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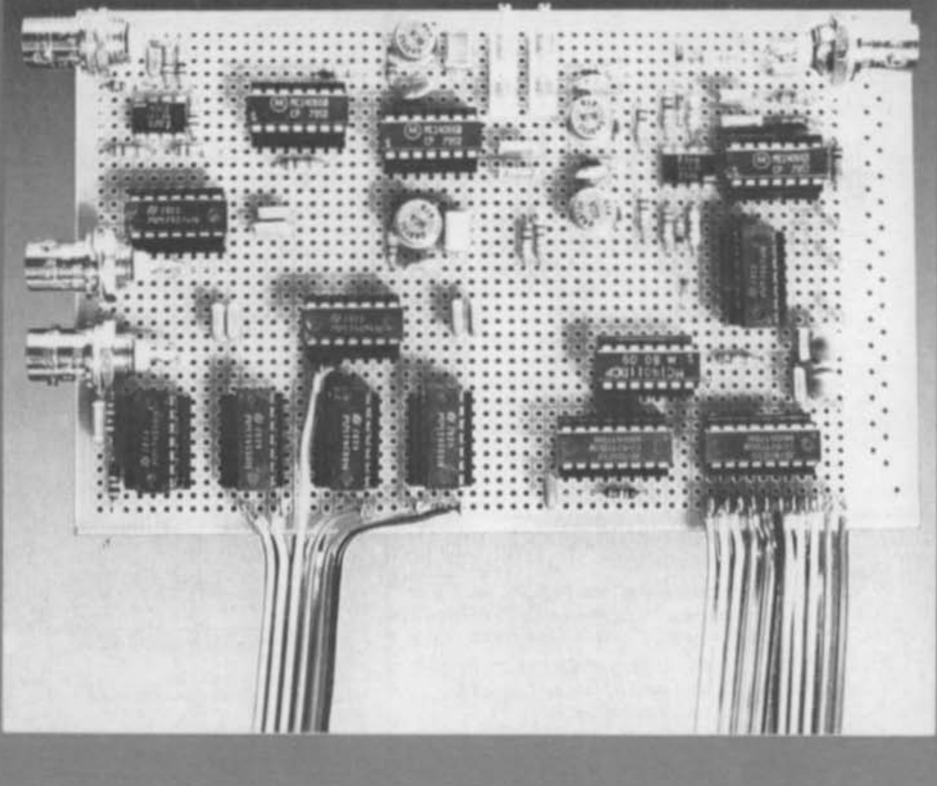
A Publication
for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

# VHF communications

Volume No.14 · Summer · 2/1982 · DM 6.00

W6 NEY: CCW-Filter



# VHF communications

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# Contents



# In the Focus

In this edition you will find the description of a low-noise preamplifier for the 2 m band using the S 3030. This transistor is not just another low-noise device but, together with the 3 SK 97, the first type of a new family of semiconductors:

Dual-Gate Gallium-Arsenide Field-Effect Transistors (DG GaAs-FETs), also called MESFET from **me**tal **s**emiconductor **f**ield-**e**ffect **t**ransistor, have a metal gate, separated from the bulk of the semiconductor by a Schottky barrier. This semiconductor family highlights the increasing use of the »3-5 technology« which implies the use of semiconductor materials made from one or more of the group 3 elements of the Periodic Table (boron B, aluminium Al, gallium Ga, indium In) with one or more of the group 5 elements (nitrogen N, phosphorus P, arsenic As, antimony Sb).

Whereas microwave GaAs-FETs exhibit the lowest possible noise figure, they are fragile and expensive (ceramic case !); the new DG-MESFETs come in a plastic case which leads to a lower price and a lower frequency limit. The manufacturer of the S 3030 (TI) gives data at 1 GHz (NF typ. 1.7 dB, gain typ. 20 dB).

In our next edition we will publish a preamp for the 70 cm band with the S 3030. Who wants to try it at 23 cm, and for the METEOSAT frequencies ?

With kind 73s

your's

Robert E. Lentz

DL 3 WR

*Charles Woodson, W 6 NEY*

# Coherent Telegraphy Transmissions

# Part 2: Practical Aspects

**This two-part article is published with the kind permission of the ARRL, and is based on an article in the May and June editions of QST. Part 1 of this article described the concept of coherent telegraphy transmissions and Part 2 is to describe how this can be made in practice.**

Coherent CW operation imposes two basic requirements at the transmitting end. First, the keying must be done within the time frames established by a stable frame reference. These frames must be sufficiently regular to enable the receiving station to determine accurately when they occur. Second, the carrier frequency must be stable within a hertz or so during the contact, including all keying periods. The time frames can be stablished by a frequency standard with the reference signal being divided by CMOS or TTL to produce pulses which define the frame. Many CCW stations use standards such as those described by Kelley, although any comparable standard would do.

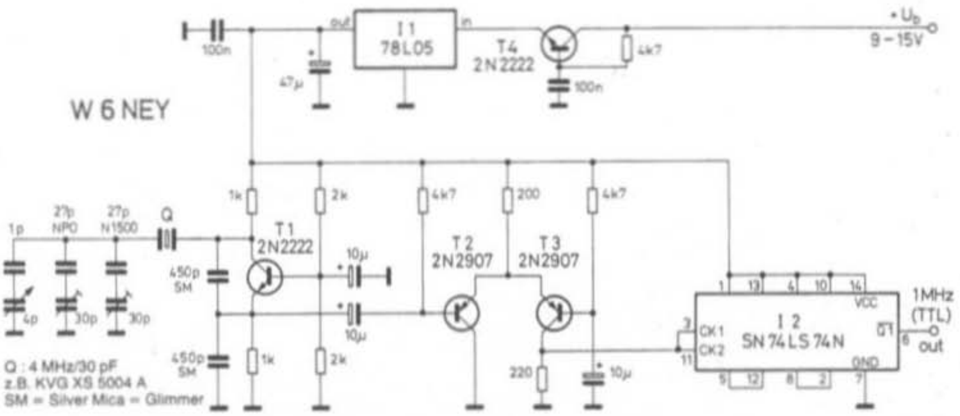


**Fig. 7: A 1-MHz frequency standard for CCW station use**

To keep the frames accuracte within 1/20 of a period for 10 »windows« per second requires a stability factor of 1/720.000 Hz per hour of contact. Since the standard mentioned is accurate and stable to less than 1 part in $10^7$ over the required period, it exceeds the required accuracy easily. A station standard suitable for supplying the 10-Hz keying reference and the CCW filter frame reference is shown in **Fig. 7**.

## KEYING

**Figure 8** shows a simple system that may be used for CCW keying. I have adapted both the Heath HD-10 (5) and the Accu-Keyer (6) for CCW operation. The Accu-Keyer is superior because of its 1-bit memory. At present, I use an AKB-1 keyboard, which is available with a

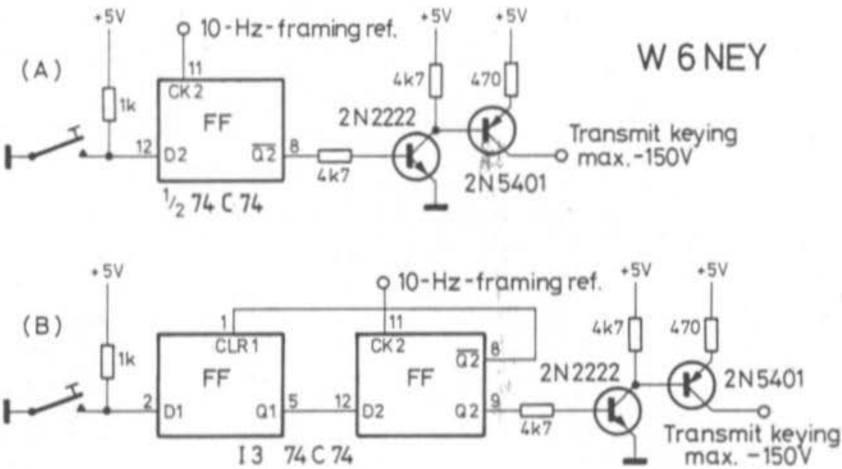


**Fig. 8: Two simple CCW hand keyers**

The key must be kept closed until the beginning of a frame, determined by the framing reference signal. Once transmissions is initiated, although the key may be opened, the transmitter will remain keyed until the end of the frame period is signaled by the framing reference. The keyer at B includes a one-bit memory which makes coherent keying by hand much more convenient. When the key is closed, the first flip-flop is set and remains set even if the key is opened. When the clock pulse arrives, the second flip-flop is set, which causes the first to be cleared unless the key is still closed. If your rig uses inverted TTL keying, use the $\bar{Q}$ output.

CCW option. I've also used a KIM-1 computer for generation of CCW and ASCII. The computer uses its internal timing clock to generate an interrupt at the beginning of each frame period. The clock frequency must be adjusted precisely for such use.

Hand sending of CCW is different from ordinary random-frame CW and takes a while to learn. This is because dots, dashes and spaces can only occur in pre-established frames and we are accustomed to initiating dots, dashes and spaces whenever we wish. With a bit of practice, the initial sending errors decrease to near that of the error rate of ordinary CW keying. You learn to hold the key down until you hear a dot or dash start and then you are able to send in rhythm with the frames for a word or phrase. A keying monitor is a must!

## TRANSMITTER STABILITY

The receiving filter passband requires that the transmitted frequency be stable during the contact period. This is perhaps the most difficult parameter to be met for CCW operation. For a CW signal time frame of 0.1 second, a 14-MHz signal must be stable to within 1 or 2 Hz. High-quality crystal oscillators have such stability except when a varying load is placed upon them, as when a transmitter is keyed. During keying, the frequency of a typical transmitter crystal oscillator will shift approximately 50 Hz. Under ordinary circumstances this wouldn't be noticed, but for a CCW signal, this would mean loss of reception because the shift is more than five times the receiving filter passband and would equate to a 20-kHz shift of a regular CW signal. Such shifting produces an amusing situation. When copying with the CCW filter in the presence of strong interference, the interfering signals sometimes appear to swish up and down the band during keying. Even if they cross the CCW frequency, the time they are in the filter passband is small. The result is that they have relatively little effect on the CCW signal itself. However, these interfering signals – through cross-modulation, overloading early receiver stages, and their effect on AGC – can (and often do) cause problems.

Transmitter stability has been achieved by using high-quality crystal oscillators which are not keyed and which are followed by several stages of amplifiers and buffers to nullify the loading effects of keying. A schematic diagram of such a transmitter-exciter is shown in **Figure 9**. The power output of this exciter is about 0.1 W and it has been used by itself (with an antenna matching network and keyer in the final stage) and as a VFO replacement. Tests have shown that after a 30-minute warm-up period the oscillator is stable within a hertz during keying and remains so for over an hour. The crystal tuning allows VFO-type operation over a 20-Hz range. To facilitate stability, very little power is drawn from the oscillator and two stages of isolation are used to minimize the load on the oscillator by later stages. In most situations, particularly when the rig is left on all the time, the N1500 compensation capacitor and corresponding trimmer may be omitted and a fixed capacitance value added in parallel with the rest of the units. When the temperature compensation trimmers are used, they are adjusted while measuring the operating frequency at two different temperatures, say, 68 and 86° F (20 and 30° C). One trimmer is adjusted to decrease capacitance and the other to increase capacitance by a like amount. The frequency is measured at the two temperature extremes again and this process continued until the oscillator frequency is the same at both temperatures.

Another method of transmitter frequency stabilization is to use PLLs to control the frequency of oscillators and use a highly stable oscillator as a reference for the PLL. A direct-conversion receiver employing this technique was described by McCaskey (7). Maynard used a 5.0- to 5.5-MHz synthesizer output and a 9-MHz-frequency standard to control an HW-8 (8). I have used a method which mixes the HFO, BFO and VFO frequencies of a double-conversion transceiver (SB-303/SB-401 combination), locking the result by controlling the VFO frequency (9). A simple scheme (shown in **Figure 10**) is used for locking the VFO

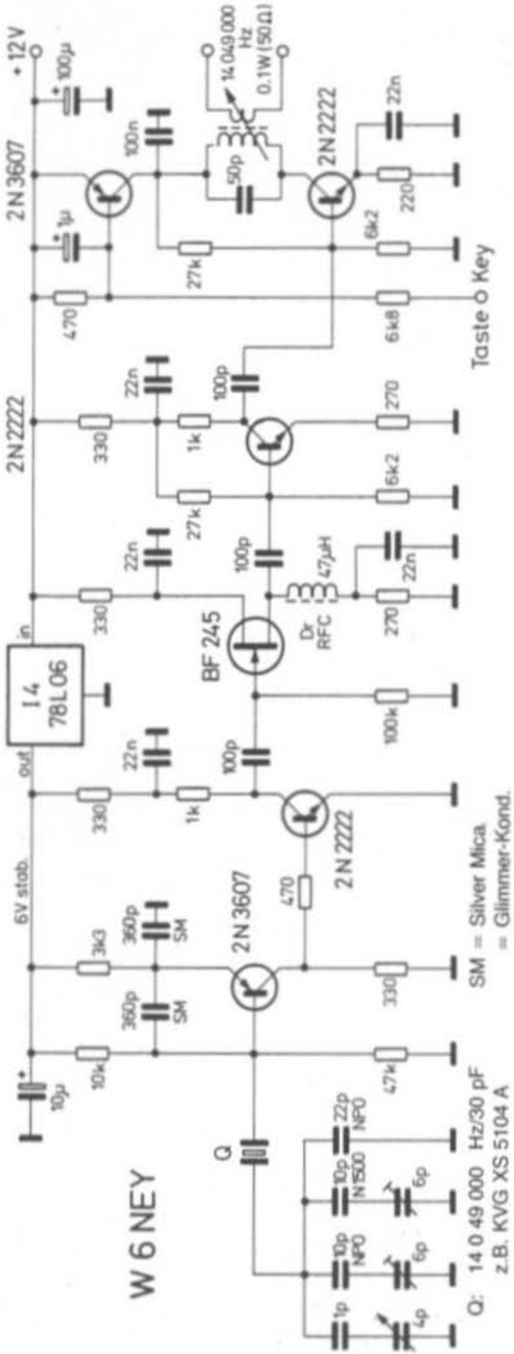


Fig. 9:
**Schematic diagram of a low-power CCW exciter/transmitter. The capacitance from the crystal to ground should total about 30 pF.**
**Fixed capacitors should be high quality ceramic or silver mica types.**
**C 1 is used to adjust the frequency of operation and affects the total capacitance by less than a picofarad once temperature compensation has been achieved with the other trimmers (see text).**
**Output inductance approx. 2.5 µH, with 2 turn over cold end to the output.**

(LMO) of an SB-303 receiver by using the built-in variable capacitive diode circuit employed for FSK operation. A high-impedance voltmeter connected to point C can be used to monitor the lock condition. During operation, the VFO is tuned slowly across the frequency of the standard; frequency lock occurs about 250 Hz above and below the reference frequency. Once locked, the crystal oscillator controls the receiver frequency and it can be set more accurately than the VFO. The crystal oscillator can be replaced by a 5.0- to 5.5-MHz synthesizer which is controlled by a suitable reference frequency; Petit has designed such a synthesizer which operates in 100-Hz steps (10).

A block diagram of the transmitter currently in use at my station is shown in **Figure 11**. The 12.9-MHz crystal oscillator is designed for high stability. Similar oscillators are used for operation on 21 and 28 MHz. The synthesizer is controlled by a 1-MHz oscillator similar to that described in Fig.7. The two oscillators run continuously and are connected to the doubly balanced mixer, but the 14-MHz stage following the mixer is keyed. This allows break-in operation on the same frequency.

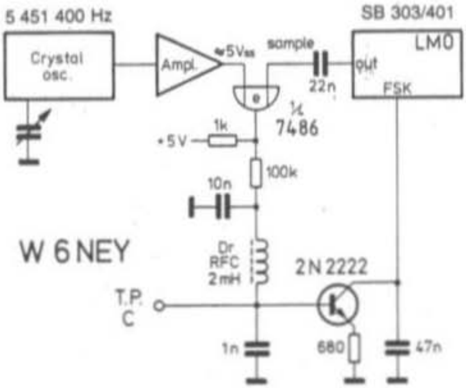


**Fig. 10:**
**This method may be used to lock the LMO of the popular Heath SB series of equipment. Point C is a test point which is used to monitor the lock condition.**

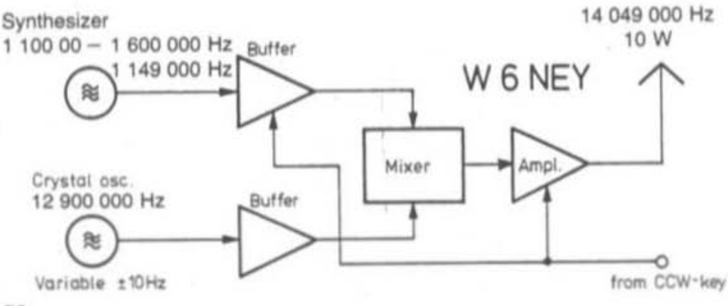


**Fig. 11:**
**Block diagram of the transmitter used by the author. The two oscillators run continuously for improved stability**

## RECEIVER REQUIREMENTS

In addition to the CCW filter, the receiver must exhibit stability on the order of 1 Hz over the length of a contact and have a tuning resetability which is less than the bandwidth of the filter. Searching for a signal while using a filter bandwidth of only 10 Hz requires almost 200 times as long as it takes to tune a band using a filter with a bandwidth of 2.1 kHz. If the phase and frame size were also unknown, it would take over 1000 times as long to tune a band searching for a CCW signal as it takes to look for an ordinary CW signal. That is why current practice involves agreeing on a precise frequency and frame length in advance. Adequate stability is easy to obtain with good crystal oscillators in receivers when temperature has been stabilized by a long warm-up period and a stable environment exists.

**Figure 12** is a block diagram of the receiver currently in use at my station. Rough tuning is done by adjusting the HF crystal oscillator and the BFO, which have ranges of about 800 Hz, to the desired frequency. The VFO of the CCW filter center frequency reference (four times the center frequency) is used for fine tuning over a range of about 25 Hz. An IF strip similar to one designed by Hayward (11) provides performance superior to others I have used. Best results are obtained when the AGC is controlled by the AGC output of the CCW filter.

Keitaro Sekine, JA 1 BLV, uses a crystal-controlled FT-901 and also has built a 2980- to 3080-Hz RC VFO for use as the reference for the center frequency of the CCW filter. Oscillators in the transceiver have been stabilized by using temperature compensation methods and high-stability crystals.

## THE FILTER

A practical coherent digital filter may be seen in **Figures 13, 14, and 15**. The first CD4066 A6 is used as a switching mixer while the second controls the sample and dump functions. An audio signal output may be derived from a digital mixer (such as shown in Fig.14) driven

