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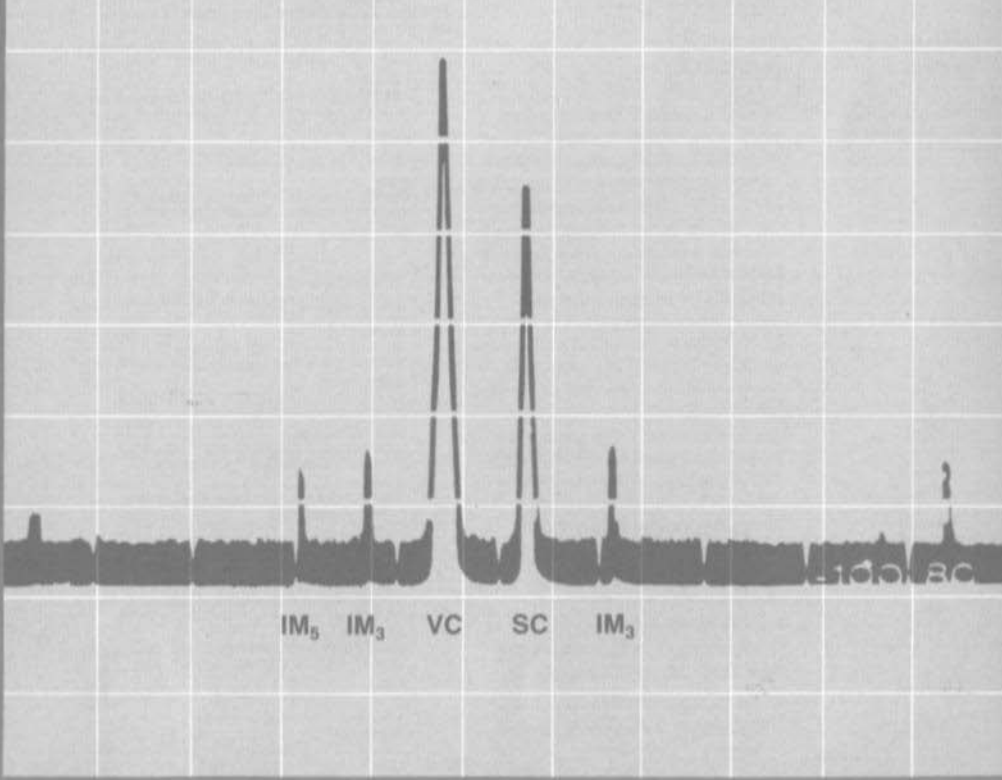
A Publication
for the Radio-Amateur
Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 13 · Autumn · 3/1981 · DM 6.00

Clean ATV Signals on the 70 cm Band with good Intermodulation Rejection



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Published by: Verlag UKW-BERICHTE, Terry Bittan, Jahnstrasse 14, D-8523 BAIERSDORF, Fed. Rep. of Germany, Telephones (09133) 855, 856.

Publisher: Terry Bittan, DJ 0 BQ

Editors: Terry D. Bittan, G 3 JVQ / DJ 0 BQ, responsible for the text
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Advertising manager: Terry Bittan

VHF COMMUNICATIONS The international edition of the German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is published in Spring, Summer, Autumn, and Winter. The subscription price is DM 20.00 or national equivalent per year. Individual copies are available at DM 5.50 or equivalent, each. Subscriptions, orders of individual copies, purchase of PC-boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative.

Printed in the Fed. Rep. of Germany by R. Reichenbach KG, Krelingstr. 39 · 8500 Nuernberg.

We would be grateful if you would address your orders and queries to your representative.

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Germany
Verlag UKW-BERICHTE, T. Bittan, Jahnstr. 14,
D-8523 BAIERSDORF, Tel. (09133) 855, 856.
Konten: Postscheckkonto Nürnberg 304 55-858,
Commerzbank Erlangen 820-1154

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Editorial

Wonder antennas with outstanding gain figures are always appearing on the market. Since the new antennas must be always better than previous antennas, one only needs to add a few dB to the gains quoted by other companies ! This has led to an impossible situation on the antenna market in nearly all countries. VHF COMMUNICATIONS has tried to bring some form of serious judgement into the situation by publishing articles giving the gain value as a function of the beamwidths. Furthermore, our authors have brought graphs giving the gain as a function of the length of a Yagi antenna; however, these are very good for estimating the gain of such antennas, but do not give any information regarding quads, skeleton slots, and other types of arrays.

We feel that it is time for the antenna manufacturers to get together and agree to give gain figures measured according to a common method or according to a neutral body.

However, gain figures are not everything, and price, life and durability are also factors that must be taken into consideration.

