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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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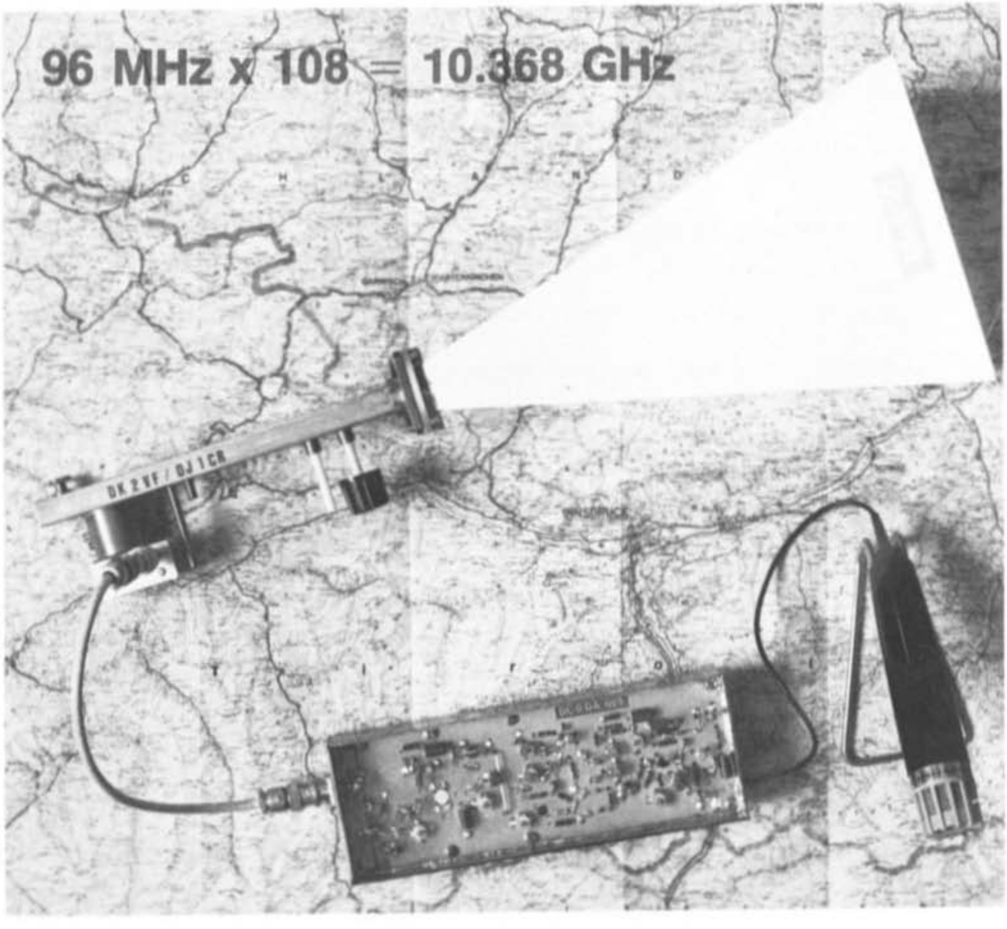

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A PUBLICATION FOR THE RADIO AMATEUR
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The following articles are to be featured in the next editions of VHF COMMUNICATIONS :

DC 3 NT:	A System for Reception and Recording of METEOSAT Images
DB 2 GM:	A 15 W Power Amplifier for 2 m
DF 7 QF:	A Frequency Doubler for 13 cm
DJ 8 IL:	A 20 W Power Amplifier for 2 m using an Integrated Module
DK 1 AG:	Design of Crystal Oscillators
DL 6 WU:	Optimum Stacking of Directional Antennas
DC 3 QS:	A Simple Radiator for 3 cm Parabolic Dishes
DC 7 MA:	Coincidence Demodulators
DK 2 RY:	A Microcomputer for Amateur Radio Applications

A FREQUENCY MULTIPLIER FOR NARROW BAND 3 cm BAND COMMUNICATIONS

by R. Griek, DK 2 VF and M. Münich, DJ 1 CR

The described varactor multiplier is designed to multiply a crystal-controlled signal by nine to produce a narrow-band transmit signal in the 3 cm band. This multiplier can be driven, for instance, by module DC 0 DA 005 (1) that is able to provide 100 to 200 mW at 1152 MHz. According to the diode, construction and alignment, the multiplier will produce 10 to 40 mW in the 10 GHz band at this drive level. If the 96 MHz crystal of this module is frequency-modulated with a deviation of approximately 0.5 kHz, an FM-signal of approximately 50 kHz deviation will be obtained at 10368 MHz.

A transmitter constructed using this frequency multiplier represents an alternative to the wideband Gunn oscillator technology, and possesses the advantage that the transmit frequency is adequately known and stable without control loops and calibrated scales. It is possible when using several, switchable crystals, whose frequencies are distributed according to the band plan given in (2), to have somewhat limited communication possibilities with wideband stations.

If a receiver is available that can be tuned over the whole 3 cm amateur band, communication will be possible without being limited to a 30 MHz or 100 MHz system. There will be no problems with compatability, and duplex operation will be possible when two antennas are used (or one antenna with circulator) as long as this is not to be made on the same frequency. A system using only one antenna and a simple waveguide switch (which is to be described in one of the next editions of VHF COMMUNICATIONS), will be smaller and more portable. This will, of course, limit communication to the simplex mode, which is usual in amateur radio communications anyway.

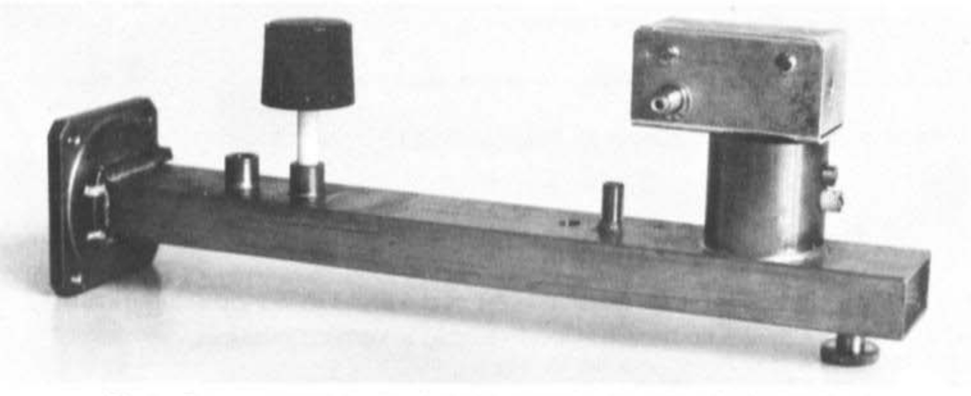


Fig. 1: Frequency multiplier (by nine) with output bandpass filter for the 3 cm band

2. CONSTRUCTION

Figure 1 shows the complete varactor multiplier for the 3 cm band. It is constructed from an approx. 190 mm long piece of R 100-waveguide (WR 90 or WG 16) with the following parts:

A variable short-circuit seals the waveguide at the right-hand side. The multiplier is built up coaxially and is located to the left of the short-circuit plunger. It is provided with a small box for the matching trimmer and a connector for the input frequency. Two metallic matching screws are to be found further to the left of this. After a further piece of waveguide, the two tuning screws of a bandpass filter and the output flange will be seen. In the photograph of the author's prototype given in Figure 1, only one of the matching screws and one of the tuning screws were required. However, in the VHF COMMUNICATION's laboratory it was found that maximum output power could only be achieved when using all four screws.

Figure 2 gives mechanical details of the multiplier. With the exception of the few given components, all parts are constructed using brass, which has not been silver-plated. The parts are assembled using soft solder. Attention should be paid that no solder is present on the inner surfaces of the waveguide.

The main body of the multiplier: the waveguide with flange, is firstly prepared. The overall dimension of 195 mm is not critical. All lengths are given from the center line of the inner conductor of the coaxial multiplier. This inner conductor protrudes into the waveguide through a hole of 6.4 mm diameter. This hole is marked firstly, after which the slot of 7 x 2 mm is measured on the lower side, as well as the holes for the two matching screws, the filter and its two tuning screws on the upper side.

It is especially important for the filter to be constructed in a precise manner. This filter was taken from (3) where it is mentioned that a magnifying glass should be used for marking and countersinking the positions for these tuning screws. The following values can be achieved: Insertion loss: 0.3 dB; bandwidth with a constant ripple: 80 MHz; 3 dB bandwidth: 120 MHz; attenuation at a spacing of 1 GHz: approx. 40 dB.

After drilling, it is possible for the threaded bushings, the eight filter pins and the flange to be soldered in one process. The author did not provide any threaded bushings or counter nuts for the matching screws in his prototype, but used a threading in the waveguide (M 5 x 0.5) which provided a tight fit to the screw.

A short-circuit plunger is made from a brass block of 15 mm in length. It is not provided with contact springs, but made so that it fits tightly. The fit should be so good that a certain amount of suction is felt when moving it in the waveguide. The plunger is fixed in position using a M 2 screw.

Figure 3 gives details for constructing the coaxial multiplier and its input matching. Electrically speaking, the construction is very simple: A matching circuit comprising a trimmer to ground, a series trimmer, and a disk capacitor to ground is to be found between the coaxial input connector for 1152 MHz and the inner conductor of the multiplier. This is followed by the diode which is connected between the inner conductor and the ground outer body. The inner conductor is isolated for DC-voltages. A 4.7 kΩ resistor between inner conductor and ground is used for selecting the operating point of the diode. It is accommodated in the box together with the two matching trimmers. An alignment screw is mounted between the diode and the waveguide which forms a capacitance to the inner conductor. The inner conductor itself protrudes 4 mm into the waveguide where it represents a coaxial/waveguide transition.

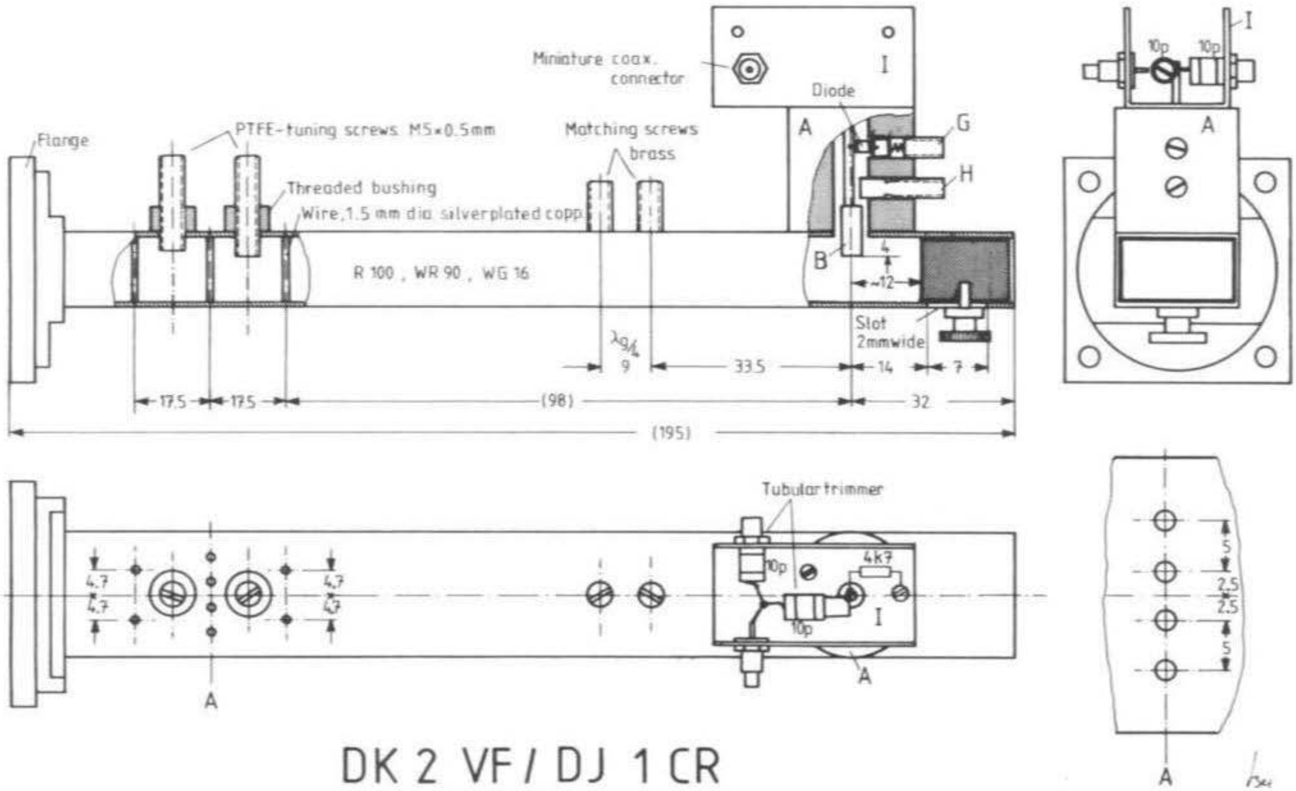
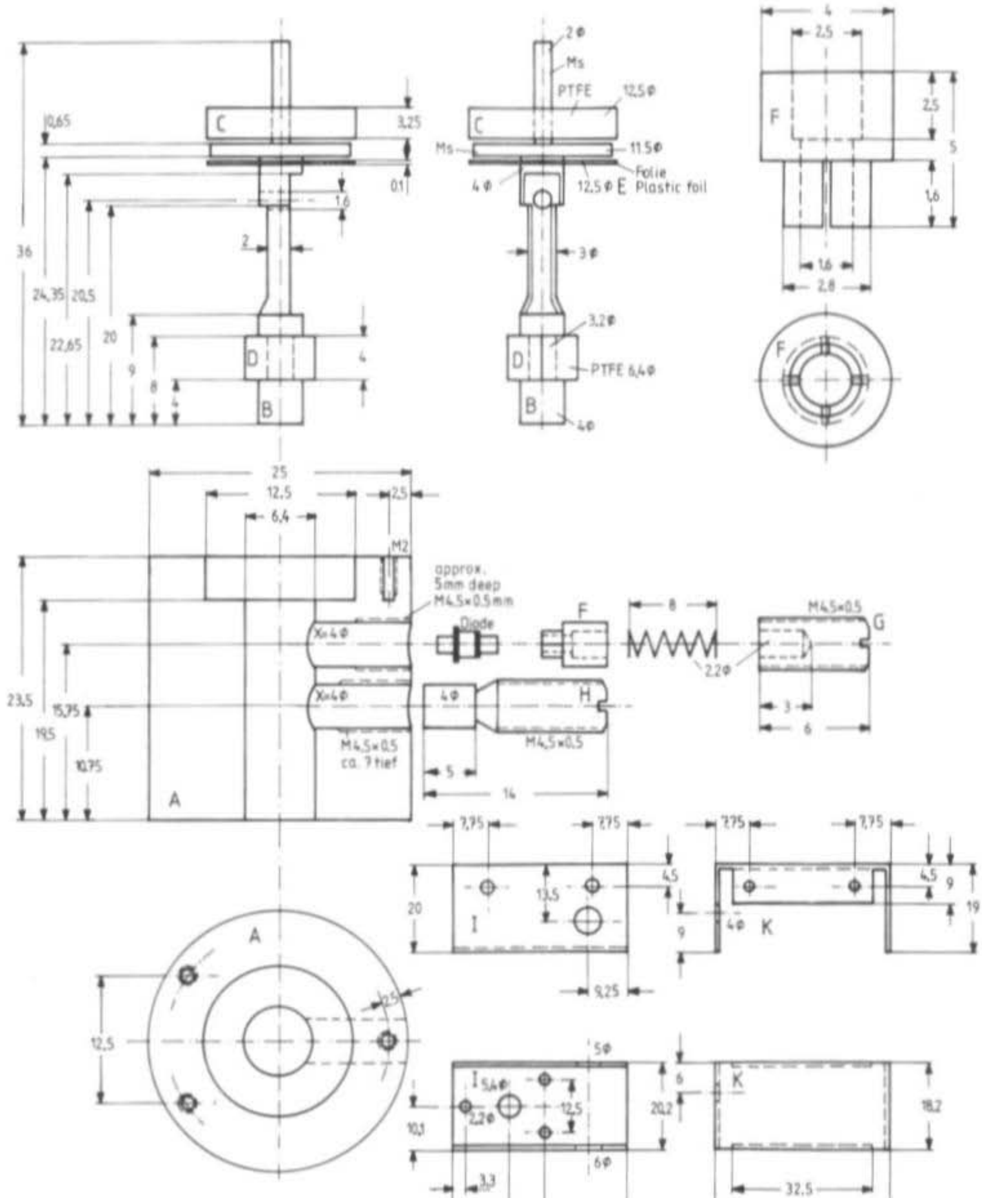


Fig. 2: Construction and dimensions of the waveguide part with matching screws and bandpass filter

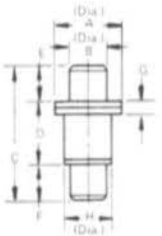


DK 2 VF / DJ 1 CR

Fig. 3: Individual parts of the coaxial multiplier

It will not be possible to produce these parts without the aid of a lathe. The outer body »A« should be made firstly by providing the central hole of 6.4 mm diameter, extending this to 12.5 mm for the disk capacitor, providing the three threaded holes for mounting the box, and the two holes for the diode mount and alignment screw.

The inner conductor »B« is lathed from a round bar of 4 mm diameter. The metal disk of 11.5 mm diameter for the capacitor is soldered into place later. The part of the bar above the disk is lathed down to 2 mm diameter, the center piece to 3 mm diameter, which is then provided with a sloping transition to the lower 4 mm portion. With this part, a 4 mm wide portion is lathed down to approximately 2.5 mm diameter, into which the slotted PTFE part »D« is to be placed later. It is now necessary for the center portion to be filed or milled flat, and for a hole to be made for the diode. The table given in **Figure 4** was taken from the MICROWAVE ASSOCIATES Data-Book and gives the dimensions of the diode. This information should be used when manufacturing the parts so that the diode has a tight fit, however, that no force is necessary since it is very sensitive to lateral and pulling forces.



DIM	INCHES MIN.	INCHES MAX.	MM MIN.	MM MAX.
A	0.119	0.127	3.02	3.22
B	0.060	0.064	1.52	1.63
C	0.205	0.225	5.21	5.72
D	0.085	0.097	2.16	2.46
E	0.060	0.064	1.52	1.63
F	0.060	0.064	1.52	1.63
G	0.016	0.024	0.41	0.61
H	0.079	0.083	2.01	2.11

Fig. 4: Dimensions of the diode used

The two PTFE parts »C« and »D« are provided to mount the inner conductor concentrically within the outer body, and the plastic foil E (triacetate or Hostaphan, 0.1 mm thick) forms the previously mentioned disk capacitor.

This is followed by manufacturing parts F, G and H, as well as an approximately 8 mm long spring. The diode mount part F is slotted four times in the thinner part and should fit well into the hole in the outer conductor A which is half provided with a flat surface, and half with a fine thread. Part G should also fit tightly into the cylindrical hole in part A. If no tool is available to provide the M 4.5 x 0.5 thread, a thread of M 5 x 0.5 mm can be used. It is then necessary for the dimensions of parts A, F, G and H to be changed correspondingly; the diameter of part X remains, however, at 4 mm.

The only parts that are now missing are the two parts made from metal plate: I and K. These parts are not critical. It is, of course, also possible to use an available case. The lower portion I is mounted, using three screws, to the outer body A; it is provided with the input conductor (BNC or smaller) and the grounded matching trimmer. Part K is the cover, which should be provided with a hole for alignment of the series trimmer.

Type	Breakdown voltage min.	Breakdown voltage max.	Junction capacitance min.	Junction capacitance max.	Capacitance ratio max.	Minority carrier life t_L min.	Minority carrier life t_L max.	Snap time t_s max.	Cut-off frequency f_c min.	Heat resistance max.	T_j max.
MA-44140	25 V	45 V	0.5 p	1.5 pF	1.5	10 ns	30 ns	100 ps	170 GHz	70°C/W	200°C
MA-44150	15 V	40 V	0.2 p	0.6 pF	1.5	8 ns	30 ns	90 ps	200 GHz	100°C/W	200°C

3. FURTHER INFORMATION

3.1. Suitable Trimmers and Diodes

Two tubular trimmers having a maximum capacitance of 10 pF are used in the 1152 MHz matching circuit. In the author's prototype, air-spaced types were used:

VOLTRONICS VOL AT 14 L.

However, the following are also suitable:

JFD Electronics Components Co., type MVM 0.8 - 10 pF;
or type AT 5200 0.8 - 10 pF.

It should be, however, possible for it to be constructed using ceramic tubular trimmers (Philips), since these were used in the 1152 MHz module described in (1).

The diodes used were manufactured by MICROWAVE ASSOCIATES, which designate them as SNAP VARACTOR diodes. The previous table gives full specifications of the two types that can be used. At the end of 1978, the cost of these diodes without tax were:

MA-44140	DM 131,— each
MA-44150	DM 166,— each

It is, of course, possible with the aid of these specifications to select equivalent diode types from other manufacturers, e.g. as mentioned in (4): BXY 41 (Philips) or DH 292 (Thomson-CSF). Of course, it is then necessary for the different case shape to be taken into consideration, when constructing the multiplier.

