifier with good overload characteristics, which is in turn followed by an 80 MHz crystal filter with a bandwidth of 15 kHz. The signal is then fed via a noise blanker to the mixer where it is converted to 9 MHz with the aid of a local-oscillator signal of 71 MHz.

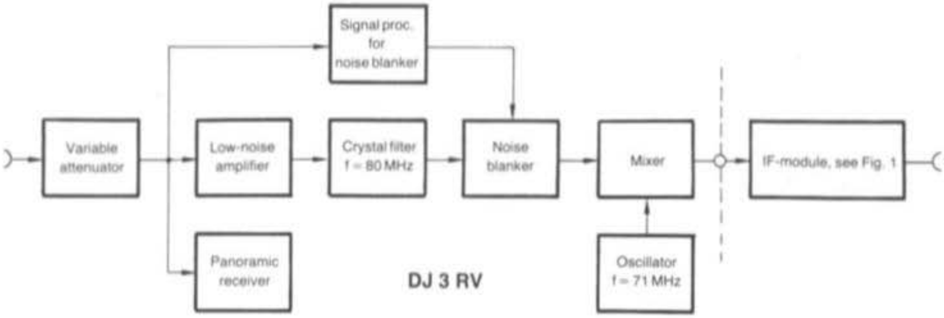

Fig. 2: Block diagram of the 80 MHz input module

An output signal for the noise blanker, or for a panoramic receiver is tapped off after the attenuator. The signal processing for these two modules is made at 10.7 MHz and 455 kHz. This avoids any possible interference to the 9 MHz circuit. The image frequencies of the 9 MHz or 10.7 MHz conversion are within TV-band I, and are suppressed with the aid of bandpass filters.

This results in a versatile IF-amplifier. It can form the basis of a microwave receiver, or a variable shortwave receiver from 100 kHz to 30 MHz.

2. MODULES OF THE INPUT CIRCUIT

The original modules of the input circuit of the receiver shown in Figure 1, were slightly modified. However, the data and values given in the original descriptions (6) - (9) are still valid, which means that only the modifications must be discussed.

2.1. Input Module DJ 7 VY 002

The input module described in (6) is constructed without the planned noise blanker and without the crystal-controlled oscillator and doubler. The output coupling is made directly at the mixer with the aid of a 50 Ω coaxial connector. The input coupling of the local oscillator signal is made at the input of filter F 4. The modified circuit originally given in (6) is given in **Figure 3**. A gain of 14 dB was measured on the author's prototype at the center of the band.

2.2. Oscillator DJ 7 VY 004

Since the receiver is to be driven from a power supply that provides a stabilized voltage of 15 V, the oscillator described in (7) is constructed without the voltage stabilizer I 7. The interconnection from Pt 1 to RFC 7 is made with the aid of a wire bridge.

2.3. Frequency Readout DK 1 OF 044/045

The frequency counter described in (8) is used as described by DJ 7 VY in (7). In the case of the author's prototype, it was found that the interference from the multiplex signal was quite strong, which means that it is necessary to well screen the counter board and readout.

The reference crystal oscillator is used as originally described; an improved oscillator with better temperature characteristics is under development and will be described later.