Satellite News

The two satellite launches mentioned in ed.2/81 were very successful, and both satellites – METEOSAT 2 and NOAA-7 – are transmitting such good images that must be seen to be believed. At present, no coastlines are being added by the computer in Darmstadt, which means that one can make really photo-like images at present. They will, however, be added next spring, once the computer software is ready.

NOAA-7 is also transmitting excellent quality images, which have more contrast than those of its predecessor TIROS-N. NOAA 7 is responsible for the afternoon passes and provides information in the visible and infrared spectrums. The APT frequency is 137.620 MHz.

NOAA-6 (morning passes) had some problems a few weeks ago both in the primary (SHF) and secondary (VHF) modes. However, these problems seem to have been solved, since NOAA-6 provides good-quality images again. Its APT frequency is 137.500 MHz.

73's de DJ Ø BQ / G 3 JVQ !

A 1.3 GHz Prescaler and Preamplifier for Frequency Counters

by J. Grimm, DJ 6 PI

Many prescalers have been described for frequency counters in VHF COMMUNICATIONS: 180 MHz (1), 250 MHz (2), and 500 MHz (3), and (4). It is now time to increase the frequency range of frequency counters up to 1.3 GHz, so that measurements can also be made on the 23 cm amateur band.

1. AVAILABLE FREQUENCY DIVIDERS FOR FREQUENCIES ABOVE 1 GHz

1.1. Non-Decade Dividers

UHF TV-band V goes up to just below 1 GHz. Many modern TV-receivers use a frequency synthesizer principle for channel selection. The leading semiconductor manufacturers offer frequency dividers that convert frequencies of up to 1 GHz to a lower RF-range. The following types are to be given as an example:

CA 3197:	Motorola, ÷ 64 or 256
U 264 B:	Telefunken, ÷ 64
SAB 1018:	Philips, ÷ 256

Experiments made with type U 264 B showed that this divider was able to operate up to 1.3 GHz with a slight reduction in sensitivity, even though an upper frequency limit of 1 GHz was given.

In order to obtain a correct frequency readout when using these non-decade dividers, it is necessary for either the counter readout to be multiplied by the frequency division factor, to use a programmable divider in the frequency counter, or to divide the time base of the frequency counter by 64 or 256.

The majority of frequency counters offered on the market or described in magazines are not programmable. In order to avoid a multiplication of the counter readout to the value of the frequency division ratio, or to avoid a considerable modification of the time base, it was decided to experiment with decade dividers.

1.2. Decade Dividers

The following decade dividers are the only ones known to the author at this time:

SP 8665:	Plessey 1.0 GHz ÷ 10
SP 8666:	Plessey 1.1 GHz ÷ 10
SP 8667:	Plessey 1.2 GHz ÷ 10

According to the frequency and price relationship it seems that these three types use a common chip, and that the various types are selected according to the guaranteed upper frequency limit.

Experiments made with type SP 8667 showed that it is at least able to operate up to 1.324 GHz. Unfortunately, the author did not have any frequency source available in excess of this frequency. In addition to this, a large number of non-decade dividers are available that could be combined in a suitable manner to obtain a decadic division ratio.

The following types are known to the author to operate in excess of 1 GHz:

SP 8617:	Plessey 1.3 GHz ÷ 4
SP 8619:	Plessey 1.5 GHz ÷ 4
MC 1697:	Motorola 1.0 GHz ÷ 4 *)
11 C 05:	Fairchild 1.0 GHz ÷ 4

*) according to (5), and (6) it operates up to 1.5 GHz.

In order to obtain a decade division ratio, it is necessary to provide two frequency dividers ÷ 5 subsequent to the ÷ 4-divider. The overall division ratio then amounts to 100.

The following ÷ 5-dividers are available at present for frequencies up to 400 MHz:

SP 8620:	Plessey
95 H 91:	Fairchild

This means that the ÷ 4-divider must always be followed by a 400 MHz ÷ 5-divider due to the cut-off frequency. The second ÷ 5-divider must also be one of these types, since the considerably cheaper SN 74 196 can only divide up to a maximum of 25 MHz. In our case, either 65 MHz (1.3 GHz ÷ 20) or 75 MHz (1.5 GHz ÷ 20) must be processed.

2. PREAMPLIFIER

The non-decade frequency dividers used in TV-receivers use an integrated preamplifier. In the case of the U 264 B, a sensitivity of approximately 13 mV was measured at 1.324 GHz. The unbalanced 50 Ω input was matched to the balanced input of the U 264 B with the aid of a ferrite transformer.