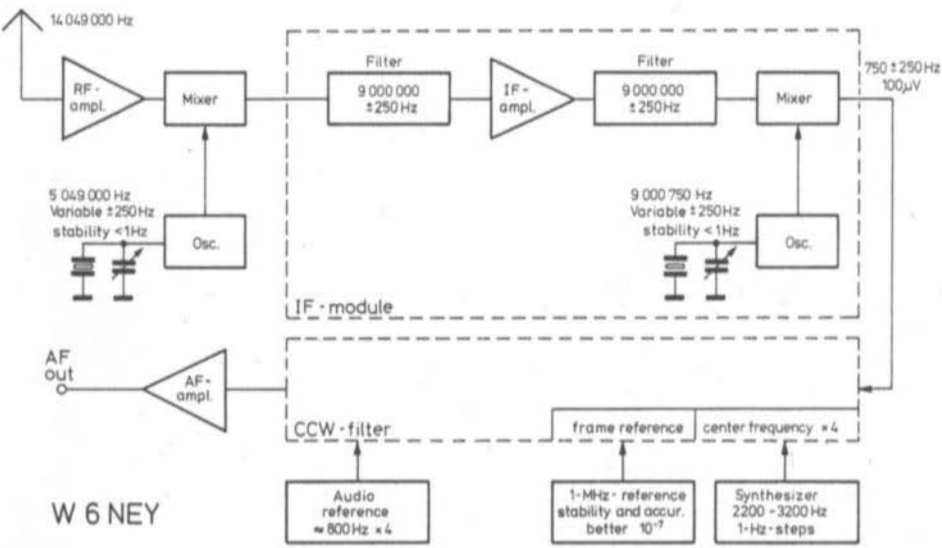


**Fig. 12: A block diagram of the receiver used by the author**

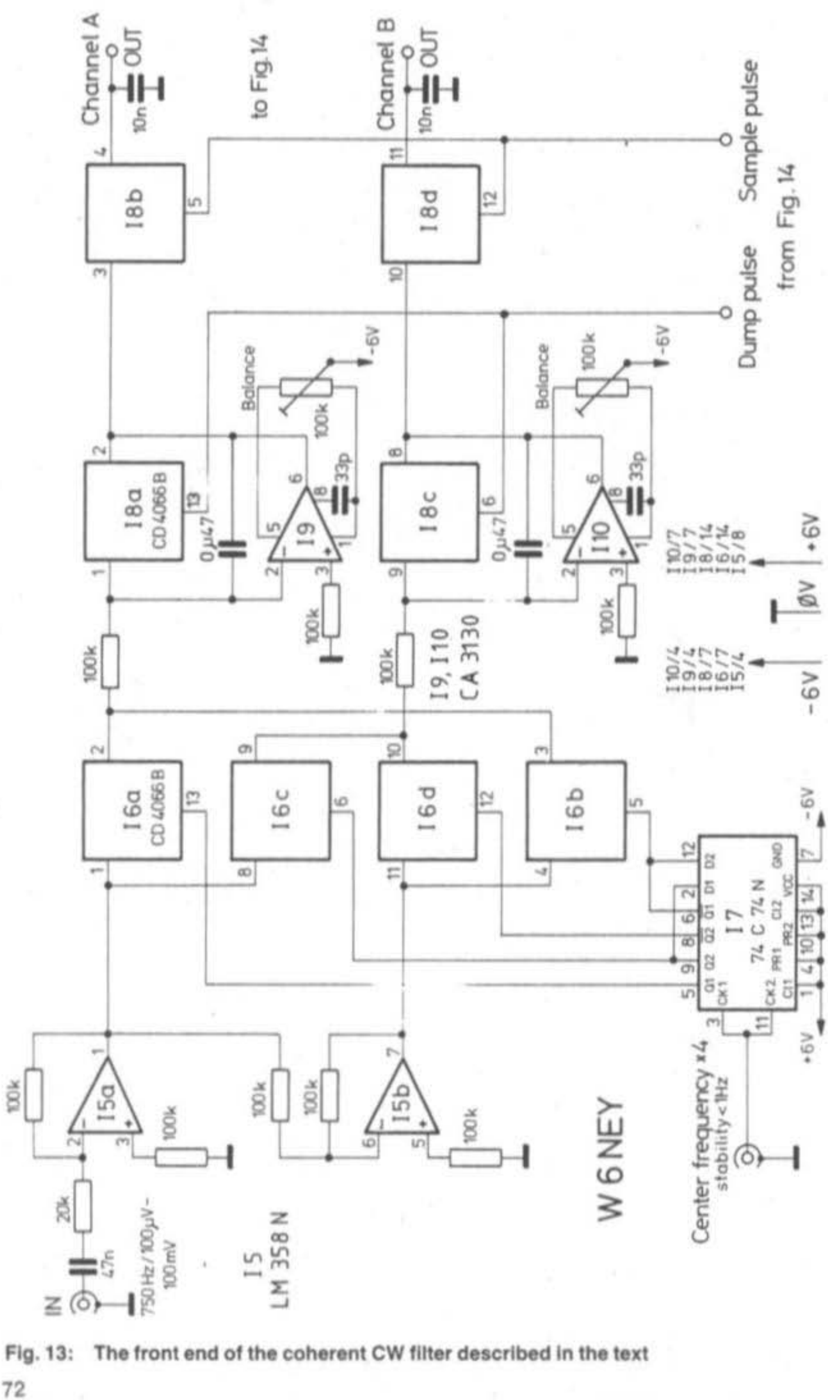


Fig. 13: The front end of the coherent CW filter described in the text

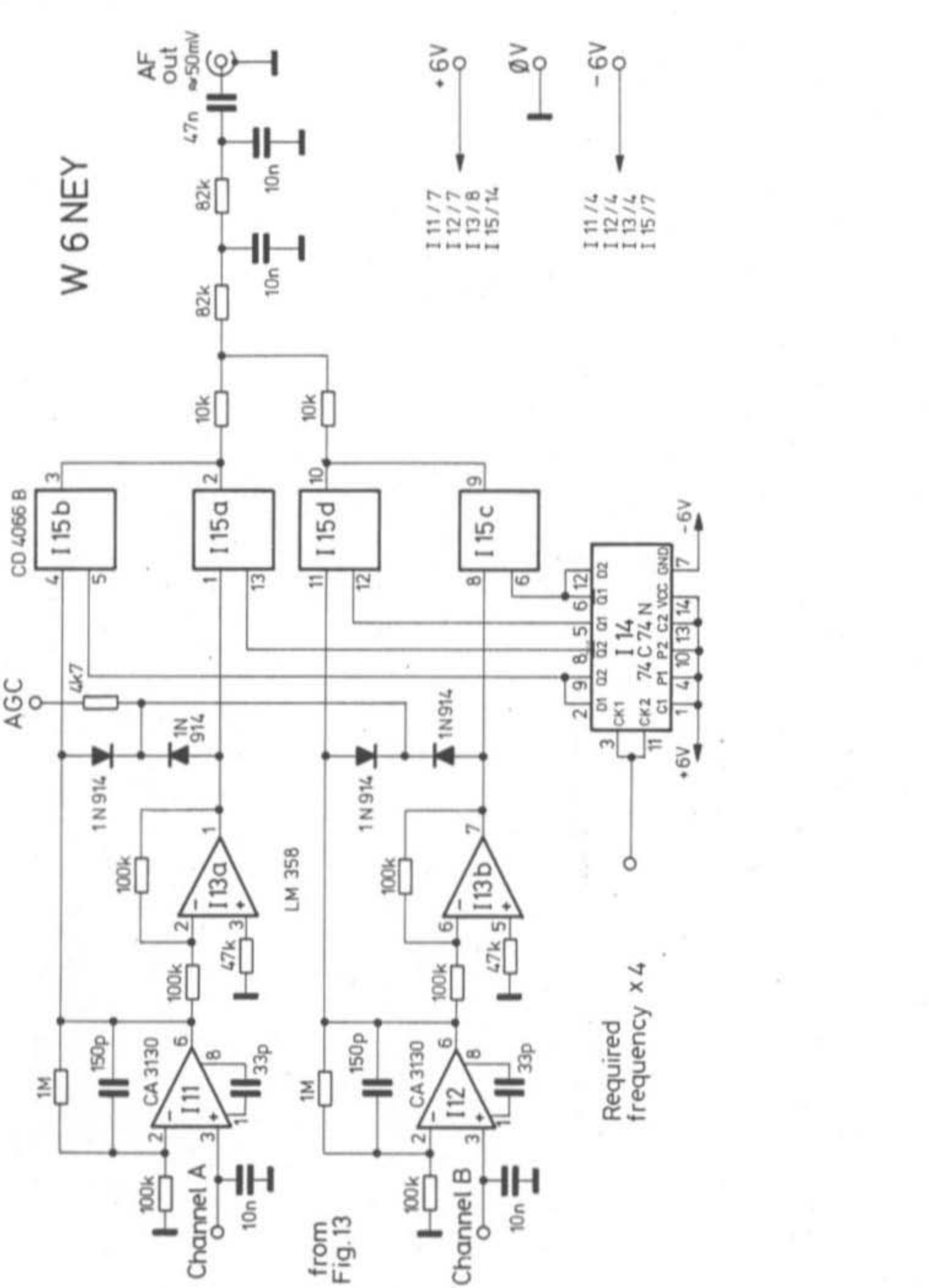


Fig. 14: This portion of the CCW filter employs digital mixing to generate an audio output

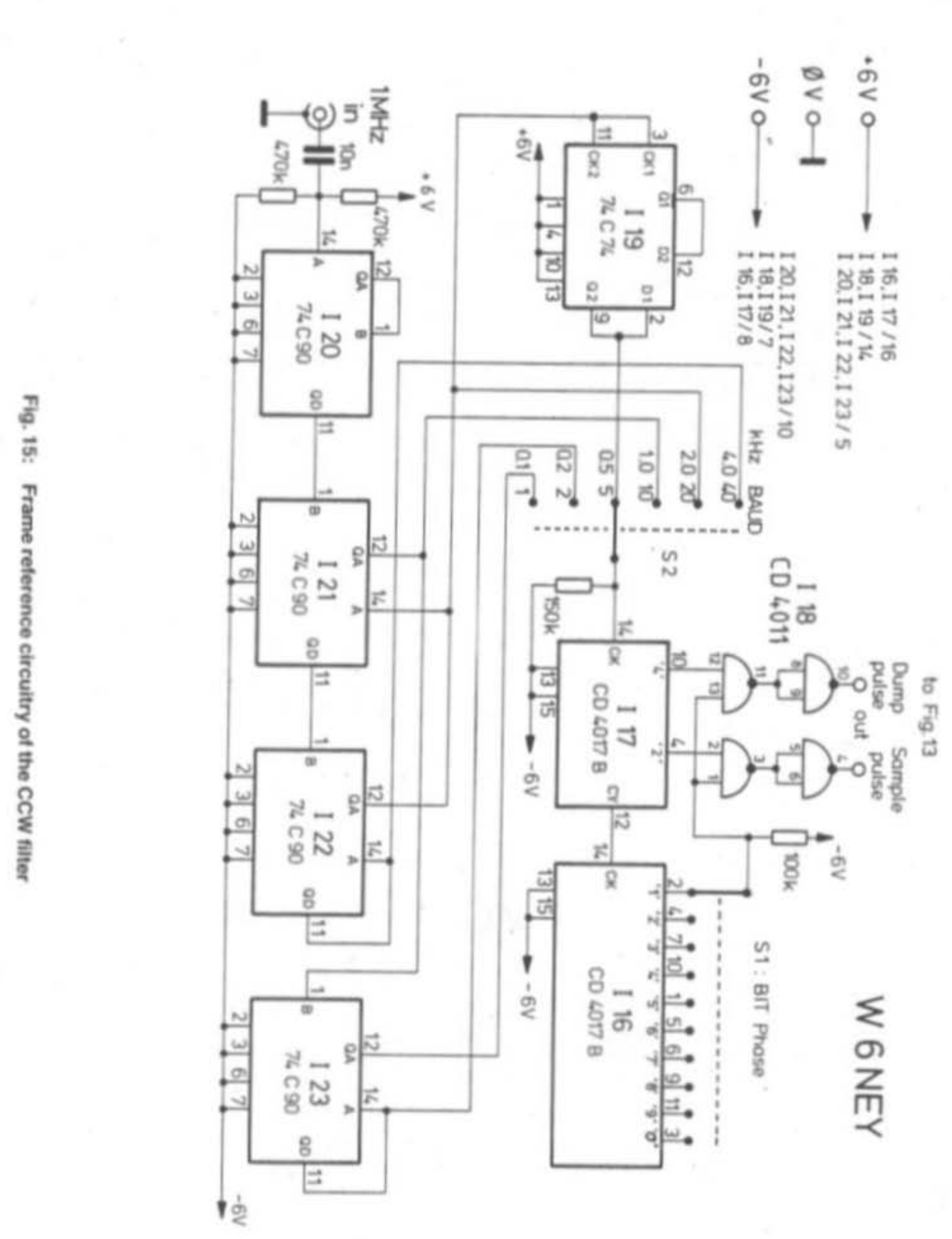


Fig. 15: Frame reference circuitry of the CCW filter

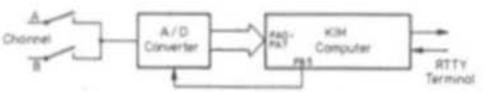


**Fig. 16:**
**An experimental computerized CCW filter system used by the author. The filter used is similar to that of Figs. 13, 14 and 15, except that all switches are computer operated.**

by the output from the two channels. The signal is the difference between the two and can be made single-ended by using an op amp, or both channels may be fed to A/D converters for computer input. A frame reference for the CCW filter is shown in Figure 15.

## A MICROPROCESSOR-CONTROLLED FILTER

The logic diagram of **Figure 16** is that of the computerized system which has been used at my station. The switching mixers are essentially the same as those used in the filter described previously. A computer program controls the A/D conversion and dump functions. Computer control of the mixer has been employed, but use of an operator-controlled VFO is a convenience. The 1-MHz internal clock is stabilized and used to define the CCW frames. Phase is adjusted by the computer program. This is done by operator command. The operator indicates an advancement or retardation of the framing phase in 10 ms increments by pressing a computer key. I have experimented with a computer program to adjust framing phase automatically, but have not yet found a satisfactory way to maintain framing phase lock during breaks in the QSO caused by QRM or pauses. Between control of the sample and dump functions, the computer also converts the received Morse signals to ASCII code and transfers the ASCII code to a CRT character display terminal or printer.

## WEAK SIGNALS AND NOISE

The reception of weak CCW signals is quite different from that of ordinary weak CW signals. Under standard conditions, as the CW signal gets weaker, QRN or QRM remain as »no signal» output and we eventually end up with a noise level dependent upon the bandwidth of the filter. With CCW, noise is a series of »dots« in frames and varies randomly in intensity. With the CCW filter, output is limited by design to one frequency and a weak signal is characterized by missing and extra dots randomly mixed with the desired signal.

Frame phase adjustment is important because if it is not accurate, a blurring of the dots and dashes into adjacent frames occurs. This makes the signal unreadable and it might go unnoticed if it is weak. When receiving a series of dots (a standard part of a CCW CQ), you can tune for maximum contrast between dots and spaces. With a strong signal, even a 10 % phase error can be noticed. A slight lead error causes a weak mark just *before* each dot or dash while a lag error results in a weak mark just *after* the dot or dash.

## OPERATING PRACTICES

Under favorable conditions, it is often convenient to operate the CCW filter at shorter than optimal frame periods. With 0.01-second frames, the bandwidth is around 100 Hz and phase adjustment makes little difference. Although selectivity is reduced and signal level decreased by 10 dB, this method is used during initial signal detection. Once a signal is located, phase adjustment and longer frame periods may be used to optimize reception.

Phase tuning may be used instead of tuning a band of frequencies. This is accomplished by using an agreed-on frame length and frequency of operation and tuning for proper phase by adjusting the filter phase. Once phase adjustment is close, the frequency may be fine tuned as well. Present practice calls for sending a 15-second stream of dots to help in frame acquisition. A steady carrier of 10 seconds duration is an aid when fine tuning to frequency.

Time-reference signals from station such as WWV may also be used to adjust the keying and reference frames of CCW receiving filters. Such adjustment must take into account the electromagnetic distance of the standards station to the receiving station as well as the electromagnetic distance between communicating stations. This procedure allows phase to be fixed and the operating to be the primary parameter which must be considered. Communication between stations located in Japan and California has been successfully accomplished using this technique. It is, however, a more difficult procedure to follow than phase tuning.

## CONCLUSIONS

CCW offers the possibility of employing some interesting operating techniques. Suppose Amateur Radio stations of the world agreed to operate at frequency multiples of 10 Hz. This would provide 20.000 channels at the bottom 200 kHz of a band. If operators further agreed on sending in frames synchronized to 0.1-second UTC time pulses, you could set the framing (about a 0.03 second delay) to correspond to the distance of the stations you wish to contact, say 6200 mi (10.000 km). Once this is set, a check of the channels may be made for a station at the desired distance. Generally, you could detect signals at distances of 5000 to 7500 mi (8000 to 12000 km) without further adjustment. Imagine microprocessor control over the entire procedure and the automatic detection of stations a particular distance away !

Coherent CW is a useful technique which improves communications effectiveness in excess of 20 dB. This factor can be used to offset poor propagation conditions, small or poorly located antennas, or low-power operation. It has the potential to be as revolutionary to CW as SSB has been to phone communication.

## ACKNOWLEDGMENT

Many of the ideas presented here are based on ideas of Ray Petit, W 7 GHM. It is impossible to discuss CCW without mentioning him. This article has benefited considerably from critical comments and suggestions from Jim Maynard, K 7 KK, and Ray.

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*Editors*

# Using the Dual-Gate GaAs-FET S 3030 in a Low-Noise Preamplifier for 144 MHz

**This new dual-gate gallium-arsenide FET-transistor has been manufactured by Texas Instruments in an X-pack plastic case. These transistors combine the advantages of the low-noise GaAs semiconductor material with the versatile, and easy to use circuitry of dual-gate FETs. Another advantage is that they are inexpensive.**

**The transistor was introduced in this magazine as the S 3000, which was specified up to 800 MHz. This transistor will no doubt soon make its appearance in TV-tuners. A professional version specified with guaranteed data to 1 GHz is the S 3030 which is to be used in this article to construct a low-noise preamplifier for the 144 MHz band. Two such amplifiers were constructed, and the measured noise figures were 0.9 dB and 0.5 dB respectively.**

## 1. CIRCUIT DESCRIPTION

At first sight, the circuit given in **Figure 1** could be a preamplifier equipped with a BF 900 or BF 981. However, the symbol shows that it is somewhat different, being a type of junction FET with two gates. The required negative gate voltage for gate 1 is obtained as the voltage drop across the source resistor R 3: Approximately 1.8 V across 180 Ω, which means that a drain current of 10 mA is flowing. Gate 2 is provided with a positive bias voltage, which has little effect on the noise figure. The value of R 1 can be changed if the source voltage should vary more than 0.3 V from its nominal value. For safety reasons, the operating voltage is limited to 10 V using a zener diode. Since a higher operating voltage does not decrease the noise figure, it is not worthwhile endangering the transistor by overvoltage (spikes), just to save the cost of the zener diode. If required, the operating voltage can be fed to the preamplifier via the coaxial cable.

With respect to the RF-circuitry it should be mentioned that it is only possible to achieve the low-noise figure when using high-Q resonant circuits, short straight lines to the transistor, and providing a good UHF-bypass of source and gate 2. As can be seen in the construction instructions, the amplifier is a hybrid between PC-board and chamber construction.

## 2. COMPONENTS

| | |
|---|---|
| T 1: | S 3030 (TI) |
| D 1: | 10 V zener diode |
| L 1: | 7 turns of 1 mm dia. silver-plated copper wire wound on a 10 mm dia. former. Self-supporting, and spaced at least 5 mm from the PC-board. Soldered directly to trimmer C 1 and ground. Coil tap direct at the connector at approx. 1.5 turns. |

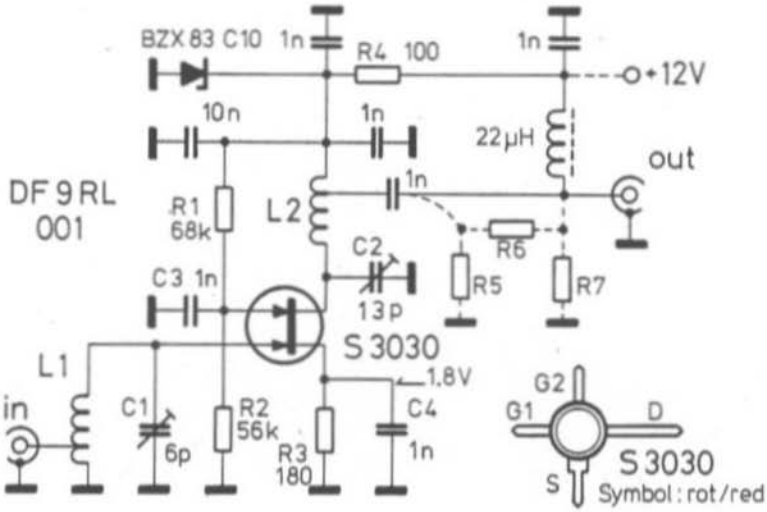


**Fig. 1: This 144 MHz preamplifier provides a noise figure down to 0.5 dB. It can be fed remotely or direct, and can be provided with an attenuator to reduce excessive gain**

L 2: 5 turns of 1 mm dia. silver-plated copper wire wound on a 10 mm former. Coil length approx. 12 mm. Self-supporting, spaced approx. 5 mm from the PC-board. Coil tap approx. 1.5 turns from the cold end.

C 1: Air-spaced trimmer with two connections: Value 6 pF

C 2: Air-spaced trimmer with two connections: Value 13 pF

C 3,C 4: 1 nF ceramic disc capacitors (Chip), value uncritical.

The other three capacitors are ceramic disc types for a spacing of 5 or 2.5 mm.

Resistors: composite carbon, rating 0.33 W

Case: Tinned metal case of 74 x 74 x 30 mm

Connectors: Preferably N-Conn., at least BNC.

## 3. CONSTRUCTION

Please read this part of the article before commencing construction, since we do not wish you to copy the negative experiences we have had during the design of this preamplifier. It is necessary for this transistor to be handled with the same care as is valid for other MOS components and GaAs-FET's. The greatest danger is from high-voltage static charges, which can occur in conjunction with synthetic flooring, clothing, and other materials. Another danger is an excessive heating during the soldering process. For this reason, the transistor is soldered into the completed circuit as the last component, and soldered into place quickly using a minimum of solder.

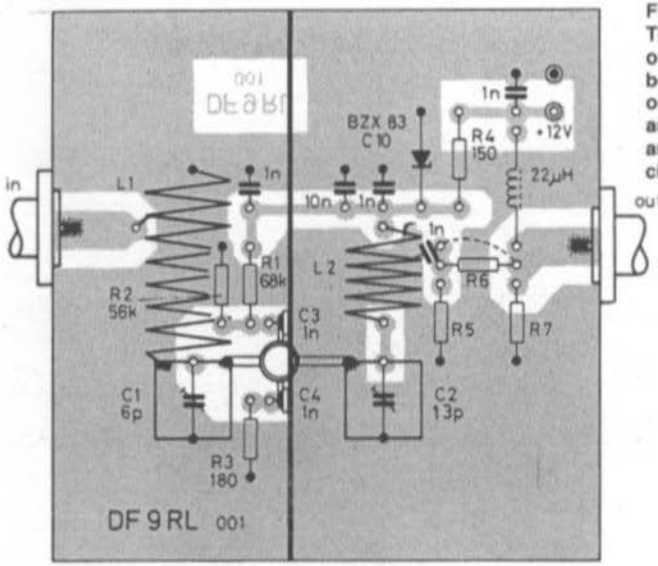


**Fig. 2: The GaAs-FET is mounted on the screening panel between the input and output circuits; source and gate 2 connections are made to chip-capacitors on the panel**

Battery or low-voltage soldering irons should be used without danger. If such types are not available, it is advisable to disconnect the soldering iron from the power supply before soldering the transistor into place.

It is good practice to connect the ground of the PC-board to the ground line of the AC-outlet, and to ground oneself, tweezers, soldering iron etc. before touching the FET.

The double-coated PC-board shown in **Fig. 2** is designated DF 9 RL 001. The dimensions are 72 mm by 72 mm, which means that it can be enclosed in a matching metal case. A screening panel should be provided between the two inductances; this is also constructed from 0.5 mm tin plate. This panel is soldered to the side panels and along the PC-board. Before installing the panel into place, a small hole should be provided for the drain connection (2.5 mm dia.) approximately 11 mm high on the panel, and between both trimmer connections. The two chip capacitors for source and gate 2 are now soldered to the panel so that the transistor can be soldered into place with direct, short connections. It should be possible for the drain and gate 1 connections to be made directly to the air-spaced trimmers without bending. A wire connection is made from the chip capacitors to the PC-board.

All ground connections are made so that the component leads pass through the board, after which they are soldered on the upper **and** lower side of the board. The flanges of the coaxial connectors should also be soldered into place to ensure a good ground connection.

## 4. ALIGNMENT AND MEASURED VALUES

Firstly check the voltage drop across the source resistor R 3. If the measured value differs by more than 0.3 V from the given value (approx. 1.8 V), resistor R 1 should be exchanged: If the voltage is too low, the value of R 1 should be reduced, and vice versa.