3.2. Maximum Power

If the construction and alignment (as well as the 1152 MHz signal !) are correct, the multiplier will operate with an efficiency of between 10 and 15 %. In the VHF COMMUNICATIONS's laboratory, an output power of 15 mW was obtained with an input power of 100 mW, and the author was able to measure up to 40 mW at 10368 MHz at an input power of approximately 250 mW.

The question of the maximum power that can be used, cannot be answered easily since the maximum permissible input power is dependent on several factors. The two most important are: Power dissipation and breakdown voltage of the diode.

The maximum permissible power dissipation P_{diss} of a semiconductor can be calculated as follows:

$$P_{diss} = \frac{T_j - T_{case}}{\Theta_{jc}}$$

where: T_j is the maximum permissible junction temperature,
T_{case} the case temperature which is dependent on cooling and ambient temperature,
Θ_{jc} the heat resistance between junction and the case.

In the case of diode type MA-44140, T_j is given as 200°C, and Θ_{jc} as 70 °C/W. If one assumes that it is possible to maintain the case temperature at 60 °C by cooling, the maximum power dissipation can be calculated as:

$$P_{diss} = \frac{200\ ^\circ C - 60\ ^\circ C}{70\ ^\circ C/W} = 2\ W \qquad (MA\text{-}44150\text{: } 1.4\ W)$$

The question is now whether it is possible to dissipate nearly 2 W of dissipation heat via the thermo-isolated inner conductor on one side, and parts F and G on the other side, so that (only) 60 °C are maintained as case temperature of the diode. Not only the construction but also the material (copper is a better heat conductor than brass) and the quality of construction play their part.

The last point is illustrated by the following: When Gunn oscillators firstly became popular, the editors received a large number of complaints regarding reported poor Gunn elements, that exhibited deteriorating values after a short time, and finally complete failure. These were typical cases of poor installation and/or unfavorable construction, where the Gunn element became excessively hot and finally failed.

We hope that the microwave specialists among our readers will excuse this somewhat extensive description, however, the author and the editors wish to take this opportunity of not only introducing this technology to newcomers to this field, but also to offer them the advantage of the experience (and mistakes) that have been made in the past.

A further limit of the maximum permissible drive power is the breakdown voltage of the diode. It is not possible – at least for radio amateurs – to calculate the maximum permissible input power from the data sheet values. This is because, firstly, the voltage value differs from diode to diode (between 25 and 45 V for diode type MA-44140), and secondly since the impedance of the diode is not known. This is very dependent on the tuning and drive power, which means that one can only make the following conclusion:

It is necessary for the first alignment to be made with a drive power that is far below the estimated maximum. With improved alignment, the drive power can be increased and the efficiency measured at each step. The limit is achieved when the efficiency or output power drops without carrying out any alignment.

It was not possible to establish the maximum permissible input power for the described multiplier.

3.3. The Exciting Crystal Oscillator

The importance of a stable crystal oscillator is illustrated clearly when one considers that the crystal frequency is multiplied by 108 times. A frequency variation of 100 Hz at 96 MHz will cause a deviation of 10 kHz in the 10 GHz band. This means that the oscillator circuit must be examined with respect to the four following characteristics:

- Microphonics (mechanical stability)
- Temperature behaviour (thermal stability)
- Interference noise across the varactor diode (interference modulation)
- Oscillator noise

From the application side, we are to differentiate between two quality steps:

- Operation in wideband FM and under line-of-sight conditions
- Narrow-band FM or CW and/or non-line-of-sight communications

In the first case, it has been found that it is only microphonics that can cause interference. It is sufficient for the oscillator and modulator to be separated from the rest of the transmitter and for it to be constructed in a mechanically stable manner in its own case with some thermal insulation. The thermal stability is limited to ensuring that the oscillator does not cease operation due to low temperatures. Unwanted modulation is avoided by ensuring that the varactor diode is only loosely coupled to the oscillator, so that several volts of AF-voltage are required to obtain the required frequency deviation. This means that interference voltages in the mV-range will have no effect. The non-linear modulation characteristics caused by this is not important for voice communications.

On the other hand, all possibilities must be used for the second application. In this case, the crystal should be placed in a crystal oven and should be aged for several weeks of operation in an oven. With respect to the noise, two recent publications should be mentioned:

In (5) the crystal oscillator described by DC 0 DA in (1), is used as an example for a high-noise oscillator, and the author gives a description of measuring equipment for experimentally measuring sideband noise.

Mr. B. Neubig, DK 1 AG (of KVG) is to discuss crystal oscillators in details in one of the next editions of VHF COMMUNICATIONS. The most important points are: The crystal loading should not be as low as for oscillators having a maximum long-term stability; the oscillator circuit must be isolated from the limiting function – one should therefore use two-stage circuits;
the most favorable transistors are high-power junction FETs such as the P 8000.

This concludes this article, even though the author was not able to provide as complete an article as he would like. However, the following references allow interested readers to obtain further details for construction of a high-quality 10 GHz transmitter.

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RSGB-Publications, London, 1978

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VHF COMMUNICATIONS 10, Edition 3/1978, pages 145 - 153

(5) U. L. Rohde: Evaluating Noise Sideband Performance in Oscillators
Ham Radio Magazine 1978, Edition 10, page 51

A 3 CM PRIMARY RADIATOR FOR PARABOLIC ANTENNAS

by R. Griek, DK 2 VF, and M. Münich, DJ 1 CR

The drawing given in **Figure 1** shows all required dimensions of the dipole radiator and reflector which are placed directly in front of the open-end of a so-called X-band waveguide R 100 (or WR-90 or WG-16). This radiator is placed through the center of the parabolic reflector from the back so that the shorter dipole is placed at the focal point. This is not only simpler mechanically, but is far more elegant than when using a horn radiator and waveguide fed around the parabolic reflector.

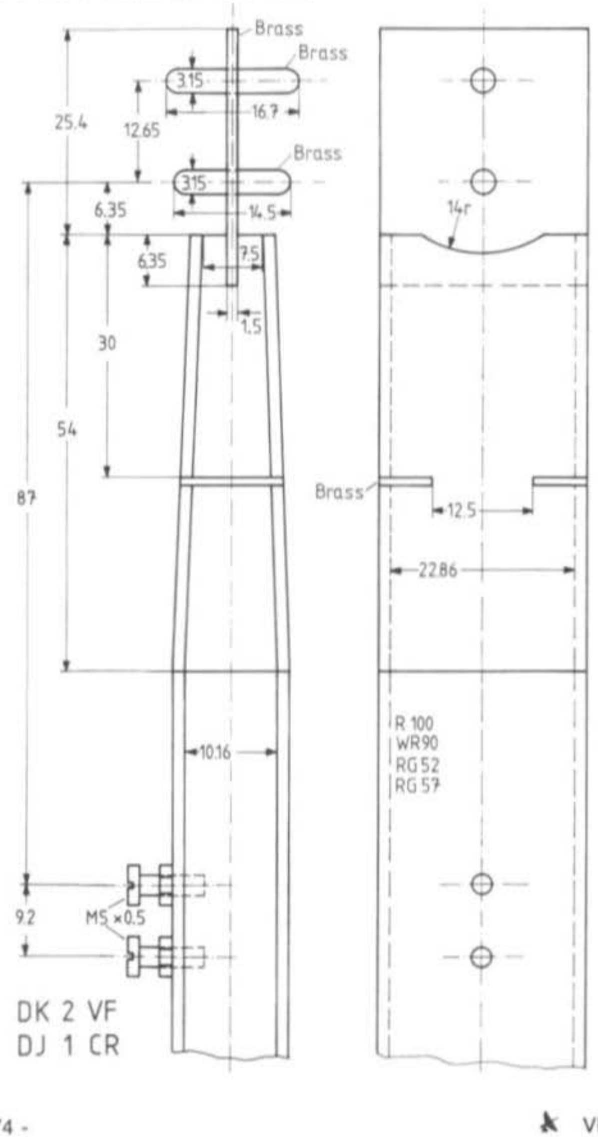

Fig. 1: Primary radiator for parabolic antennas

The 10 dB beamwidth (1) of the radiator assembly amounts to approximately 60°, which makes it more suitable for parabolic reflectors having a focal angle between 60 and 80°. The author uses it in parabolic reflectors of 50 cm in diameter, having a focal distance of 17 to 18 cm (f/d approx. 0.3) **(see Figure 2)**.

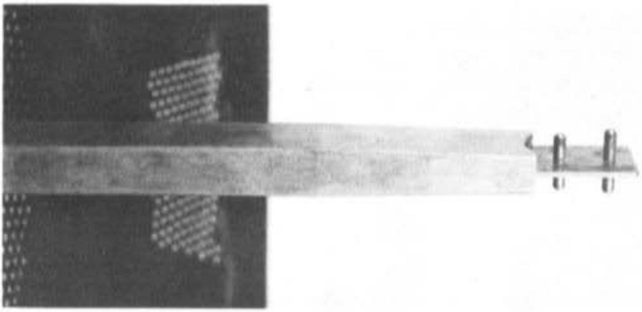

Fig. 2: Photograph of the prototype radiator

The described primary radiator is taken from (2) and has been further developed in discussions with M. Hamel, DJ 8 VY, and Dr. D. Evans, G 3 RPE, and recalculated for the amateur frequencies in the 3 cm band.

The operation of this radiator is simple: the two dipoles (»radiator« and »reflector«) are soldered to a metal surface that is placed in the open end of the waveguide parallel to its wide side. When using the usual TE_{10}-mode in the waveguide, the E-vector is parallel to the dipoles which means that they will be energized by the radiation. If the metal surface is mounted symetrically in the waveguide, both dipole halves will be energized equally. The narrowing of the waveguide in one plane is used for impedance matching; it also improves the radiation diagram be decoupling the walling of the waveguide from the dipoles. The impedance of the system is also determined by the insertion depth of the metal surface into the waveguide, and by the spacing of the dipoles from its open-end. Furthermore, an iris in the narrow part of the waveguide, and two matching screws in the straight part optimize the matching. The two matching screws are aligned for lowest VSWR, or maximum field strength in conjunction with a local station.

It should be mentioned that all parts are made from brass, and can be silver-plated, although this is not absolutely necessary. It is important, however, that any residual solder between the dipoles and the metal surface is removed, as well as between this surface and the waveguide. Also, it is necessary for the narrowing of the waveguide to be exactly symmetrical.

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VHF COMMUNICATIONS 8, Edition 4/1976, pages 207 - 214

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McGraw-Hill Book Company, New York

AN SSB TRANSMITTER FOR THE 13 cm BAND

EXPERIMENTAL SYSTEM USING ENVELOPE ELIMINATION AND RESTORATION

by R. V. Galle, VK 5 QR

The technology of SSB-signal processing using the envelope elimination and restoration method was described for VHF amateurs in (1). Dr. Karl Meinzer, DJ 4 ZC, extended the required adapter using a frequency divider, which then allowed a subsequent frequency multiplication to the required UHF or SHF band. This could be made with the aid of conventional varactor multipliers (2). The author used this system to obtain a SSB transmitter for the 13 cm band and was able to obtain 4 W of SSB-signal at 2304 MHz in this experimental system.

This was sufficient to cover the 1885 km path from Adelaide to Albany (VK 6 WG) using the summer »duct-conditions« present on February 17, 1977. Of course, this amazing distance has nothing to do with the signal processing method, but there is hardly a simpler method of obtaining SSB-signals at a useful power level than when using this system. For this reason, this article is to provide a refresher regarding the principle and to describe the transmitter used. It should be underlined, that this is only an experimental system, and this is the reason why no attempt has been made to provide full constructional data. The most important details are to be shown and discussed; and several photos that were made by VK 5 RT provide an impression of the author's experimental construction.

1. BLOCK DIAGRAM

The block diagram given in **Figure 1** shows the modules and filters used. A transceiver KWM 2 or a FT 101 B are used as SSB exciter at 21 MHz. The heart of the construction, the processor, is to be described in detail in section 2. The frequency of the subsequent crystal oscillator is selected so that a frequency of 2304 MHz is obtained in the 13 cm band after multiplying by six. The linear power amplification is made at a frequency of 384 MHz, where it can still be made easily. It is possible, for instance, to use the linear amplifier described by G. Freytag DJ 3 SC (3). The output power depends on the specifications of the subsequent frequency multiplier. A MICROWAVE MODULES varactor tripler MMV 1296 can be used for the tripling from 384 MHz to 1152 MHz. It is only necessary for it to be tuned somewhat lower. This varactor tripler can be driven up to an input power of 20 W, so that an output level of 10 W at 1152 MHz can be obtained.

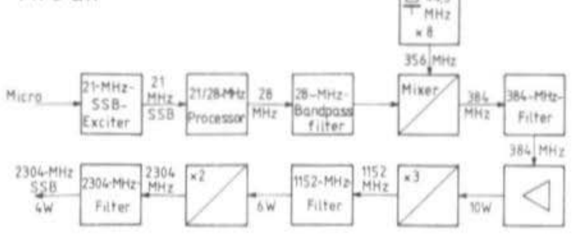

Fig. 1: SSB transmitter for the 13 cm band with envelope elimination and restoration

The following frequency doubler from 1552 to 2304 MHz is to be described in detail in section 3, since it is assumed that any additional information in this area is still very welcome. It is extremely important that bandpass filters are provided in front of and between the frequency multipliers so that these are only driven with a single frequency.

It should be mentioned that module DJ 6 ZZ 006 (4) can be used as transmit mixer from 28 to 384 MHz, and an interdigital converter (5) used as receive converter.

2. THE PROCESSOR

The principle of operation was described in detail in (1) and (2). Basically, the input SSB-signal at 21 MHz is split into its amplitude-modulated component (envelope) and its phase-modulated component (FM). Both these signal components are now separately processed: The PM-signal is divided by six, and the resulting frequency of 3.5 MHz possesses only 1/6 of the original phase deviation. This signal is amplitude-modulated with the separately amplified envelope and finally mixed with a crystal-controlled frequency of 31.5 MHz to obtain the output frequency of 28 MHz. After this, the new SSB-signal that possesses only 1/6 of the phase deviation, is amplified selectively until the required level is obtained for the subsequent module.

Figure 2 shows the circuit diagram of the processor. The 21 MHz SSB-signal (the frequency may be in the range of 21.0 to 21.5 MHz) is fed to the input at a level of approximately 100 mV and is amplified to approximately 5 V by transistor T 1.

An envelope demodulator equipped with diode D 2 removes the AM component, and the AF signal is fed to the AF amplifier (T 5), which modulates the PM-signal in the stage equipped with T 4. Diode D 1 is provided for bias voltage generation for transistor T 5.

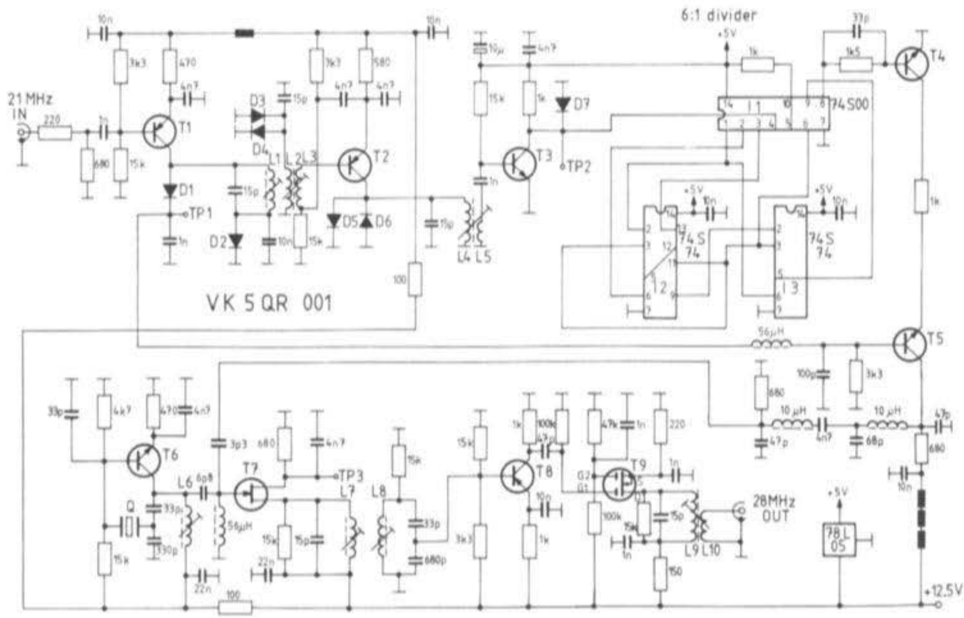

Fig. 2: The 21/28 MHz SSB processor with a division factor of six

The signal amplified in T 1 is also fed via a bandpass filter (L 1, L 2) to the first limiter (D 3, D 4). After subsequent amplification (T 2) and limiting (D 5, D 6), the 21 MHz PM-signal is fed to a pulse shaper (T 3). The square wave signal is now divided by six. In actual fact, standard TTL circuits should be sufficient at an input frequency of 21 MHz. However, the author used Schottky-TTL circuits to be on the safe side.

Transistor T 4 operates as an electronic switch that is driven by the 3.5 MHz square wave voltage. The current flowing via T 4 drives the AF voltage at the base of T 5, so that the envelope is modulated onto the 3.5 MHz signal.

The combined signal is now fed via a lowpass filter for suppression of the sixth harmonic to the mixer equipped with the FET T 7. The crystal oscillator equipped with T 6 provides a frequency of 31.583 MHz, which allows a difference frequency of 28.083 to 28.0 MHz to be filtered out via the bandpass filter comprising L 7 and L 8, which is then amplified in the last two stages.

2.1. Components for the Processor

T 1, T 2, T 8:	RS 2003 (Japan), AF 106, AF 127 or other Germanium-PNP-RF transistors
T 3 - T 5:	2 N 706 or similar Silicon-NPN-VHF transistor
T 6:	BF 173, BF 224, BF 199 or similar VHF transistor
T 7:	MPF 102, BF 245 or similar FET
T 9:	MPF 121, 40673, 40841 or similar DG-MOSFET
D 1, D 3 - D 7:	1 N 914, 1 N 4148 or similar Silicon-planar switching diode
D 2:	AA 112, AA 118 or similar Germanium diode

All inductances are wound on a 6 mm coil former with HF core, using 0.4 mm dia. enamelled copper wire.

L 1:	20 turns, L 2: 22 turns, spaced 15 mm from another
L 3:	4 turns on L 2
L 4:	22 turns, L 5: 4 turns on L 4
L 6:	15 turns
L 7, L 8:	22 turns; spaced 12 mm from another
L 9:	22 turns; L 10: 3 turns on L 9

2.2. Construction

A PC-board was developed for the processor circuit given in Figure 2; this PC-board is shown in **Figure 3**. Its dimensions are 145 mm x 70 mm and it is double-coated. Only a few through-contacts are required, and these are made during the mounting of the components by soldering to the upper and lower side of the board. These positions are marked in the component location plan **(Figure 4)** by small crosses. Screening of the whole processor and between the individual stages is essential to ensure that no components of the original SSB-signal are present at the output. For this reason, the PC-board is divided into six chambers using screening panels **(Figure 5)** and is then soldered into an RF-tight case. This is done as follows:

The side of the PC-board having the large ground surface is to be designated as upper or component side. The lower side is thus the side with the few conductor lanes. All connection holes are drilled from the lower side.