The decade dividers or ÷ 4-dividers, on the other hand, require an input voltage of 150 to 600 mV at 1.3 GHz according to type. In order to obtain a frequency readout at the lowest possible input voltage, it is necessary to use a sensitive preamplifier in front of the first divider. Experiments made with the thin-film hybrid amplifier OM 335 described in (4) were not successful. It seems that the gain falls off rapidly in excess of 860 MHz. The wideband amplifier type UTO 1521 manufactured by Avantek was very suitable (1 MHz to 2.3 GHz) but not acceptable due to its enormously high price. Descrete wideband amplifiers equipped with four BFW 92, BFR 34 or BFR 90A provided excellent gain characteristics, but had the disadvantage of high component and room requirements, and a poor repeatability.

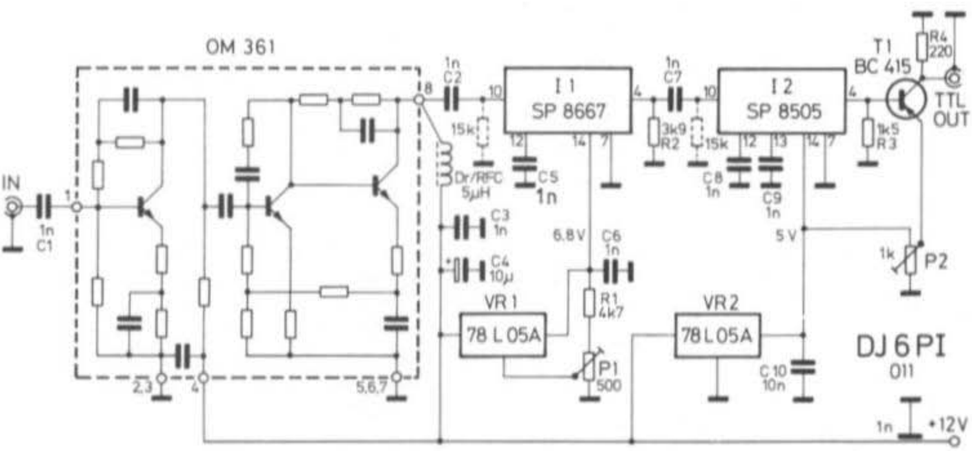

Fig. 1: Complete circuit diagram of the 1.3 GHz ÷ 100-divider which can be directly connected to the TTL counting gate. T 1, I 2, and VR 2 can be deleted when used in conjunction with a 250 MHz counter

The most favorable performance-price compromise was the newly developed OM 361 manufactured by Philips. In contrast to type OM 335, this type only requires an operating voltage of 12 V, however, the input and output must be isolated from the subsequent circuitry using external capacitances. The voltage connection to the third amplifier stage must be made using an external inductance of approximately 5 μH.

PRACTICAL CIRCUIT

As can be seen in **Figure 1**, the input voltage is amplified in the thin-film hybrid amplifier OM 361 to the required input level of the first divider. The Plessey-IC SP 8667 divides input frequencies up to 1.3 GHz by 10. In order to allow the frequency divider to be directly connected to the counting gate, the first divider is followed by a Plessey SP 8632 or SP 8505 as second ÷ 10-divider which is then followed by a subsequent level converter ECL/TTL (T 1).

If the 1.3 GHz divider is to be connected directly in front of an available 250 MHz or 500 MHz counter, it is possible for I 2 and T 1 to be deleted.

The OM 361 operates with 12 V, the SP 8667 with 6.8 V, the SP 8632 and the level converter transistor T 1 with 5 V. In order to use an operating voltage of 12 V, two 5 V voltage stabilizers type 78 L 05 A were used. The voltage control VR 1 is adjustable to 6.8 V with the aid of trimmer potentiometer P 1; resistor R 1 ensures a limiting to this value.

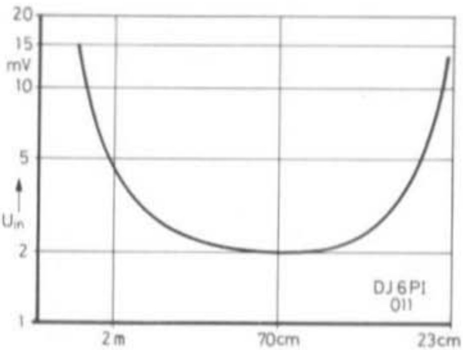

Fig. 2:
The circuit requires less than 20 mV

Figure 2 gives the minimum input voltage values as a function of frequency for this module in excess of which correct operation is possible. The maximum permissible input voltage amounts to approximately 0.5 V.