It is now possible to connect the preamplifier between antenna and receiver. Align the pre-

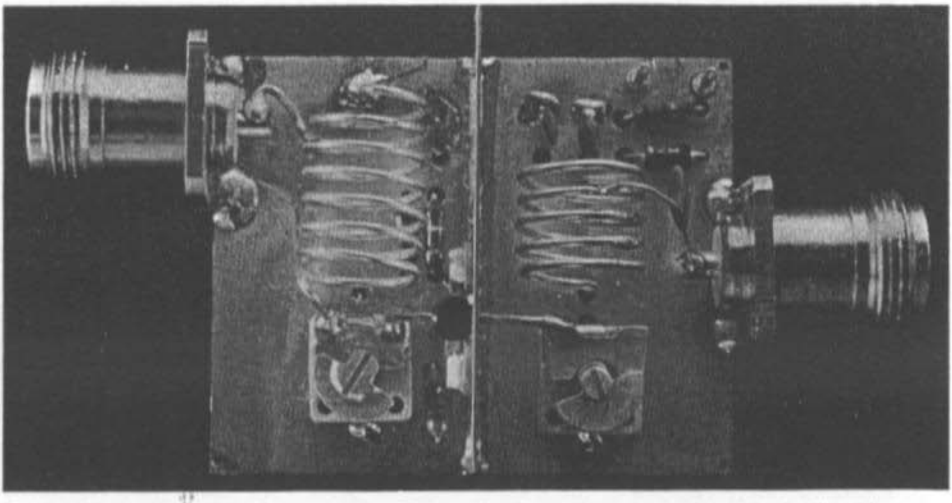

**Fig. 3: Photograph of the first prototype. Remote-feeding, attenuator and case had still not been provided. This prototype was used for the preliminary measurements**

amplifier firstly for maximum noise (if no signal is available). After this, align the preamplifier carefully for best signal-to-noise of a weak beacon signal. If the preamplifier breaks into oscillation, this will probably be caused by the poor matching between the output of the preamplifier and the input of the receiver. Since the gain is usually too great for most applications, the attenuator shown as dashed lines in the circuit diagram can be provided, and this connected into the signal path between the output of the preamplifier and the input of the receiver. It is usually sufficient to provide an attenuator of 3 dB to avoid any tendency to oscillation. In practice, the coaxial cable between preamplifier and receiver usually possesses enough attenuation in order to neutralize the preamplifier. This is especially the case when the preamplifier is mounted at the antenna.

Too much gain will deteriorate the large signal handling capabilities of the receiver without bringing any advantages. This is especially the case when the preamplifier is connected to the receiver with a short cable or direct. In such cases, an attenuator should be connected between preamplifier and receiver, and this can usually be upto 10 dB without reducing the noise figure of the system. **Table 1** shows suitable attenuators for an impedance of 50 Ω constructed using standard values.

| Attenuation dB | R 6 Ω | R 5 = R 7 Ω |
|---|---|---|
| 3,15 | 18 | 270 |
| 6,14 | 39 | 150 |
| 9,54 | 68 | 100 |
| 12,4 | 100 | 82 |

**Table 1:**
**Suitable attenuator four-poles in pi-circuits using standard resistor values**

After good construction and optimum alignment, the noise figure of the preamplifier will be only dependent on the tap on inductance L 1. If the required measuring equipment is available, the most favourable tapping point can be determined. In all cases it will be between turn 1 and 2 from the cold end.

The following values were measured on the prototypes (without attenuator):

| | |
|---|---|
| Gain: | 28 dB |
| 3 dB-bandwidth: | 13 MHz |
| NF: | 0.5 - 0.9 dB |

*Jan Martin Nøding, LA 8 AK*

# Switching Logic for Feeding Preamplifiers

**The application that led to the development of these circuits was the control of my 144 MHz preamplifier without making any modification to my transceiver. Such a circuit can be very simple, but due to the bad experience made by many friends, it was decided to use a more extensive circuitry. The circuit should make it virtually impossible for the transmit energy to reach the preamplifier, whatever faults occur. With simpler circuits, such accidents can occur occasionally if the operating voltage should fail, if the PTT-line is broken, or when a relay is defective.**

Two circuits are to be described from the large number considered. The switching circuit shown in **Figure 1** is part of the two subsequent circuits (Figures 2 and 3.) The circuit is designed that the preamplifier is switched on when relay A is not energized. When the PTT-line from the transmitter is grounded, relay A will be energized and will connect the antenna relay to the transceiver output connector. In this case, the transceiver output can be connected via a power amplifier, or direct to the antenna via a break contact on the preamplifier relay.

If the line to the transceiver is broken and relay A is not energized, the transmit energy will be converted to heat in resistors R 1 and R 2 **(Figure 2)**, or directly fed to the output in the circuit given in **Figure 3**. The latter circuit is to be preferred in practice, although there is no reason why the circuit given in Figure 2 should not operate correctly.

Resistors R 1 and R 2 should have the lowest inductivity possible. A rating of 0.25 W should be sufficient for the few seconds required before noticing the fault. The meter reading and LED will indicate when the circuit is not operating correctly.

A Reed-relay should not be used as relay A in Figure 1, but a conventional telegraphic relay type such as a Tris 154 or similar. Resistor R 1 is a shunt resistor for the meter, whose value should be selected to obtain the desired deflection on the meter. If, for instance a current of 60 mA is required, it is advisable to shunt the meter for 100 mA full scale deflection.

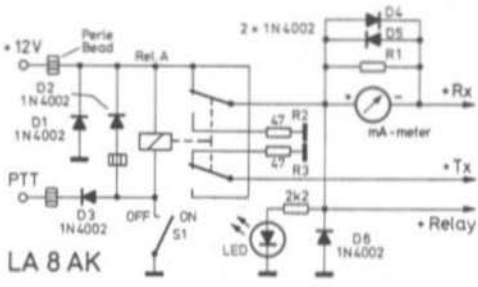


**Fig. 1:**
**This control and monitoring circuit can be used to achieve the relay switching shown in Fig. 2, or the diode switching given in Fig. 3. R 1 is the shunt resistor for the meter. R 2 and R 3 ground the DC when the circuit is switched off.**

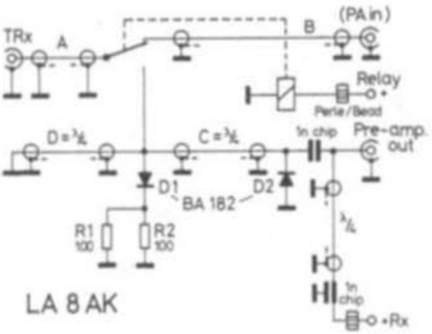


**Fig. 2:**
**In this coax-relay switching, coaxial cables A and B can be of any required length. The other three cables are λ/4 in length. The maximum power rating is dependent on the coaxial relays used.**

The following should be noted with respect to the circuit shown in Figure 3, which is used with success by the author:

The circuit is enclosed in a case made from PC-board material. The dimensions are 100 mm by 60 mm by 35 mm. As can be seen in **Figure 4**, the case is divided into three chambers. Whereas ferrite chokes are good for low RF-voltages, λ/4 lines are preferable for high RF-levels. These are made from thin 50 Ω coaxial cable RG-174/U and have the following lengths:

| Line | Length |
|---|---|
| A | as required |
| B | λ/4 |
| C | λ/4 |
| D | λ/2 |
| E | λ/4 |
| F | λ/2 |
| G/H | as required |

Lines A and B are enclosed in chamber 1, lines C, F, and H in chamber 2, and lines D, E, and G in chamber 3. A low-capacitance PTFE feed-through is provided on the panel between chambers 1 and 2 on to which the components D 1, RFC 1, and C 3 are soldered.

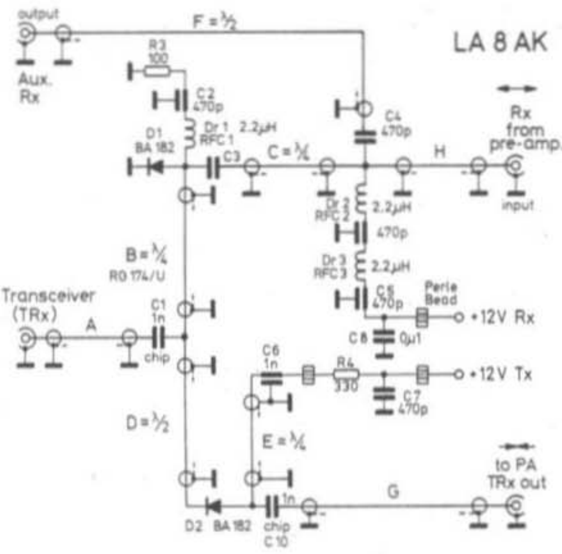


**Fig. 3:**
**This elegant diode antenna switching was tested upto a power level of 10 W. The BA 182 is an older switching diode which can be replaced by the more modern types BA 243 or BA 282.**

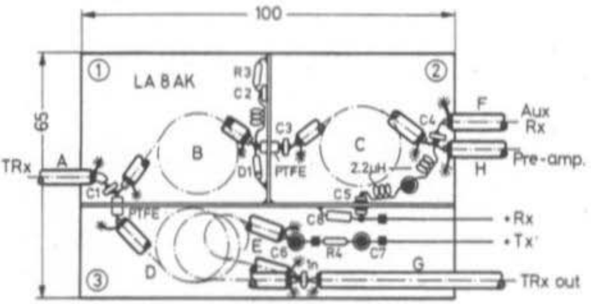


**Fig. 4: Recommended construction of the circuit given in Fig.3**

The described diode switching and preamplifier feed can be used for RF-power levels upto approximately 3 W. It provides an isolation of about 30 dB at the receiver input. A loss of approximately 0.1 W (at 2.8 W) is present in the transmit path before switching. These values were measured with a Bird-Wattmeter 43 in the 10 W range.

Since the output level of the BFT 66 preamplifier used by the author is high due to the gain of between 20 and 30 dB, it is possible for a second receiver to be connected to the second output »Aux.RX«.

The circuit was tested with a maximum of 10 W, and several faults were simulated by breaking the 12 V and PTT-lines. No incorrect operation could be observed.

The antenna relay of the mast-head preamplifier should be switched in parallel with the operating voltage as can be seen in **Figure 5**. A second relay K 2 is used in order to keep the transmit power level at the preamplifier input below the limit value of 10 dBm (10 mW).

An isolation of at least 40 dB will be necessary at a transmit power level of 100 W. This cannot be achieved by a short-circuit (especially at VHF/UHF) but with a good 50 Ω termination.

Finally a note regarding the overall gain: If the feeder following the mast-head preamplifier is not too long, the overall gain is often high enough that local stations overload the receiver. This can be avoided by reducing the gain to a suitable level or by providing an attenuator in front of the receiver, or by disabling one of two preamplifier stages.

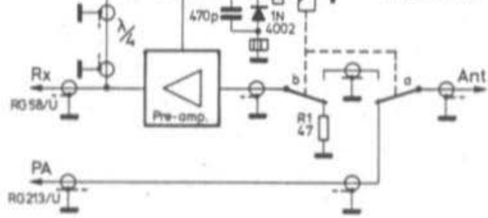


**Fig. 5: Typical circuit for switching a mast-mounted preamplifier; contacts a and b should preferably be in two different relays**

*Klaus J. Schoepf, DB 3 TB*

# A VXO-Local-Oscillator for 144 MHz Transceivers

**A local oscillator frequency of either 135 to 137 MHz (9 MHz IF), or 133.3 to 135.3 MHz (10.7 MHz IF) is required for 144 MHz transceivers. Several different methods are available for achieving this: Superhet VFO, phase-locked oscillator, synthesizer, or VXO. The first oscillators have the disadvantage that a temperature-compensated, and mechanically stable VFO is required. Both PLL-VFO and synthesizers exhibit the well-known problems envolved with VCOs. Variable crystal oscillators (VXO) do not possess these problems, and are also very low-noise oscillators. Until now, they have not been used extensively, probably because of the high cost of the crystals. However, the problem of cost can be avoided by using cheap CB-crystals. The low cost of the crystals themselves makes it worthwile to carry out a frequency multiplication by five to obtain the required frequency.**

## 1. CIRCUIT DESCRIPTION

As shown in **Figure 1**, 8 crystal oscillators are built up using a Clapp-circuit, and the required frequency range is selected by connecting the operating voltage to the appropriate oscillator. Of course, the operating voltage must be well stabilized and be hum-free in order to obtain a clean output signal.

The crystals are pulled by the series-impedance, comprising the pulling inductance Z and the parallel capacitor of 6.8 pF, far enough from the nominal frequency so that the required pulling range of 50 kHz can be obtained with the aid of the varactor diode. The pulling ranges of the individual oscillators overlap by at least 25 kHz.

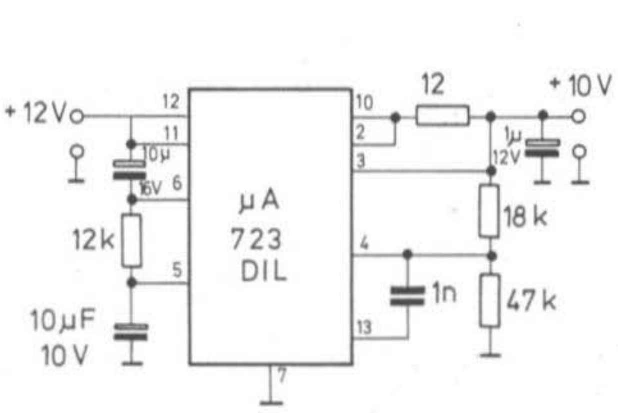


**Fig. 2:**
**A stable, and hum-free tuning voltage can be obtained using a µA 723**

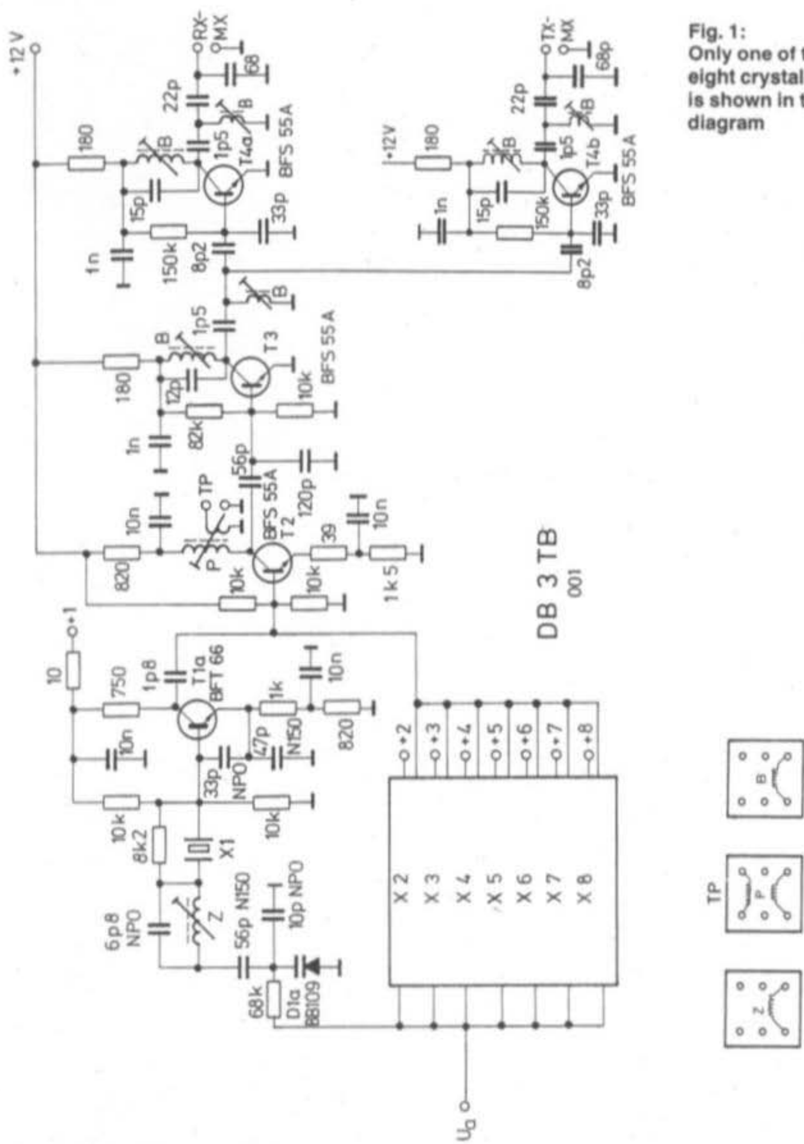


**Fig. 1: Only one of the total of eight crystal oscillators is shown in the circuit diagram**

The tuning is made with the aid of varactor diodes type BB 109. The tuning voltage $U_t$ is in the range of approx. 2.5 V to 8 V, and must also be stable and hum-free. This can be obtained using the stabilizer circuit comprising the µA 723 stabilizer shown in **Figure 2**. This circuit is only used for the tuning voltage and the power supply of the oscillators.

The oscillators are followed by a buffer (T 2), which allows a very weak coupling to be made to the crystal oscillators via 1.8 pF, and also ensures that there is no reaction from the subsequent frequency multiplier. The active frequency multiplier is equipped with transistor T 3. A capacitively-coupled bandpass filter is provided in the collector circuit in order to ob-

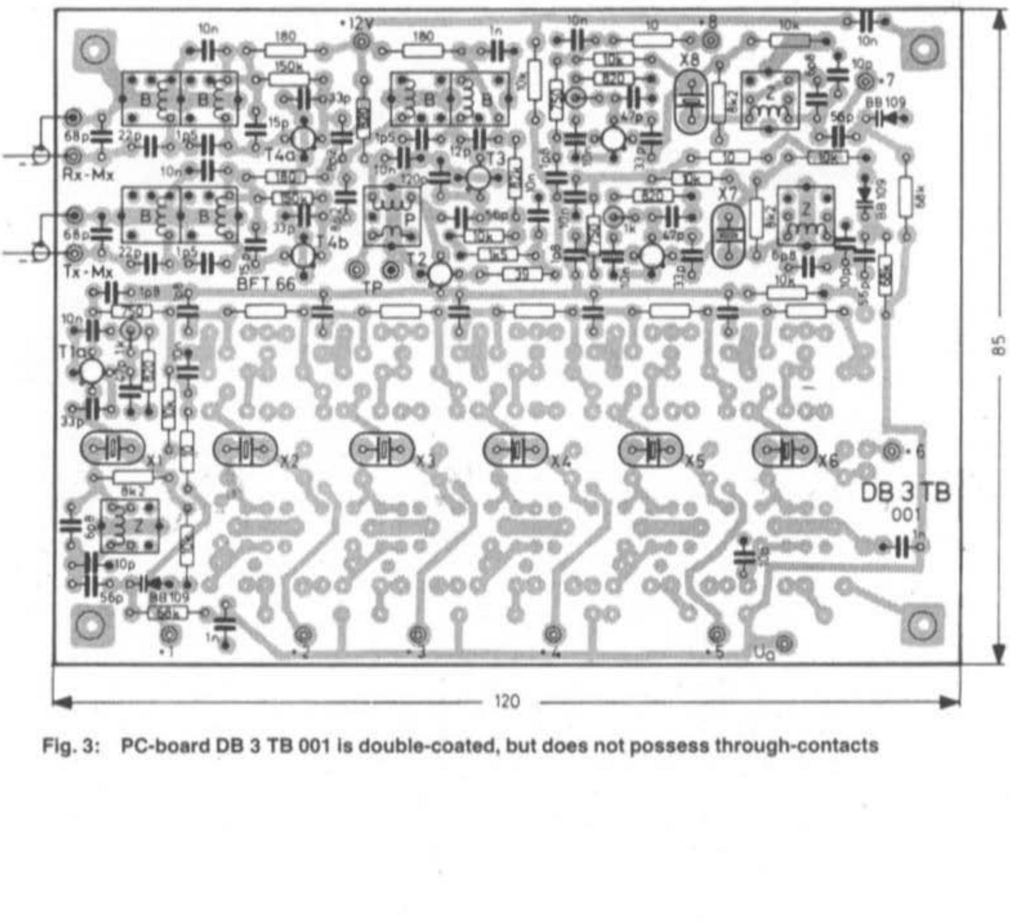


**Fig. 3: PC-board DB 3 TB 001 is double-coated, but does not possess through-contacts**

tain the required bandwidth of 2 MHz, and to suppress the unwanted spurious lines at a spacing of approximately 27 MHz.

The signal path is separated after the frequency multiplier for the transmitter and receive mixers. Each is provided with its own amplifier stage having a bandpass filter at the output for further spurious rejection. The output level is then approximately + 10 dBm (across 50 Ω). This is usually sufficient for driving most mixers.

## 2. CONSTRUCTION

The eight crystal oscillators, buffer, frequency multiplier, and the two output amplifiers can be accommodated on a common PC-board as shown in **Figure 3**. The dimensions of the PC-board are 120 mm x 85 mm, and the upper surface is in the form of a continuous ground surface. All connections that are grounded must therefore be soldered to the upper side of the board.

Measurements have shown that the noise of the output signal could be reduced by approximately 3 dB by using a transistor type BFS 55 A. For amateur applications, however, the cheaper transistor complements BF 224/BF 199 will be sufficient. Attention must be paid that these types are correctly installed.

The coils are wound according to the following information and connection drawings. The beginning and end of the winding are coated with solder, wound around the appropriate connections after which they are soldered into place. The soldering process must be rather quick so that the coil formers are not damaged.

## 2.1. Components

T 1a - T 1h: BFT 66 (Siemens), or, if not available, BF 224 or BF 199

T 2 - T 4 b: BFS 55 A (Siemens), if not available, BF 199 (see text)

D 1a - D 1h: BB 109 (Siemens)

1 piece integrated stabilizer µA 723 (DIL)

Ceramic disk capacitors for 5 mm spacing; the temperature compensation values given in the circuit diagram must be observed.

Non-ribbed coil formers with an outer diameter of 4.3 mm are used for all coils. The cores are M 3.5 x 0.5, and screening cans of 7.5 x 7.5 x 12 mm high are used.

8 pieces pulling coils »Z«: 23 turns of HF-stranded wire 8 x 0.03. Core: orange

1 piece coil »P«: 13 turns of 0.2 mm dia. enamelled copper wire; coupling winding: 2 turns, core: orange

6 pieces coil »B«: 4 turns of 0.4 mm dia. enamelled copper wire. Core: violet

All resistors for 10 mm spacing.

Crystals: Case HC-25/U or HC-18/U.
Frequencies: see Table in section 3.