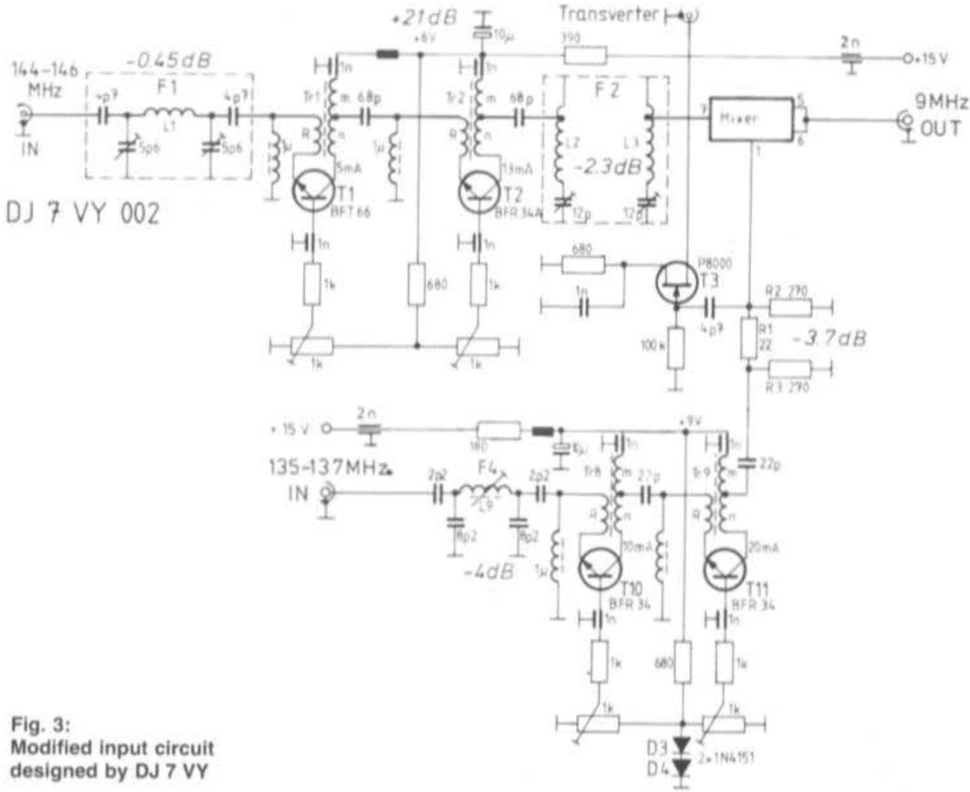

Fig. 3: Modified input circuit designed by DJ 7 VY

2.4. Noise Blanker DJ 7 VY 003

The noise blanker was used as described in Figure 10 of (9) without modifications. The gain in the required passband was 5 dB in the author's prototype, and − 0.4 dB with the noise blanker switched off.

The following editions of VHF COMMUNICATIONS are to bring the following sections:

3. IF-AMPLIFIER

3. 1. Matching stage for the crystal filter
3. 2. Crystal filter with switching
3. 3. Controlled IF-amplifier with S-meter
3. 4. Notch-filter
3. 5. Limiter-Amplifier
3. 6. FM-demodulator
3. 7. AM-demodulator
3. 8. SSB/CW-demodulator
3. 9. BFO
3.10. AF-amplifier
3.11. IF-output coupling

4. CONSTRUCTION DETAILS

4. 1. Input circuit
4. 2. IF-module
4. 3. Power supply

5. ALIGNMENT INSTRUCTIONS

6. MEASURED VALUES

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A Converter for the 13 cm Band Equipped with Two Preamplifier Stages and an Active Mixer
VHF COMMUNICATIONS 8, Edition 4/1976, pages 194-201

(2) R. Heidemann, DC 3 QS:
Receive Mixer for the 6 cm Band
VHF COMMUNICATIONS 12, Edition 1/1980, pages 46-50

(3) G.Börs, DB 1 PM, H.Fleckner, DC 8 UG:
SSB on the 10 GHz Band
Part 1: Generation of the Local Oscillator Frequency
VHF COMMUNICATIONS 12, Edition 3/1980, pages 130-138
SSB on the 10 GHz Band
Part 2: Waveguide Modules
VHF COMMUNICATIONS 13, Edition 1/1981, pages 2-12
SSB on the 10 GHz Band
Part 3: Intermediate Frequencies in the 2 m or 70 cm Band
VHF COMMUNICATIONS 13, Edition 1/1981, pages 13-17

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A Simple Converter for Reception of Weather Satellites in Conjunction with 2 m FM-Receivers
VHF COMMUNICATIONS 12, Edition 4/1980, pages 211-214