4. CONSTRUCTION AND ALIGNMENT

The frequency divider is accommodated on a double-coated epoxy PC-board which has been designated DJ 6 PI 011 **(Figure 3)**. In order to avoid unwanted injection, the PC-board should be installed in a sealed casing. For simplicity, the prototype was accommodated in a standard tinplate case whose dimensions were 72 mm x 72 mm (see **Fig.4**). The ground surface of the board is soldered to the case on both sides. In the case of boards without through-contacts, the ground connections of C 3, C 4, C 5, C 6 and pin 7 of I 1 should be soldered to the ground surface on both sides of the board.

Capacitors C 1 and C 2 are 1 nF chip capacitors for which narrow slots should be cut into the board. The capacitance surfaces should be soldered to the conductor lanes on both sides. The choke RFC is accommodated on the lower side of the board. The input must be made via an SHF-connector with the shortest possible connections to C 1. In the case of the prototype, a BNC-connector was directly soldered to the case. The output can be connected using a RF-connector or coaxial cable to the counting gate.

Before connecting the operating voltage, both trimmer potentiometers should be adjusted to their full, anticlockwise position. After connecting an operating voltage of 12 V (approx. 200 mA), potentiometer P 1 is adjusted so that 6.8 V is present at pin 14 of I1. Ifa voltage of 6.8 V cannot be obtained, it will be necessary for R 1 to be reduced to 3.9 kΩ.

Potentiometer P 2 is aligned after connecting a signal source of approximately 20 mV at 1.3 GHz so that a stable counter readout results.

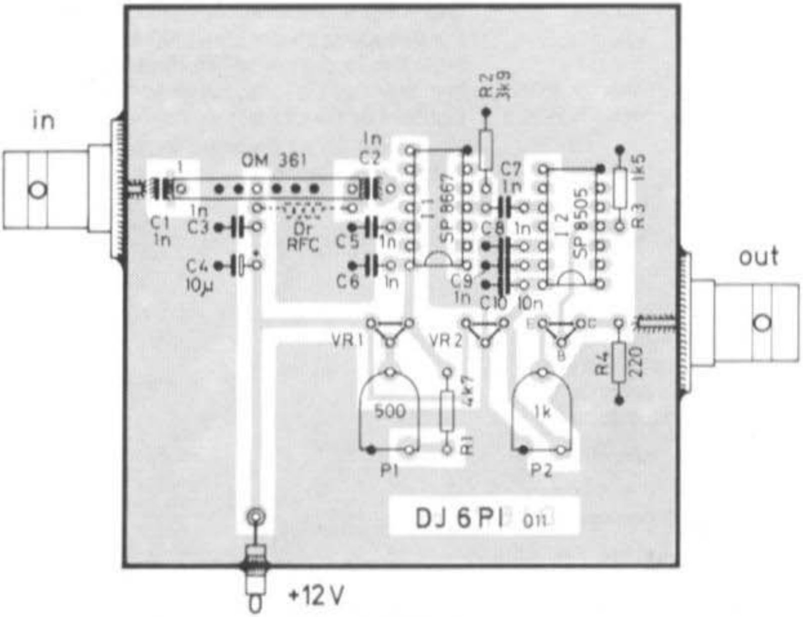

Fig. 3: Component locations on the double-coated PC-board DJ 6 PI 011 (72 mm x 72 mm)

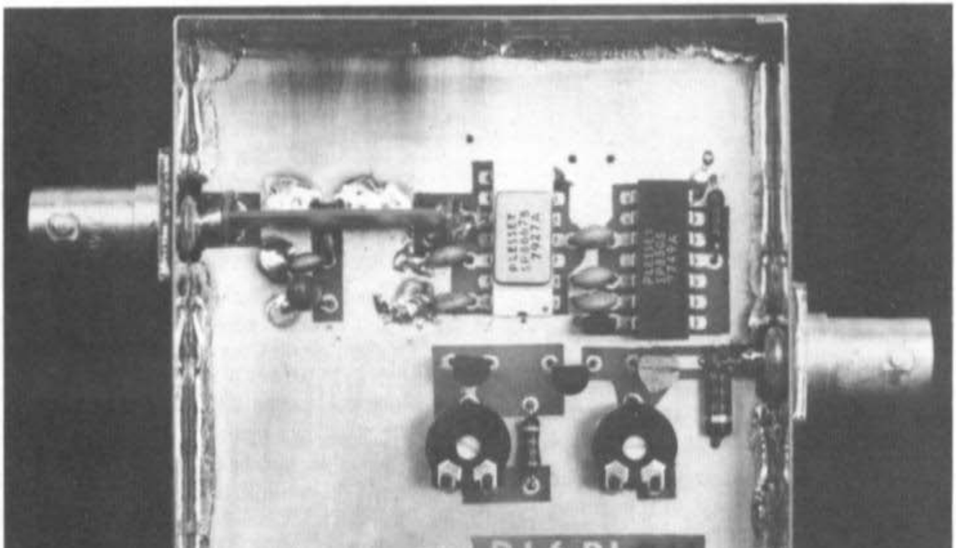

Fig. 4: Photograph of a prototype fitted into a standard metal case

4.1. Components

Preamplifier:

OM 361: Ensure that it is installed correctly! Pin 1 is designated with »1« and should be connected to C 1.