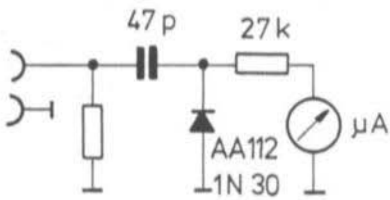


**Fig. 4:**
**A simple RF-indicator suitable for alignment**

## 3. ALIGNMENT

A frequency counter and two simple VHF-indicators as shown in **Figure 4** are required. The frequency counter is connected to testpoint TP, which can be used later for connecting an internal frequency counter. The wiring is shown in **Figure 5**.

The frequency ranges are selected one after another and are aligned for the lower frequency limit with minimum tuning voltage (approx. 2.5 V) according to the frequency table. The frequency counter remains connected during the subsequent alignment.

The frequency range 145 to 145.25 MHz is now selected and the tuning voltage reduced to minimum. An RF-indicator is now connected to the outputs RX-MX and TX-MX, and all resonant circuits are aligned for maximum output. If available, the bandpass filters can be aligned with the aid of a sweep generator to the required frequency band.

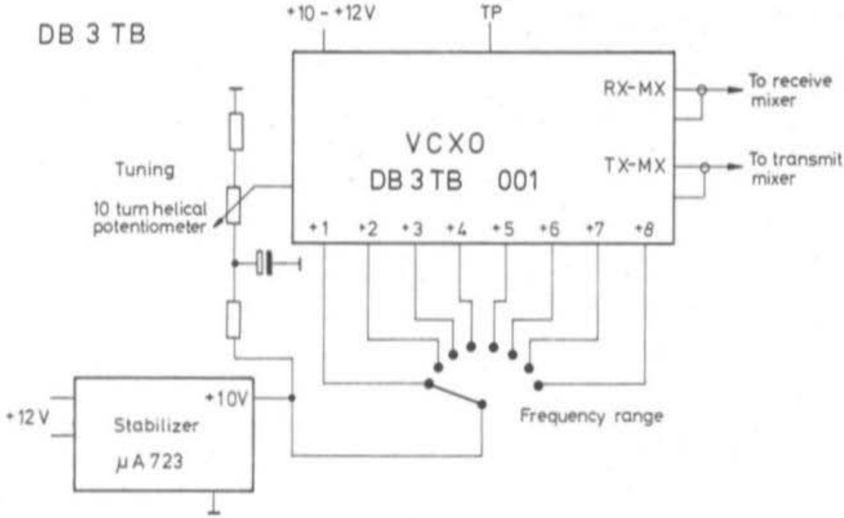


**Fig. 5: Wiring of the whole local oscillator**

| Range (MHz) | Crystal Frequencies (MHz) IF = 9 MHz | IF = 10.7 MHz | Frequency at $U_t$ (min) IF = 9 MHz | IF = 10.7 MHz |
|---|---|---|---|---|
| 144.000 – 144.250 | 27.085 | 26.750 | 26.995 | 26.655 |
| 144.250 – 144.500 | 27.135 | 26.800 | 27.045 | 26.705 |
| 144.500 – 144.750 | 27.185 | 26.850 | 27.095 | 26.755 |
| 144.750 – 145.000 | 27.235 | 26.900 | 27.145 | 26.805 |
| 145.000 – 145.250 | 27.285 | 26.950 | 27.295 | 26.855 |
| 145.250 – 145.500 | 27.335 | 27.000 | 27.245 | 26.905 |
| 145.500 – 145.750 | 27.385 | 27.050 | 27.295 | 26.955 |
| 145.750 – 146.000 | 27.435 | 27.100 | 27.345 | 27.005 |

**Table 1: Frequency Plan**

## 4. REFERENCES

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Temperature-Compensated Oscillator with Varactor Tuning
VHF COMMUNICATIONS 5, Edition 2/1973, pages 116-122

*Thomas Morzinck, DD Ø QT*

# A Receive Converter for the 6 cm Band

**»How was it during the contest ?«**
**»Phantastic ! Especially the enormous pile-up on 6 cm !«**
**»Pile-up on 6 cm ???«**
**»Of course, two stations in only nine hours .....«**

## INTRODUCTION

This fictitious conversation clearly describes the present situation on the so-called 6 cm band, whose narrow-band section is from 5760 to 5762 MHz.

The reasons for the low activity – and not only on 6 cm – are, of course, well known, and are not to be discussed in this article, with the exception of one of them. If someone is interested in this frequency range and goes looking for some constructional articles, he will find very little, whether this is for converters, or for frequency multipliers (4). For this reason, the references at the end of this article will give several useful publications for constructions on this band.

## MUST IT ALWAYS BE WAVEGUIDE ?

Mechanical work is very difficult for many readers, especially when such things as a lathe are not available. On the other hand, a mechanical construction is often the only possibility in order to become active in the SHF-range at an acceptable price. This is especially valid for the reproducibility of constructional articles. The question whether a converter functions is usually decided by the noise figure, which is the most important criterium for converter constructors. Unfortunately, it increases on increasing the frequency, especially when one attempts to reduce the size of converters using »normal« technology to achieve higher frequency ranges. More regarding this problem was described by Dahms in (7), and Heidemann in (3).

For this reason, the constructional article published by Neie (1) was not suitable for me, although no external oscillator chain was required and one could be able to avoid the »two-box« construction with intermediate connections.

The remaining waveguide-constructions described in (2) and (3) have been combined by me, and the results of this combination are now to be described in detail. The construction is mainly based on details given by Kuhne (2), and it is only the input coupling that has been replaced by the wideband waveguide-coaxial transition described in (3).

The operation will be seen easily in **Figure 1**: Both the input signal from the N-connector, and the oscillator signal generated in the varactor multiplier are filtered out in filter 1, or 2 respectively and fed to the central mixer diode. The IF-output coupling is made – as usual – via a bypass capacitor (C 1). A suitable IF-amplifier is, for instance, one of those described in (4) or (7).

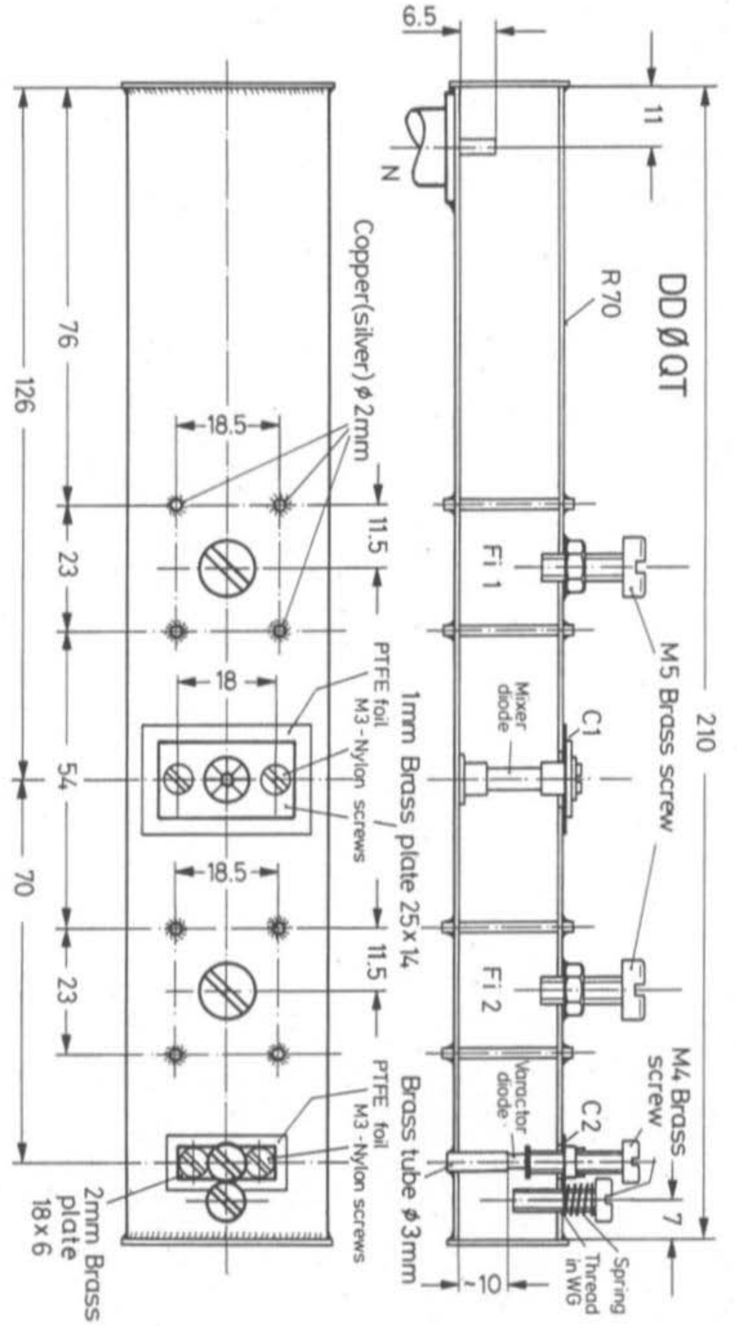


**Fig. 1: The converter is built into a piece of waveguide. It consists mainly of: wideband coax/WG transition, signal frequency filter (Fi 1), mixer diode, LO-frequency filter (Fi 2), and varactor multiplier**

## SPECIAL FEATURES

In contrast to the construction in (3), the varactor stage is integrated into the waveguide. The most important modification is the circuit of the multiplier stage, which multiplies by five. This is now very simple since it can be driven by a standard 1152 MHz oscillator chain, which is most certainly already available in the shacks of most UHF-SHF amateurs. In order to obtain the required local oscillator frequency of 5616 MHz necessary for an IF of 144 MHz, it is only necessary to exchange the 96 MHz crystal for one of 93.6 MHz.The required alignment corrections at 1123.5 MHz are very small, and will always be within the alignment range of the trimmers and inductances.

A further difference to (2) is the tuning screw between varactor diode and waveguide termination. This considerably improves the efficiency and provides a higher mixer diode current. A further alignment possibility using a screw between filter 2 and the varactor diode was found to be without effect, and can therefore be deleted.

## CONSTRUCTION

A short piece of R 70 (WR 137 or WG-14) was used as waveguide. The inner dimensions are now to be given so that any interested constructor can make it from 1 mm to 2 mm thick brass plate, if he is not able to obtain a suitable waveguide. The inner dimensions are: 34.8 mm x 15.8 mm. All important dimensions can be taken from Figure 1, and need not be discussed here.

## IMPORTANT COMPONENTS

Varactor diode: BXY 28 or BXY 38 (Philips)

Mixer diode: BAW 95 (Philips)

C 1,C 2: Home-made plate capacitor with PTFE-foil (maybe with glass fibre), approx. 0.13 mm thick size: 30 x 18 or 22 x 10 mm

C 3: 6 pF plastic foil trimmer, 7.5 mm dia. (Philips: grey)

C 4,C 5: 6 pF ceramic tubular trimmer (Philips)

C 6: Chip capacitor 100-1000 pF

L 1: 1 turn of 1 mm dia. silver-plated copper wire wound on a 4 mm former, self-supporting

L 2: Brass strip 14 x 5 mm 0.5 mm thick, mounted approx. 5 mm over the ground surface

## FURTHER DETAILS

The eight filter wires should be soldered into place. This is followed by soldering the nuts for the M 5 tuning screws to the outside of the waveguide. This is made as follows:

A M 5-thread is cut into the waveguide and a nut together with an **aluminium** screw screwed into place so that it fits tightly. It is now possible for the soldering process to be carried out without danger.

The greatest problem is the mount for the mixer diode in the waveguide, which is not too accessible due to the filter.

However, this difficulty can be solved quickly using the following trick:

A hole of 2 mm in diameter is drilled directly opposite the hole for C 1, which must be central to the later position of the diode. It is now possible for the diode mount to be screwed tight with the aid of an M 2 screw and nut through this hole. The waveguide can now be heated from below without worrying about the shifting of the diode mount. Of course, it is necessary for the corresponding surface in the waveguide and the lower side of the diode mount to be solder-plated beforehand, and this should not be too thick, otherwise the mount could be tilted ! The

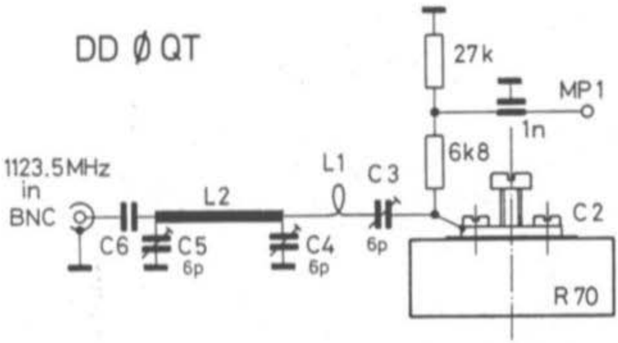


**Fig. 2: The varactor multiplier is driven with approx. 100 mW at 1123.5 MHz via this circuit**

screw can be removed after the soldering process – the mount should then be in its correct position !

For mounting the modules »IF-preamplifier« and »varactor multiplier/input circuits« **(Figure 2)**, it is necessary for two brass plates of 0.5 - 1.0 mm thickness to be hard-soldered to the waveguide for use as mounting flanges. After carrying out the metal work, it is possible for the PC-boards to be placed slightly under these brass plates and soldered into place using a 100 W soldering iron. This can be seen in the photograph of the author's prototype in **Figure 3**.

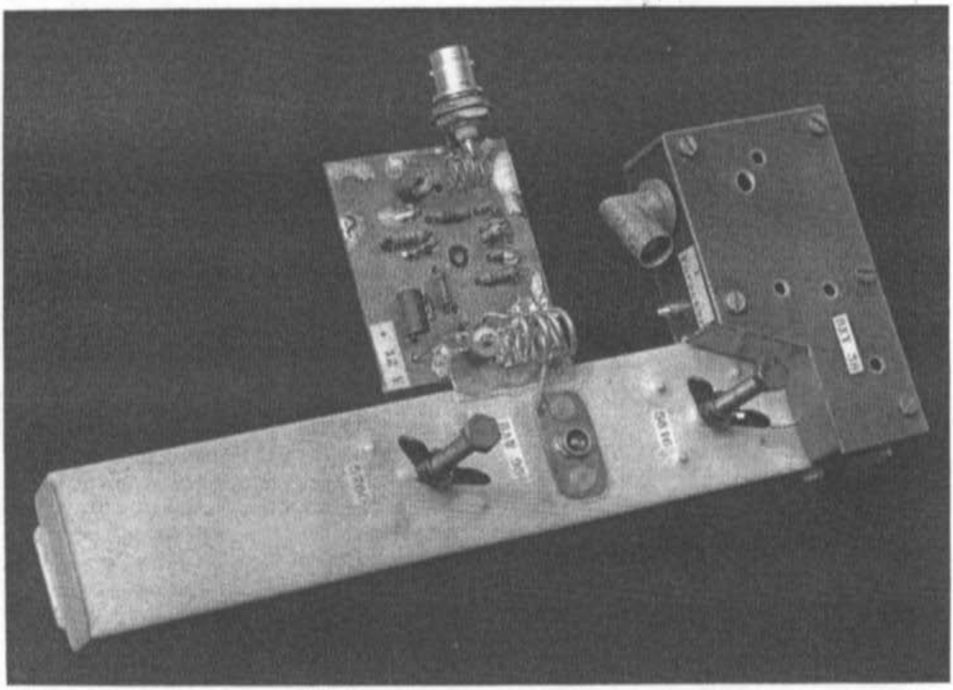

**Fig. 3: Photograph of the author's prototype converter**

## ALIGNMENT

Inject an oscillator power of approximately 100 to 200 mW at 1123.5 MHz. The input circuits of the varactor multiplier should be aligned with the aid of a μA (mA)-meter connected between testpoint 1 and ground. All further alignments are made for maximummixer diode current.

Several indications of mixer diode current will result on tuning the screw of filter 2. The correct maximum resulted in the author's prototype at a slight insertion depth of the M 5-screw into the waveguide.

When correctly aligned to 5760 MHz, the position of the screw of filter 1 will only be slightly different from that of filter 2 !

## MEASURING RESULTS

The noise figure of the converter was measured to be 9 to 10 dB when using a random diode type BAW 95. Further measured values can be seen in **Table 1**.

| Input power 1123.5 MHz | Mixer diode current BXY 38 | Mixer diode current BXY 28 |
|---|---|---|
| 120 mW | 4 mA | 2.5 mA |
| 700 mW | 45 mA | 25 mA |

The measurement up to 120 mW was made after reducing the input power, but **without** realignment.

After removing the mixer diode and tuning both filters to 5616 MHz, an output power of approximately 65 mW was measured at the input socket (1123.5 MHz in: 700 mW, varactor: BXY 38).

## FINAL NOTES

Several things can be improved on this converter and the main thing would be the varactor multiplier. If a short-circuit plunger and a diode mount that can be shifted vertically were to be used, I am sure that the efficiency could be increased to more than 10 %. The IF-output coupling using a bypass capacitor is also lossy, and the choke output coupling used in (3) should be able to increase the sensitivity.

The author would like to thank Rolf Heidemann, DC 3 QS for his assistance during the measurements.

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DUBUS-Information, Edition 2/79

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*Uli Mallwitz, DK 3 UC*

# Experiments with a 10 GHz Frequency Multiplier with Interdigital Filter Coupling

**In (1) DK 1 UV described a frequency tripler from 3456 GHz to 10368 GHz. Unfortunately, this tripler exhibited high conversion loss. The author then carried out experiments to improve the poor efficiency.**

**In addition to this, attempts were made to also use the improved frequency multiplier simultaneously as up-converter for a 10.3 GHz narrow-band communication system. The 144 MHz band was to be used as intermediate frequency range.**

## 1. IMPROVEMENT OF THE FREQUENCY TRIPLER

The conversion loss of over 10 dB could have several causes:

1) Unsuitable diode
2) Unsuitable matching
3) Incorrect operating point
4) High losses at 3.4 GHz

The nameless cheap diode manufactured in England that looks like the 1N23 is quite popular and is used extensively. It should operate well in the 10 GHz band (DC 3 QS). Due to the wide-spread use of this diode, the description is to be based on it.

Since the matching and operating points have been solved in a similar manner to other descriptions, one must assume that the values are approximately correct.

With respect to the high losses at 3.4 GHz, a comparison showed that this is where the problem lies. Namely, an interdigital filter mixer for the 9 cm band as described in (2) was equipped with a Schottky diode HP 2800 and driven with a known power; the resulting diode current was noted. The same diode was then installed in the interdigital filter arrangement described in (1) after which it was driven with the same RF-power. The resulting diode current was more than ten times less !

The experiments were repeated with various drive powers and using other diodes. They showed clearly that the losses in the signal path from the input connector to the diode were too high.

In order to avoid a recalculation and/or a silver-plating of the filter parts, the design described in (2) was used to drastically reduce the losses. After carrying out a few modifications, which are to be described later, this resulted in a design that produced a diode current as was exhibited in the original 9 cm interdigital converter.

### 1.1. CONSTRUCTION OF THE TRIPLER

The photograph of the author's experimental prototype **(Figure 1)** shows the waveguide part at the top, the interdigital filter with input connector for the 3456 MHz signal at the center, and the 144 MHz circuit for the up-conver-

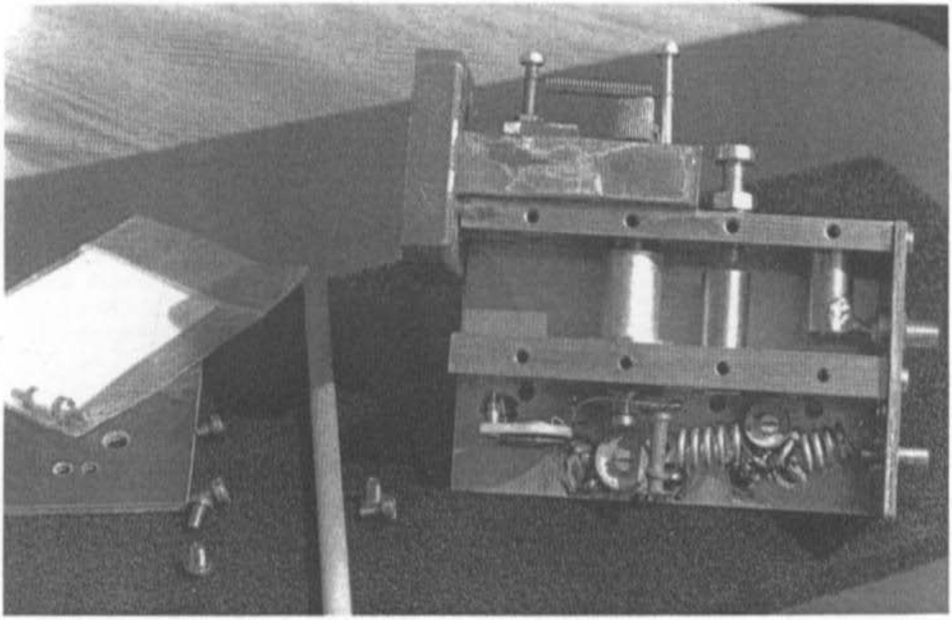

Fig. 1: Photograph of the author's prototype with cover removed

ter below. Further details are given in **Fig. 2.** All three fingers are made from non-silver-plated, polished brass; two are of 10 mm diameter, and are 16 mm in length, and the third is 12 mm in diameter and 16.5 mm long.

The spacing from the SMA-connector to the third finger – which supports the tripler diode – was taken from the original design. It may be noticed that the finger for the diode has been

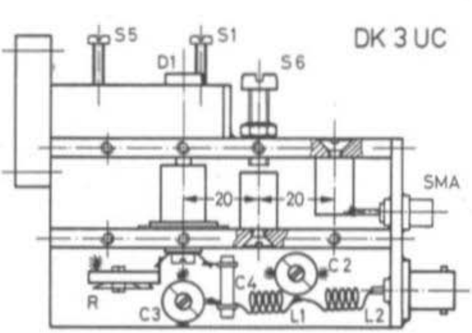


**Fig. 2:**
**Overall construction and important dimensions**

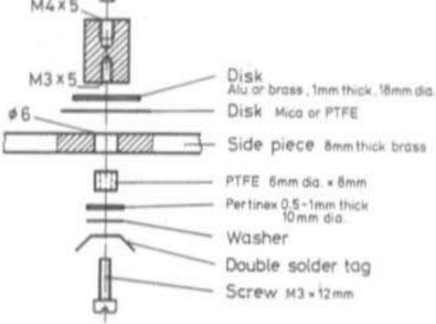


**Fig. 3:**
**The finger supporting the diode is capacitively bypassed to ground**

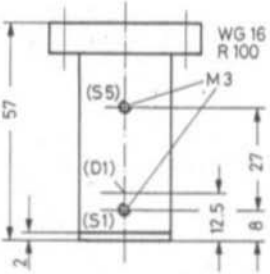


**Fig. 4: Waveguide part with choke flange**

turned around in order to have all alignment screws on the same side. Since this finger is dampened considerably, this has no adverse effect. This finger is also insulated galvanically and is bypassed to ground using a home-made bypass capacitor **(Figure 3)**. A capacitor made in this manner will have a capacitance of approximately 10 to 100 pF according to the dielectric. The finger is accessible for DC-currents via the M 3-screw, so that a trimmer potentiometer can be connected for adjusting the operating point. Usual values are between 20 and 100 kΩ.