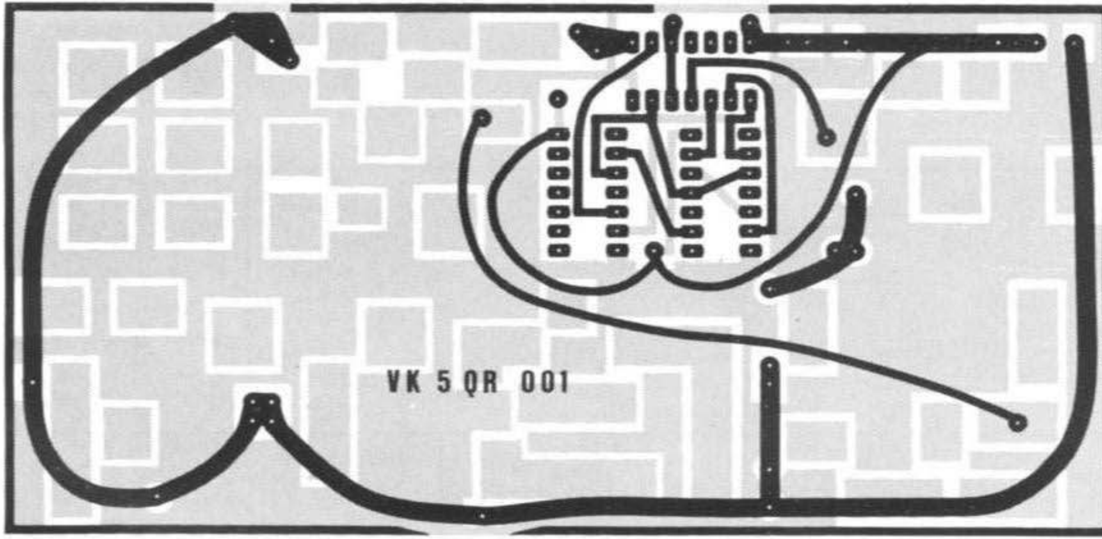

Fig. 3:
The double-coated PC-board for the 21/28 MHz SSB processor with frequency division factor 6 (VK 5 QR 001)

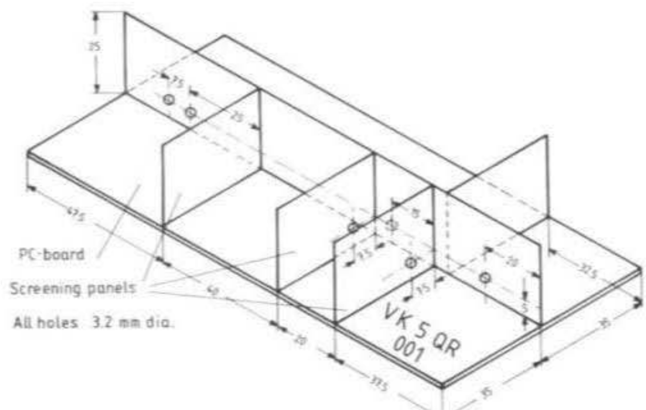

Fig. 5:
Mechanical construction of the processor

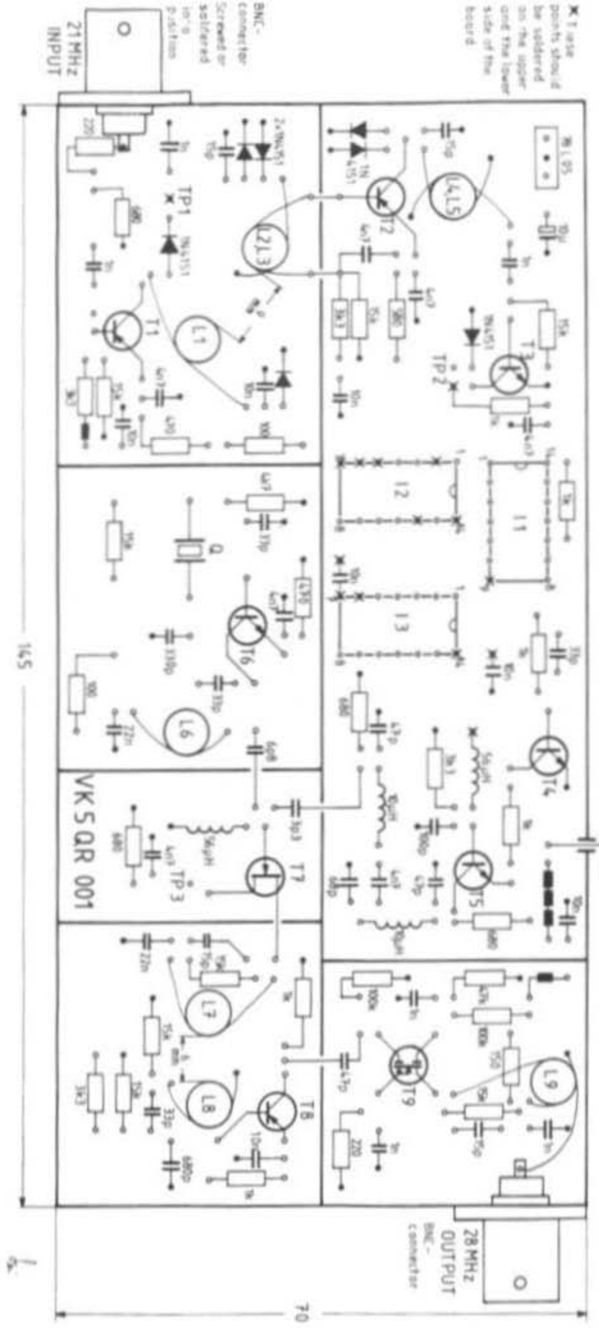

Fig. 4: Component location plan for the processor PC-board VK 5 QR 001

After this, the screening panels are manufactured as shown in Figure 5, they are then provided with the given holes, and then soldered onto the component side of the PC-board.

This is followed by winding the coils (coil data given in section 2.1.) and the windings fixed with a little fast-drying adhesive. Before fixing the coils with a dual-component adhesive, one should ensure that the given spacings are correct.

This is followed by equipping the stages: input amplifier, limiter, and frequency divider, connecting the operating voltage, and injecting a 21 MHz signal. After coarsely aligning the inductances, the operation of the divider is checked using a counter or by monitoring.

It is now possible for the rest of the circuit to be constructed.

After mounting all the components, the module is installed into a metal framework equipped with BNC-connectors and an operating voltage feedthrough. This is followed by the alignment.

2.3. Alignment of the Processor

A constant input level at 21 MHz is important so that the limiter operates correctly and that the modulator thus receives an approximately constant voltage. After aligning L 1, a voltage of at least 5 V and max. 6 V should be measured at TP 1 with the aid of an RF- (tube) voltmeter.

The alignment of the circuits comprising L 2 and L 4, which are dampened by the limiter diodes, can be made with the aid of the voltmeter connected to test point TP 2, and reducing the 21 MHz input voltage temporarily to a lower value in order to obtain a reading that is not limited.

The crystal oscillator is aligned with the voltmeter connected to test point TP 3, and the transient performance checked.

Attention should be paid during the subsequent alignment of the mixer and output amplifier that they are not aligned to the crystal oscillator frequency by mistake. After this, it is possible for the signal to be fed to the 28/384 MHz transverter.

3. ANOTHER VERSION FOR THE 23 cm BAND

Since the same principle and described construction can be used with a few modifications for the 23 cm band, a few details are now to be given regarding this. It is, of course, far easier to obtain the linear power amplification in the 70 cm band and subsequently triple the frequency than to generate the SSB-signals directly at 23 cm.

The same concept remains, and it is only necessary for the frequencies to be changed:

In the processor, one will no longer divide by six, but by three. The corresponding extract from the circuit diagram is given in **Figure 6**, which is inserted between T 3 and T 4 in Fig. 2.

The crystal oscillator is provided with a 35.166 MHz crystal so that the »intermediate frequency« of 7.0 to 7.166 MHz is converted to 28.166 to 28.0 MHz. No further modifications must be made to the components or inductances.

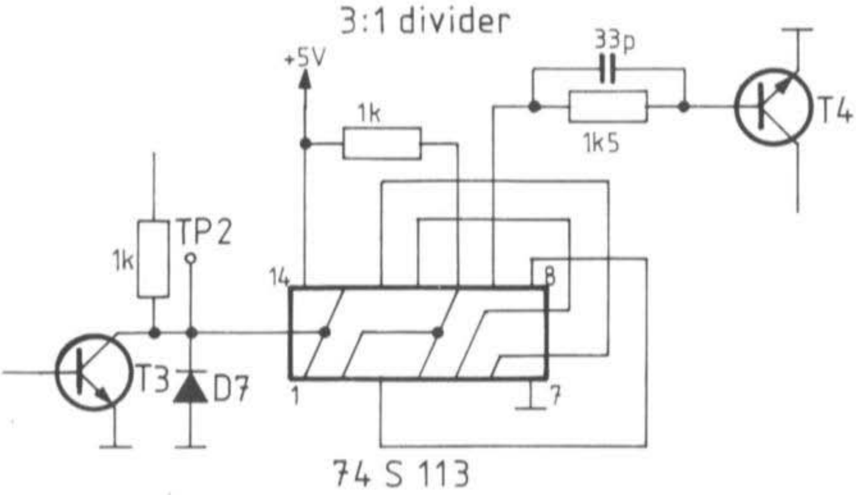

Fig. 6: A 3:1 divider replaces the 6:1 divider when the processor is to be used for the 23 cm band

In the following transverter, this signal is mixed with the local oscillator frequency of 404 MHz to the normal 70 cm range of between 432 and 434 MHz. The output power is then amplified linearly, and then fed to the frequency tripler.

If there is sufficient interest, a PC-board will also be described for this version.

4. DOUBLER 1152/2304 MHz

The main dimensions of this module are given in **Figure 7**. The resonator panels are made from 4 mm copper plate, base and cover from 2 mm copper plate. The whole doubler can be mounted in a cast aluminium case similar to that used by MICROWAVE MODULES. The author used BNC-connectors, but according to his experience at 2300 MHz, it is advisable to use N-connectors.

A $\lambda/4$ coaxial circuit is provided for the input and output frequency. They are connected together with the aid of a coupling link which is provided with a 4 mm wide brass strip as coupling capacitor. This coupling link is passed through a third chamber in which the multiplier diode type VSE 66 P (Mullard/Philips) is accommodated. The diode is mounted with one flange in a heat-conductive pin, and the other flange is provided with a cap having a strip-shaped connection to the coupling link.

The input coupling at 1152 MHz is made galvanically with the aid of a tap. The output coupling at 2304 MHz is made capacitively using a disk of approximately 6 mm diameter to the inner conductor of the connector. The degree of coupling is varied by screwing the conductor in or out. The photograph given in **Figure 8** gives an approximate impression of the construction of this module.

The author hopes that this method of SSB-processing for microwave frequencies will increase activity on these bands, and would like to hear your experiences.

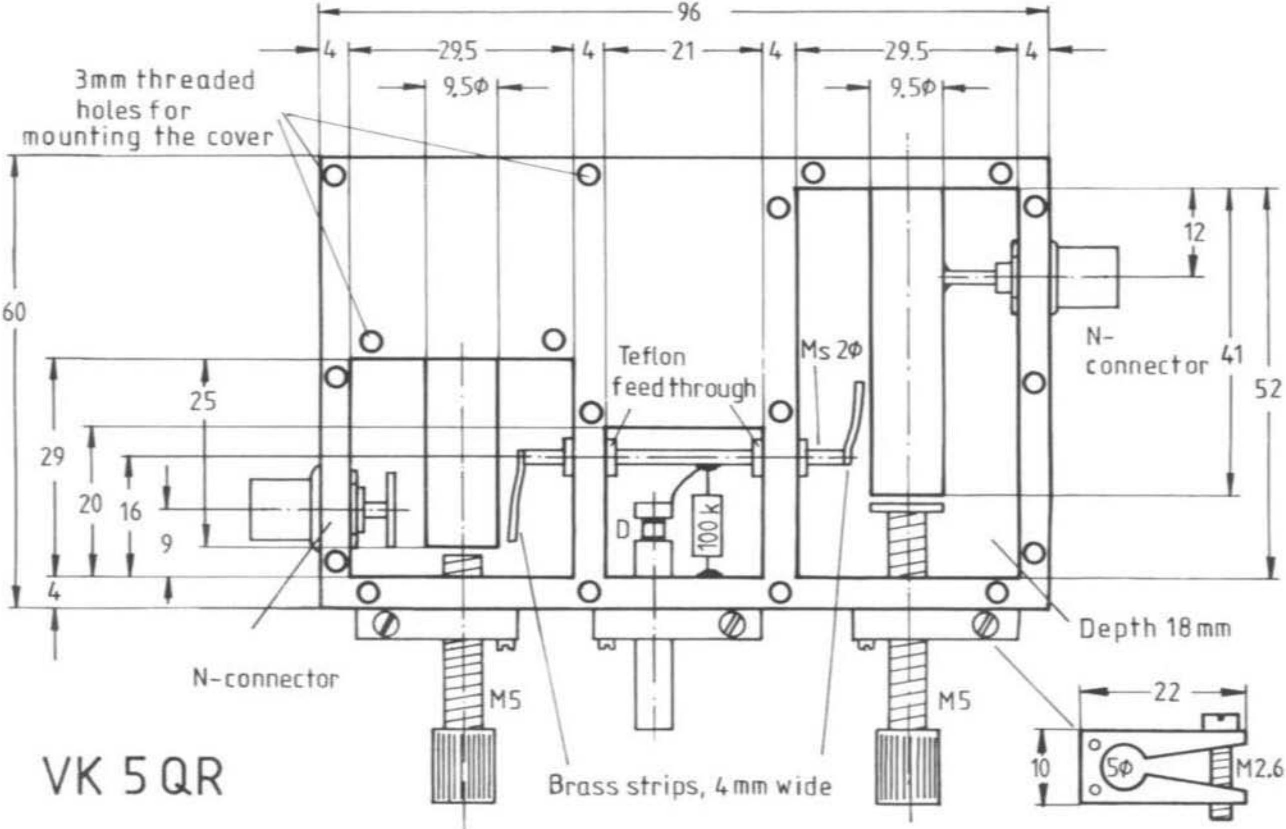

Fig. 7: Efficient frequency doubler from 1152 MHz to 2304 MHz

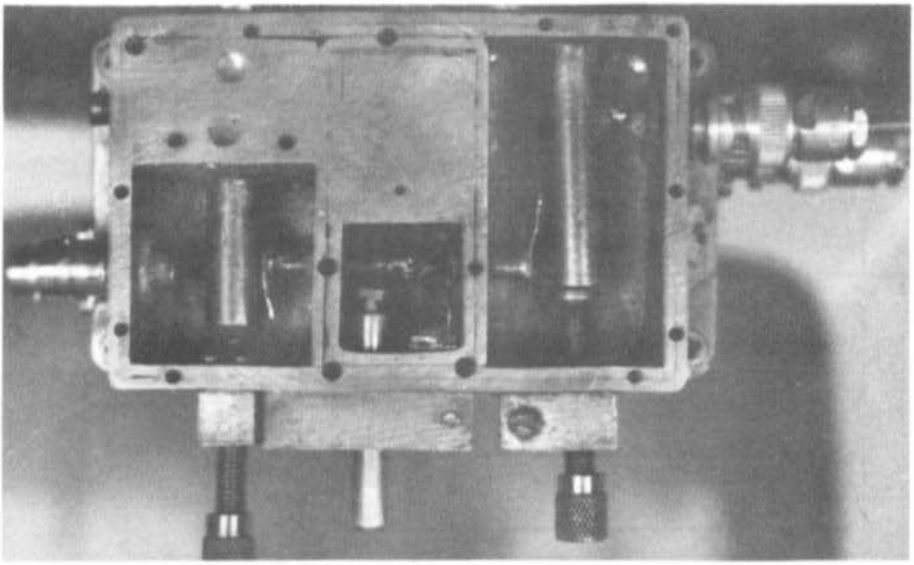

Fig. 8: Author's prototype of the frequency doubler with 4-5 W output power

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Interdigital Filters Extended to Form Receive Converters
VHF COMMUNICATIONS 10, Edition 3/1978, pages 154 - 168

INTERDIGITAL CONVERTERS AS TRANSMIT MIXERS

by U. Mallwitz, DK 3 UC

Receive converters equipped with passive mixers are also suitable for use as transmit converters. An example of this based on the interdigital converter (1) is now to be described for the 13 cm band.

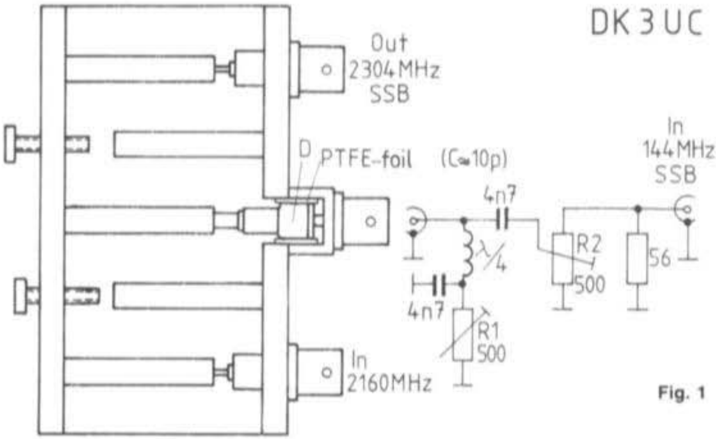

Fig. 1

The oscillator power at 2160 MHz is a decisive factor for the output power. An oscillator power of 25 mW was used by the author but several hundred mW can be used. The second criterion is the type and installation of the mixer diode D. A diode type P 081 B is shown in **Figure 1** which is built into a diode type 1 N 21 case. This diode is manufactured by Parametric Ind.; however, other Schottky diodes such as BAW 95 (Philips) should be suitable. The IF-side of the diode must be bypassed with a SHF bypass-capacitor of approximately 10 pF (PTFE-foil). The bias of the diode is aligned using R 1 for maximum output power, and the drive using R 2. It is possible for the coaxial cables to be directly soldered to the appropriate resonator to save connectors.

The author also tried other types of diodes in this mixer: BXY 38 C, BXY 39, and 153/24 (10 GHz diode manufactured by Marconi); however, the output power levels were approximately 4 dB less than that obtained with Schottky diodes.

Of course, it is not possible for long-distance communication to be made with output power levels of between 0.1 and 100 mW – according to the diode type and oscillator power. However, it represents a good start on this band, and such interdigital mixers are suitable for alignment applications and for communications over line-of-sight paths of several tens of km.

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SSB-TRANSMIT MIXERS FOR THE SHF BANDS

by R. Heidemann, DC 3 QS

Part 1: 13 cm Band

This and the following articles are to describe a simple method of generating an SSB signal for the SHF amateur bands at 13 cm, 9 cm, 6 cm, and 3 cm. In order to obtain reproducible modules that are also suitable for portable operation, the mixer is to be built up using a PC-board construction that does not require any extensive alignment. The complete circuit is equipped with semiconductors. In addition to this, no expensive SHF transistors are used in the frequency multipliers and mixers, but only cheap power varactors, which are available from the author. For this reason, this concept is most certainly not the most favorable for all conditions and it will not be able to replace a 13 cm fixed station equipped with the well-proved 2 C 39 tubes (1).

No information is to be given regarding the local oscillator since a sufficient number has already been published such as (2), (3), and (4). However, after many discussions, the author has got the impression that it would be advisable to describe suitable varactor multipliers. For this reason, further details are to be given regarding the construction of frequency doublers for 2160/2304 MHz, 3312/3456 MHz, and 2592 MHz (= 10368 ÷ 4), as well as triplers for 3312/3456 MHz, 10368 MHz, and triplers for 3456 MHz, 10368 MHz output frequency.

1. FUNDAMENTALS

In order to offer the reader a better understanding of the operation of the described mixer, this section is to give a short description of the theory in an unmathematical manner.

The phase speed of a wave on a line (e.g. coaxial or micro-stripline) is dependent on the longitudinal inductivity and lateral capacitance component (L' or C'). The following is valid for the so-called propagation constant β:

$$\beta = \frac{2\pi}{\lambda} = \sqrt{L' \times C'}$$

where λ is the line wavelength. If $v = f \times \lambda$, the following will be obtained:

$$v = \frac{2\pi f}{\sqrt{L' \times C'}}$$

This means the larger the inductivity and capacitance components, the lower will be the phase speed v.

The phase shift $\Delta\varphi$ caused by a line of length l is as follows:

$$\Delta\varphi = 2\pi f t = 2\pi f \frac{l}{v} = l\sqrt{L' \times C'}$$

It will be seen from this equation, that the phase shift of a wave is also dependent on the capacitance component. The effective capacitance component of a micro-stripline becomes voltage-dependent when used with varactor diodes. This means that the phase shift $\Delta\varphi$ is also voltage-dependent. It is possible when using a line with voltage-dependent capacitive component (line with one or more varactor diodes, varactor in the form of a stripline) for the phase of a carrier wave of frequency f_0 (e.g. 2160 MHz) to be shifted in time with a connected modulation signal of frequency f_m (e.g. 144 MHz), which means that this line can be used as phase modulator.

The conversion of this phase modulation into the required amplitude modulation (modulation = conversion) can be made most easily in the form of a decoupled line (3 dB hybrid, magic-Tee). For this purpose, the carrier wave is fed to a 3 dB divider, after which both signals are shifted by a constant phase angle. In addition to this, the phase modulation is added to **one** of these signals. Finally, the unmodulated and phase-modulated signal components are added. According to the phase shift between these two signals, an amplification ($\Delta\varphi = 2\,n\pi$, $n = 0.1,...$) or rejection ($\Delta\varphi = (2\,n + 1)\,\pi$, $n = 0.1,...$) will take place. This means that an amplitude modulated signal is present at the output of this circuit as the conversion product $mf_0 \pm nf_m$ ($m = 1.2, ...; n = 0.1, ...$). For explanation, the vector diagram valid for the output voltage is given in **Figure 1**.

a is the vector of the non-phase modulated signal component

$$\vec{a} = A \exp(j\,2\,\pi\,f_0\,t)$$

$\vec{b}_0$ is the vector of the signal component that is shifted with respect to $\vec{a}$ by the constant phase angle φ_0

$$\vec{b_0} = A \exp(j\,(2\,\pi\,f_0 t + \varphi_0))$$

$\vec{b}$ is the vector phase-modulated with f_m

$$\vec{b} = \vec{b}_0 \exp(j\,\Delta\varphi \cos(2\pi\,f_m t))$$

The following is valid for vector $\vec{a} + \vec{b}$ of the output voltage:

$$\vec{a} + \vec{b} = 2\,A\,\{\cos\frac{\varphi_0 + \Delta\varphi\cos(2\,\pi\,f_m t)}{2}\}\,\exp\,[\,j\,(2\,\pi\,f_0 t + \frac{1}{2}(\Delta\varphi\cos(2\,\pi\,f_m t) + \varphi_0))]$$

The amplitude modulation, which is the variation of the output voltage component $\Delta/\vec{a} + \vec{b}/$ as function of the phase angle $\Delta\varphi$, can be seen easier in the vector diagram given in Figure 1 than in this equation.