I 1:	SP 8667 (Plessey)
I 2:	SP 8632 or SP 8505
T 1:	BC 415 or similar silicon PNP transistor
VR 1, VR 2:	78 L 05 A
C 1, C 2:	1 nF chip capacitor
C 3, C 5 - C 9:	1 nF with connection leads
C 4:	10 μF / 25 V tantalum
C 10:	10 nF ceramic capacitor

Trimmer resistors, horizontal mounting, spacing 10/5 mm

Carbon resistors 0207 for 10 mm spacing

RFC: Choke, approx. 5 μH, e.g. 2 turns of enamelled copper wire placed through a ferrite bead, with the shortest possible connections. This is accommodated on the lower side of the board.

Solder R 4 on the upper and lower side of the board.

5. FINAL NOTES

A flickering counter readout will be seen when the input of the counter is not terminated, or when the input level is too low. This oscillation, which usually takes place around the frequency of highest sensitivity, can be neutralized by loading the divider inputs with 15 kΩ. However, this has an adverse effect on the sensitivity of the counter.

In order to avoid overloading or destroying the OM 361, the frequency to be measured should be fed via an attenuator so that it is only strong enough to provide a stable readout. It would have been possible to avoid this limitation and the flickering readout without sufficient input signal using an PIN-diode control and a comparator circuit. However, this would have complicated the construction and increased the price considerably.

Experiments made with the OM 361 in front of the frequency divider chain SP 8619, 2 x SP 8620 provided an excellent input sensitivity of only 1 mV at 1.3 GHz, however, the gain adjustment of the OM 361 was very critical, and no reliable circuit could be described by the author at this point.

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

Summarized from a Lecture at the VHF-UHF Convention 1980 in Weinheim, West Germany

by B. Neubig, DK 1 AG

1. IMPORTANCE OF SHORT-TERM STABILITY AND PHASE NOISE IN TRANSMITTER AND RECEIVER APPLICATIONS

The output frequency of UHF and SHF equipment is to be obtained from a stable crystal oscillator with the aid of frequency multiplier chains.

Figure 1 shows the various frequency plans that can be used. All bands up to 3 cm can be obtained from a basic frequency of 1152 MHz. This is obtained with the aid of frequency doublers, and triplers from the given frequencies. Since it is difficult to manufacture crystal oscillators operating at a frequency of 192 MHz, and because 72 MHz and 144 MHz will cause unwanted spurious signals in the 2 m and 70 cm band (converter), the most suitable frequency seems to be 96 MHz.

The resulting total multiplication factor n_{tot} is given in the illustration, and is between 12 (23 cm) and 108 (3 cm). This results in very high demands on the frequency stability of the master oscillator. Whereas insufficient long-term stability will be noticeable as drift that can be avoided using a stable crystal oscillator circuit in a crystal oven and using aged crystals of sufficiently high Q, short-term stability is far more critical.