**Figure 4** shows the output waveguide with its three holes, the short-circuit plate and the standard flange (the length of 57 mm is valid when using a choke flange). The circuit diagram given in **Figure 5** shows the double Pi-

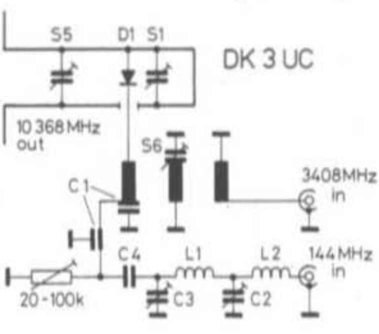


**Fig. 5:**
**Circuit diagram of the frequency tripler/ up-converter**

link for injecting the IF-signal. This is required for the up-converter, which is to be described later.

Special care must be taken when assembling the parts shown in Figure 3: The surfaces that form the bypass capacitor must be flat; the finger must be at a right angle to the side pieces so that the diode is not damaged later. The same is valid for the transition finger/diode. This finger has a dia. of 12 mm and may not be longer than given. Several constructions showed that the diameter was somewhat critical; this was caused by the impedance of this finger.

## 1.2. ALIGNMENT OF THE FREQUENCY MULTIPLIER

A 3456 MHz signal at a power level between 50 and 500 mW is now fed to the SMA-connector (see Figure 2), and the interdigital filter is aligned as described in (2). After inserting screws S 1 and S 5 into the waveguide by approximately 5 mm, it should be possible for a power level to be indicated when using a diode probe at the output of the waveguide. The interdigital filter resonator should be now optimized by alternate alignment, in conjunction with screws S 1 and S 5, and the trimmer R. The prototype did not show any abrupt tuning behaviour; the output power could be aligned continuously with all of the tuning elements. However, this can be different when using other diodes.

An output power of 27 mW was measured at 10368 MHz with a drive power of 120 mW.

If this multiplier is to be installed in a waveguide group, it is advisable for a fine alignment to be made on screw S 5 after connection. Screws 3 and 4 that were given in (1) have been found to be superfluous in all test series, whereas screw 2 was necessary with several examples. It is therefore advisable to provide these holes, and to use them as required.

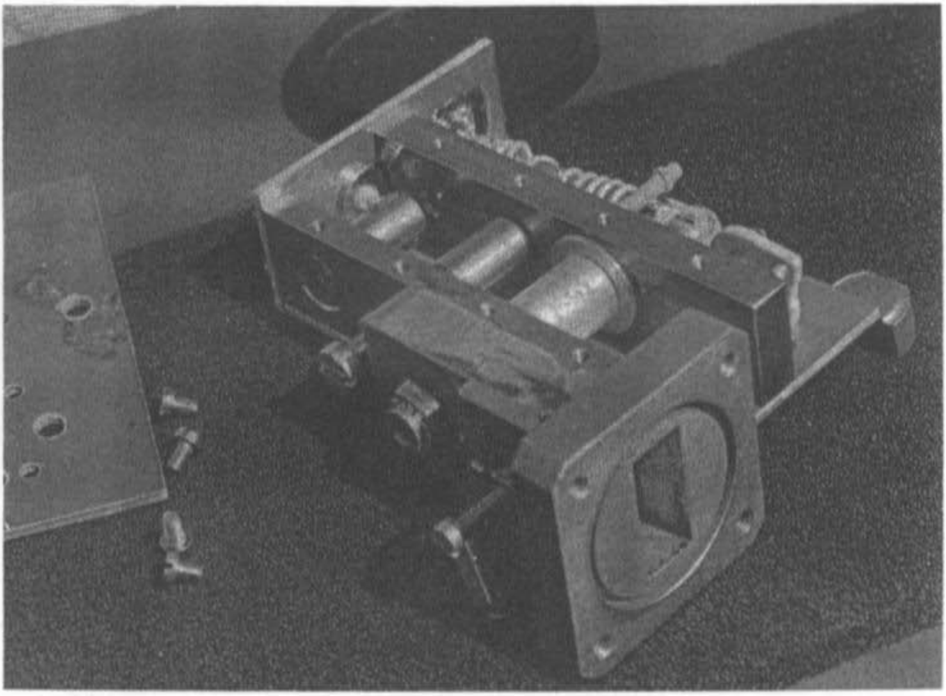

**Fig. 6: The tuning screws and the bypass capacitor can be seen clearly**

## 2.
## FREQUENCY TRIPLER / UP-CONVERTER

This extension of the tripler has already been shown in Figure 2 and 4. **Figure 6** shows the opened prototype from the other side.

The components of the 144 MHz circuit are as follows:

L 1, L 2: 5 turns of 0.8-1 mm dia. silver-plated copper wire wound on a 6 mm former, self-supporting

C 1: Home-made disk capacitor as shown in Figure 3

C 2: Plastic foil trimmer, 60 pF (Philips, red, large type)

C 3: Plastic foil trimmer, 45 pF (Philips, yellow, large type)
– may not be required ! –

C 4: Ceramic disk capacitor 4.7 nF

R: Trimmer potentiometer 100 kΩ

Coaxial connector for 3 GHz:
SMA, SMC, or other suitable connectors

## 2.1.
## ALIGNMENT OF UP-CONVERTER

Firstly carry out the alignment described in 1.2. The output frequency of 10368 MHz can be indicated by using a previously aligned filter for this frequency together with a detector, or with

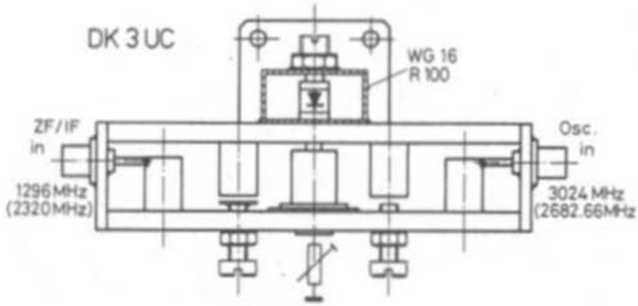


**Fig. 7:**
**Proposal for a frequency multiplier/mixer with an IF in the 23 cm or 13 cm band**

the aid of an existing receiver (e.g. a »Gunnplexer«) together with an attenuator of approximately 80 to 140 dB.

The 144 MHz IF-signal is now fed to the module at a power level of approx. 150 mW, after which screws S 5 and S 1, and the potentiometer R should be aligned for maximum output power. The trimmer capacitors C 2 and C 3 should be inserted to half capacitance during this alignment. Finally, C 3 should be aligned for maximum required signal, and C 2 for best matching (lowest VSWR) between 2 m-transmitter and mixer module.

The following values were measured on the author's prototype:

| | |
|---|---|
| Oscillator power (3408 MHz): | 120 mW |
| Drive power (144 MHz): | 150 mW |
| Tripled frequency (10224 MHz): | 27 mW |
| 10368 MHz subsequent to DB 6 NT-filter: | 8 mW |

Experiments made with higher powers at 3408 MHz have shown until now that saturation effects occur at approximately 0.7 W. DK 1 ZD carried out experiments with a diode type DH 636. This diode went into saturation at approximately 2 W at 2 GHz, which means that hardly any conversion gain was present, and only 7 mW could be measured at 10368 GHz. After reducing the drive power to 0.5 W, the output power increased to approximately 20 mW ! These experiments are to be continued !

## 2.2. ANOTHER IF

The up-converter can also be operated with an intermediate frequency in the 70 cm or 23 cm band. For the latter, an interdigital filter can also be used at IF-level as a selective coupling to the mixer diode. Such a circuit is indicated in **Figure 7**. The frequencies for an IF in the 13 cm band are given in brackets.

The minimum complement is one coupling and one tuneable finger, each. In order to use only one profile, the fingers for the IF are greatly shortened. This means that a plate capacitor is required for tuning it to resonance.

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(2) J. Dahms, DC Ø DA:
Interdigital Converters for the GHz Amateur Bands – Interdigital Filters Extended to Form Receive Converters
VHF COMMUNICATIONS, Volume 10, Edition 3/1978, pages 154-168

*Sepp Reithofer, DL 6 MH*

# A Straight-Through Mixer for 24 GHz

**In recent years, a large number of radio amateurs have become active on the 10 GHz band. It will be seen, especially during contests, that there is a considerable trend to these higher frequencies. Of course, not all equipment is home-made. The 24 GHz band has been available to amateurs for some years now, but does not enjoy this popularity.**

**In the case of 10 GHz, the relatively inexpensive »Gunnplexer« has done much to increase activity on this band. In the case of home-made equipment, the so-called straight-through mixer has become popular due to the surprisingly good results obtained with it. The author described such a mixer for the 10 GHz band in VHF COMMUNICATIONS (1). This is now to be followed by a similar design for a transceiver for the 24 GHz band, which has proved itself in the field.**

## CONCEPT

Several different concepts for the 24 GHz band were built up and tested by the author. When making a comparison between expense and efficiency, the described straight-through mixer concept has been found to be the most favorable. **Figure 1** shows the complete, ready-to-operate module. The case is made out of PC-board material and accommodates all components necessary for operation except the battery. The waveguide module is shown in **Figure 2**.

### Positioning of the Mixer Diode

Experiments with straight-through mixers for the 10 GHz band have shown that a very high sensitivity can be achieved in the receive mode, however, the transmit power remains low. This ratio is even more critical when using such a mixer for the 24 GHz band, if a special construction is not used. The reason for this relatively low transmit power is the mixer diode, mounted between oscillator and antenna.

If the diode is placed out of the E-field, which is at its strongest at the center of the waveguide, it will not absorb so much power, which means that more is available for transmit applications. One must only pay attention that the mixer diode receives sufficient oscillator injection for reception so that the receive sensitivity is not deteriorated. This means that a compromise must be found for each oscillator power level between transmit power and diode current (approx. 0.5 to 3 mA).

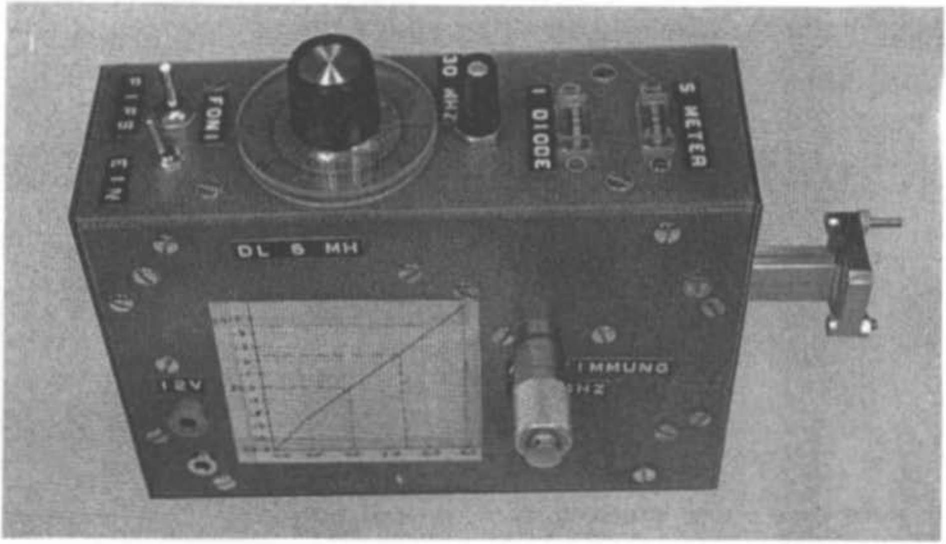



Fig. 1: The dimensions of this 24 GHz transceiver (without waveguide connection) are: Length = 150 mm, width = 50 mm, height = 100 mm

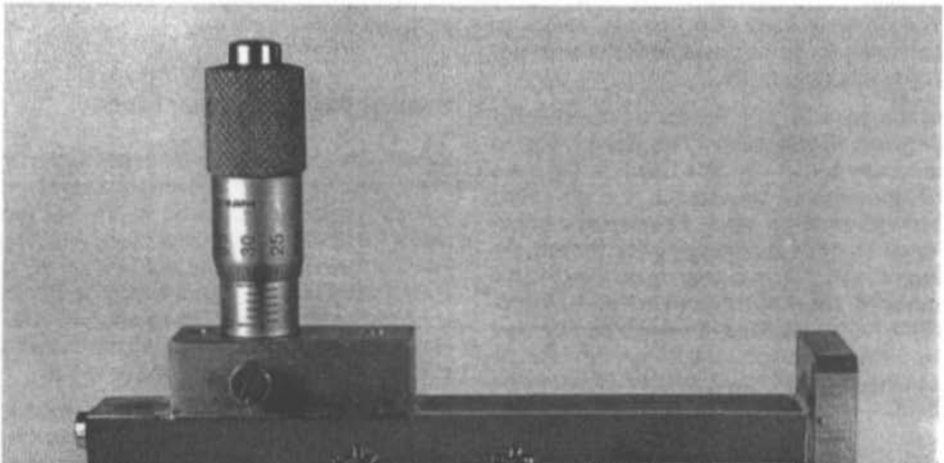



Fig. 2: The waveguide module from the side; the Gunn oscillator is tuned with the aid of a micrometer

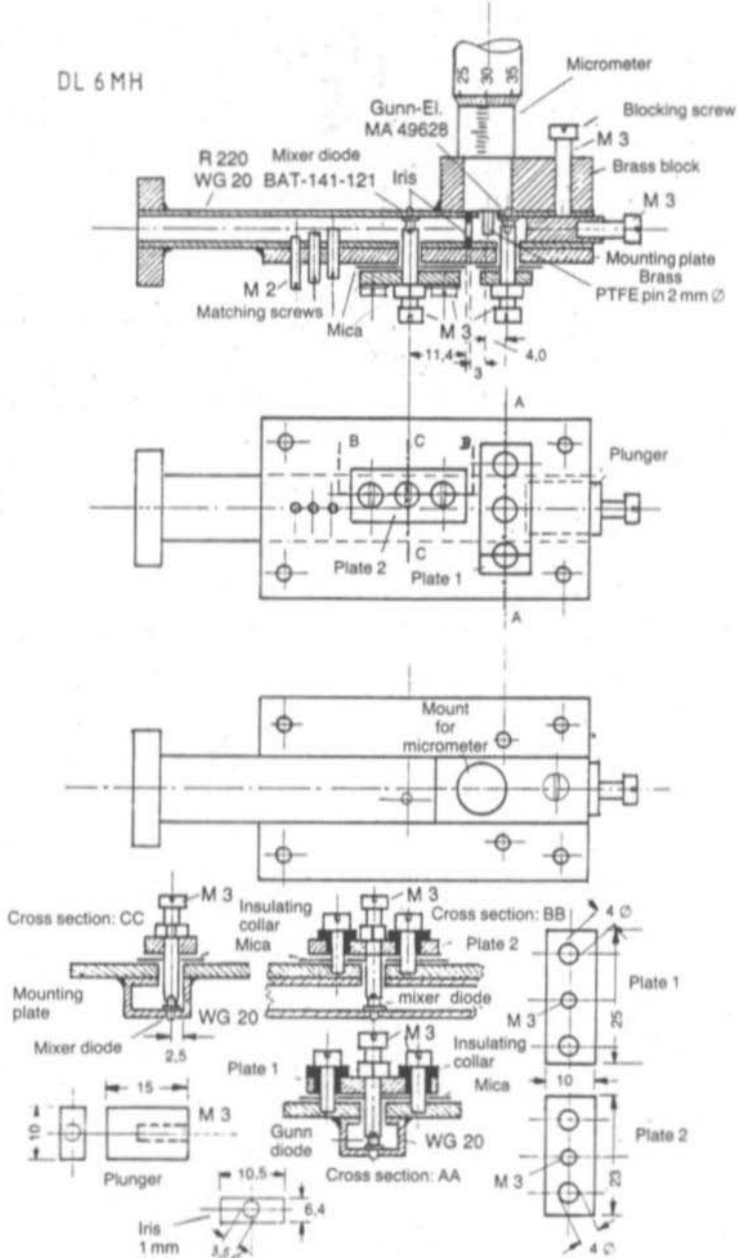



**Fig. 3: Cross sectional drawing, from below and above, as well as detail drawings of the 24 GHz straight-through mixer**

This seems to have been achieved in the case of the described mixer. When using a 12 mW Gunn diode, 4 to 5 mW are available in the transmit mode. During experiments made by the author on straight-through mixers where the diode was in the center of the waveguide (in other words at the position of maximum E-field), a higher conversion current was generated. However, only transmit power levels in the order of 100 to 200 μW were available. Even with these low transmit powers it was possible for communication to be made over distances of 30 km (line-of-sight), even if the signals were weak.

## CONSTRUCTION

**Figure 3** shows the cross section of the 24 GHz straight-through mixer; the cross-section drawing is not quite correct, but allows the construction to be described easily. The end of the waveguide is terminated with a short-circuit plunger. The plunger is provided with a layer of Sellotape in order to avoid intermittent contacts within the waveguide. The maximum RF-output of the oscillator is adjusted with this.

### Tuning

A PTFE tuning pin can be inserted into the resonator chamber with the aid of a micrometer screw between the Gunn diode and the iris at the other end of the resonant chamber.

The steel pin of this micrometer is softened and drilled out in order to insert the tuning pin. A brass block is soldered onto the waveguide for mounting the micrometer screw. This block is provided with a suitable hole for this. The micrometer is clamped using 3 mm screws at the side.

### Iris

The iris with an aperture of 3.5 mm diameter is mounted directly in the waveguide. In order to do this, slots are sawn into the broadsides into which the iris is placed and subsequently soldered. The slots may only be as thick as the metal plate of the iris. Normal metal sawblades provide a cut width of approximately 1 mm, which means that the iris can be just as thick. If it cannot be inserted tightly, it should be rubbed down with emery cloth. Attention should be paid during the soldering process that no solder can flow into the waveguide itself.

A rectangular slot can be used in the iris instead of the round hole. In this case, a slot is sawn into the narrow sides of the waveguide and a metal plate soldered in from both sides. The spacing between both metal plates should amount to 3 mm in the waveguide.

### Length of the Resonator

The spacing between center of the Gunn diode and the iris is selected to be somewhat less than $\lambda_l/2$. Wavelength $\lambda_l$ of the waveguide R 220 (WG 20) amounts to 15.25 mm at the center frequency of the band 24.1 GHz; this results in the theoretical value of 7.625 mm for $0.5 \times \lambda_l$. However, in order to provide a certain tuning range for the PTFE pin, a spacing of 7 mm has been selected. In the drawing, the spacing from Gunn diode to PTFE pin is 4 mm, and 3 mm from the PTFE pin to the iris. This is thus a total of 7 mm. This dimension is critical, since the frequency range to be achieved is dependent on this.

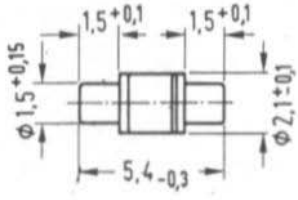
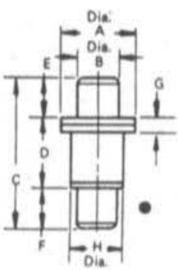



**Fig. 4:**
**Dimensions of the mixer diode BAT 14-121, which can be used for the frequency range between 26.5 and 40 GHz. Optimum noise figure is provided at an oscillator power of 2 mW.**




TYPICAL
$L_p$ = .42 nH
$C_p$ = .20 pF

| DIM. | INCHES MIN. | INCHES MAX. | MM MIN. | MM MAX. |
|---|---|---|---|---|
| A | .119 | .127 | 3,02 | 3,23 |
| B | .060 | .064 | 1,52 | 1,63 |
| C | .205 | .225 | 5,21 | 5,72 |
| D | .085 | .097 | 2,16 | 2,46 |
| E | .060 | .064 | 1,52 | 1,63 |
| F | .060 | .064 | 1,52 | 1,63 |
| G | .016 | .024 | 0,41 | 0,61 |
| H | .079 | .083 | 2,01 | 2,11 |

**Fig. 5: Dimensions of the Gunn diode MA-49628. The dot marks the heat-sink side (+ $U_B$)**

## Mixer Diode

The mixer diode is spaced 3/4 $\lambda_l$ from the iris, which amounts to 11.4 mm. As previously mentioned, it is not mounted in the center of the waveguide. The spacing from the inside of the waveguide to the center of the diode amounts to 2.5 mm.

Three matching screws are provided in front of the mixer diode with which the diode current of the mixer can be adjusted. The author used a mixer diode type BAT 14-121, which is manufactured by Siemens **(Figure 4)**. A good receive sensitivity was provided at a mixer current of approximately 200 μA. Under laboratory conditions, it was found that an input power of – 100 dBm provided a good indication on the S-meter.