A considerable advantage of this mixer with respect to other circuits is that the carrier frequency f_0 can be suppressed by correct selection of φ_0 to π. In practice, a compromise must be found between carrier suppression and efficiency. It will be seen in the vector diagram and in the above equation that the output signal $\vec{a} + \vec{b}$ also possesses a phase modulation in addition to the amplitude modulation. This phase modulation can be suppressed by extending the single-pole modulator to the push-pull mode. The distortion factor of push-pull modulators is also less, which has been confirmed in the author's own experiments. In the case of a push-pull modulator, vector $\vec{a}$ also has a phase modulation component. The associated vector diagram is given in **Figure 2**.

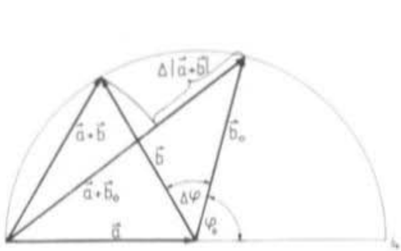


Fig. 1: Vector diagram for the output voltage of the single-pole modulator

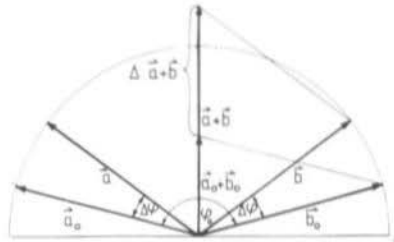


Fig. 2: Vector diagram for the output voltage of the push-pull modulator

The following is valid for the output signal of the push-pull modulator:

$$\hat{a} + \hat{b} = 2\,A\,(\cos \frac{\varphi_0 + 2\,\Delta\varphi \cos(2\,\pi\,f_m t)}{2})\,\exp\,[\,j\,(2\,\pi\,f_0 t + \frac{\varphi_0}{2})]$$

It will be clearly seen in Figure 2 and the above formula that the phase modulation of the output signal has been removed. However, for cost reasons only single-pole modulators are to be described whose 13 cm and 9 cm versions have been in operation with great success over a considerable period of time. To close this theoretical consideration, **Figure 3** shows a block diagram of both types of mixer.

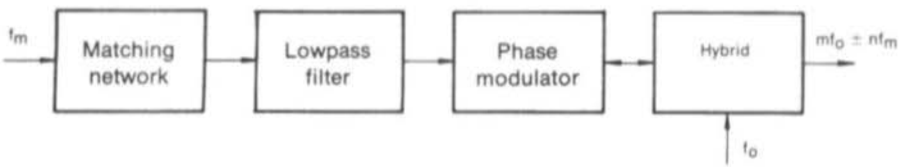


Fig. 3 a: Block diagram of the single-pole mixer

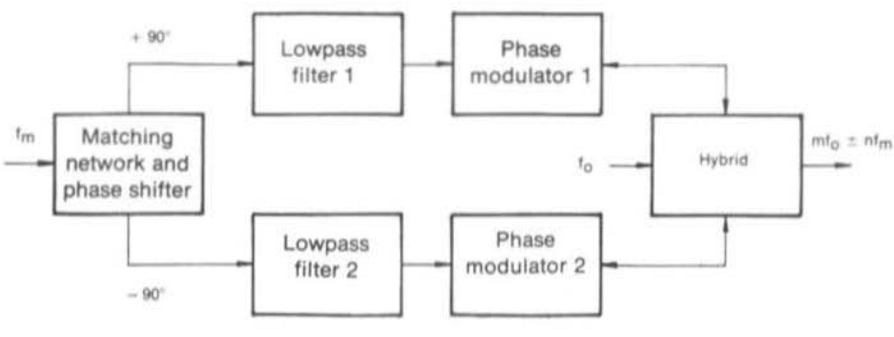


Fig. 3 b: Block diagram of the push-pull mixer

2. LINEAR TRANSMIT CONVERTER FOR THE 13 cm BAND

After considering the theoretical considerations of simple hybrid power-mixers in the previous section, the author is now to describe the practical construction of such a 13 cm transmit converter. For selectivity reasons, a local oscillator frequency of 2160 MHz is to be used together with an intermediate frequency of 144 MHz. For reproducibility reasons, it is advised for the transmit converter to be constructed in PC-board technology, using the low-loss PTFE glass fiber material Duroid RT 5870, with a thickness d = 1.57 mm as dielectric (5). Experiments using normal epoxy PC-board material resulted in considerably poorer results, and the author cannot recommend use of this material. The stripline circuits can easily be recalculated for other thicknesses (e.g. d = 0.79 mm). The complete circuit of the transmit converter is given in **Figure 4**; **Figure 5** shows the PC-board.

The components are as follows:

C 1:	90 pF plastic foil trimmer (red)
C 2:	1 nF ceramic disk capacitor
C 3:	Home-made capacitor, see **Figure 6**
Tr:	5 pF ceramic tubular trimmer
R:	100 kΩ
L 1:	2 turns of 1 mm dia. silver-plated copper wire wound on a 6 mm dia. coil former with core
D:	BXY 28 (Philips)

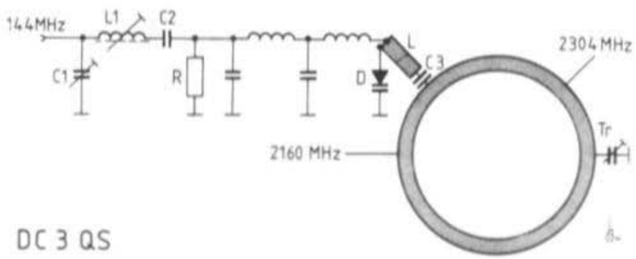


Fig. 4: Circuit of a linear power mixer for the 13 cm band

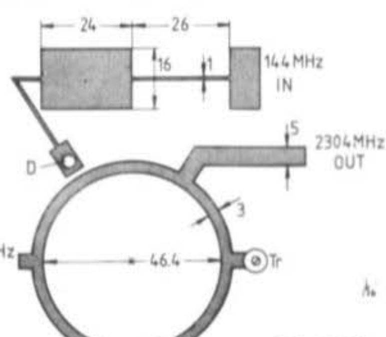


Fig. 5: PC-board of the linear power mixer for the 13 cm band

The LC-network comprising L 1 and C 1 is used as matching network for the intermediate frequency of 144 MHz. The subsequent series circuit of λ/4 lines operates as lowpass filter, and ensures that no 2160 or 2304 MHz energy is passed from the phase modulator to the IF circuit. The phase modulator comprises line L and varactor diode D. A BXY 28 diode is used here and diode types BXY 27, 38, 39, 40 provide far inferior results. The conversion of the phase modulation to the required amplitude modulation (modulation = conversion !) with suppressed carrier (2160 MHz) takes place in the connected hybrid. The carrier suppression and depth of modulation as a function of the phase angle φ_0 is aligned using Trimmer Tr. The fixed capacitors C 2 and C 3 are used as DC-isolating capacitors in order to provide the most favorable DC operating point of the varactor with the aid of resistor R. In addition to this, capacitance C 3 provides a defined termination of the IF network due to its high reactive impedance at 144 MHz. The PC-board construction of the mixer considerably simplifies construction. **Figure 6** shows details regarding the mounting of the varactor and the transition from BNC to micro-stripline; **Figures 7, 8, and 9** show photographs of the author's prototype.

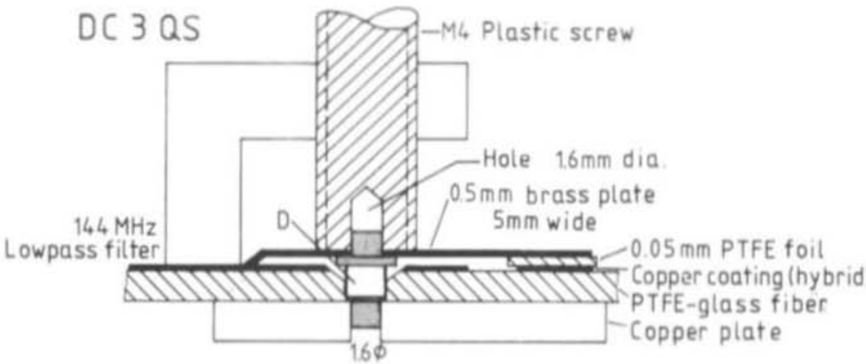


Fig. 6 a: Mounting of the varactor, line L and capacitor C 3

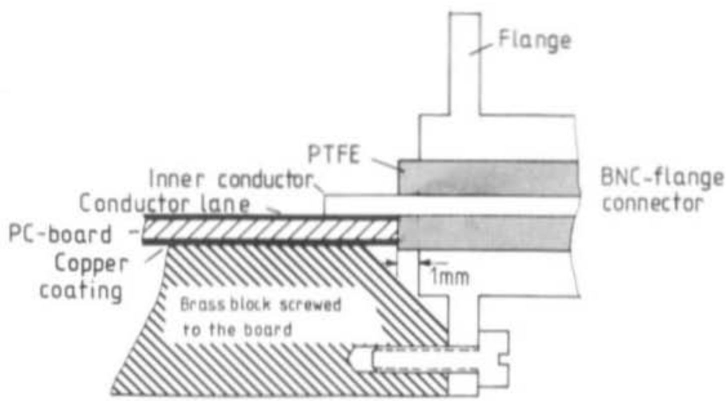


Fig. 6 b: Mounting of a BNC connector to the micro-stripline

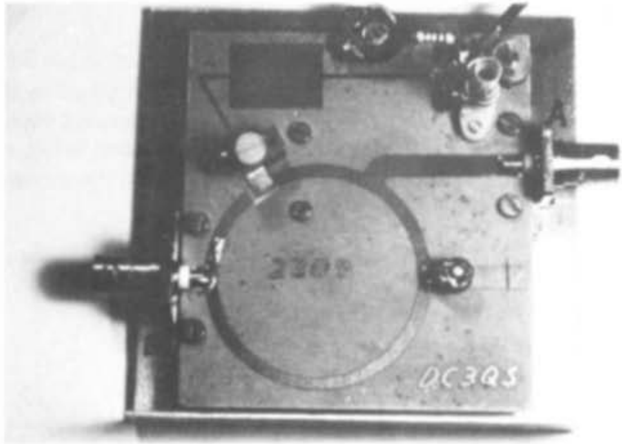


Fig. 7:
Photograph of the author's prototype transmit converter 144/2304 MHz

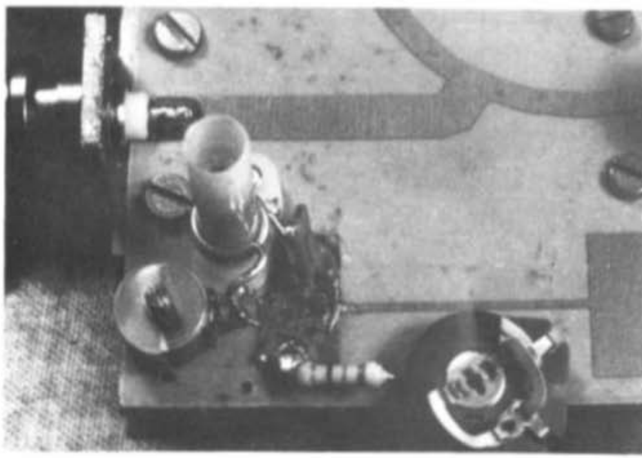


Fig. 8:
The 144 MHz matching network

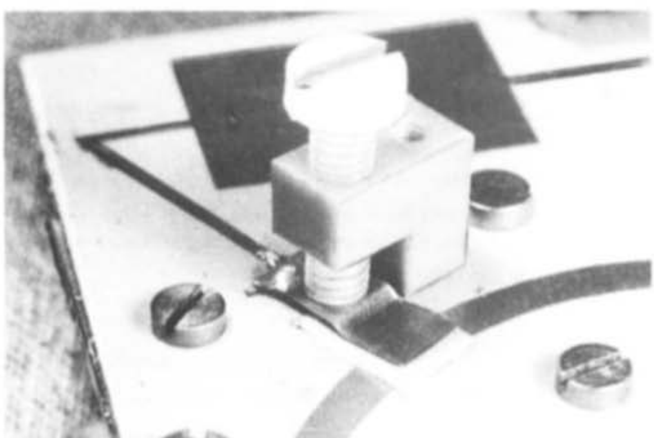


Fig. 9:
L and C 3 are depressed onto the varactor using a plastic screw

3. LOCAL OSCILLATOR SIGNAL FOR 2160 MHz

The most favorable relationship with respect to efficiency and spurious signal suppression is obtained when the 2160 MHz oscillator signal is obtained by doubling a clean 540 MHz signal twice. Only the two power-doublers are to be discussed here, since the 540 MHz signal was obtained in a DJ 4 LB oscillator circuit as described in (2), and amplified to 2 to 5 W in a modified DJ 3 SC strip similar to (6). The circuit of the subsequent doubler from 540 MHz to 1080 MHz is shown in **Figure 10**.

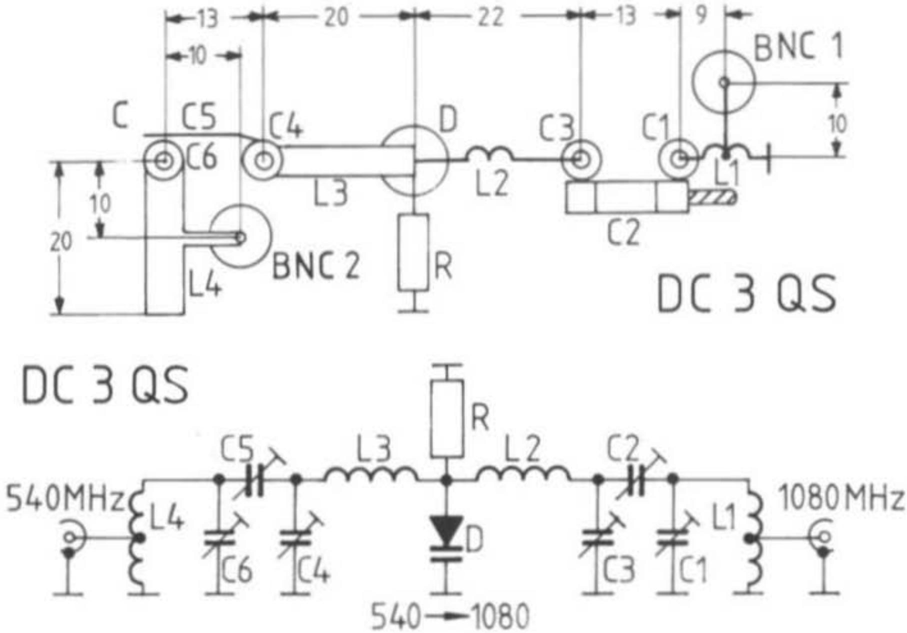


Fig. 10: Frequency doubler 540/1080 MHz, view from above and circuit diagram

The components are as follows:

C:	Coupling plate from 0.5 mm thick, 4 mm wide copper plate
L 1:	2 turns of 1 mm dia. silver-plated wire on a 6 mm former, self-supporting
L 2:	2 turns of 1 mm dia. silver-plated wire on a 5 mm former, self-supporting
L 3:	0.5 mm thick copper plate, width 4 mm, height 9 mm
L 4:	0.5 mm thick copper plate, width 5 mm, height 7 mm
R:	100 kΩ
D:	VEC 8711, BXY 35, BAY 66 (Philips)
All trimmers:	0.5 - 3.5 pF ceramic tubular trimmers
BNC 1, BNC 2:	BNC connectors for single-hole mounting

According to the varactor diode used, efficiencies in the order of 50 to 60% can be expected.

Two circuits are now to be described as doubler from 1080 MHz to 2160 MHz. In the case of the first version **(Figures 11 and 12)**, where coaxial input and output circuits constructed from PC-board epoxy material are used, efficiencies in the order of 50 to 80% can be obtained. The second version **(Figures 13 and 14)** is far simpler to construct, but the efficiency is only 30%.

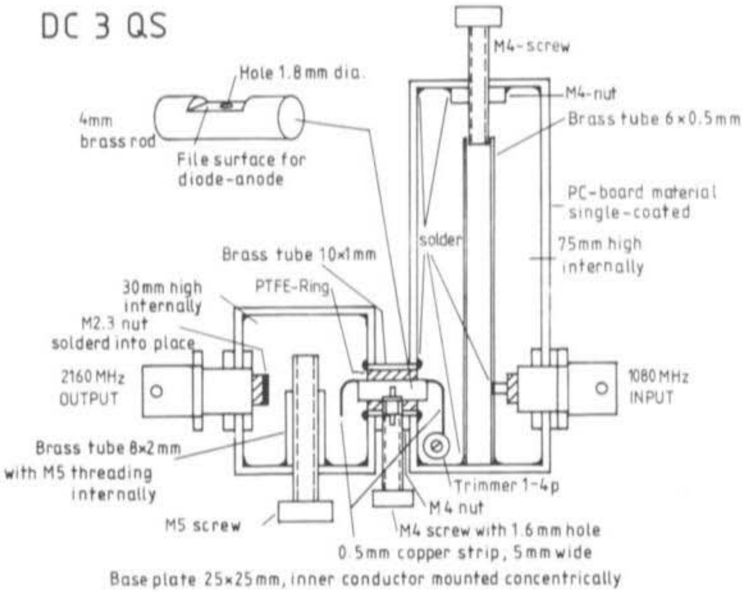


Fig. 11: Construction of a coaxial doubler 1080/2160 MHz

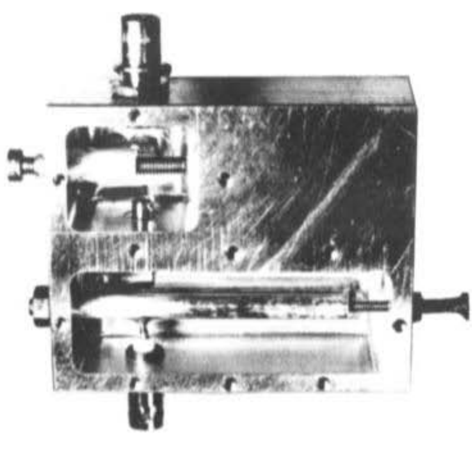


Fig. 12: Photograph of the author's prototype of a doubler constructed according to Fig. 11

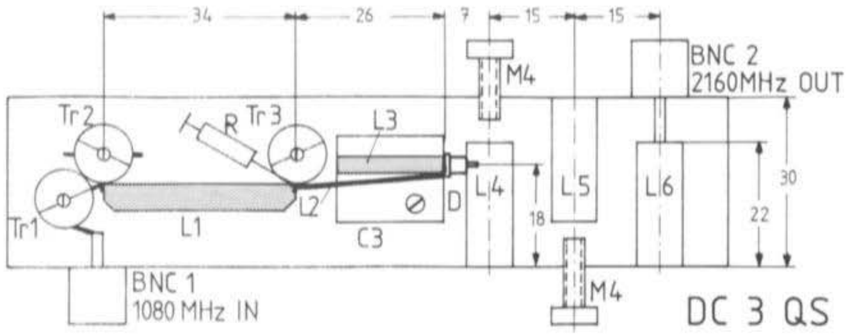
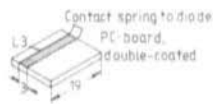


Fig. 13 a: Construction of the frequency doubler 1080/2160 MHz



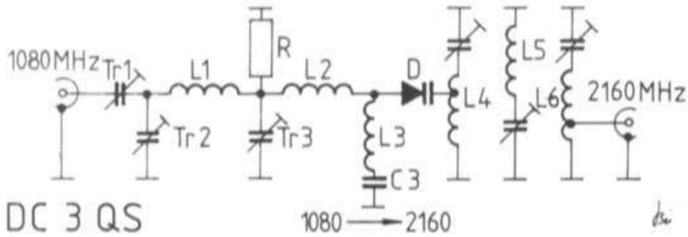


Fig. 13 b: Circuit diagram of the varactor doubler 1080/2160 MHz

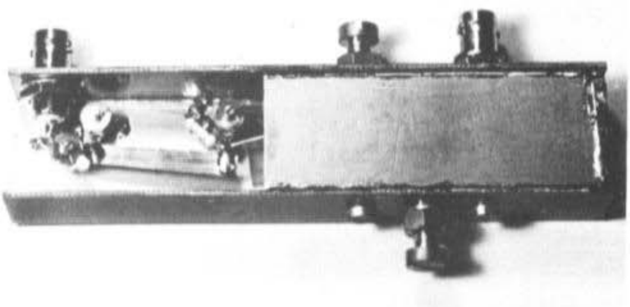


Fig. 14
Photograph of the author's prototype constructed as shown in Figure 13

Parts for the doubler given in Figure 13:

L 1:	0.5 mm thick copper plate, length = 34 mm, width = 5 mm, height = 6 mm
L 2:	1.5 mm dia. silver-plated copper wire, length = 25 mm, height = 6 mm
L 3:	see Figure 13c
L 4 - L 6:	8 mm dia. brass, length = 22 mm
Tr 1 - Tr 3:	Plastic foil trimmers; Tr 1: 22 pF (green), Tr 2: 12 pF (yellow), Tr 3: 6 pF (grey)
R :	approximately 20 kΩ
D :	BXY 27
BNC 1, BNC 2:	BNC connectors for single-hole mounting

Attention should be paid during construction of varactor multipliers that the currents of all required harmonics of the fundamental frequency are able to flow via the varactor diode. This demand is automatically fulfilled when using multipliers in a so-called parallel circuit (version 1). This is not the case with series doublers (version 2). In this case, the LC-networks comprising Tr 1, Tr 2, L 1, Tr 3, and L 2 are used to match the diode D to the input impedance of 50 Ω at the input frequency of 1080 MHz. Since this circuit represents a lowpass filter, no current can flow through the diode at a frequency of 2160 MHz. A considerable improvement in efficiency resulted after the series resonant circuit L 3 / C 3 was connected to the input side of the diode by using an open λ/4 stripline tuned to 2160 MHz. The most favorable tapping point on the first line of the three-stage interdigital filter should be selected according to the diode type used.