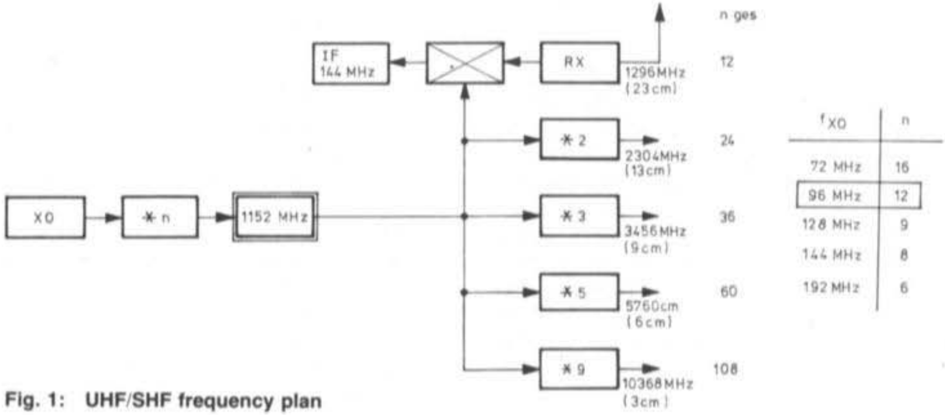

Fig. 1: **UHF/SHF frequency plan**

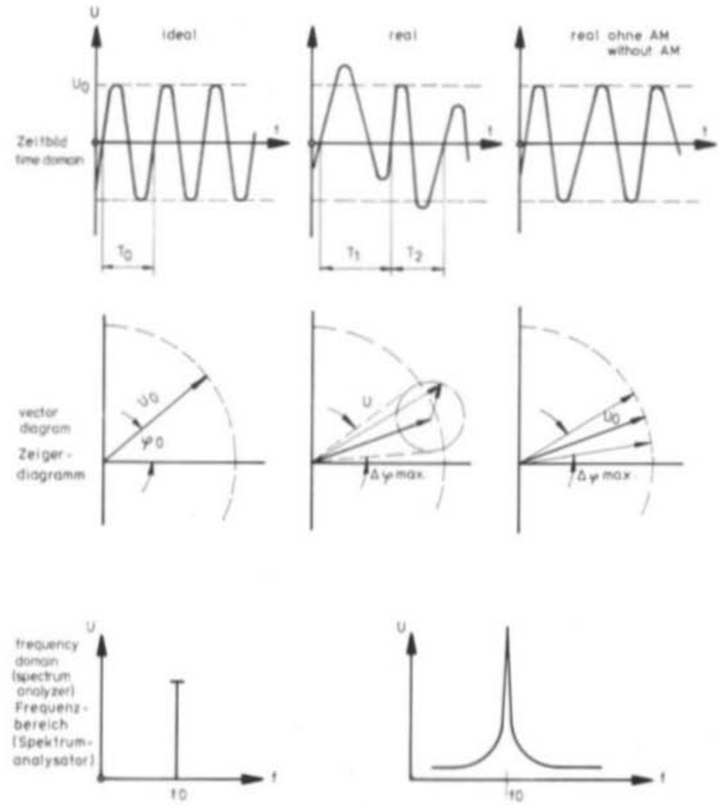

Fig. 2: Frequency stability in various presentations

Short-term stability is observed as »short-period« fluctuations of the selected frequency, or phase within a period of seconds. **Figure 2** shows this with the aid of various diagrams.

The upper diagram shows the oscillator signal in time domain, that is as can be observed on an oscilloscope. The center diagram shows the same effect in the form of a vector diagram: The amplitude is shown as the length of the arrow, and the phase angle by the angle φ_0. The sinewave oscillation corresponds to a rotation of the arrow, and the momentary value is the projection of the arrow onto the reference axis. The lower row shows the same effect in the frequency domain, that is when observing the signal on a spectrum analyzer.

An ideal oscillator possesses a purely sinusoidal output voltage of constant amplitude U_0, the point of the arrow will move at a constant speed in a circular direction, and the spectrum will consist of a single line with a frequency $\omega_0 = 2\pi \times f_0$

$$U(t) = U_0 \times \sin(\omega_0 t + \varphi_0) \qquad (1)$$

The output signal of a real oscillator – which has been rather exaggerated in Figure 2 – will exhibit short-term amplitude and phase (frequency) fluctuations that are more or less statistically distributed. In the case of the vector diagram, this means that the length of the arrow will fluctuate, and will also not be constant in its circular path, but will swing around the ideal position. In the spectral display, this will be observed as a wider line in conjunction with amplitude fluctuations.

Such a »noisy« oscillation can be determined mathematically as:

$$U(t) = [U_0 + \varepsilon(t)] \times \sin \underbrace{[\omega_0 t + \varphi(t)]}_{\Phi(t)} \qquad (2)$$

This means that the amplitude U_0 is provided with a noise component due to a momentary value ε (t), and the total phase Φ (t) by the component φ (t).

Since an amplitude limiting takes place within the oscillator chain (oscillator, multiplier stages, receiver-IF in the FM-mode) it is possible to neglect the »amplitude noise« (right-hand column in Figure 2), and the output signal will be

$$U(t) = U_0 \times \sin[\omega_0 t + \varphi(t)] \qquad (3)$$

The momentary frequency at any time-point t will amount to:

$$f(t) = \frac{1}{2\pi}\frac{d}{dt}\Phi(t) = f_0 + \frac{1}{2\pi}\frac{d\varphi(t)}{dt} \qquad (4)$$

This equation shows the relationship between the frequency and the phase noise, which cannot be separated from another.

The output signal consists of a spectral line of finite width. Away from the main line, a virtually »white« wideband noise base exists, whereas the main line itself is increased in its direct vicinity by an increasing noise level. In addition to this, weaker lines may be observed from harmonics, and other spurious signals.

The deviations from the constant frequency can be split into two types: determinable ambient effects, and statistic effects. Both types cannot be clearly separated from another experimentally.

The ambient effects are influences of the operating voltage, the ambient temperature, mechanical and electrical loading, as well as aging of the crystals and other electrical components.