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An Extremely Low-Noise 96 MHz Crystal Oscillator for UHF/SHF Applications
VHF COMMUNICATIONS 13, Edition 3/1981, pages 135-143

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

(7) M. Martin, DJ 7 VY:
Low-Noise VHF Oscillator with Diode Tuning, Digital Frequency Control, and Frequency Indicator
VHF COMMUNICATIONS 13, Edition 2/1981, pages 66-82

(8) J. Kestler, DK 1 OF:
A Settable Up-Down Frequency Counter
VHF COMMUNICATIONS 13, Edition 2/1981, pages 83-94

(9) M. Martin, DJ 7 VY:
A Noise Blanker for Large-Signal Conditions Suitable for Shortwave and VHF-Receivers Having a Large Dynamic Range – Part 2
VHF COMMUNICATIONS 12, Edition 2/1980, pages 96-106

The Output Power of a Transceiver can be Reduced Automatically on Switching on the Power Amplifier

by J.M. Noeding, LA 8 AK

It is usually necessary to reduce the output power when a power amplifier is to be connected to a transceiver or transmitter, in order to ensure that it is not overdriven. For instance, a power amplifier equipped with a QQE 06/40 only requires a drive power of 1 to 3 W for an output power of 100 W. Many transmitters or transceivers provide an output power of 10 to 20 W. The described circuit is able to reduce the output power automatically when a PA is used. In this case, the exciter (transceiver) can only provide its full output power when the PA is switched off.

Suitable circuits for this mode of operation have already been described in (1), however, they do not operate automatically, but require a short-circuit plug when the PA is not connected. The advantage of the circuit given in **Figure 1** is that the power is only reduced when the transceiver (transmitter) is connected to the power amplifier, and when the modifications shown in Figure 1 have been made. Without PA, the transceiver will operate at full output power.

Only three resistors, a capacitor, and a diode must be installed into the transceiver (to the

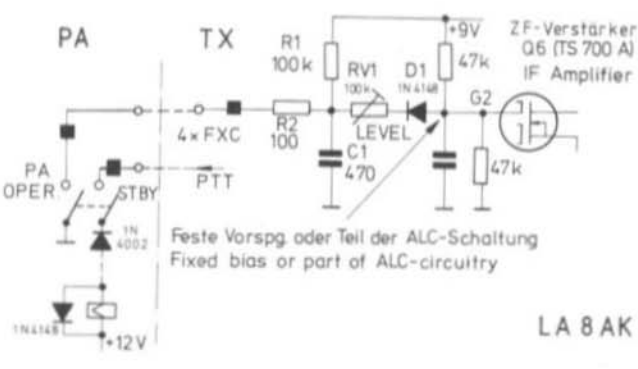

Fig. 1: Only a few components and slight modifications are required in the PA (left) and transceiver (right) for automatic reduction of the output power level

right of the dashed line), whereas it is only necessary for the single-pole switch to be exchanged for a double-pole one in the power amplifier (to the left of the dashed line). This is the »OPERATE« switch of the PA; when this switch is closed, the PA can be driven, and the described circuit will reduce the drive power automatically to a predetermined value.

The required output power of the transceiver is selected with the aid of RV 1. This is made by starting at the lowest value and increasing slowly until the power amplifier is just able to give its maximum output power. As soon as the PA is switched off with the aid of the left-hand contact, or when the interconnection is broken, diode D 1 will isolate the additional circuit, and the amplifier stage in the IF-module of the transmitter will once again receive its original G 2-voltage, and thus provide full gain.

The circuit diagram shows that the PA is switched on by the PTT-line of the transceiver. This line can, however, be switched to ground using any relay contact. It is not important for the operation of the power-reduction circuit.

It is sometimes a problem to bypass the DC-lines at VHF and UHF – especially at high power levels. For this reason, it is advisable for ferrite beads to be provided on all connection leads between transceiver and PA in order to ensure that the sensitive, low-signal stages of the transceiver are protected against high RF-voltages.