4. CONNECTION AND ALIGNMENT

The alignment of the mixer and multiplier will not be difficult if a clean 540 MHz signal is available. Using a simple power meter (terminating resistor with diode and/or coupling for a wavemeter), the first doubler (540 MHz to 1080 MHz) is aligned, followed by the second doubler (1080 MHz to 2160 MHz). The alignment of the first doubler should not be altered during the alignment of the second (parasitic oscillations !). For spurious signal considerations (spectral components due to the phase modulation), the mixer should not be used without a suitable filter (7) at the output. Simple, single-stage interdigital filters have been found very suitable for this. For alignment of the mixer, the local oscillator signal of 2160 MHz is connected, together with approximately 1 W power from a 2 m transmitter, and a suitable power meter connected via the output filter to the output. With the local oscillator signal at 2160 MHz present, the 2 m matching network comprising L 1 / C 1 is firstly aligned for minimum VSWR. A 2304 MHz signal should now be present at the output. Trimmer Tr is now aligned for maximum output power. Finally, the output circuit and the output coupling of the second doubler should be corrected. The alignment of the output filter can only be optimized when a spectrum analyzer is available. In the case of the author's 13 cm transmitter aligned without the aid of a spectrum analyzer, communication with DC 0 DA over a distance of 62 km took place immediately without any difficulties. Later measurements showed that the following values were exhibited when using the single-stage filter:

Input power:	540 MHz	33.0 dBm
Output power 1st doubler:	1080 MHz	30.0 dBm
Output power 2nd doubler:	2160 MHz	28.5 dBm
Output power mixer:	2304 MHz	24.8 dBm
Spurious signal:	2160 MHz	− 0.5 dBm
Spurious signal:	2016 MHz	− 15.0 dBm
Spurious signal:	2448 MHz	5 dBm

It will be seen from these specifications that an additional, or better filter (7) must be used when the mixer is to be directly connected to the antenna. If the mixer is to be followed by a selective power amplifier equipped with tubes (1), a good single-stage filter will be sufficient.

Finally, it should be mentioned that this mixer is also very suitable as ATV modulator due to its large IF bandwidth from DC to several hundred MHz. Successful A 5 / F 3 experiments made by the author in conjunction with the 10 GHz version of this mixer (which is to be described in a later edition) confirm this.

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TECHNOLOGY AND FREQUENCY PLAN FOR REPEATERS IN THE 23 cm BAND

by T. Morzinck, DD 0 QT

The West-German PTT has been issuing licences for 23 cm repeaters since 1976. Although these repeaters are still not very well known, they are of interest technically. They allow less qualified amateurs, and those with a limited budget to take part in microwave communications and gather experience with relatively simple means.

Such activity is also very important with respect to defending these amateur bands.

1. FREQUENCY PLANNING

The selection of the input frequencies was made in conjunction with the European band plan for the 70 cm band. The task was to allow communications using a varactor tripler in conjunction with a 70 cm FM transceiver. For example, repeater channel R 74 (70 cm: 431.150 MHz) is tripled to channel R 24 (23 cm: 431.150 MHz x 3 = 1293.450 MHz). This means that the channel spacing in the 23 cm band is three times that of the 70 cm band.

The repeater output frequencies are 33 MHz lower, in the range of 1260 MHz to 1262 MHz. This relatively large frequency spacing has a number of advantages in practical operation. It is possible, for instance, to monitor one's own signal over the repeater without problems.

2. ANTENNAS AND POLARIZATION

Until now, horizontal polarization has been selected for 23 cm repeaters, since such horizontally polarized omnidirectional antennas such as slots, Big Wheel (2), and halos are so compact at this frequency that they can easily be mounted on the roof of a car. However, horizontal polarization was mainly selected for operation together with fixed stations, whose antennas are usually horizontally polarized.

Most antennas on this band are usually sufficiently wideband, that they provide a good match in the transmit mode even at 1293 MHz. In the case of narrow-band Yagis, a drop-off of gain may be noticed in the receive mode at 1260 MHz.

3. COMMUNICATION

Mobile operation in the 23 cm band does offer some new characteristics, especially since the fading due to multi-path communications is extremely fast. Virtually no interference was found in the vicinity of Muenster in Northern Germany, and what little there was, was caused by sporadic interference from radar stations. With the exception of this, the band is very »peaceful«. It was, however, necessary for the repeater DB 0 YM to be provided with a »radar trap« to ensure that the repeater is not blocked for hours by such interference.

Communications over this repeater are very leisurely due to the relatively low activity. It is unfortunate that the repeater is not popular with many serious 23 cm amateurs who mainly operate SSB in a frequency range of between 1296 and 1298 MHz, since especially these amateurs would be able to help others to become active on this band, and give others the benefit of their experience.

4. ACCESSORIES

4.1. Transmitter

4.1.1. The most simple means of obtaining a transmit signal is to use a varactor tripler in conjunction with a 70 cm transmitter. Such varactors have been described in this magazine, e.g. the varactor tripler by DK 1 PN (1) which was very simple to construct. In addition to this, such triplers are also available on the market ready-to-operate.

4.1.2. If no 70 cm transmitter is available, it would be possible to firstly triple a 2 m signal in order to obtain the input frequencies on the 70 cm band. However, special attention must be paid to screening and filtering, since the fundamental frequency at 2 m will be 143.700 MHz, which is outside the amateur band. It is necessary to provide a filter (3) after the 70 cm/23 cm multiplier in order to ensure that any residual signal will not cause interference to 70 cm repeaters.

4.1.3. The most elegant manner would be to use a transmit mixer that converts the frequency range 144 to 146 MHz linearly to 1293 to 1295 MHz (4). It is only necessary to change the usual 96.000 MHz crystal to one of 95.75 MHz. It is also possible in this case for the oscillator frequency to be coupled out for a receive mixer, so that it is possible to monitor the input frequencies.

Usually, the first two possibilities will be sufficient for most applications. The three possibilities are given in the form of a block diagram in **Figure 1**.

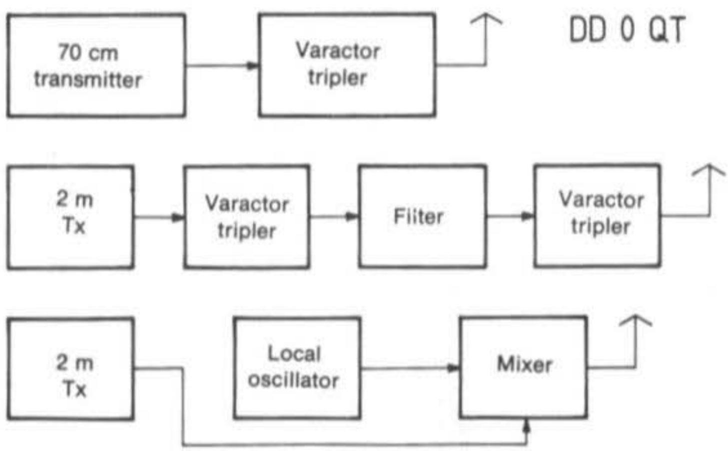

Fig. 1: Three possibilities of producing the transmit signal in the 23 cm band

4.2. Receiver

4.2.1. The most simple means is to use an available 23 cm/2 m converter. The somewhat lower frequency does not usually cause any problems since it is within the trimmer range of the inductances or capacitances.

A tremendous number of converters has been described, and it is only necessary to look through the last few years of VHF COMMUNICATIONS to find a suitable circuit.

However, it is advisable to redesign the hybrid rings for 1260 MHz.

The interdigital filter converter (5) is recommended for newcomers, since it will not offer any difficulties with respect to alignment, operation, and sensitivity. For those amateurs who are a little frightened of the highly mechanical concept, it is possible for the base plate and cover to be made from PC-board material, the side panels from brass plate (0.8 mm or thicker), and 10 mm holes drilled for the fingers. The fingers themselves are cut 5 mm longer than given, are placed through the holes in the side panels, and soldered on the outside (!) so that the finger is soldered to the side panel around the edge of the bar.

When using this method, it is very easy to orientate the fingers correctly, and it solves a lot of problems that would otherwise require lathing. The cover is then mounted into place and fixed, using a large soldering iron opposite to the fingers or tuning screws, and on the opposite side, e.g. at approximately 10 positions. This means that it is possible to remove it if it becomes necessary to exchange the diode.

4.2.2. If a 2 m receiver is not to be used, and one wishes to use a 70 cm receiver also for reception, a 23 cm/70 cm converter will be required. This converter can be combined with the varactor tripler to construct a complete transverter.

As far as this is known by the author, no converter has been described for conversion of 23 cm to 70 cm. A local oscillator frequency of 828 MHz (92.000 MHz x 9 = 828 MHz) is required to convert 1260 to 1262 MHz to 432 to 434 MHz. It may be of interest to know that the same local oscillator can be multiplied by four to obtain: 828 MHz x 4 = 3312 MHz: the required local oscillator frequency for 9 cm converters. It may not be possible to easily convert the well-known 23 cm/2 m converters for a 70 cm output. Those readers with a certain amount of experience could modify the DL 9 JU / DC 8 XB (6), which has been found very suitable for this application. Fundamentally speaking, attention should be paid that the bypass capacitors at the IF-side of the diode are not too large, in order to ensure that the intermediate frequency of 432 MHz is not bypassed. Only a low-noise amplifier with Pi-matching (BFR 91, BFT 66) is suitable as IF-preamplifier. By the way, the output frequency in the 23 cm band will not correspond to that of the 70 cm channel. It would therefore be necessary to provide the 70 cm receiver with a further crystal in the frequency range of 432 to 434 MHz. This crystal can be saved when the transverter is only to be used on one frequency. The crystal for the local oscillator frequency can be found using the following equation:

$$\text{Crystal frequency for local oscillator} = \frac{\text{Required 23 cm frequency} - \text{Output frequency 70 cm}}{9}$$

If there is sufficient interest, the author would be willing to describe a suitable converter for 23 cm / 70 cm operation. If you are interested in such a converter, please inform the editors, our your national representative.

4.3. Transmit Switching

Of course, it is also possible for a good coaxial relay to be used, however, they tend to be very expensive in this frequency range. Another possibility is to use a second antenna, which is in many cases cheaper than the relay and is very compact at this frequency.

For those readers that already possess an antenna, it is possible to provide a simple antenna splitting circuit. This will provide sufficient isolation between transmitter and receiver at the relatively low transmit powers used for repeater operation. This is constructed using an inter-digital filter and providing a further output connector at the center finger where the diode is usually connected. This is then connected to the antenna, one resonator is then aligned to the transmit frequency, and the other to the receive frequency. Transmitter and receiver are then connected to the associated finger. The principle of operation is shown in **Figure 2**.

In addition, the antenna splitting filter also operates as bandpass filter, which means that it may even be possible to delete the required bandpass filter. Of course, such a simple filter cannot be compared with the diplexers used in repeater stations. However, it has been found more than suitable for replacing a coaxial relay.

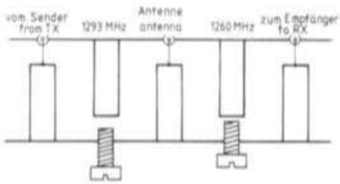

Fig. 2:
A simple transmit/receive splitter

5. BRIEF DESCRIPTION OF THE 23 cm REPEATER DB 0 YM

The basic conception of a repeater is well known. The concept of DB 0 YM will be seen in **Figure 3**. The repeater has been constructed with amateur means. The repeater was designed especially for its application and does not possess any modified professional communication modules. The following information is to be given for those readers that may be interested in using the receiver or transmitter concept for their own applications:

5.1. Receiver

The principle used in the receiver of obtaining the two required frequencies from a single oscillator, has proved itself in practice. In this manner, one is able to save several stages and the alignment is simpler. In addition to this, the first IF is spaced quite some way from any high-level frequencies in the 2 m band. Although the receiver in DB 0 YM is provided with a wideband crystal filter suitable for a 50 kHz spacing, it was found that this was not sufficient for the instability and frequency drift of 70 cm transceivers used in conjunction with varactor triplers due to temperature fluctuations.

This meant that it was necessary to provide an AFC-circuit in the repeater receiver despite use of the wide IF.

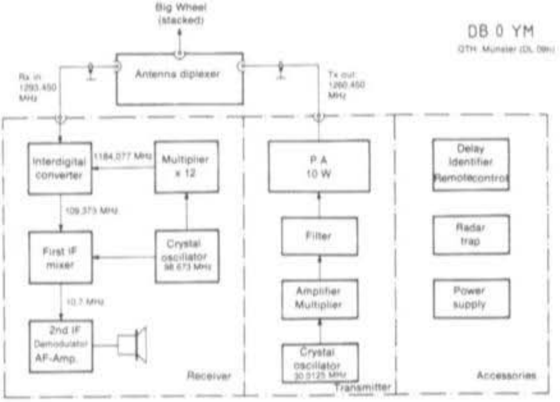

Fig. 3:
The given frequencies refer to those used at the repeater DB 0 YM (channel R 24)

Other frequencies can be calculated as follows:

Receiver: Input frequency minus 10.7 MHz : 13 = Crystal frequency.

Transmitter: See notes given in section 5.2.

Constructors should therefore examine the temperature response of their local oscillator and, if necessary, install an AFC circuit, especially when filters for a 25 kHz spacing are used in the 2nd IF. Considerable temperature fluctuations are present for mobile and portable communications, which can cause considerable drifts !

An available 2 m receiver strip without local oscillator can be used for the first IF-mixer and 2nd IF-demodulator modules (Figure 3). It is only necessary for the circuits aligned to 144 MHz or 135 MHz to be realigned for 110 or 100 MHz. This is achieved by increasing the number of turns by one or two, or increasing the parallel capacitors. This can easily be achieved with the aid of a dip-meter. The IF-circuits in the converter can also be modified in the same manner.

The crystal oscillator frequency can be coupled out after the crystal oscillator from the local oscillator module and fed via a buffer and – if required – a further amplifier stage to the previous 135 MHz input of the receive module.

Special attention must be paid to the crystal oscillator. If this oscillator is only 500 Hz from its nominal frequency, this frequency error will be 6.5 kHz in the receive mode ! This is far too high when a crystal filter for a bandwidth of 20 kHz is used in the receiver.

5.2. Transmitter

A 70 MHz crystal was used in order to keep the circuit as simple as possible, and to provide a clean, low-harmonic output signal. A clean output signal was obtained without problems using bandpass filter coupling. It was somewhat more difficult to obtain a clean frequency modulation of the crystal oscillator. For this reason, the transmitter has been modified in the meantime to use a 35 MHz crystal. It is now possible to obtain sufficient frequency deviation.

5.3. Accessories

The only special accessory is the so-called »radar trap«. This trap does not blank radar interference, but only differentiates between a radio carrier and interference, and only switches on the transmitter when the former is present.

5.4. Antenna Diplexer

The design of this diplexer was based on the simple antenna splitter described in section 4.3., and was further developed in several prototypes. The filter consists of several resonators in each passband direction and possesses considerably higher stopband values. These repeater diplexers are, of course, much more extensive, and are only designed for the special applications in a repeater station. For this reason, they are not to be described in detail here.

At this point, the author would like to thank the following amateurs for their assistance and advice during the construction of this repeater: D. Aland, DB 1 QX; R. Heidemann, DC 3 QS, B. Lenz, DC 5 DF; and W. Franke, DF 5 QW.

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A FM TRANSCEIVER FOR THE 2 m BAND
Part 2: The Transmitter

by J. Kestler, DK 1 OF

4. THE TRANSMIT AMPLIFIER MODULE

According to (1), the output power of the synthesizer is 14 mW, which means that a three-stage amplifier would be sufficient to obtain an output power of 10 W. This assumes that an average power gain of 10 dB per stage can be obtained. The author decided, however, to use a four-stage amplifier in order to compensate for transistor spread and to allow the provision of neutralizing networks. It is then possible for higher power transistors to be used, and to obtain output power levels in the order of 25 to 40 W.

4.1. Circuit of a 10 W Class-C Amplifier

The circuit diagram of the FM transmit amplifier is given in **Figure 11**. The 145 MHz signal to be amplified is fed to the 50 Ω connection Pt 361 and is fed via the transformation link C 1 - L 1 - C 2 to the first stage of the amplifier. Resistor R 1 is connected at the point of highest impedance and is provided to accept any excess drive power. The selection of its resistance value is discussed in the »Alignment Procedures« in section 4.4. The input transistor T 1 operates in class A and thus provides the maximum possible gain, as well as a relatively drive-independent input impedance. The network comprising L 2, C 3, and C 4 is used to match the second amplifier stage; transistor T 2 is biased to class B with the aid of a base-voltage divider, which reduces its drive power requirement (compared with class C). The operating voltage is fed from Pt 362 via L 3.

The subsequent matching link L 4 - C 5 - C 6 then transforms the signal to an output impedance of 50 Ω at Pt 363. This 50 Ω point between the two portions of the amplifier has the following advantages:

Firstly it is thus possible for both portions to be used separately, since each has a 50 Ω input and output (the PC-board is designed so that it can be separated at this point). Secondly, the alignment process is simplified greatly (as will be described later). The disadvantage is the extra components required: one inductance and two trimmers, which could be saved if a direct coupling between the stages was used.

The transformation link comprising C 7 - C 8 - L 5 matches the 50 Ω input Pt 364 to transistor T 3. Resistors R 2 and R 3 are provided for neutralization; although they reduce the gain of the stages somewhat, they do avoid pulling and hysteresis effects and provide neutralization.

The matching between the two class-C power amplifier transistors T 3 and T 4 is made with the aid of L 7, C 9 and C 10. The operating voltage is fed from Pt 365 and Pt 366 and is by-passed and filtered with the aid of chokes RFC 2 and RFC 4. It is not necessary for these voltages to be fed via the contact on the transmit-receive relay, since T 3 and T 4 will not have any current drain under non-drive conditions. The output network comprises L 9, C 11 and C 12; the task of this network is to transform the output impedance of T 4 to 50 Ω at output Pt 367 of the amplifier. Resistor R 4 is provided to limit any collector voltage peaks at T 4 to permissible levels if the transmitter were to be used accidentally without antenna.