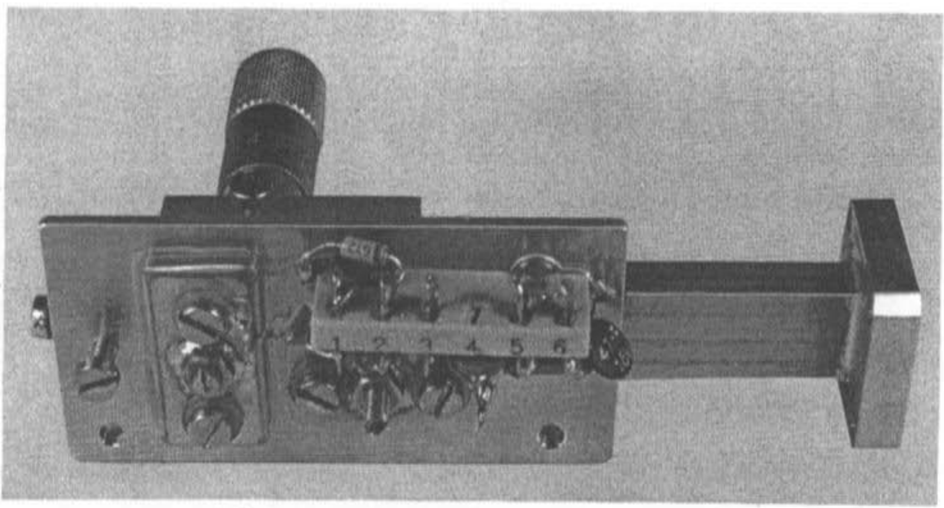



**Fig. 6: The bypass capacitors for the Gunn oscillator (left) and the mixer diode (below center), as well as the solder tag strip can be seen clearly.**

## Bypass Capacitors

The RF-bypass of the Gunn voltage, and the intermediate frequency of 30 MHz taken from the mixer diode is made with brass plates which are insulated from the waveguide with the aid of mica layers.

A mounting plate is provided on the lower side of the waveguide. Threaded holes of 3 mm in diameter are drilled in this plate for mounting the bypass capacitors. The insulating collars are usually used for mounting power transistors.

## FURTHER DETAILS

As can be seen in Figure 3, the screws for mounting the Gunn diode and the mixer diode have been drilled at their ends.

A Microwave Associates – Gunn diode type MA 49628 was available to the author.
Its main specifications at 22 GHz are:
$P_{min}$ = 10 mW/$U_{op}$ = 5.0 V (typ), or
8.0 V (max)/$I_{op}$ = 200 mA (max);
case: see **Figure 5**.

As can be seen in **Figure 6**, a solder tag strip is provided on the lower side of the mounting plate for all connections. The wiring of the transceiver for 24 GHz is given in the block diagram given in **Figure 7**. **Figure 8** shows finally the internal construction.

Horn radiators or parabolic dishes can be used as antennas. The author described suitable parabolic antennas for 10 GHz and 24 GHz in (2).

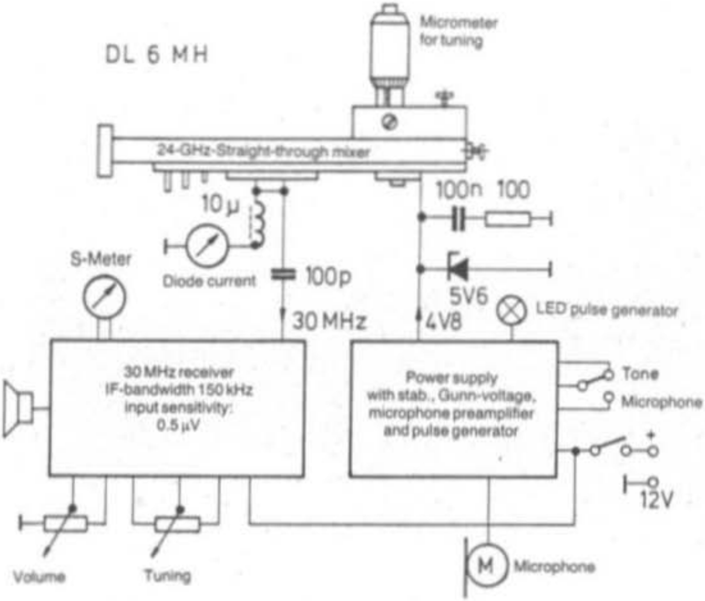



**Fig. 7: Block diagram of the 24 GHz transceiver**

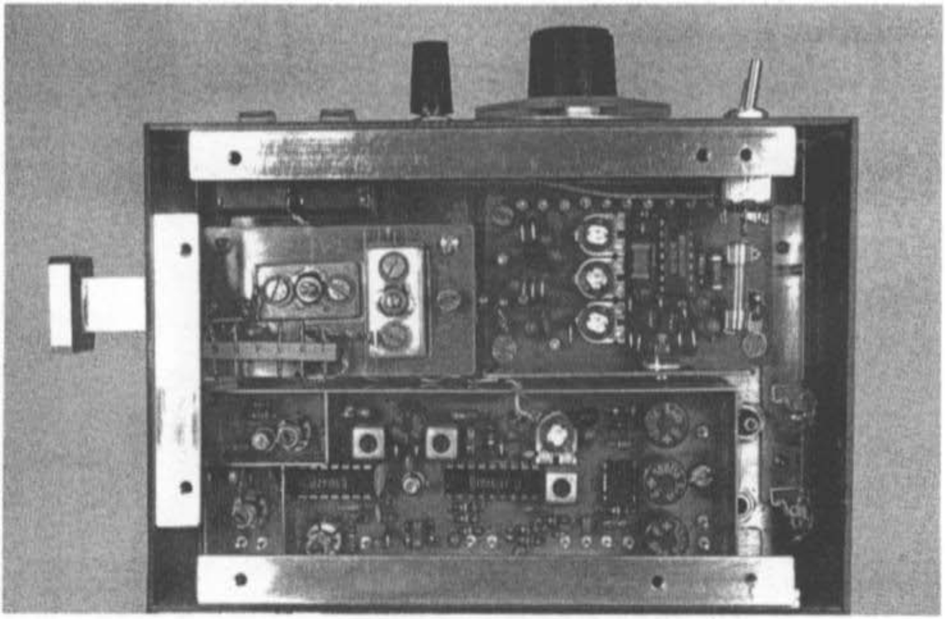


**Fig. 8: Internal view of the transceiver showing the straight-through mixer (upper left), board for voltage stabilizer, modulator, and tone generator (upper right), as well as the IF-board (below). The loudspeaker is mounted on the side panel that has been removed.**

## REFERENCES

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A Transceiver for the 10 GHz Band
VHF COMMUNICATIONS, Volume 11, Edition 4/1979, pages 208-215

(2) S. Reithofer, DL 6 MH:
Home-made Parabolic Dishes for Microwave Applications
VHF COMMUNICATIONS, Volume 12, Edition 3/1980, pages 139-145

*Leif Åsbrink, SM 5 BSZ*

# Dynamic Range of 2 m Transceivers

# Part 3: Modifications to the IC 211 and IC 245

**The introduction to this series of articles was given in Edition 1/82 of VHF COMMUNICATIONS. This explained in detail why the dynamic range of transmitters must be as great as possible, and explained this with the aid of examples. A comparison of values measured on well-known commercial 2 m transceivers showed that there was a lot to be desired in this respect. Part 2 of this series of articles, which was published in Edition 1/82, gave a number of modifications to the TS 700. In the TS 700 the problems were caused by undesired AM-modulation of the carrier. In ICOM IC 211 and IC 245, like in most other transceivers, the noise is caused by undesired phase- or frequency-modulation.**

The measured noise sidebands of a number of IC 211 and IC 245 transceivers are given in **Figure 1**. The continuous lines show the measured values before modification, and the dashed lines afterwards. Transceivers A and B were measured both before and after the described modifications. In the case of transceiver C, the VCO was replaced by a high-quality (!), home-made oscillator. The other transceivers: D, E, F, and G were only measured before or after the modification.

The VCO of the transceiver series IC 211 and IC 245 exhibit the usual weakness of commercially available transceivers: The varactor diode is fed via a 47 kΩ resistor. The output of the phase comparator is low-impedance and has a relatively low noise component. Howver, a considerable noise voltage is present at the varactor diode, which is caused by the leakage current. This voltage has a 1/f component, which causes a correspondingly varying voltage drop across this resistor.

The easiest solution to this problem is to feed the varactor diode from a low-impedance source, which can be easily achieved. It is only necessary to connect an RF-choke in parallel with the 47 kΩ resistor. After carrying out this modification, the noise sidebands will be considerably reduced, and the main component will now come from the phase comparator. In order to suppress this noise, it is necessary to build-up a passive filter as shown in **Figure 2**, and to insert it between phase comparator and VCO.

The filter shown has an output impedance of approximately 1 kΩ at 10 kHz and will completely short out the noise caused by the leakage current. It will also suppress the noise from the phase comparator sufficiently to ensure that it will have no effect on the noise sidebands.

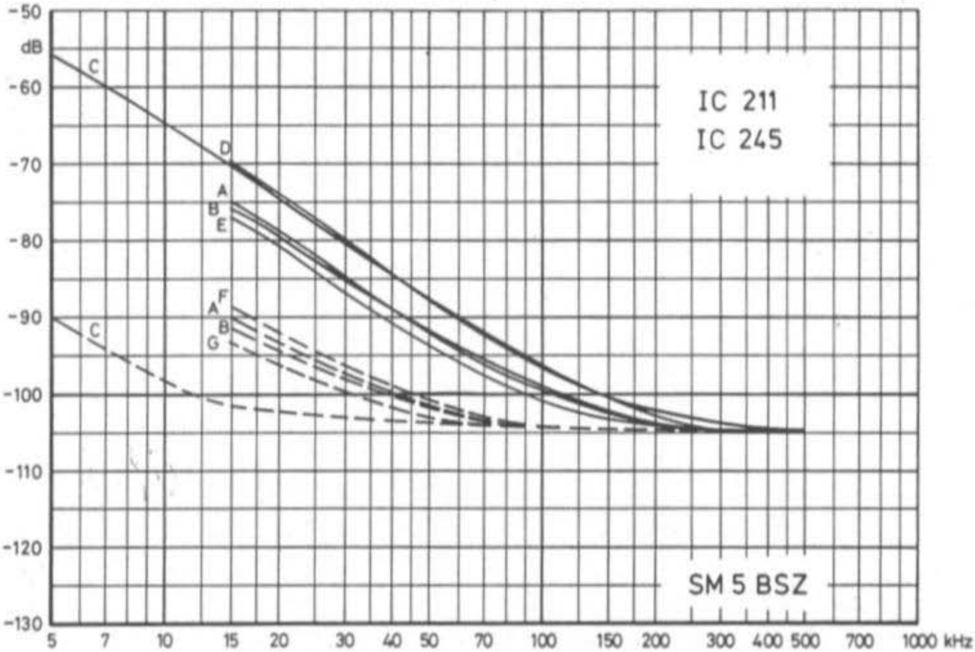



**Fig. 1: Sideband noise of various transceivers IC 211/IC 245. Measuring bandwidth: 3 kHz. Continuous lines: Original state, dashed lines: after modification**

The filter causes an additional phase shift in the control circuit which can lead to instability and poor lock-in characteristics. However, this phase shift can be compensated for by realigning the trimmer potentiometer in the active loop filter so that the control circuit locks in correctly. There are various different versions of the IC 211 and IC 245 transceivers and this potentiometer is to be found in different positions in the unit. In order to identify the correct potentiometer, one should study the circuit extract given in **Figure 3**. It is designated R 28 in this circuit, and often it is found that the best lock-in characteristics are obtained with this potentiometer adjusted to one of its limit positions. The active filter should be soldered with short connections directly to the VCO-module, as can be seen in **Figure 4**.

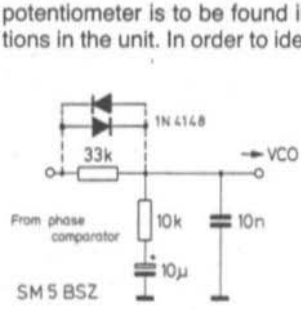



**Fig. 2:**
**This filter should be installed between phase comparator and VCO**

The VCO of many transceivers will have been modified already by the manufacturer. This is in the form of a RC-network of 470 Ω/1 nF, which will have been inserted into the source-circuit of the oscillator transistor. These two components are accommodated on the conductor side of the PC-board, after breaking the required conductor lane. **Figure 5** shows this ICOM-modification, as well as the modification recommended by the author, in the form of a circuit diagram.

The manufacturer's modification reduces the values of the noise sidebands by approximately 5 dB when compared with the original state.

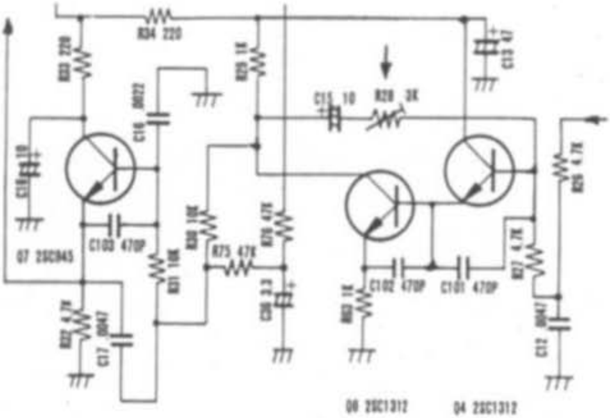


**Fig. 3:**
**The circuit extract from ICOM shows where the potentiometer is to be found that allows the phase shift to be adjusted**

This can be seen from the measured sidebands of transceivers A, B, and E in Figure 1. ICOM probably wanted to reduce the oscillator power with this modification in order to reduce the RF-voltage across the varactor diode. This in turn will reduce the leakage current through the diode, and subsequently the voltage drop across the 47 kΩ resistor. After carrying out the recommended modifications listed in this article, the ICOM-modification will no longer be required, since the voltage source of the varactor diode will exhibit such a low-impedance for lower frequencies that the 1/f component of the leakage current will not be able to cause any voltage drop.

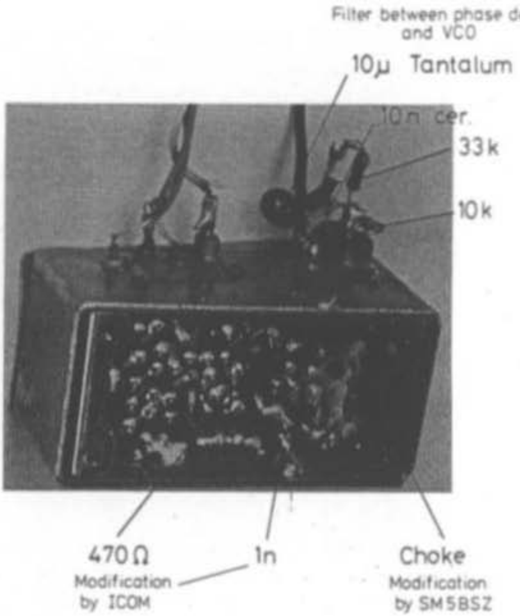



**Fig. 4:**
**This photo shows where the parts should be located**

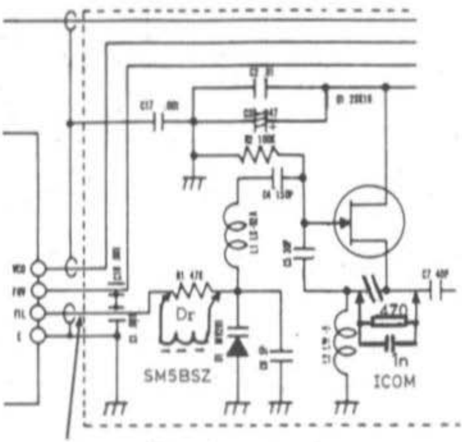



**Fig. 5:**
**This partial diagram of the VCO shows both the ICOM modification, and the modification recommended by the author**

It should be noted that two components are somewhat critical: The 10 µF capacitor should have a low leakage current, since such leakage currents can have a high noise component. The author uses tantalum electrolytics. The RF-choke should be a good VHF-type, in other words, it should have a high Q so that it does not deteriorate the Q of the resonant circuit. The inductance and capacitance values are not critical, since any detuning of the oscillator resonant circuit can be compensated for by correcting the core of L 1. A choke made from one layer of enamelled copper wire wound on a 1.5 mm diameter ferrite rod of 10 mm in length has been found to be very suitable.

The author has taken such chokes from transistorized VHF/UHF TV-tuners. A $\lambda/4$ coil would also be suitable.

A disadvantage of the recommended modifications should also be mentioned: The lock-in time is increased for large frequency variations. This will be noticed as a short delay between transmit and receive when operating split frequency operation (repeater operation).

This disadvantage can be avoided by connecting a pair of IN 4148 diodes across the 33 kΩ resistor (dashed lines in Figure 2). The author has not tested such diodes in amateur transceivers, but in a different application where fast locking was required. Such solution worked excellently. For these diodes to work, the output impedance of the phase comparator must be much smaller then 33 k, which is the case for IC 211/IC 245. It is wise to connect an oscilloscope to the comparator output when all modifications are done. The AC component of the output from the phase comparator should be below 0.5 V (peak), also when a considerable audio level is present at the built-in loudspeaker, or knocking at the transceiver to simulate mobile usage.

The RF-choke should be constructed in a mechanically stable manner, since a microphonic effect will be caused if its hot end is vibrated by a high pressure from the built-in loudspeaker, or during mobile operation. The slow reaction of the phase control will then lead to an unreliable lock-in characteristic. It is therefore advisable to glue the choke with an adhesive having good RF-characteristics (low tan delta). The adhesive should be approximately 1 mm thick. It represents the dielectric of a stray capacitance, and if it is made too thin, the high capacitance, and high electric field strength in the dielectric will cause losses and thus increased noise sidebands.

Several dozen IC 211/245 were modified according to this description in Sweden and Finland. According to my knowledge, no problems have been encountered. The average improvement at small frequency spacings from the carrier is approximately 15 dB. This means that the described modifications reduce the interfering noise sidebands by the same value as would be the case when switching off a 300 W linear amplifier and operating the station »barefoot« with only 10 W !

In the case of the IC 211, an improvement of approximately the same value will be present in the receive mode. This improvement is less pronounced in the case of the IC 245 due to its simpler input circuit of the receiver.

**Part 4 of this series of articles will describe improvements to the FT 221 transceiver.**

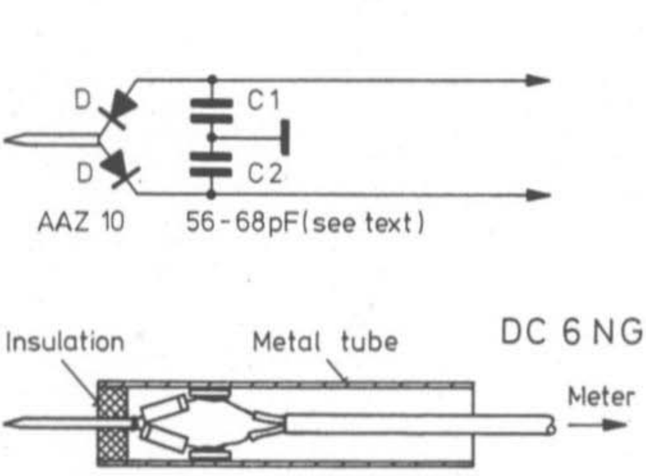
*Dr. Siegfried Behrens, DC 6 NG*

# An RF-Probe for Test and Measurement Purposes

**It is sufficient to use a simple RF-probe together with a subsequent DC-meter to establish whether an active four-pole is oscillating or not. The application range is from VLF up to UHF. If the voltmeter has a high impedance, it is possible to indicate unmodulated RF-voltages quite accurately. One great advantage is that no ground connection is required, since the hand capacity of the user is sufficient for grounding.**

The concept is based on the circuit given in (1). Further information regarding the use of semiconductors for rectification and as a meter amplifier were found in (2). However, a meter amplifier will not be necessary as long as the DC-meter has an impedance of more than 50 kΩ/V.

The simple circuit is shown in **Figure 1**. The current circuit is completed by the input impedance of the meter, which is not shown.



Germanium diodes are used to rectify the RF-voltage. These are types used for low-impedance rectifier circuits. The following types are especially suitable:

AAZ 10 (Telefunken)
AAY 27, **AA 116** (Siemens)
OA 90 (Philips)
1 N 40

These are older diode types, and one will probably find the AA 116 easiest. If a few 1 N 21 diodes are found in the drawer, then the probe can be built-up with these.

Capacitors C 1 and C 2 should preferably be chip capacitors of maximum 100 pF. This value may seem too low, but it will guarantee that the capacitors still operate as capacitors at UHF, which would no longer be the case with capacitance values in excess of 1 nF.

A suitable construction is shown in **Figure 2**. The operation at higher frequencies depends on a careful, and stable construction and installation of the probe tip, and by maintaining the shortest connections to the diodes, and from these to the disc capacitors. All other parts are uncritical. Do not forget to use some means of fixing the cable in the probe so that the connections are not broken on pulling.

The author was able to measure RF-voltages of between 30 kHz and 430 MHz with an error of only 10 %, even though construction was not perfect. Unfortunately, defined voltages of higher frequencies were not available, and it was not possible to determine the upper frequency limit.

## REFERENCES

(1) Information on the Siemens HF-Multizet

(2) Information on the Resonance Frequency Meter WAM (Rohde u. Schwarz)

*Friedrich Krug, DJ 3 RV*

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

# Part 2: Matching Stage for the Crystal Filter

**Part 1 of the article was published in edition 4/81 of VHF COMMUNICATIONS and described the concept of the receiver, with characteristics and possibilities of extension, as well as the modules of the input circuit. The first module to be described in detail is the matching stage for the crystal filters. This is a very important stage for determining the dynamic range of the receiver, and for this reason it is to be described in more detail; a large number of measured values are to be given for various types of transistors and crystal filters.**

## 3. IF-AMPLIFIER

The complete IF-module comprises a number of individual stages, which are given in **Fig.4** in the form of a block diagram. The circuits are divided onto four PC- boards and screened in conventional metal boxes. This allows the IF-module to be used in a versatile manner, since the individual PC-boards can be used separately as required.

The first PC-board contains the matching amplifier, crystal filters for the various bandwidths, as well as a low-noise amplifier, which can be provided with a Notch filter or a further IF-filter for improving the ultimate selectivity.

The second board contains the variable IF-amplifier, the control voltage generator, and the output coupling stages.

The third board contains the demodulators and an AF-amplifier; and the fourth board contains the oscillators for the BFO, as well as the auxiliary oscillators for the alignment of the IF-module.

The overall circuit, and component location plans for the boards will be given in Section 4 of this article, whereas Section 3 is to describe the individual stages and their operation so as to obtain a better understanding of the circuit. A large number of different circuits were examined, and the most favourable circuit with the best reproducibility (when using amateur measuring equipment) was then selected for the application. The demands given in Section 1 were to be fulfilled as closely as possible.