The statistic effects on the oscillator frequency are added to the ambient frequency variations. Such effects are caused by noise of the active circuit components, by additional noise of the components, including the crystal. The crystal noise originates from effects in the resonator structure, as well as from physical surface effects between crystal disk and electrodes. This error causes a dependence of the resonant frequency on the loading of the crystal, as well as variations of the electrical equivalent values on changing the exciting amplitude.

If one observes the common form of an oscillator, the main points where the noise is generated can be localized according to **Figure 4**:

The main components of an oscillator are:

- The amplifier for commencing and maintaining oscillation
- A limiter circuit, which is usually realized by the amplifier itself. (The task is to stabilize the maximum amplitude after commencing oscillation).

Furthermore:

- The resonant circuit – in our case the crystal, which may be provided with phase-shift, LC-reactive components.
- Finally, a buffer for isolation of the oscillator from the subsequent stages.

The level of the white wideband noise will be determined by the noise contribution of the buffer and subsequent amplifier stages; the increase of noise in the vicinity of the carrier frequency is formed by the selective oscillator network. The finite line width of the carrier itself is dependent on the Q of the crystal, the type of network, and the low-frequency noise characteristics (1/f-range) of the oscillator stages. The low-frequency noise can affect the oscillator in two different ways:

a) Directly on the frequency-determining parameters (crystal) which is called parametric noise
b) It can be multiplied and mixed up in the range of the carrier frequency indirectly via non-linearities.

The Q of the crystal is dampened in any oscillator circuit down to the so-called effective Q_{eff}, which means that the crystal and oscillator will possess a total Q_{eff} that may be (considerably) lower than the Q of the crystal on its own.

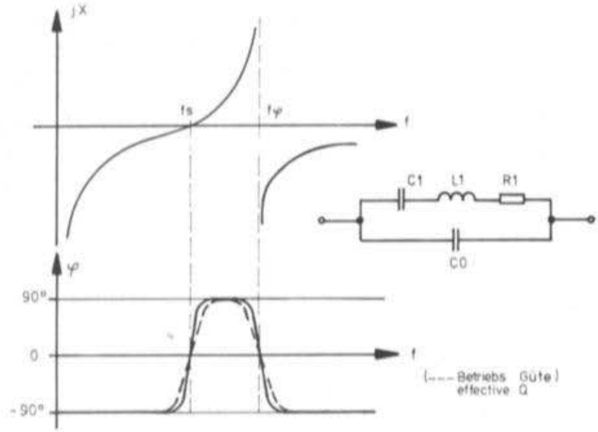

Fig. 3:
Reactive impedance and phase characteristic of quartz crystals

The relationship between phase noise and frequency noise is determined by the effective Q:

$$\varphi = \arctan\left(-2 \times Q_{eff} \frac{\Delta f}{f_0}\right) \qquad (5)$$

The following results in the vicinity of the oscillator frequency f_0:

$$\frac{d\varphi}{df} = -2 \frac{Q_{eff}}{f_0} \qquad (6)$$

Example: $Q_{eff} = 50000, \quad f_0 = 96$ MHz

$$\frac{\Delta\varphi}{\Delta f} \approx -1 \times 10^{-3} \frac{rad}{Hz} \approx -0.05°/Hz$$

This means that the oscillator must offer constant phase conditions with an accuracy of 1/20° in order to obtain a short-term stability of 1 Hz (corresponding to 100 Hz at 10 GHz)!

Figure 3 illustrates equation (5); the dashed line is the phase characteristic of a dampened crystal.

What are the effects of short-term stability and phase noise in practice ? Noise sidebands of a transmit signal will accommodate a frequency range in the vicinity of the transmit frequency whose width is dependent on the frequency multiplication factor, and they will also superimpose themselves on any weak signals that appear. In the case of receivers, these noise sidebands will be superimposed in the mixer on each receive signal and will cause a desensitization of the receiver in the vicinity of strong stations (1).

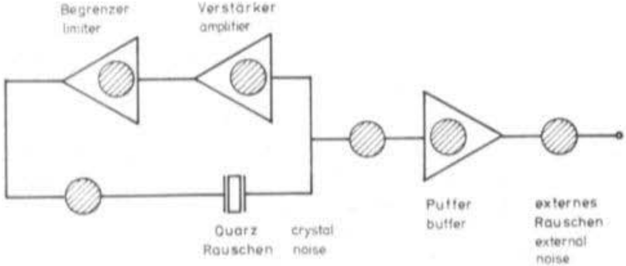

Fig. 4:
Noise sources in a crystal oscillator

2. MEASUREMENT OF THE SHORT-TERM STABILITY

2.1. Definitions, Measurement and Evaluation Magnitudes

2.1.1. Measurements in the Frequency Domain

If the noise spectrum in the vicinity of the carrier is studied more accurately, the following will be observed:

As was given in equation (4), frequency and phase variations are directly dependent on another. It is therefore sufficient when evaluating the short-term stability, to either measure the relative frequency variation (in ppm) with respect to a very stable reference frequency, or for the phase variations to be compared with a source having a quasi-stationary phase.