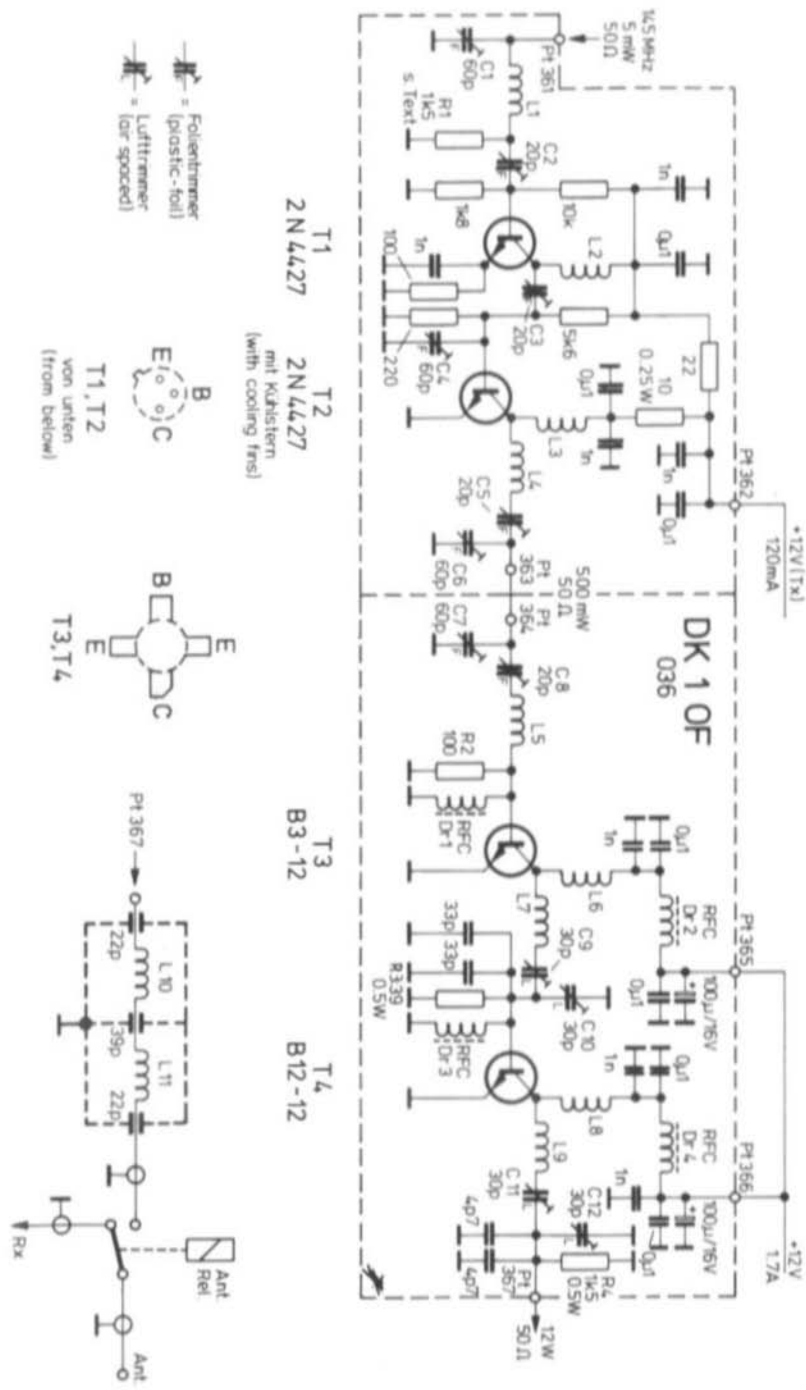

Fig. 11: Circuit diagram of the FM transmit-amplifier

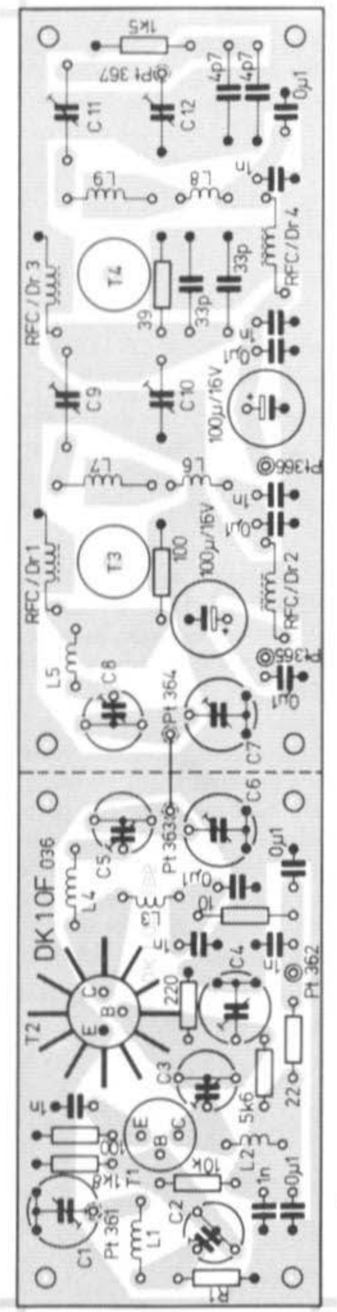

Fig. 12: Component locations on PC-board DK 1 OF 036

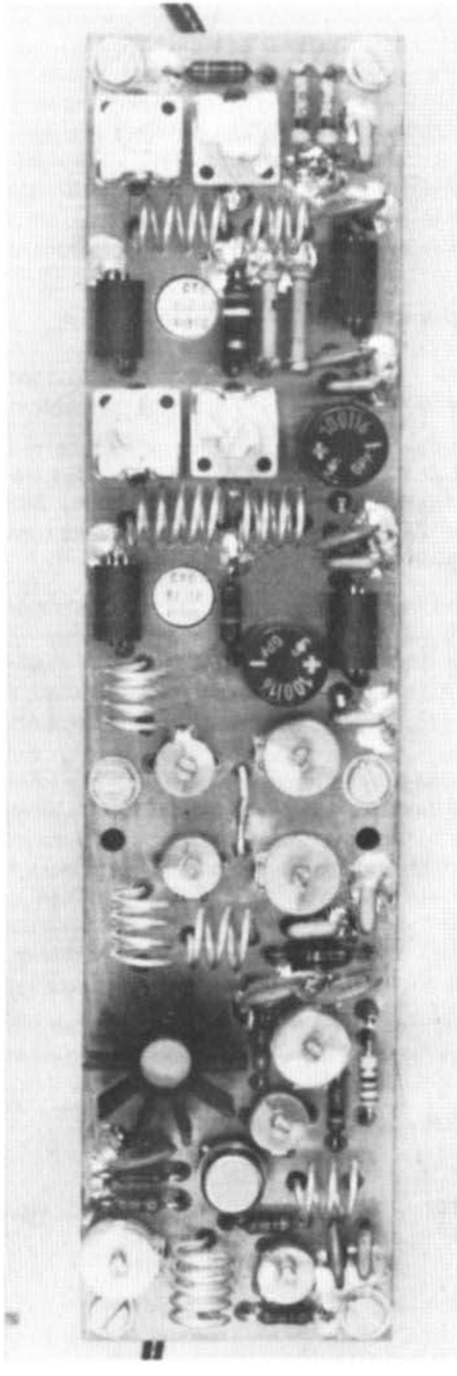

Fig. 13: Photograph of the author's prototype

Of course, a considerable harmonic spectrum will be generated due to the non-linear, class-C operation of the power amplifier. This will not be sufficiently attenuated in the output network. Since the first harmonic (290 MHz) is in the military aircraft communication band of 225 to 400 MHz, the second in the 70 cm amateur band, and the third and fourth in the UHF-TV band between 470 MHz and 860 MHz, this means that an additional low-pass filter is necessary. A suitable filter is given in the lower right-hand corner of Figure 11, however, no room is provided on the PC-board for this. It is advisable for this filter to be combined with the antenna relay to form a separate module, since the mismatch of the transmit-receive relay can also be compensated for during the alignment of the filter.

4.2. Construction of the Transmit Amplifier

To aid construction, the author has designed the double-coated PC-board DK 1 OF 036, whose dimensions are 170 mm x 40 mm; no through-contacts are provided.

The upper side of the board is in the form of a continuous ground surface. All connection wires of the components that are grounded are soldered to the upper and lower side of the board (exception: the plastic foil trimmers). All other holes are countersunk slightly (approx. 3 mm dia.), in order to avoid short-circuits to the ground surface. The component locations are given in **Figure 12**.

Due to the relatively high RF currents, ceramic disk capacitors of at least 8 mm in diameter and 0.8 mm leads should be used for the 1 nF bypass capacitors in the driver and output stage. Such capacitors may be difficult to obtain due to the trend to miniaturization. As alternative, chip-type disk capacitors can be used, and provided with connection wires of 1 mm diameter, silver-plated wire. A capacitance of 470 to 680 pF is sufficient.

Although the described amplifier does not indicate any tendency to oscillation even without being completely screened **(Figure 13:** Photograph of the author's prototype), it is advisable to do so. Not only would this avoid a direct radiation of unwanted harmonics, but also such RF radiation could cause interference to the synthesizer. It is therefore advisable to provide the board with an approximately 30 mm high screening panel around the board made from 0.5 mm thick metal plate, and to feed in the operating voltages via feed-through capacitors ($\geqq$ 1 nF). The spacing between the lower side of the board and the lower edge of the screening is given by the dimensions of the driver and output transistor, and amounts to 4.5 mm.

Thin coaxial cable (RG-174) is used for the input and output connections of the amplifier. This cable is soldered to the corresponding points on the lower side of the board.

4.3. Special Components

T 1, T 2:	2 N 4427 (Motorola, RCA); T 2 with cooling fins
T 3:	B 3-12 (CTC)
T 4:	B 12-12 (CTC)
C 1, C 4, C 6, C 7:	Plastic foil trimmer 60 pF, 10 mm dia. (yellow)
C 2, C 3, C 5, C 8:	Plastic foil trimmer 20 pF, 7.5 mm dia. (green)
C 9, C 10, C 11, C 12:	Air-spaced trimmer 30 pF (Tronser)

L 1, L 7:	5 turns of 1 mm dia. silver-plated copper wire, inner diameter 6 mm, self-supporting, length to match holes on the PC-board
L 2, L 3, L 8:	3 turns, otherwise as L 1
L 4, L 5, L 6, L 9:	4 turns, otherwise as L 1
L 10, L 11:	72 nH; 3 turns, 10 mm long, otherwise as L 1
RFC 1, RFC 4:	Six-hole core, 2.5 turns (VK 200)

4.4. Alignment of the Transmitter

The wire bridge between Pt 363 and Pt 364 should be left disconnected at first. Instead of this, a piece of coaxial cable is connected from Pt 363 to a VHF power meter, or, if not available, via a SWR-meter to a 50 Ω terminating resistor for power indication. Connection Pt 361 is now connected with the output of the synthesizer and Pt 362 via an ammeter to the operating voltage of + 12 V. Trimmers C 1 to C 6 are now adjusted in sequence alternately for maximum output power. The resistance value of R 1 is selected so that approximately 500 mW is provided at output Pt 363 at the given input power level. This is the case at a current drain at Pt 362 of approximately 120 mA. If R 1 is deleted, an output power of 1 W will be obtained at a drive power of 5 mW.

The SWR-meter is now connected between the synthesizer and input of the power amplifier. Trimmers C 1 and C 2 are now aligned alternately so that a minimum VSWR, and maximum output power (or current drain) occur at the same position.

It is favorable for alignment of driver and power amplifier if the output power can be measured selectively since the high harmonic content will cause the power meter or SWR-meter to indicate incorrect values. For this reason, it is advisable to align the lowpass filter comprising L 10 and L 11 at this time.

To achieve this, the input of the lowpass filter is connected via a SWR-meter to the output of a 145 MHz source having a low-harmonic content (e. g. signal generator, or another 2 m transmitter). It is also possible to use the synthesizer described in (1), if the SWR-meter is sufficiently sensitive. The output of the antenna relay should be connected to a 50 Ω terminating resistor or a good antenna (VSWR $\leq$ 1.2). L 10 and L 11 are now aligned alternately for best input-VSWR by pulling or depressing the windings of these inductances.

The output of the synthesizer is now again connected to Pt 361, the bridge Pt 363 - 364 connected, and the air-spaced trimmers C 9 to C 12 aligned in the following manner:

C 9: 80 % capacitance, C 10: 45 %, C 11 and C 12: 60 %.
Connection Pt 367 is now connected to the input of the low-pass filter, and a VHF-power meter or SWR-meter/50 Ω termination is connected to the antenna output. Connections Pt 362, Pt 365 and Pt 366 are now connected in parallel and to a power supply, which should, if possible, possess current limiting at approximately 2 A. If this is not the case, the alignment should be commenced at an operating voltage of 8 V (only Pt 365, and Pt 366). Trimmers C 7 and C 8 are aligned alternately for maximum current drain of the driver, and C 9 to C 12 in sequence alternately for maximum output power.

This is followed by carrying out the same process at an operating voltage of 10 V, and finally again at 12 V. The adjustment of C 1 to C 6 is not altered.

Of course, attention must be paid during the alignment process that the two power transistors T 3 and T 4 are sufficiently cooled. This means that an appropriate heat sink and heat-conductive paste must be used.

5. THE MODULATOR

Since PLL synthesizers possess a voltage-controlled oscillator (VCO), whose output frequency is more or less linearly dependent on the tuning voltage, it is easy to directly modulate this oscillator. This is achieved by either providing a separate, additional varactor diode, or to superimpose the modulation (AC) voltage on the tuning voltage (DC) from the phase-comparator stage. The first method provides a relatively constant frequency deviation over the whole frequency range to be selected, and the second has the advantage that it can be obtained with a minimum of components. Of course, it should be considered that any modulation represents an interference to the balance of the phase-control circuit and will cause balancing (transient) processes. It is now a question of the design of the control circuit, especially the loop filter, to determine the effects of such balancing processes, i.e. which »loop response« the PLL-circuit possesses. Generally speaking, the following points are valid:

- The shorter the time constants in the control circuit, the better the lower modulation frequencies will be controlled (limited).
- The PLL-loop should react aperiodically in the case of rapid interference voltages, which means that it should not require a long transient time with a dampened periodic oscillation, since this would emphasize certain modulation frequencies.
- High-frequency (and noise) components on the tuning voltage line to the VCO will generate a large interference-frequency deviation, since these will not be controlled. This will cause a considerable increase of the output signal spectrum.

The synthesizer described in (1) exhibits a favorable response with respect to interference of the control circuit; it does not require long transient times and is therefore suitable for direct modulation. **Figure 14** gives the measured modulation characteristic of the control circuit as given by the designer: this shows the required AF-modulation voltage as a function of the modulation frequency. The curve indicates a favorable characteristic within the AF-bandwidth of 300 Hz to 3 kHz, and no unwanted resonances are indicated.

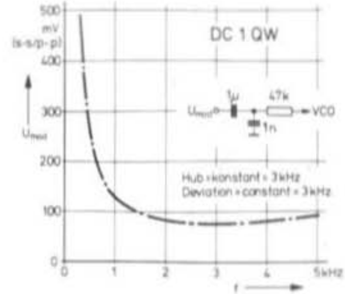

Fig. 14: Modulation characteristic of the synthesizer according to DC 1 QW

The task of the modulator is now to compensate for the frequency response of the synthesizer in the audio bandwidth of 300 Hz to 3 kHz. Frequencies in excess of 3 kHz should be attenuated. The block diagram of the modulator is given in **Figure 15**, and **Figure 16** gives a circuit of the frequency-determining networks. The calculation of the required component values was made using a BASIC-programme in conjunction with a digital computer HP 9830.

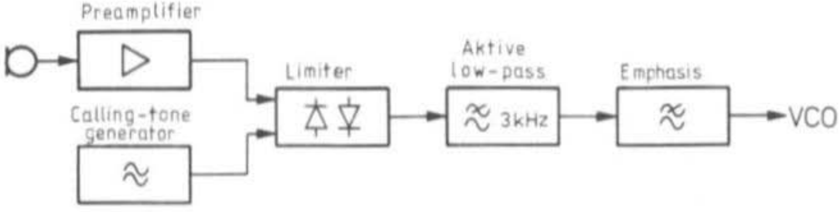

Fig. 15: Block diagram of the modulator

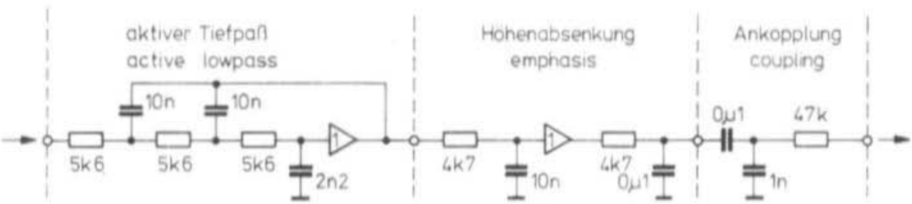

Fig. 16: The frequency-determining RC-links

5.1. Circuit of the Modulator

As can be seen in **Figure 17**, the audio voltage from the microphone is fed via connection Pt 371 to a RF-filter comprising R 1 and C 1, and then to the input stage T 1 of the preamplifier which is connected in the form of an emitter follower. This stage has an input impedance of approximately 150 kΩ, which means that it is also possible for high-impedance (crystal) microphones to be used. The other three stages of the preamplifier equipped with T 2, T 3, and T 4 are DC-coupled and provide a voltage gain of approximately 150 times, which means that an input sensitivity of approx. 5 mV (RMS) results in a frequency deviation of 5 kHz. The amplified audio voltage and the signal from the calling-tone generator are combined in the network via R 12 - C 5 and R 13 - C 6 and then fed to the subsequent limiter.

The circuit of the calling-tone generator seems somewhat extensive (Wien-bridge oscillator), however, only one inductance is required and it possesses a very good frequency and amplitude stability, which is especially important in conjunction with mobile operation with its extreme temperature fluctuations. The frequency-determining components are C 7, C 8, R 15 / R 16, and R 17 / R 18. The sine-wave signal is fed via C 11 to diode D 1 and D 2, which form a proportional DC-voltage which influences to a larger or lesser degree FET T 5, which is switched as a variable resistor.

This ensures that the feedback of amplifier I 1 is adjusted automatically so that oscillation amplitude is held constant. Furthermore, this results in less distortion of the output voltage.

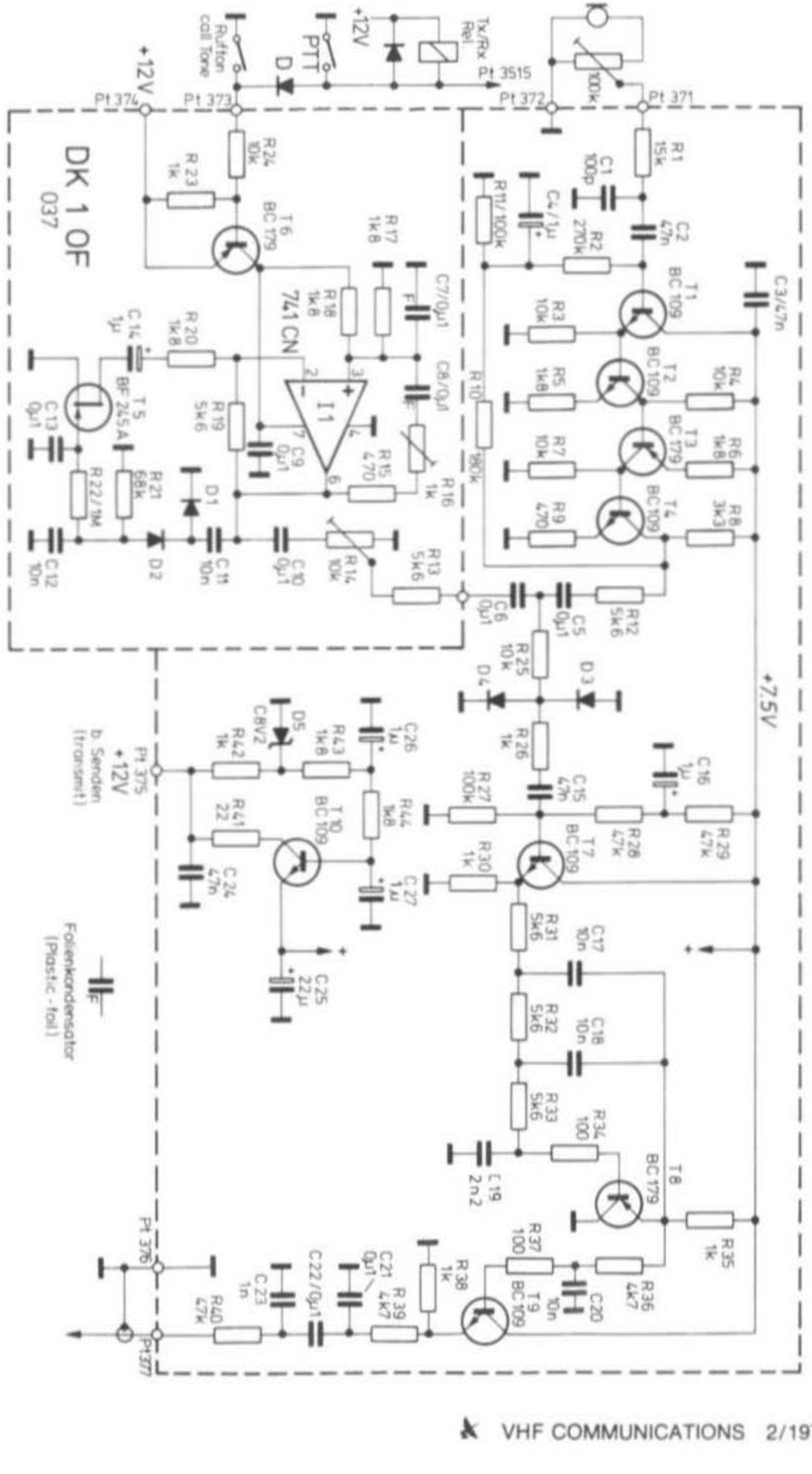

Fig. 17: The complete modulator comprising preamplifier, calling-tone generator, limiter and AF-filter

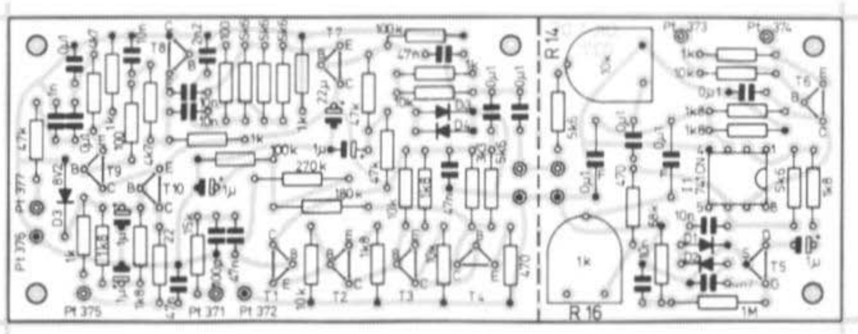

Fig. 18: Component locations on PC-board DK 1 OF 037

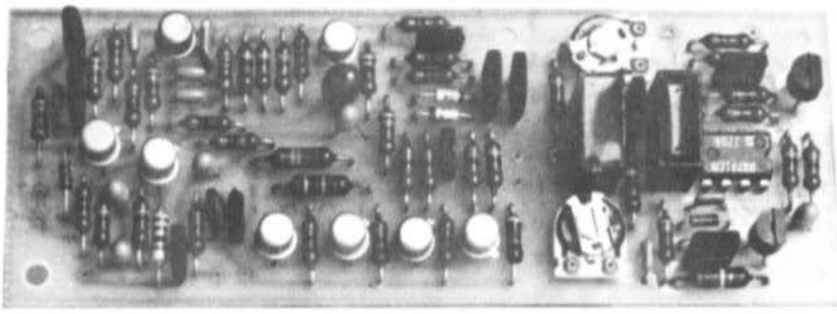

Fig. 19: Photograph of the author's prototype

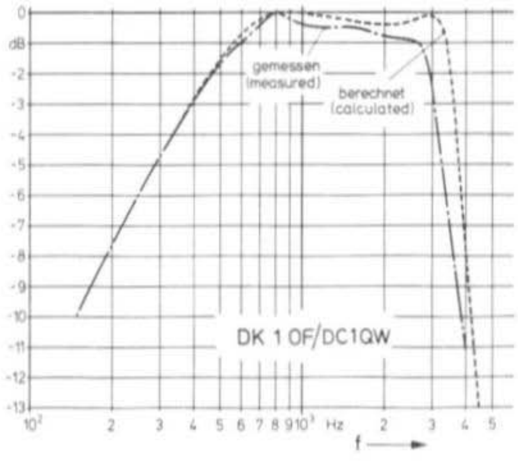

Fig. 20: Audio frequency response of modulator and synthesizer

The most favorable frequency deviation can be adjusted with the aid of R 14, and the exact frequency with R 16.