## 3.1. MATCHING STAGE FOR THE CRYSTAL FILTER

In the case of a Superhet receiver as shown in Figure 1, the required IF-signal is filtered out from the signal spectrum at the output of the mixer with the aid of a crystal filter, which provides the main selectivity. The signal is subse-

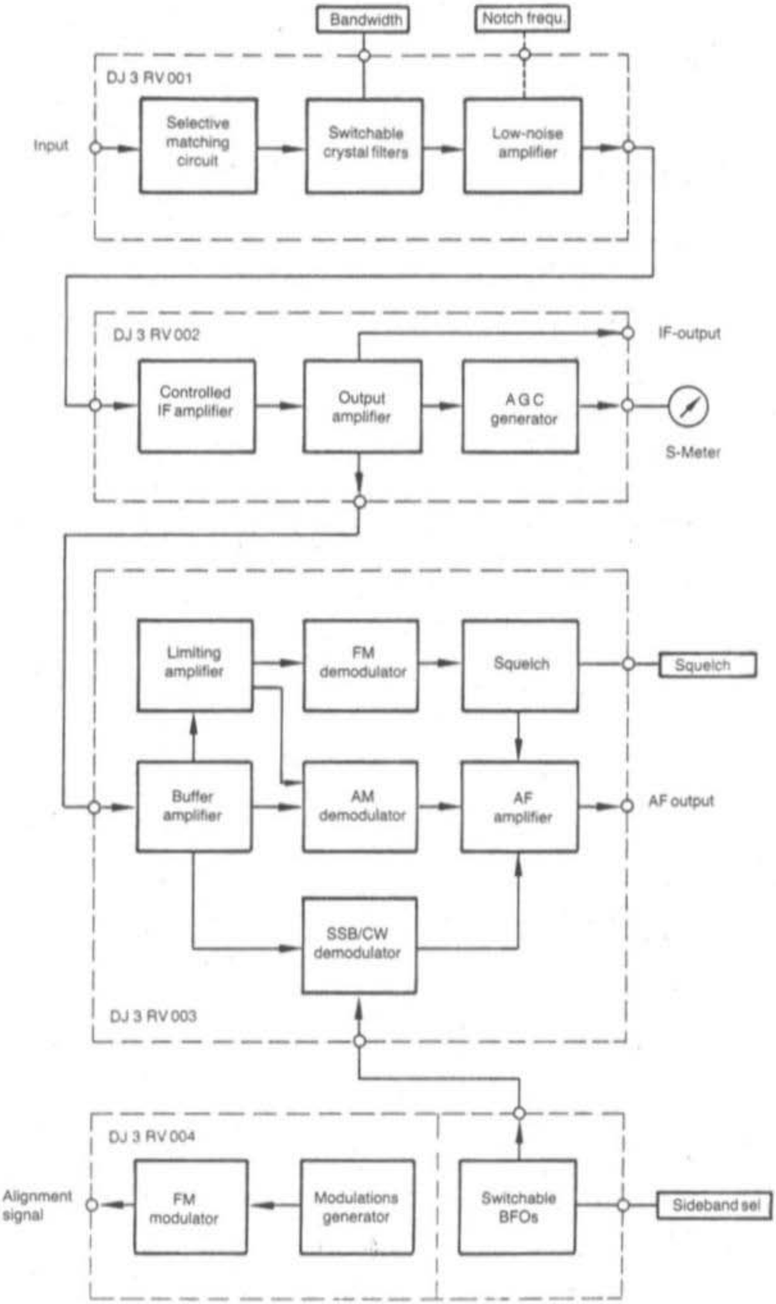



**Fig. 4: Block diagram of the IF-module**

quently amplified in the IF-module. The crystal filter has, however, only a defined impedance within the passband range, whereas a low-intermodulation mixer requires a wideband ohmic termination – usually 50 Ω. For this reason, a matching amplifier is required between mixer and filter, which must have a frequency-independent input impedance and whose output impedance must match the impedance of the filter.

The input impedance of the amplifier must be as independent as possible of the output impedance, which is mainly given by the input impedance of the filter. This can only be achieved using low-reactive circuits, which means that feedback circuits such as used in the input circuit (Figure 3) cannot be used. Furthermore, the amplifier and the filter determine the overall intermodulation and noise characteristics of the whole IF-module, which means that only low-intermodulation, and low-noise components can be used.

These problems were described very extensively by DJ 7 VY in (10); further information was given by DL 7 AV in (11) and also in (12).

### 3.1.1. Dynamic Range

According to the demands given in Section 1, the IF-module should process signals within a dynamic range of 100 dB without intermodulation, and be able to handle a maximum input voltage of $U_{in\,max}$ = 50 mV without limiting. The latter corresponds to an input power of $P_{in\,max}$ = – 13 dBm.

The intermodulation-free dynamic range ID is considered to be the input level range that is between the noise floor and the input power $P_{in}$ at which the intermodulation products are equal to the noise power $P_n$.

$$ID = P_{in} - P_n$$

For those readers that only know noise to be something that comes out of the loudspeaker or headphones, but still want to understand the following article, it is suggested that they read through the DL 3 WR article in (13).

In the case of a minimum input voltage of $U_{in\,min}$ = 0.5 µV for a usual signal-to-noise ratio of 10 dB used in amateur communications, the noise floor will be as follows when referred to the noise voltage at the input:

$$U_n = \frac{U_{in\,min}}{\sqrt{10}} = 0.16\,\mu V$$

This corresponds to a noise power $P_n$ across 50 Ω of:

$$P_n = \frac{U_n^2}{50\,\Omega} = 5 \times 10^{-16}\,W \triangleq -123\,dBm.$$

This means that the maximum permissible noise figure NF for the IF-module at the largest bandwidth of 15 kHz will be

$$NF = \frac{P_n}{kTB} = \frac{5 \times 10^{-16}}{4 \times 10^{-21} \times 15000} = 8.33 \triangleq 9.2\ dB.$$

Since the IF-module is designed to follow a passive mixer, the noise figure should be even lower.

The largest input power $P_{in}$ at which the intermodulation rejection amounts to IM = 100 dB for the maximum permissible noise figure of NF = 9.2 dB, can be calculated according to the following equation:

$$P_{in} = P_n + IM = -123\,dBm + 100\,dB = -23\,dBm.$$

The third-order intercept point referred to the input of the IF-module must then be at least:

$$IP = 0.5\,IM + P_{in} = 0.5 \times 100\,dB - 23\,dBm = +27\,dBm.$$

Unfortunately, this intercept point could not be achieved using the matching stage in the input circuit of DJ 7 VY. An IP of 19 dBm and a noise figure of NF = 4.4 dB were measured when using the FET P 8002. In order to obtain better specifications, various matching circuits were examined systematically.

In principle, an input stage of an IF-module consists of bandpass filter, amplifier, and crystal filter. This is shown in the block diagram given in **Figure 5**. The attenuation or gain $G_p$, noise figure NF, and the intercept point IP

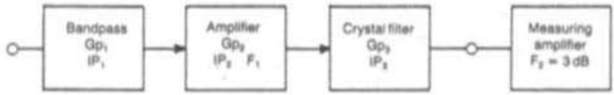



**Fig. 5: Block diagram of the matching circuit**

were examined at the individual components, and in conjunction with the circuits and subsequently compared.

From the individual specifications, it is possible to obtain an approximation of the overall values of the noise figure $NF_{tot}$, and the intercept point $IP_{tot}$, referred to the input of the IF-amplifier, which are then valid for the intermodulation-free dynamic range ID.

These result from the connection of the individual elements as shown in Figure 5 and can be calculated according to the following equations from (12) and (13). The attenuation of the bandpass coupling ($Gp_1$), the gain of the stage ($Gp_2$), and the attenuation of the crystal filter ($Gp_3$) have been considered. A noise figure $NF_2 = 2 \mathrel{\hat{=}} 3$ dB is assumed for the amplifier following the crystal filter.

The following is valid for the total noise figure $NF_{tot}$:

$$NF_{tot} = \frac{1}{Gp_1} + \frac{NF_1 - 1}{Gp_1} + \frac{\frac{1}{Gp_3} - 1}{Gp_1 \times Gp_2} + \frac{NF_2 - 1}{Gp_1 \times Gp_2 \times Gp_3}$$

For the calculation of the dynamic range, it is advisable to calculate the noise power $P_n$ referred to the input in dBm:

$$P_n = 10 \lg \frac{NF_{tot} \times B \times kT}{1 \text{ mW}}$$

$$= 10 \lg (4 \times 10^{-18} \times NF_{tot} \times B)$$

$NF_{tot}$ = Noise figure
B = Bandwidth of the filter in Hz
K = Boltzmann constant
= $1.38 \times 10^{-23}$ Ws/K
T = Ambient temperature in Kelvin (K)

The following is valid for the intercept point $IP_{tot}$:

$$IP_{tot} = -10 \lg (10^{-IP_1/10} + 10^{(Gp_1 - IP_2)/10} + 10^{(GP_1 + GP_2 - IP_3)/10})$$

Since the intercept point is always given in dBm, all values should also be in dB in the last equation.

The intermodulation-free dynamic range ID then results as:

$$ID = \frac{2}{3} (IP_{tot} - P_n)$$

directly in dB.

### 3.1.2. Bandpass Filter

The matching amplifier should only be driven by signals in the IF-range. A filter circuit that also provides a wideband mixer termination is in the form of a bandpass filter and was calculated in (14). This bandpass filter [also used in (6)] is given in **Figure 6**. It only allows signals in the IF-range to be passed to the subsequent amplifier. All other signals such as the image and the residual oscillator frequency at the output of the mixer will be suppressed.

In an experimental circuit constructed by the author, a residual oscillator signal of − 14 dBm was measured at the IF-output of the mixer SRA-1H at an oscillator power of + 17 dBm, which corresponds to an attenuation of only 31 dB. The bandpass filter will attenuate this signal by approximately 30 dB, whereas the attenuation of the required signal only amounts to $Gp_1 = 0.87 \mathrel{\hat{=}} -0.6$ dB.

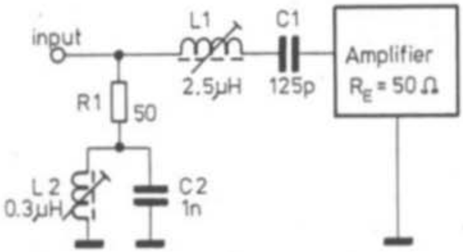



**Fig. 6:**
**Bandpass filter as termination for a ring mixer**

This insertion loss will deteriorate the noise figure, whereas inductances L 1 and L 2 will deteriorate the intercept point. The inductances are constructed using the special coil set available from the publishers. According to the position of the core in the coil, an intercept point of $IP_1$ = 42 to 44 dBm resulted. Experiments made with other ferrite materials did not provide any better values. This shows the importance of the non-linear behaviour of ferrite materials for the IP.

### 3.1.3. Amplifier

The noise of the amplifier should be as low as possible, and the gain of the amplifier should be such that the filter loss and noise figure of the subsequent stages have little effect on the overall noise figure. Furthermore, the intercept point should be as high as possible. Since all amplifier elements possess a non-linear behaviour, a poorer intermodulation rejection will result at high-gain levels due to the higher output amplitude, in other words, the result will be a lower IP.

The higher output amplitude means that the subsequent filter will be driven at higher level, which means that the IP of the filter can no longer be neglected. The IP of the filter should always be greater than or equal to the IP of the amplifier referred to its output.

$IP_3 \geq IP_2 + Gp_2$ (values in dB)

For the comparison measurements, the amplifiers were operated without filter and terminated with the ohmic value of the nominal filter impedance $Z_L$.

Various high-current FETs, and the FET P 8002 also in push-pull were examined for the amplifier stage. The circuits are given in **Figure 7**. The measured values for the intercept point IP referred to the input, and the noise figure are given in **Table 2**.

The values for the IP were measured on several examples using two signals having a level of – 10 dBm each, and the spread is given.

Power transistors with better intermodulation behaviour have, however, usually poorer noise figures, and were not examined here.

The input impedance $Z_{in}$ of the circuits should amount to 50 Ω and can be selected using the transformation ratio tr of the input transformer and the transadmittance slope $y_{21}$.

$$Z_{in} = \frac{tr^2}{y_{21}}$$

For transistors with $y_{21}$ = 20 mS such as the P 8002 and CP 643, no input transformer will be required.

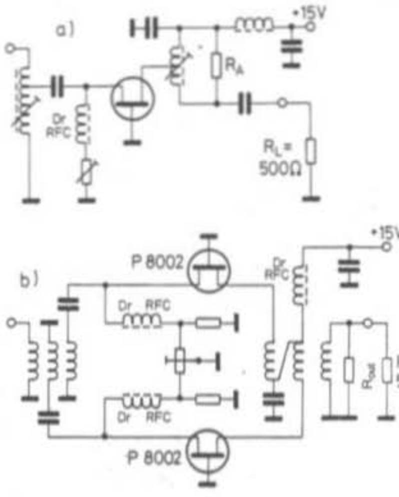



**Fig. 7:**
**Measuring circuits for matching amplifiers**

| Components | Noise Fig.NF$_1$ | IP |
|---|---|---|
| P 8002 circuit a | 2 ≙ 3 dB | 26 - 28 dBm |
| 2 x 8002 push pull circuit a | 2.3 ≙ 3.6 dB | 28 - 31 dBm |
| CP 643 circuit a | 2.5 ≙ 4 dB | 25 - 28 dBm |
| CP 640 circuit a | 2 ≙ 3 dB | 29 - 31 dBm |
| P 8002 circuit b | 2.2 ≙ 3.4 dB | 29 - 31 dBm |

**Table 2**

The transadmittance slope $y_{21}$ is dependent on the drain current and has a relatively large spread in the case of FETs. The author measured the drain current required for $y_{21}$ = 20 mS at 9 MHz and a drain-gate voltage of $U_{DG}$ = + 15 V on 20 different P 8002 transistors. The values spread from $I_D$ = 19 mA to $I_D$ = 36 mA, whereas the majority was in the order of 26 mA.

In the case of the push-pull circuit, it is necessary for the transistors to be selected, since the measured values of the intermodulation rejection can only be obtained when using two transistors having identical dynamic behaviour. Otherwise, the obtained values will be only that obtained when using individual transistors.

All amplifiers were designed to have the same output power during these comparative measurements.

The power gain is given by the transformation to the resistance $R_{out}$, and the parallel-connected input resistance $R_L$ of the IF-filter. The resistance $R_{out}$ is the source resistance required for matching the filter. It must be included, because the output resistance of the transistor is far higher than the nominal impedance of the filter.

The following is valid for matching: $R_{out} = R_L$.

The filters used exhibit various impedances within the passband range. For this reason, the output resistance of the circuit was designed for the lowest filter impedance Z = 500 Ω II 30 pF.

The filters with higher impedance are matched with the aid of impedance-converter circuits. In the case of a load resistance of $R_L$ = 500 Ω, a transistor P 8002 will exhibit a power gain of 7 dB. The power fed to the filter is 3 dB lower than the output power of the transistor, since half the power is consumed in the source resistance $R_{out}$. This means that the gain per stage only amounts to $Gp_2$ = 4 dB.

The diagrams given in **Figure 8** show that a higher gain will deteriorate the intercept point. The IP referred to the input, and the $IP_{out}$ referred to the output, as well as the gain-per-stage Gp are given as the function of the collector resistance R. The collector resistance is:

$$R = \frac{R_{out} \times R_L}{R_{out} + R_L} = \frac{R_L}{2}$$

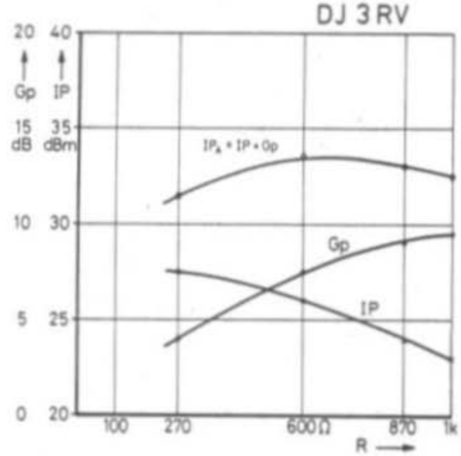
since matching of $R_{out} = R_L$ is assumed.

Measurements were made on transistors P 8002, CP 640, and with two P 8002 connected in parallel in circuit a.

The push-pull circuit was not considered in these tests and in the subsequent considerations because very few amateurs have access to measuring systems to select the parameters of the transistors.

The parallel-connected transistors were selected to have the same gate-source voltage $U_{GS}$ at a drain current $I_D$ of 25 mA. This measurement can be carried out by most constructors.

Two components that are important in obtaining a good intermodulation rejection are the input transformer Tr 1, and the chokes. Transformer Tr 1 must be in the form of a wideband transformer with 3 x 8 turns of enamelled copper wire of 0.35 mm diameter wound on a 6 to 10 mm toroid core from highly permeable, lossy ferrite material. Suitable material is, for instance, N 30 (Siemens) or 3 D 3 (Philips).



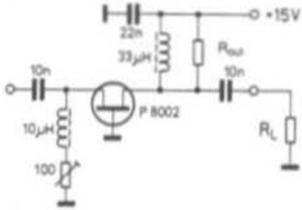
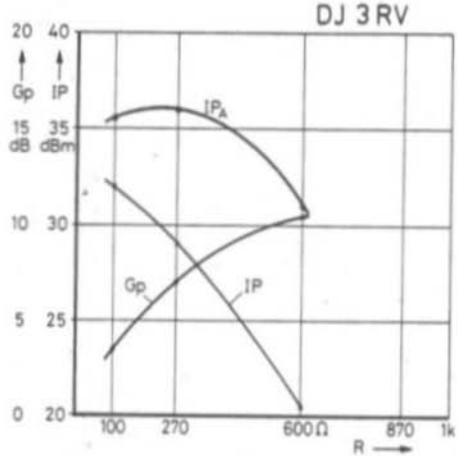




**Fig. 8a:**
**Intercept point and gain per stage for the P 8002**



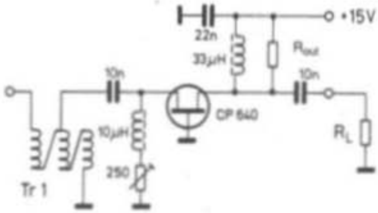
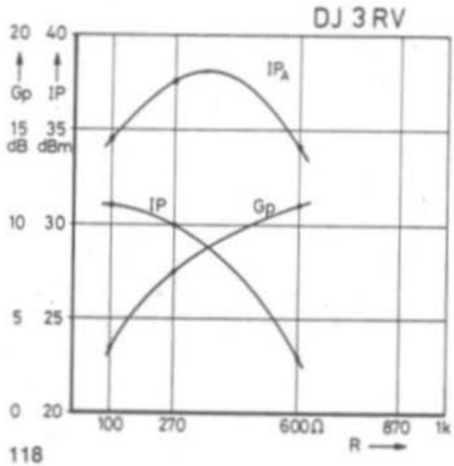




**Fig. 8b:**
**Intercept point and gain per stage for CP 640**



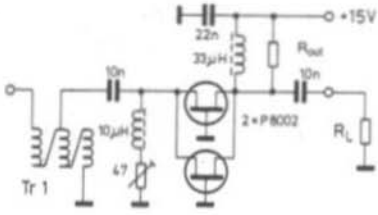




**Fig. 8c:**
**Intercept point and gain per stage for 2 x P 8002**

Transformers using higher-Q material are not suitable since their permeability varies considerably with the degree of magnetization. For the same reason, fixed inductances wound on a ferrite core are used for the chokes.

### 3.1.4 Filter

Four different filter bandwidths are planned for the IF-module. Monolithic crystal filters are used:

XFM – 9S05 with 12 kHz bandwidth
XFM – 9S01 with 5 kHz bandwidth
XFM – 9S03 with 2.4 kHz bandwidth

as well as the conventional crystal filter:

XF – 9 NB with 500 Hz bandwidth.

Since a large number of readers will have filters available, and may wish to use these in the IF-amplifier, the author carried out measurements on available conventional crystal filters. Of interest are the insertion loss $Gp_3$ and the intercept point $IP_3$ referred to the input of the filters.

The insertion loss of the filter specified by the manufacturer can be measured from the input and output voltage of the filter as

$$Gp = 20 \lg \frac{U_{out}}{U_{in}}$$

However, most filters require a transformation circuit for matching it to the amplifier, which causes additional losses. This must be taken into consideration when calculating the insertion loss $Gp_3$. For this reason, $Gp_3$ was calculated from the power ratio at the output of the filter to power $P_{in}$ at the measuring input.

$$Gp_3 = 10 \lg \frac{U_{out}^2}{R_F \times P_{in}}$$

where $R_F$ is the real part of the filter impedance.

The IP of the filters was measured in the test circuit given in **Figure 9** using signals in the stopband range above and below the passband frequency. The passband curves of the filter were aligned for the lowest possible ripple by adjusting L 3, C 3, and C 4. The nominal impedance of 500 Ω for the filter is obtained at the input with R 1, R 2, and C 3, or with the transformation of L 3 and C 3 in the case of filters with a nominal impedance of more than 500 Ω.

The following is valid according to the KVG-catalogue:

$$L\,3 = \frac{\sqrt{R_{in}\,(R_F - R_{in})}}{2\pi f}$$

$$C\,3 = \frac{\sqrt{\frac{R_F}{R_{in}} - 1}}{2\pi f \times R_F}$$

with a filter impedance $R_F$ // $C_F$.

The output of the filter is connected to a parallel LC-circuit comprising L 4 and C 4, with which the reactive component of the filter termination can be realized. The ohmic component is mainly obtained using resistors R 3 and R 4. This circuit also allows the unwanted circuit capacitances and the input capacitance of the source follower to be compensated for. The source follower operates as impedance converter to the required impedance of 50 Ω. The measured values are given in **Table 3**.