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Reducing the Output Power of Transistorized SSB Transmitters and Transverters
VHF COMMUNICATIONS 9, Edition 1/1977, page 37

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 4/1981 of VHF COMMUNICATIONS

DJ 1 JZ 003	**METEOSAT CONVERTER:** **Interdigital Mixer and IF-Preamplifier**		**Ed. 4/1981**
PC-board	DJ1JZ 003	Double-coated, undrilled	DM 25.—
Minikit	DJ1JZ 003	1 Schottky mixer diode hp 2817, 1 DG-MOSFET, 1 coilformer, 1 m silver-plated wire, 3 plastic foil trimmers, 2 chokes, 6 resistors, 9 cer. caps. Mechanical parts: side pieces, cover, resonators, screws (silver-plated and protected), as well as all parts for C 1	DM 164.—
Kit	**DJ1JZ 003**	**complete with above parts**	**DM 189.—**

DC 7 EP 001	**Inexpensive TV Pattern Generator**		**Ed. 4/1981**
PC-board	DC7EP 001	Single-coated, with plan	DM 14.—
Semiconductors	DC7EP 001	1 IC MM 5322 N, 1 transistor	DM 126.—
Minikit	DC7EP 001	1 res.circuit 455 kHz, 2 Al. electrolytics, 2 Styroflex, 2 cer., 5 feedthrough capacitors, 1 trimmer pot., 7 resistors, 1 metal case, 13 solder pins, 1 IC socket	DM 23.—
Kit	**DC7EP 001**	**complete with above parts**	**DM 159.—**

1.1 m Parabolic Antenna Kit	**Ed. 3/1979**
12 Segments for screwing or riveting together. Complete with center plate and mounting hardwave for feed	DM 180.—
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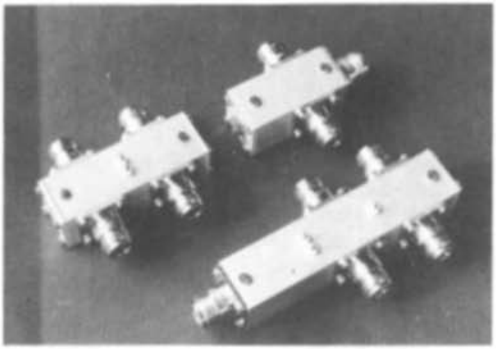

Fig. 1

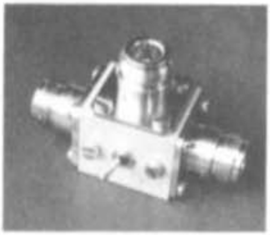

Fig. 2

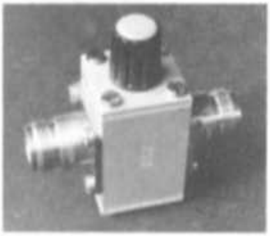

Fig. 3

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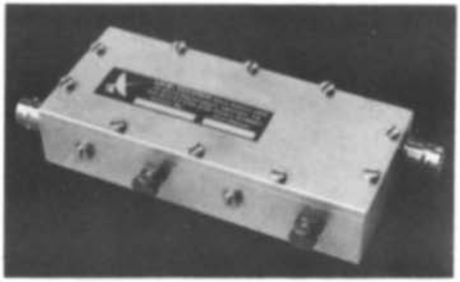

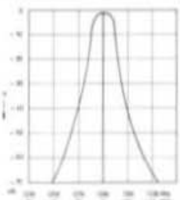

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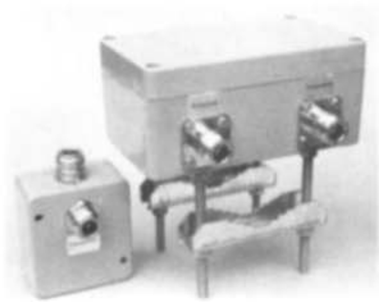

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