The operating voltage for the calling-tone generator is switched on via T 6 on depressing the calling-tone button; at the same time, the transmitter is switched on via diode D.

The limiter comprises R 25 and diodes D 3 and D 4 (Figure 17). The threshold voltage of the diodes will not be achieved under normal drive conditions; only when an excessive input voltage is obtained will the limiting be actuated, and will avoid an excessively high frequency deviation of the transmitter. The harmonics of the AF-signal generated in this manner are suppressed in the subsequent active low-pass filter which is decoupled with the aid of T 7. This filter comprises C 17 to C 19, and R 31 to R 33 as frequency-determining components, as well as the emitter follower T 8.

Components R 36 and C 20 present the first RC-link for suppression of the high-frequency AF components; this is decoupled via T 9 from the second link comprising R 39/C 21. Capacitor C 22 provides a galvanic insulation of the modulator from the tuning line to the VCO, and resistor R 40 provides a sufficient output impedance of the modulator so that the dynamic behaviour of the synthesizer is not affected.

The voltage supply to the modulator is made via Pt 375 and pass transistor T 10, which is provided with a base voltage that is stabilized in D 5 and well filtered in R 43 / C 26 and R 44 / C 27. It is not necessary for Pt 375 to be connected to the operating voltage in the receive mode; all time constants in the modulator are designed so that it is ready for operation a maximum of 0.1 s after depressing the PTT-button.

5.2. Construction

For construction of the modulator, the author has designed the single-coated PC-board DK 1 OF 037. The dimensions of this board are 110 mm x 40 mm. The component locations are given in **Figure 18**. If the calling-tone generator is not required, it is possible to cut the PC-board along the dashed line. **Figure 19** shows a photograph of the author's prototype.

5.3. Special Components

T 1, T 2, T 4, T 6, T 7, T 9, T 10:	BC 109, BC 413 or similar
T 3, T 6, T 8:	BC 179, BC 415 or similar
T 5:	BF 245 A
I 1:	741 CN or TBA 221 B
D 1, D 2, D 3, D 4:	1 N 4151 or similar silicon diode
D 5:	BZX 97 / C 8 V 2 or other 8.2 V zener diode
R 16, R 19:	Trimmer potentiometers, for horizontal mounting, spacing 10/5 mm
Resistors ≤ 100 kΩ:	10 mm spacing
Resistors > 100 kΩ:	12.5 mm spacing

Electrolytics:	Tantalum drop types ≧ 10 V DC
C 7, C 8:	0.1 μF plastic foil capacitor, spacing 10 mm
All other capacitors ≦ 0.1μF:	Ceramic disk, or multi-layer capacitors

5.4. Final Notes

The alignment of the described module is limited to adjustment of the amplitude and frequency of the calling-tone generator. Otherwise, the circuit should operate immediately. Although a considerable amount of calculation was used to obtain a correct, i.e. flat frequency characteristic of the modulator plus synthesizer, this is mainly dependent on the subjective consideration of the transmitted signal, and is largely dependent on the microphone used. Ceramic or crystal microphones often tend to emphasize the higher frequency components, whereas dynamic types often provide too little of the higher frequency components, which often leads to the comment that the partner station reports a loud signal that is, however, difficult to understand.

The described modulator can be matched relatively easily to an available microphone. If more low-frequency components are required, the values of the coupling capacitors C 2, C 5, C 15 and C 22 should be increased; the higher modulation frequencies can be emphasized by varying the value of C 21.

Finally, the calculated and measured overall frequency response is given in **Figure 20**; the deviations between these are caused by component tolerances of the frequency-determining components.

6. REFERENCES

(1) G. Heeke: A Synthesizer for the 2 m Band in C-MOS Technology
VHF COMMUNICATIONS 10, Edition 3/1978, pages 130 - 144

(2) J. Kestler: A FM-Transceiver for the 2 m Band – Part 1: The Receiver
VHF COMMUNICATIONS 11, Edition 1/1979, pages 44 - 53

»SUEDWIND« – FOR MOBILE AND DF-APPLICATIONS

by J. Becker, DJ 8 IL

1. MOBILE APPLICATIONS

A simple ⊔-shaped bracket is sufficient for use as mounting clamp for mobile operation. This metal plate should be covered with plastic or material to avoid scratching the case. The mechanical stop on inserting the transceiver into the bracket is the 13-pin connector (1). Very little room is required, which means that the mounting bracket can be mounted in the dashboard behind a suitable cutout. The transceiver can be inserted and removed easily if it protrudes by approximately 3 to 4 cm.

Figure 28 gives the wiring diagram together with connection details for a mobile microphone and shows how a car radio can be used as audio amplifier. Interference was not present inspite of the fact that the transceiver was mounted in the direct vicinity of the ignition key, windscreen wiper motor, and cable harness. Special means were taken during the construction of the »SUEDWIND« transceiver against such interference: A high-quality monolythic crystal filter mounted on the RF/AF board in a carefully screened position is able to suppress the wideband interference so that no problems are presented despite the simple LC-discriminator (2, 3). The voltage stabilizer with a very high stabilizing factor, and the RF filtering (1) at various points in the circuit also filter out any interference signals that are fed in via the operating voltage.

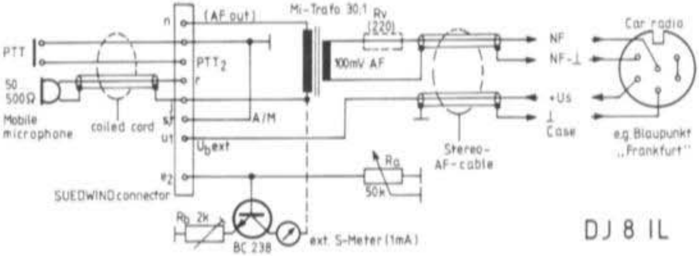

Fig. 28: Wiring diagram for using the »SUEDWIND« transceiver as mobile station. It is also possible for the other sensor keys A and Z for squelch, calling tone and scanning to be remote-controlled in the same way as the PTT.

The 30 : 1 microphone transformer is driven back to front (see Fig. 28) and is not only used for level matching to the AF input of the car radio, but also used for isolation of the audio circuits of both units and thus avoiding any unwanted ground loops. The transformer should be magnetically screened. In the case of some car radios, it is possible for the volume control to also be used to fade in another loudspeaker. The base of the volume control is grounded via a resistor (22 Ω). This point is often fed to the AF input connector for the cassette recorder, or traffic information decoder. A resistor (R_V of e.g. 220 Ω) can be used to match the AF level when connecting in this manner.

The interference level provided by the mobile antenna is greater than would be the case in open country or when operating from the fixed station. This is especially the case in dense city traffic. It is therefore advisable to use an additional squelch trimmer on the mobile mount, especially when a fixed resistor $R_S = 22\ k\Omega$ (1) was installed into the transceiver. The dashed lines given in Figure 28 also show how a larger, external S-meter can be connected, if required.

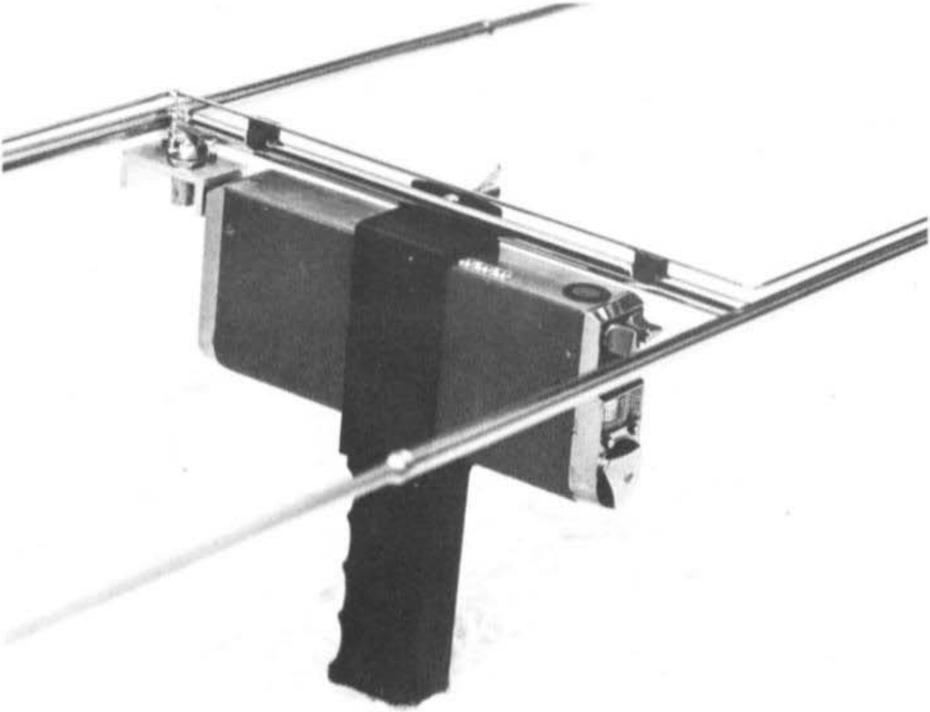

Fig. 29: Photograph of the »SUEDWIND« when used as DF-receiver together with a HB 9 CV antenna

2. DF-RECEIVER

The compact dimensions of the »SUEDWIND« transceiver allow a handy DF-receiver to be constructed. **Figure 29** shows a photograph of a simple construction for this application: The transceiver is mounted in a 40 mm wide framework made from 2 x 1 mm aluminium plate, covered with plastic foil. This framework provides an eyelet for mounting the HB 9 CV antenna. Below the transceiver, a grip from a film-camera light is mounted with the aid of a photostand nut.

At the director end of the HB 9 CV antenna, the SO 239 connector was exchanged for a 40 x 25 x 17 x 3 mm large aluminium bracket which is then provided with the interconnection plug for the antenna connection of the »SUEDWIND« transceiver. As can be seen in Figure 29, an additional BNC-connector was provided in parallel, so that the antenna could still be used for other applications. In order to obtain the same frequency response of the antenna, it was necessary to decrease the value of the 18 pF capacitor to 15 pF.

A good DF-receiver must either have a very large S-meter range of approximately 110 dB (4), or the S-meter should be adjusted electronically so that various field strength ranges can be adjusted (5). The range of the S-meter in the »SUEDWIND« transceiver (2) can be increased by 3 to 4 S-points (21 dB) for direction finding under high signal levels, if transistor T 104 (1) is short-circuited, thus reducing the sensitivity of the receiver.

The free, sixth position of the volume switch can be used for this as is shown in the circuit given in **Figure 30**: The contact wafer of the multi-position switch is disconnected at position »x«. A miniature resistor R_h on the conductor side of the board allows the transmitter to be heard, even in this position.

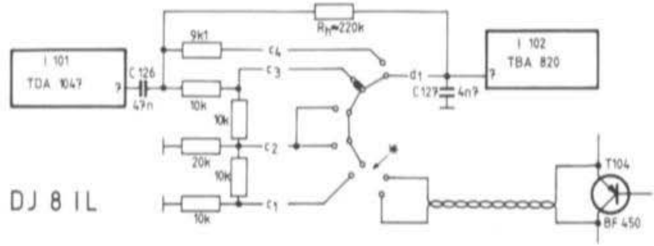

Fig. 3: Installation of an electronic 21 dB attenuator for DF at high signal levels

A power amplifier equipped with the integrated module Philips BGY 35 is to be described in VHF COMMUNICATIONS. This module allows the output power of the »SUEDWIND«, or other similar FM walky-talky to be increased to 15 to 20 W, and, at the same time increases the spurious and harmonic rejection to more than – 60 dB.

3. REFERENCES

(1) J. Becker: »SUEDWIND« – A 2 m FM Hand-Held Transceiver with 80 or 396 Channel Synthesizer and Touch-Key Operation – Part 1: Circuit Description VHF COMMUNICATIONS 10, Edition 4/1978, pages 194 - 212

(2) J. Becker: »SUEDWIND« – A 2 m FM Hand-Held Transceiver with 80 or 396 Channel Synthesizer and Touch-Key Operation – Part 2: Construction, Wiring and Alignment VHF COMMUNICATIONS 11, Edition 1/1979, pages 2 - 16

(3) J. Kestler: A FM-Transceiver for the 2 m Band – Part 1: The Receiver VHF COMMUNICATIONS 11, Edition 1/1979, pages 44 - 53

(4) DF-Receiver »HUNTER 2« manufactured by Semco

(5) M. Schmaußer: A DF-Receiver for the 2 m Band Equipped with Integrated Circuits, Crystal Filter and S-Meter VHF COMMUNICATIONS 10, Edition 4/1978, pages 213 - 217

ATTENUATORS FOR POWER MATCHING

by E. Wiedemann, DL 8 XI

One often faces the problem that the output power from an exciter is too high to drive the transverter, power amplifier, or frequency multiplier. When no modifications are to be made to the exciter (1), attenuators will be required. The calculation and design, as well as the practical construction of unbalanced attenuators of Pi and T-types are to be described in the following article. It will be shown how the calculated resistance values having the desired power rating can be realized using composite carbon resistors with power ratings of 0.5 W which are readily available on the market.

1. VOLTAGE AND POWER RATIO, IMPEDANCES

Radio equipment with coaxial inputs and outputs usually have an impedance of 50 Ω, 60 Ω, or 75 Ω. It is necessary to know the impedance value for our application, since the attenuator must be designed for the same impedance. Any interconnection cable used must also exhibit the same impedance so that no reflections occur.

The calculation of the attenuators is made using the voltage ratio U_{in}/U_{out}, where U_{in} is the voltage at the input of the attenuator, and U_{out} is the voltage at its output. This voltage ratio can be taken from the diagram in **Figure 1** if the power ratio P_{in}/P_{out} is known.

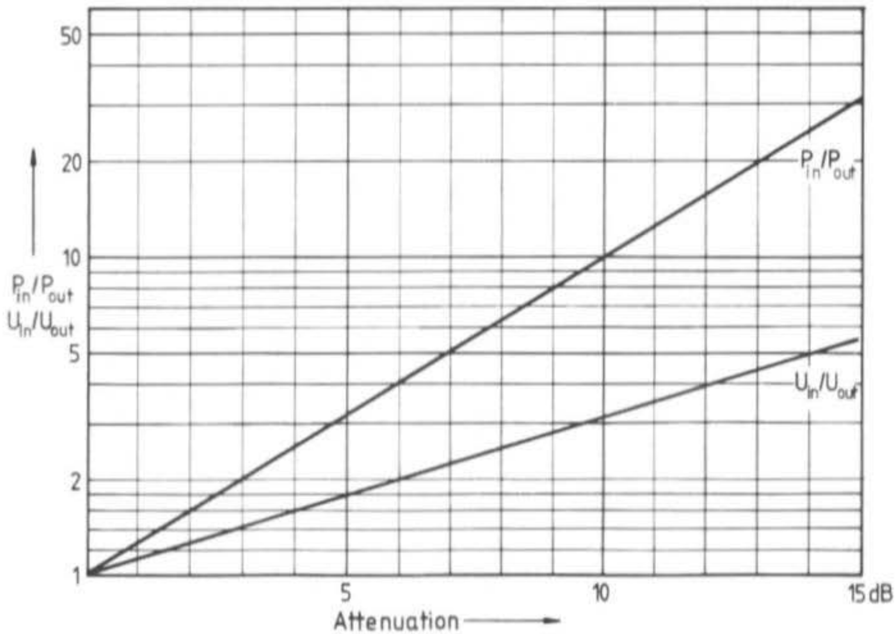

Fig. 1: Conversion of voltage or power ratios into attenuation values (in dB)

This is best seen in an example:
An exciter provides an output power of 25 W, however, the linear amplifier only requires a drive level of 10 W. The problem is thus for which ratio the attenuator should be designed ?

The power ratio is 25/10 = 2.5; this value is selected on the vertical scale (ordinate) in the diagram, and this line is followed in a horizontal direction until the P_{in}/P_{out} line is reached. From this point, a line is traced vertically down to the U_{in}/U_{out} line, from here the line is then traced back to the ordinate horizontally where the following can be read off: U_{in}/U_{out} = 1.6; furthermore, it is possible to read off the attenuation value of 4 dB.

2. DESIGN OF THE ATTENUATORS

The attenuators are passive four-poles, having a defined input and output resistance. At this point, let us speak of resistances, and not of impedances, since we are talking about real resistances – without reactive component –. The two most popular practical constructions are in the form of a Pi or T-circuit, which are now to be discussed.

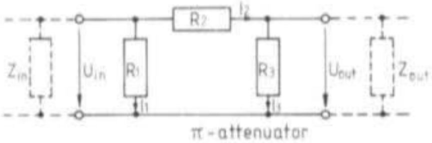

Fig. 2: A Pi-Attenuator

2.1. Pi-Circuits – Resistance Value and Rating

Figure 2 shows an unbalanced attenuator using a Pi-circuit. The resistance values can be calculated easily using the following, simple equations:

$$R_1 = R_3 = Z_{out} \times \frac{n+1}{n-1} \tag{1}$$

$$R_2 = Z_{out} \times \frac{n^2-1}{2n} \tag{2}$$

R in Ω
Z_{out} = Output Impedance = Input Impedance of the equipment to be connected (in Ω)
n = U_{in}/U_{out} = Voltage ratio (from Figure 1)

In order to calculate the voltage rating of the three resistors, it is necessary to know the input voltage U_{in}, the output voltage U_{out} and the voltage drop in the attenuator:

$$U_{in} = \sqrt{P_{in} \times Z_{in}} \tag{3}$$

$$U_{out} = \frac{U_{in}}{n} \tag{4}$$

$$\Delta U = U_{in} - U_{out} \tag{5}$$

with U in V, P in W and Z in Ω.

The rating of the resistors can be calculated as follows:

$$P_{R_1} = \frac{U_E^2}{R_1} \qquad (6)$$

$$P_{R_2} = \frac{\Delta U^2}{R_2} \qquad (7)$$

$$P_{R_3} = \frac{U_A^2}{R_3} \qquad (8)$$

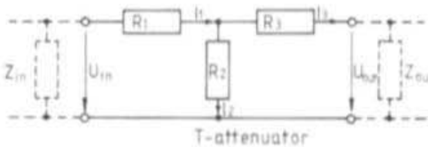

Fig. 3: A T-Attenuator

2.2. T-Circuits – Resistance Values and Rating

Figure 3 shows an unbalanced attenuator using a T-circuit. The resistance values of this circuit can also be calculated easily:

$$R_1 = R_3 = Z_{out} \times \frac{n-1}{n+1} \qquad (9)$$

$$R_2 = Z_{out} \times \frac{2\,n}{n^2-1} \qquad (10)$$

with R in Ω, $n = U_{in}/U_{out}$ = Voltage ratio,
and Z_{out} = Impedance in Ω

We require the voltages for calculation of the rating:
The input voltage U_{in} is calculated from the given power P_{in} and the impedance, or input impedance Z_{in} (in this case $Z_{in} = Z_{out}$):

$$U_{in} = \sqrt{P_{in} \times Z_{in}} \qquad (11)$$

The output voltage results from the required attenuation as:

$$U_{out} = \frac{U_{in}}{n} \qquad (12)$$

The voltage drop in the attenuator is:

$$\Delta U = U_{in} - U_{out} \qquad (13)$$

Equations 11 and 3, 12 and 4, as well as 13 and 5 are identical.

In order to keep the calculation principle as clear as possible, the currents are to be calculated firstly and the power rating afterwards:

$$I_1 = I_2 + I_3 \tag{14}$$

$$I_2 = \frac{U_{in} + U_{out}}{2\,R_2 + R_1} \tag{15}$$

$$I_3 = \frac{U_{in} - I_2\,(R_1 + R_2)}{R_1} \tag{16}$$

It is therefore necessary to firstly calculate I_2, followed by I_3, and finally I_1. It is now possible for the power dissipation ratings to be calculated:

$$P_{R_1} = I_1^2 \times R_1 \tag{17}$$

$$P_{R_2} = I_2^2 \times R_2 \tag{18}$$

$$P_{R_3} = I_3^2 \times R_3 \tag{19}$$

At this point we should mention that the calculation is not as bad as it seems. In practice, the calculation is quite easy, as will be seen in the following two examples.