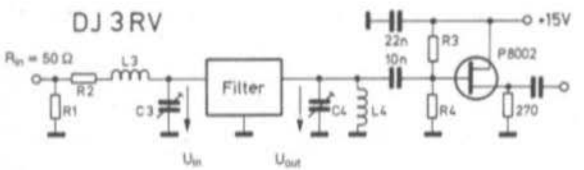



**Fig. 9: Measuring circuit to measure the IP of the filter**

| Filter | Nominal impedance | R 1 Ω | R 2 Ω | L 3 µH | C 3 pF | C 4 pF | L 4 µH | R 3 = R 4 Ω | $P_E$ dBm | $IP_3$ below | $IP_3$ above | $Gp_3$ dB |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| XF-910 | 6 kΩ | – | – | ~ 10 | 5-40 | 3-13 | 22 | 12000 | 2 x 0 | 43 dBm | 42 dBm | – 3 |
| XFM-9S 05 | 8,2 kΩ | – | – | ~ 10 | 5-40 | 3-13 | 22 | 18000 | 2 x 0 | 42 dBm | 43 dBm | – 6 |
| XFM-9S01 | 3,3 kΩ ‖ 2 pF | – | – | ~ 7 | 10-60 | 3-13 | 22 | 6800 | 2 x 0 | 40 dBm | 40 dBm | – 5 |
| XFM-9S03 | 1,8 kΩ ‖ 3 pF | – | – | ~ 5 | 10-60 | 3-13 | 22 | 3900 | 2 x 0 | (37 dBm) | (38 dBm) | – 6 |
| XF-9NB | 500 Ω ‖ 30 pF | 56 | 470 | – | 5-40 | 10-60 | 10 | 1000 | 2 x -16 | 34 dBm | 36 dBm | – 6 |
| XF-9B | 500 Ω ‖ 30 pF | 56 | 470 | – | 5-40 | 10-60 | 10 | 1000 | 2 x -16 | 36 dBm | 34 dBm | – 3 |
| XF-9D | 500 Ω ‖ 30 pF | 56 | 470 | – | 5-40 | 10-60 | 10 | 1000 | 2 x -16 | 35 dBm | 32 dBm | – 3 |
| XF-9E | 1,2 kΩ ‖ 30 pF | – | – | ~ 4 | 10-60 | 10-60 | 22 | 2700 | 2 x 0 | 35 dBm | 36 dBm | – 6 |

**Table 3:**
**IP values of crystal filters. Type XFM-S03 exhibited too high ripple values.**

The IP is extremely dependent on the ferrite material used for L 3. The same coil set was used for this inductance as used in the band-pass filter.

Since it is difficult to define the IP of the filter, the measuring method used is to be described in somewhat more detail.

Those signals that cause unwanted intermodulation products in a receiver are outside the passband range of the filter. If a good RF-preselection is assumed, the interference signals will only appear in the vicinity of the intermediate frequency and will be third-order intermodulation products. Second-order intermodulation products can be neglected.

This was described extensively by DL 1 BU in (15).

The experiments were therefore made with signals 40 kHz and 80 kHz below or above the center frequency of the filter. In the case of the 9 MHz filter, these frequencies are $f_1$ = 8.960 MHz, and $f_2$ = 8.920 MHz for the signals below the center frequency, and $f_1$ = 9.040 MHz, and $f_2$ = 9.080 MHz for the frequencies in excess of the center frequency. The third-order intermodulation product ($2 f_1 - f_2$) will then fall within the passband range of the filter. This can be measured easily as interference power $P_i$ at the output of the measuring circuit, since the original test signals will be suppressed in the filter.

The impedance is not defined in the stopband range of the filter. As approximation, it can be assumed that the filter impedance is very high with respect to the nominal impedance of the filter for the test signals. Since the amplifier in the receiver, or the generators in the measuring set-up are matched to the nominal impedance of the filter, a virtually open circuit will be present, and the test signal voltages $U_{in}$ will be greater than under matched condition. The actual test signal power into the filter is thus very low due to the high impedance.

The power present at the filter in the case of matching was therefore used as reference power $P_{in}$ for definition of the IP. The intermodulation rejection IM is therefore defined as the ratio of the power at output $P_{out}$ of the measuring circuit for a signal having the reference level in the passband range to interference power level $P_i$.

$$IM = P_{out} - P_i \quad \text{(values in dBm)}$$

The IP of the filter is thus

$$IP = \frac{1}{2} IM + P_{in} \quad \text{(values in dBm)}$$

As can be seen from the values given in Tables 2 and 3, a dynamic range of 100 dB can be achieved. With $R_{out}$ = 560 Ω, the following values will result for the circuit in Fig. 8c:

Gain per stage $Gp_2$ = 5 ≙ 7 dB, and an $IP_2$ of 30 dBm.

The following equation is valid for calculating the noise figure of the input stage with a filter loss $Gp_3 = 0.25 \triangleq -6$ dB and an $IP_3$ of 40 dBm, when using the equations given in Section 3.1.1.:

$$NF_{tot} = \frac{1}{0.87} + \frac{2.3-1}{0.87} + \frac{4-1}{0.87 \times 5} + \frac{2-1}{0.87 \times 5 \times 0.25} = 4.25 \triangleq 6.3\ \text{dB}$$

The noise power at the maximum bandwidth of 12 kHz is thus:

$$P_n = 10\ \lg\ (4 \times 10^{-18} \times 4.25 \times 12000) = -126.9\ \text{dBm}$$

and the intercept point:

$$IP_{tot} = -10\ \lg\ (10^{-44/10} + 10^{(-0.6-30)/10} + 10^{(-0.6+7-40)/10}) = 28.7\ \text{dBm}$$

This results in a theoretical intermodulation-free dynamic range of

$$ID = \frac{2}{3}\ (28.7 + 126.9) = 103.7\ \text{dB}$$

The following IP-values were measured with crystal filters XF-9NB and XFM-9S05 at a test level of 2 x − 10 dBm:

| Type of filter | $IP_{tot}$/dBm below | $IP_{tot}$/dBm above |
|---|---|---|
| XF-9NB | 26 | 27.5 |
| XFM-9S05 | 29 | 29 |

The measurement of the noise figure was not carried out, since this is only worthwile when made on the complete IF-module.

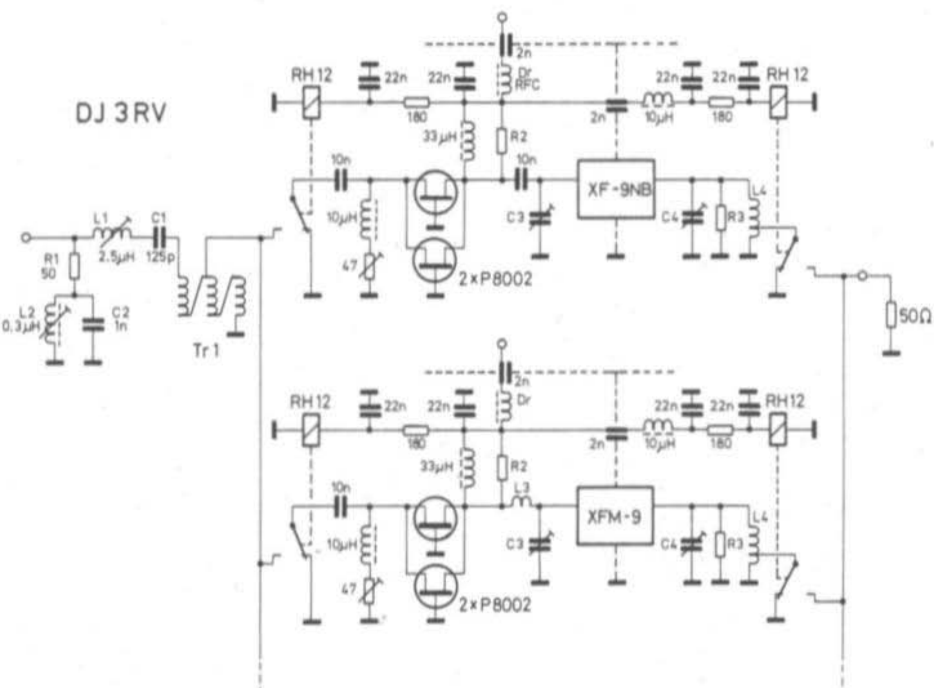



Fig. 10: **Bandwidth switching (only given for 2 channels)**

## 3.2. CRYSTAL FILTER WITH SWITCHING

The matching amplifier is followed by the crystal filters. The load impedance of the amplifier is fixed at $R_L = 500\ \Omega$, which means that the monolithic filters must be matched using a transformation circuit comprising L 3 and C 3.

Diode switches comprising the switching diode BA 283 and relays RH-12 were examined for switching the filters. If the diodes are driven with current values in excess of 10 mA in switched condition, and blocking voltages in excess of 10 V in blocked condition, no deterioration of the IP can be measured when using these switches. Since diode switches have less capacitance, and are far smaller, they were originally used. However, it was found in the final prototype that the decoupling between the switching lines was not sufficient. The additional circuitry required for improving this would be too extensive, and for this reason relays were chosen as the simplest solution to the problem. For this reason, a new development was required, which was achieved using relays.

The relays possess a relatively high capacitance of 4 pF when in their rest position, and 9 pF when energized. Since four relays are connected in parallel to the source resistance $R_{out}$, the inavoidable circuit capacitance provides an additional capacitive load of approximately 35 pF. This means that it is not possible to align all filter types to an optimum passband curve.

An additional impedance transformation of the amplifier output to a lower value, or the compensation of the capacitances using $\pi$-links will deteriorate the IP too greatly. The best results will be obtained when the complete filter and matching amplifier are switched as shown in **Figure 10**. The different insertion losses of the filter can be compensated for using R 3 and the tap on L 4, which allows the overall gain to be independent of the selected bandwidth.

A low-noise amplifier is required after the filters, which should not be overdriven by signals of 50 mV at the input of the matching stage, if it is to conform to the demands given in Sect. 1.

An output coupling such as used in the measuring circuit (Figure 9) does not exhibit sufficiently low noise characteristics. One could assume that a dual-gate MOSFET similar to the recommendation of DJ 7 VY in (6) could be used here. However, the disadvantage of this is that the filter module could not be used as an independent module due to the required control voltage. Furthermore, the dual-gate MOSFET would be overdriven in controlled condition in excess of 1 $V_{pp}$.

If one considers the gain-per-stage, the matching transformation, and filter attenuation, an input voltage of 50 mV will result in a voltage of 700 mV when using the filter XFM-9S05 into the nominal terminating impedance; this would be approx. 2 $V_{pp}$ at gate 1 of the FET ! Such levels cannot be processed from low-noise, controllable DG-FETs such as 3N211 or BF 910 in controlled condition without limiting.

Since in-band signals with 50 mV subsequent to the mixer are seldom – this would correspond to 100 mV ≙ – 7 dBm (!) at the input of the mixer assuming a conversion loss of approx. 6 dB – it is possible for a FET to be used. The advantage is the simple gain adjustment. If the FET is not connected to the control voltage, the gain-per-stage can be selected with the aid of the bias voltage at gate 2, so that the differing attenuation of the filter can be compensated for. The circuit is given in **Figure 11**.

The solution with the largest dynamic range is a transformation of the filter output impedance to approximately 50 Ω and subsequently to use a low-noise amplifier equipped with a transistor type BFQ 69 with slight feedback. The signal can then be directly taken from this stage with an impedance of 50 Ω and fed to the variable IF-amplifier.

In order to allow the first module to be used independently, it is necessary to improve the ultimate selectivity using another filter so that it is greater than the dynamic range. The filter XF-

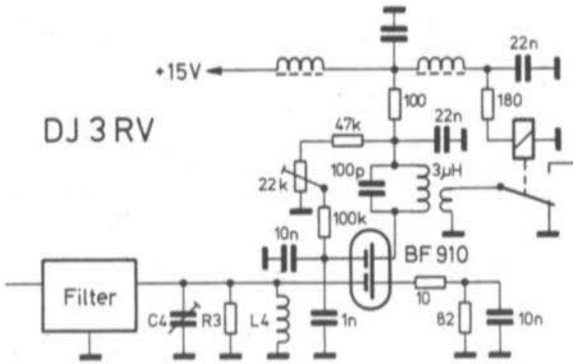



**Fig. 11:**
**Output coupling of the filter using a BF 910**

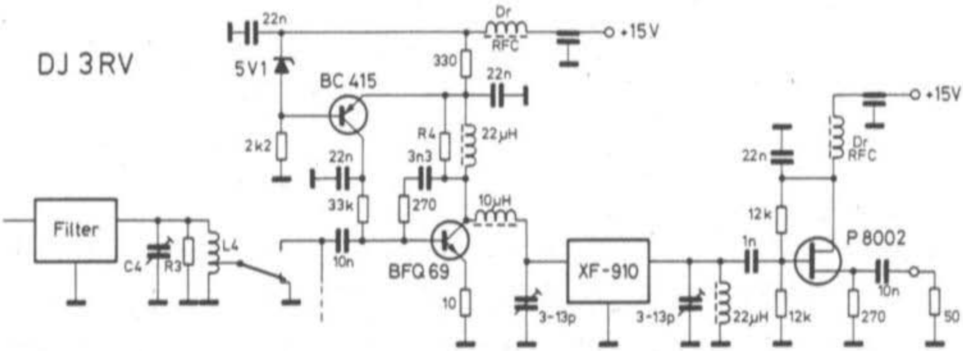



**Fig. 12a:**
**Output coupling of the filter using a BFQ 69**

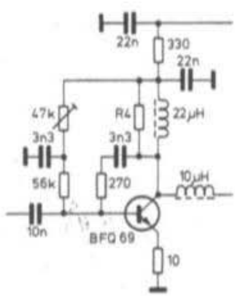



**Fig. 12b:**
**Simple operating point adjustment for the BFQ 69**

910 is sufficient for this. The circuit is given in **Figure 12**.

A Notch filter could be used instead of a crystal filter and such a filter is to be described later. However, this is only advisable at the lowest bandwidth, since the crystal used in the filter cannot be pulled over a wide frequency range.

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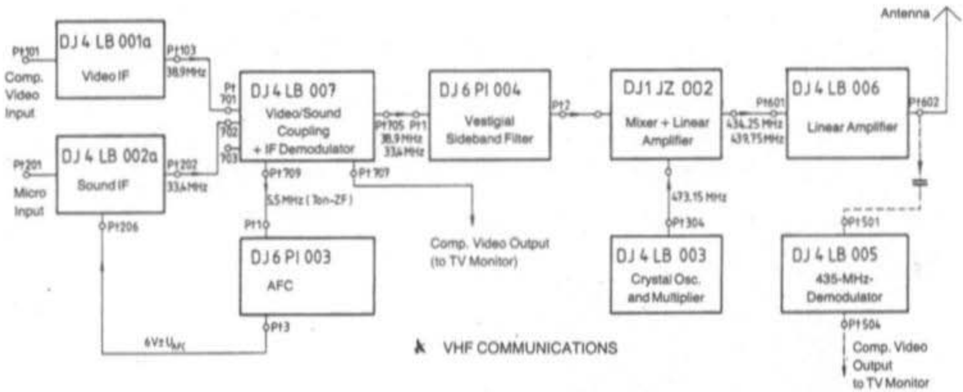
# An Amateur-Television Transmitter for Home Construction




A television transmitter built from modules described in VHF COMMUNICATIONS is shown in the above **block diagram**. Each function is realised on an individual PC-board. Each PC-board is built into its own tinned-metal box, which leads to a very clean operation without unwanted stray coupling and without problems caused by radiation. Each module may be aligned and tested on its own. All this encourages the home constructor since it makes it easy to understand the different functions, and it finally leads to a high-value ATV transmitter to which all possible video sources (black/white or color) may be connected.

For an amplification of the transmit power, a variety of linear amplifiers for the 70 cm band may be used (not FM »linears« !), whereby care should be taken to adjust the drive so that the output power does not exceed half the PEP value of the SSB mode.

The ATV modules listed have been published by three authors. The descriptions are detailed and will enable successful duplication. They are to be found in the following **editions of VHF COMMUNICATIONS:**

| | | | |
|---|---|---|---|
| VHF COMMUNICATIONS | 1/1973 | VHF COMMUNICATIONS | 4/1977 |
| VHF COMMUNICATIONS | 2/1973 | VHF COMMUNICATIONS | 3/1981 |
| VHF COMMUNICATIONS | 2/1976 | **This set of 6 editions is available** | |
| VHF COMMUNICATIONS | 1/1977 | **at DM 24.—** | |

**Individual kits:**

| | | | | | |
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| DJ 4 LB 002a | kit, complete | DM 99.— | DJ 4 LB 003 | kit, complete | DM 92.— |
| DJ 4 LB 007 | kit, complete | DM 90.— | DJ 1 JZ 002 | kit, complete | DM 131.50 |
| DJ 6 PI 003 | kit, complete | DM 50.— | | | |

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Tel. West Germany 9133-855. Representatives see cover page 2

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

In VHF COMMUNICATIONS, a system for reception and display of weather satellite images was described by Rudy Tellert, DC 3 NT, which can be used for the geostationary satellites (METEOSAT/GOES) as well as for the polar-orbiting weather satellites of the NOAA and METEOR families. Images are displayed either on a FAX machine or on a CRT screen in combination with a camera. Since other authors contributed to the system, it may be helpful to our readers and prospective customers to have a list showing the individual parts and modules required to complete a system suiting the requirements of the customer.

| Part in the system | Described in Edition | Kit designation | Kit DM | Total DM |
|---|---|---|---|---|
| 1. **Parabolic antenna,** 1.1 m diam., 12 segments to be screwed or riveted together | 3/1979 | **Parabolic antenna kit** | 180.— | |
| | | **Riveting machine + rivets** | 93.— | |
| | | **1.7 GHz Cavity radiator kit** | 90.— | |
| | | | **Total:** | 363,— |
| | | (ready-to-operate: DM 150.—) or with radiator ready-to-operate: | | 423,— |
| 2. **Low-noise amplifier for 1.7 GHz** (Originally described for use at 2.4 GHz, this unit is tuned to 1.7 GHz) | 1/1980 | **DJ6PI 010** | 225.— | 225.— |
| 3. **METEOSAT Converter,** consisting of two modules (Output: first IF = 137.5 MHz) | 4/1981 | **DJ1JZ 003** | 189.— | |
| | 1/1982 | **DJ1JZ 004** | 185.— | |
| | | | | 360.— |
| 4. **VHF Receiver,** frequency range: 136-138 MHz (Output: 2400 Hz sub-carrier) | 4/1979 | **DC3NT 003** | 225.— | |
| | 1/1980 | **DC3NT 004** | 80.— | |
| | | | | 305.— |
| 5. **Control Electronic and Video Processor** | 3/1980 | **DC3NT 005** | 125.— | |
| | 2/1980 | **DC3NT 006** | 80.— | |
| | 4/1980 | **DC3NT 007** | 98.— | |
| | 2/1980 | **DC3NT 008 Kit 1, for CRT only** | 115.— | |
| | | DC3NT 008 Kit 2, CRT + FAX | 160.— | |
| | 2/1981 | **DC3NT 009 for CRT only** | 99.— | |
| | 4/1980 | **DC3NT 010 for FAX mach. only** | 98.— | |
| | 4/1980 | **DC3NT 011** | 59.— | |
| | 4/1980 | **DC3NT 012** | 115.— | |
| | 4/1980 | **DC3NT 013** | 78.— | |
| | 3/1981 | **DC3NT 014 for CRT display** | 98.— | |
| | 3/1981 | **DC3NT 015/016 for CRT only** | 478.— | |
| Combination A: Use of FAX machine only<br>DC3NT 005-008 (Kit 2)<br>DC3NT 010-013 | 1/1981 | ........................................ | | 800.— |
| **Combination B: CRT use only**<br>DC3NT 005-008 (Kit 1)<br>DC3NT 009<br>DC3NT 011-016 | | ........................................ | | 1320.— |
| Combination C: Use of FAX machine **and** CRT<br>DC3NT 005-016 | | ........................................ | | 1450.— |
| 6. **VHF Communications** containing information on and description of the above systems | 4/1978 | ........................................ | 4.— | |
| | 3/1979 | ........................................ | 4.50 | |
| | 4/1979 | ........................................ | 4.50 | |
| | Vol.1980 | ........................................ | 18.— | |
| | Vol.1981 | ........................................ | 20.— | |
| | 1/1982 | ........................................ | 6.— | |
| | 12 Editions, complete with a binder | | | **55.—** |
| | **Set of drawings for FAX machine** | | | **8.—** |

The above kits contain all parts necessary to complete the associated PC-board. They do not contain cabinets, switches, sockets, meters, transformer and so on. You will find the complete listing of all parts in our »Pricelist of PC-boards, kits, crystals, filters and literature«, which is available free from VHF COMMUNICATIONS/UKW-BERICHTE

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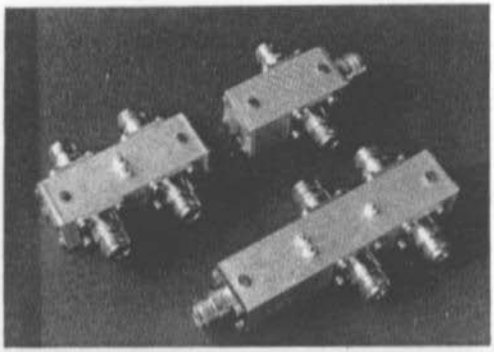

Fig. 1

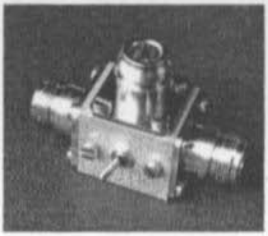

Fig. 2

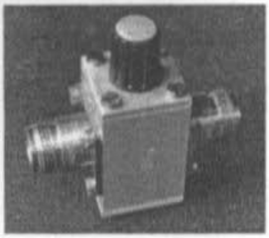

Fig. 3

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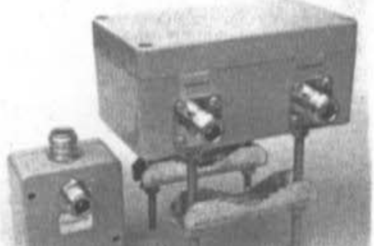

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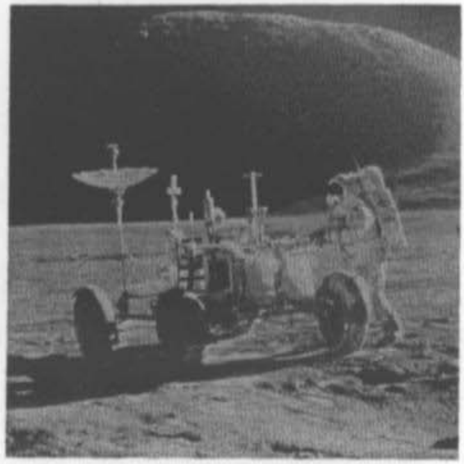

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