3. PRACTICAL EXAMPLES

3.1. Calculation of a Pi-Attenuator

Requirements: Power attenuation 10 dB
Output power from the transmitter 20 W, $Z_{in} = 60\ \Omega$
Output impedance $Z_{out} = 60\ \Omega$

Voltage ratio $U_{in}/U_{out} = n$; taken from Figure 1: $n = 3.2$
(or calculated: $U_{in}/U_{out} = \sqrt{10} = 3.16$)

$$U_{in} = \sqrt{P_{in} \times Z_{in}} = \sqrt{20 \times 60} = 34.64\ \text{V}$$

$$U_{out} = \frac{U_{in}}{n} = \frac{34.64}{3.16} = 10.96\ \text{V}$$

$$\Delta U = U_{in} - U_{out} = 34.64 - 10.96 = 23.68\ \text{V}$$

$$R_1 = R_3 = Z_{out} \times \frac{n+1}{n-1} = 60 \times \frac{3.16+1}{3.16-1} = 116\ \Omega$$

$$R_2 = Z_{out} \times \frac{n^2-1}{2\,n} = 60 \times \frac{3.16^2-1}{2 \times 3.16} = 85\ \Omega$$

$$P_{R_3} = \frac{U_{in}^2}{R_1} = \frac{34.64^2}{116} = 10.34\ \text{W}$$

$$P_{R_2} = \frac{\Delta U^2}{R_2} = \frac{23.68^2}{85} = 6.6\ \mathrm{W}$$

$$P_{R_3} = \frac{U_{out}^2}{R_3} = \frac{10.96^2}{116} = 1.04\ \mathrm{W}$$

$$\text{Total} = 17.98 \approx 18\ \mathrm{W}$$

3.2. Selection of the Resistors

Assuming that only resistors having a power rating of 0.5 W are available, the following parallel values will result:

$R_1 = 116\ \Omega$ $\qquad\qquad P_{R_1} = 10.34\ \mathrm{W}$

this will result in at least $\frac{10.34}{0.5} = 21$ resistors.

The following resistance value will result:
116 x 21 = 2436 Ω; the next highest nominal preferred value is 2700 Ω. When parallel-connecting 23 resistors of 2700 Ω, the resulting resistance value will be 117 Ω (this value is within the tolerance limits of the resistors). Since the attenuator will probably be fed from both ends, it will be necessary for R_3 to be calculated in the same way as R_1. However, if one end is defined as input, it will be possible to select R_3 for the actual power rating:

$R_2 = 85\ \Omega$ $\qquad\qquad P_{R_2} = 6.6\ \mathrm{W}$

Required parallel resistors: $\frac{6.6}{0.5} = 14$ pieces;

Required resistance value: 85 x 14 = 1190 Ω, nominal preferred value 1200 Ω;

$$\text{Resulting resistance value} = \frac{1200\ \Omega}{14} = 85.7\ \Omega$$

$R_3 = 116\ \Omega$ $\qquad\qquad P_{R_3} = 1.04\ \mathrm{W}$

Parallel resistors: $\frac{1.04}{0.5} = 2$

Required resistance value: 2 x 116 Ω = 232 Ω, nominal preferred value 270 Ω

Resulting resistance value $R_3' = \frac{270}{2} = 135\ \Omega$

This value is not sufficiently accurate.
It is necessary for a further resistor R_{P_3} to be connected in parallel:

$$R_{P_3} = \frac{R_3 \times R_3'}{R_3' - R_3} = \frac{116 \times 135}{135 - 116} = 824\ \Omega$$

A resistor of 860 Ω is selected.

This results in the following total resistance:

$$R_3 = \frac{Rp_3 \times R_3'}{Rp_3 + R_3'} = \frac{860 \times 135}{860 + 135} = 116.7\ \Omega$$

3.3. Calculation of a T-Attenuator

Requirements: An attenuator, for matching an exciter to a power amplifier: Input power 2.5 W; Z = 60 Ω

Exciter: Output power 12 W; Z = 60 Ω

The required attenuation is obtained as follows:

$$\frac{P_{in}}{P_{out}} = \frac{12}{2.5} = 4.8; \quad 10 \log 4.8 = 6.8 \text{ dB (or from Figure 1)}$$

This results in the following voltage ratio from Figure 1:

$$n = U_{in}/U_{out} = 2.19 \text{ (calculated } U_{in}/U_{out} = \sqrt{4.8} = 2.19)$$

$$U_{in} = \sqrt{P_{in} \times Z_{in}} = \sqrt{12 \times 60} = 26.8\ \text{V}$$

$$U_{out} = \frac{U_{in}}{n} = \frac{26.8\ \text{V}}{2.19} = 12.24\ \text{V}$$

$$R_1 = R_3 = Z_{out} \times \frac{n-1}{n+1} = 60 \times \frac{2.19-1}{2.19+1} = 22.38\ \Omega$$

$$R_2 = Z_{out} \times \frac{2\,n}{n^2-1} = 60 \times \frac{2 \times 2.19}{2.19^2-1} = 69.23\ \Omega$$

$$I_2 = \frac{U_{in} + U_{out}}{2 \times R_2 + R_1} = \frac{26.8 + 12.24}{2 \times 69.23 + 22.38} = 0.243\ \text{A}$$

$$I_3 = \frac{U_{in} - I_2\,(R_1 + R_2)}{R_1} = \frac{26.8 - 0.243\,(22.38 + 69.23)}{22.38} = 0.203\ \text{A}$$

$$I_1 = I_2 + I_3 = 0.243 + 0.203 = 0.446\ \text{A}$$

$$P_{R_1} = I_1^2 \times R_1 = 0.446^2 \times 22.38 = 4.452\ \text{W}$$

$$P_{R_2} = I_2^2 \times R_2 = 0.243^2 \times 69.23 = 4.088\ \text{W}$$

$$P_{R_3} = I_3^2 \times R_3 = 0.203^2 \times 22.38 = 0.922\ \text{W}$$

Sum: 9.462 W ≈ 9.5 W.

The selection of the resistors is made as in example 1 using available, or easy-to-obtain resistors.

4. PRACTICAL CONSTRUCTION OF ATTENUATORS

It is immaterial for practical operation whether the attenuators use a Pi or T-circuit. This depends mainly on the resistors available. For this reason, it is often advisable to calculate the resistance values of both types. It is usually more favorable to select a Pi-circuit since higher resistance values result, that can be obtained by parallel connection of several individual resistors.

4.1. Resistors

For attenuators that are to be used at frequencies in excess of 30 MHz, it is necessary to use capless, non-helical composite carbon resistors (low-inductance). However, since these are very expensive and not readily available, carbon resistors having the usual tolerance and power ratings are to be used for all applications, except when precision attenuators are required. In most cases, the parallel connection of several resistors usually compensates for the tolerances. Since readily available resistors are generally helically made, in other words are inductive, they are used in a bifilar manner: this means that the colour coding is mounted at one end with one resistor and on the other end with the next resistor and so on, assuming that the resistors are from the same manufacturer. This results in an inductive compensation and thus a certain amount of frequency independence of the attenuator. Of course, this becomes more and more important, the higher the frequency.

The resistors should be rated for the RMS power level. When only SSB (A 3 j) is to be used without clipper, the resistors can be rated for approximately half the PEP output power.

4.2. Construction

If the attenuator is not to be installed within the equipment it is advisable to install the resistors in a small metal box (tin plate or brass) and to use the case as ground surface. The leads of the resistors should be as short as possible. An aluminium case is not very suitable due to the difficulties of soldering.

Another possibility would be to bend a metal strip in the form of an U, provide this with the required holes and solder the connectors and resistors into place. The upper part of the case can then be provided in the form of a further metal plate and soldered into place **(Figure 4)**.

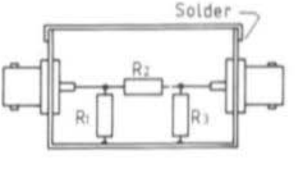

Fig. 4: Example of a practical attenuator

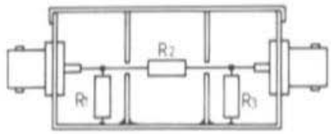

Fig. 5: An attenuator for higher frequencies with screened assembly for high attenuation values

If the attenuator is to be used at higher frequencies, it is recommended that the individual resistor networks are accommodated in different chambers in order to avoid that RF-energy is able to »jump« the series resistor. **Figure 5** shows such an attenuator. In this case it is necessary to use a completely sealed case.

4.3. Testing

The attenuation value should be measured after completing the attenuator. This can be achieved by connecting a defined DC-voltage of say 10 V to the input connector and measuring the voltage present at the output. The ratio of U_{in}/U_{out} can be found in the diagram given in Figure 1, which will then give the appropriate attenuation value.

However, it is necessary during these measurements that the output is terminated with a real output impedance Z, otherwise one will measure incorrect values.

It is not usually necessary to have high-precision attenuators, which means that deviations of ± 2 to 5 % will not be important. Further details were given in (2).

5. REFERENCES

(1) H. J. Dierking: Reducing the Output Power of SSB
Transistorized SSB-Transmitters and Transverters
VHF COMMUNICATIONS 9, Edition 1/1977, page 37

(2) J. Wasmus and G. Laufs: A Calibrated Attenuator
VHF COMMUNICATIONS 1, Edition 3/1969, pages 169 - 173

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 2/1979 of VHF COMMUNICATIONS

VK 5 QR	**SSB on 13 cm Band using Envelope Elimination and Restoration – 21 MHz/28 MHz SSB-Processor with Subsequent Multiplication by Six**		**Ed. 2/1979**
PC-Board	VK 5 QR 001	Double-coated, without thru-contacts	DM 24.—
Semiconductors	VK 5 QR 001	3 germ. and 6 silicon transistors, 7 diodes, 3 Schottky-Low Power TTL-ICs, 1 integrated stabilizer	DM 54.—
Minikit 1	VK 5 QR 001	7 coilformers w/core, 4 ferrite beads, 1 ferrite choke, 4 miniature chokes, 1 metal case, 1 feedthru cap., 2 BNC connectors	DM 34.—
Minikit 2	VK 5 QR 001	34 carbon resistors, 40 ceramic caps., 1 tantalum capacitor	DM 29.—
Crystal	31.583 MHz	series resonance, 3rd overtone, HC-6/U	DM 25.—
Kit	**VK 5 QR 001**	**complete with the above parts**	**DM 158.—**

DK 1 OF	**FM-Transmitter for 2 m Matching Receiver DK 1 OF 034/035 and Synthesizer from DC 1 QW**		**Ed. 2/1979**
Transmitter			
PC-Board	DK 1 OF 036	Double-coated, w/out thru-contacts, w/plan	DM 12.—
Semiconductors	DK 1 OF 036	2 2 N 4427, 1 each B 3-12 + B 12-12	DM 79.—
Minikit	DK 1 OF 036	8 plastic-foil trimmers, 4 airspaced trimmers, 4 chokes, 18 ceramic caps, 2 alu. electrolytics, 3 feedthru caps, 3 special feedthru caps for lowpass filter, 11 resistors, 1 cooling fins	DM 59.—
Kit	**DK 1 OF 036**	**complete with the above parts**	**DM 148.—**
Modulator			
PC-Board	DK 1 OF 037	single-coated, with plan	DM 10.—
Semiconductors	DK 1 OF 037	11 transistors, 1 IC, 5 diodes	DM 17.—
Minikit	DK 1 OF 037	2 trimmer pots., 42 resistors, 19 ceramic caps, 6 tantalum caps, 2 plastic foil caps.	DM 28.—
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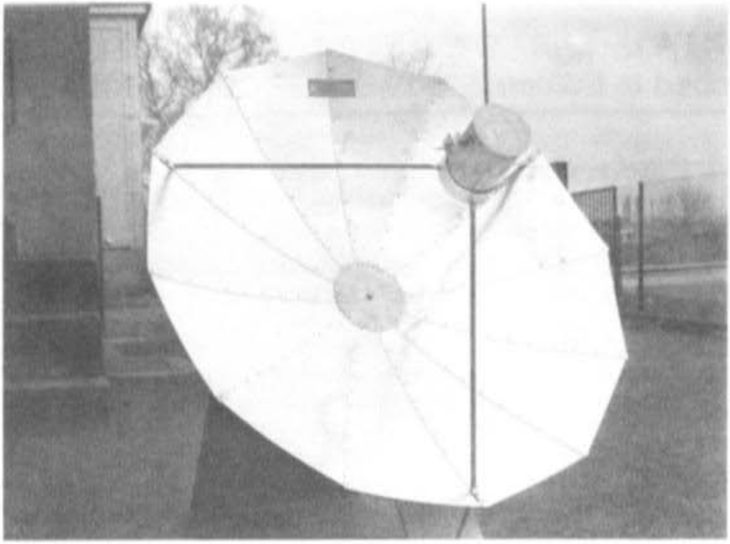

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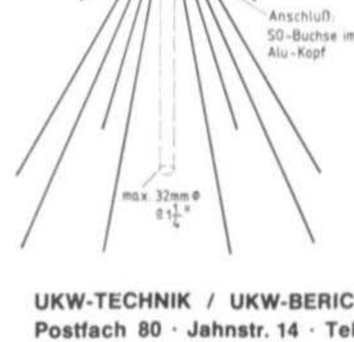
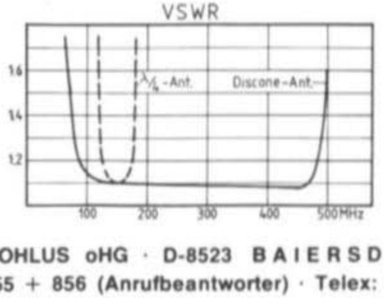
WIDEBAND OMNIDIRECTIONAL DISCONE ANTENNA



- Frequency range: 80 - 480 MHz
- Gain: 3.4 dB / λ/4
- Impedance: 50 Ω
- Power rating: 500 W
- Polarisation: Vertical
- Connection: SO 239 socket in the head
- VSWR: < 1.5 : 1
- Weight: 3 kg
- Dimensions: Height: 1.00 m / Diameter: 1.30 m
- Material: Aluminium
- Mounting: Antenna head is put onto a 32 mm (1 ¼") dia. mast and secured by a screw.



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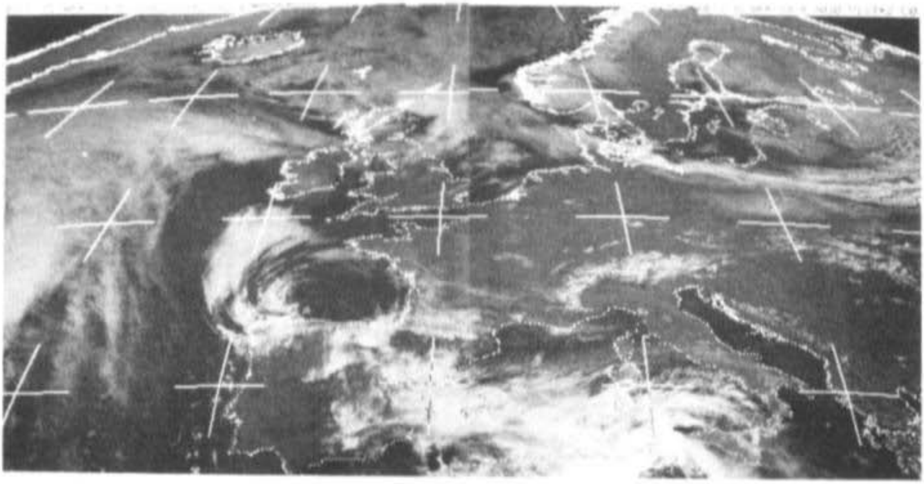

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Details on METEOSAT reception were given in edition 3/78 and 4/78 of VHF COMMUNICATIONS. The basic principle of operation of the two types of image processing were also explained in these editions.

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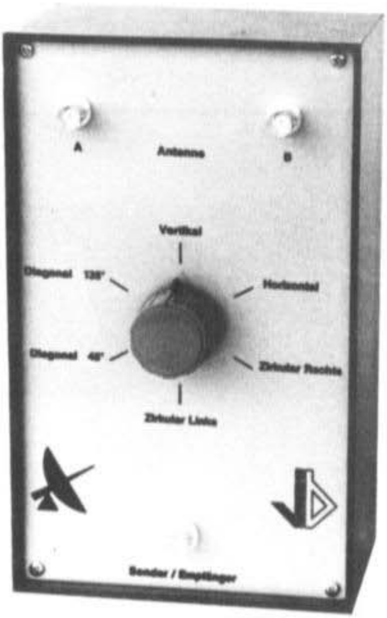

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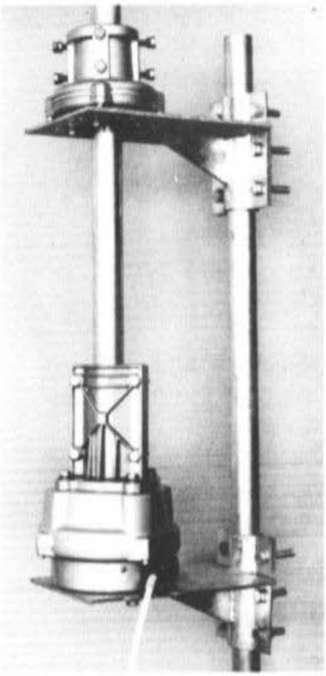
Polarisations Switching Unit for 2 m Crossed Yagis

Ready-to-operate as described in VHF COMMUNICATIONS. Complete in cabinet with three BNC connectors. Especially designed for use with crossed yagis mounted as an »X«, and fed with equal-length feeders. Following six polarisations can be selected: Vertical, horizontal, clockwise circular, anticlockwise circular, slant 45° and slant 135°.

VSWR:	max. 1.2
Power:	100 W carrier
Insertion loss:	0.1 to 0.3 dB
Phase error:	approx. 1°
Dimensions:	216 x 132 x 80 mm



Antenna rotating system as described in 1/1977 of VHF COMMUNICATIONS

We have designed an antenna rotating system for higher wind loads. This system is especially suitable when it is not possible to install a lattice mast. The larger the spacing between the rotator platforms, the lower will be the bending moment on the rotator. This means that the maximum windload of the antenna is no longer limited by the rotator, but only by the strength of the mast itself and on its mounting. Please request the prices either from your National representative, or direct from the publishers.

This system comprises:
Two rotator platforms
One trust bearing
One KR 400 rotator, or other rotator.

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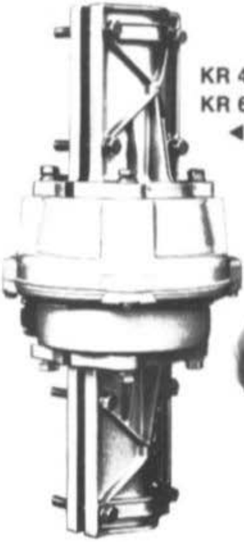
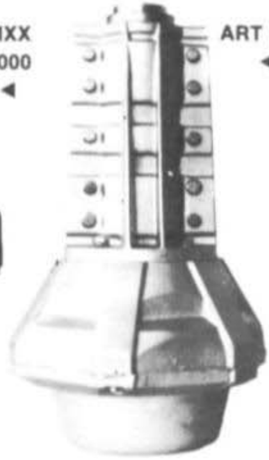
ANTENNA ROTATING SYSTEMS



KR 400
KR 600 ◄

NEW !
KR 2000 ◄

MXX
1000 ◄



ART 8000 ◄

SPECIFICATIONS

Type of Rotator	KR 400	KR 600	KR 2000	MXX 1000	ART 8000	
Load	250	400	800	1000	2500	kg
Pending torque	800	1000	1600	1650	2450	Nm *)
Brake torque	200	400	1000	1200	1400	Nm *)
Rotation torque	40	60	150	180	250	Nm *)
Mast diameter	38 - 63	38 - 63	43 - 63	38 - 62	48 - 78	mm
Speed (1 rev.)	60	60	80	60	60	s
Rotation angle	370°	370°	370°	370°	370°	
Control cable	6	6	8	7	8	wires
Dimensions	270 x 180 ∅	270 x 180 ∅	345 x 225 ∅	425 x 205 ∅	460 x 300 ∅	mm
Weight	4.5	4.6	9.0	12.7	26.0	kg
Motor voltage	24	24	24	42	42	V
Line voltage	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	220 V / 50 Hz	
	50	55	100	150	200	VA

*) 1 kpm ≙ 9.81 Nm

Vertical Rotor KR 500
Especially designed for vertical tilting of antennas for EME, OSCAR etc.

Type	**KR 500**
Load	ca. 250 kg
Brake torque	197 Nm *)
Rotation torque	40 Nm *)
Horiz. tube diam.	32 - 43 mm
Mast diameter	38 - 63 mm
Speed (1 rev.)	74 s
Rotation angle	180° (+ 5°)
Control cable	6 wires
Line voltage	220 V/50 Hz 30 VA
Weight	4.5 kg

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