The mathematical evaluation is only to be explained briefly: The frequency deviations, (or phase deviations) are measured periodically in a certain measuring period, comparing only the variation from one measurement to the next. The statistic mean value of this product is called the auto-correlation function of the frequency or phase variation. From this one can obtain the spectral density of the relative frequency variation S_y (f), or the spectral densitiy of the phase noise $S\varphi$ (f) with the aid of an integral transformation.

The spectral density of the frequency stability and the phase variation are dependent on another according to the following equation:

$$S_y(f) = \left(\frac{f}{f_0}\right)^2 S_\varphi(f) \qquad (7)$$

If one of these two experimentally measured spectral densities is drawn as a function of the frequency difference to the carrier, it is possible for the resulting noise spectrum to be differentiated according to type **(Table 1)**.

$$S_y(f) = \sum_{n=-2}^{2} a_n f^n$$

for $0 < f \leqq f_{max}$

$$S_\varphi(f) = v_0^2 \times \sum_{n=-2}^{2} a_n f^{n-2}$$

Figure 5 shows that one can draw one or more straight pieces between the measuring points when using a double-logarithmic display. These then have a slope of − 2, − 1, 0, etc. The physical nature of the individual components is summarized in Table 1. According to equation 7, the exponent of the frequency dependence is to the value of two less at S_φ (f) than at S_y (f).

Type of noise	S_y (f)	S_φ (f)
Statistic noise of the frequency	$a_{-2} \times f^{-2}$	$v_0^2 \times a_{-2} \times f^{-4}$
1/f noise of the frequency	$a_{-1} \times f^{-1}$	$v_0^2 \times a_{-1} \times f^{-3}$
White noise of the frequency Statistic noise of the phase	a_0	$v_0^2 \times a_0 \times f^{-2}$
1/f-noise of the phase	$a_1 \times f$	$v_0^2 \times a_1 \times f^{-1}$
White noise of the phase	$a_2 \times f^2$	$v_0^2 \times a_2$

Table 1

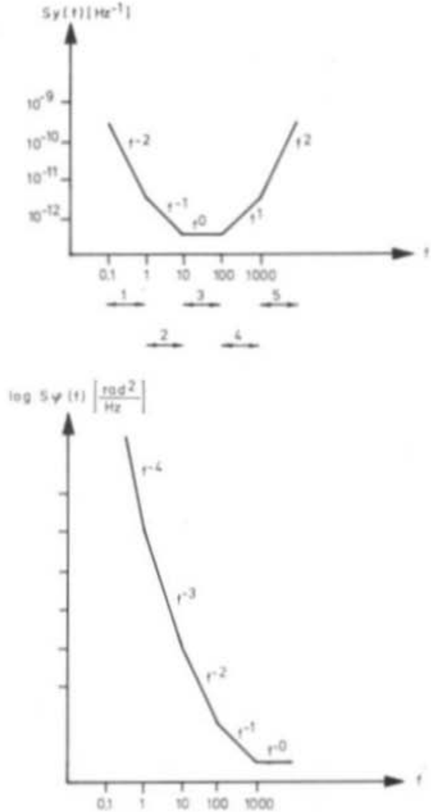

Fig. 5: Types of spectral distribution of frequency (S_y) and phase noise (S_φ)

Random noise of the frequency results in a noise spectrum proportional to f^{-2}. A noise behaviour f^{-1} represents the 1/f-noise of the frequency, and the white noise of the frequency will represent a constant contribution in magnitude S_y. It is equivalent to a random noise of the phase, i.e. S_φ (f) ~ f^{-2}.

The next line shows the 1/f noise of the phase and finally the white noise of the phase.

The most often used value for characterizing the phase noise (in the frequency range) experimentally, is the magnitude £ (f). This has the dimension Hz^{-1} and is defined as the ratio of the power at a certain point of a noise sideband to the total signal power, referred to a measuring bandwidth of 1 Hz (at the frequency ν_0 + f, i.e. f Hz from the carrier).

$$£(f) = \frac{S(\nu_0 + f)}{P_{carrier} + 2 \times P_{sideband}}$$

Since the noise power is usually equally distributed to both sidebands, the following is valid for $\Delta_{\varphi noise} << 1$ rad:

$$£(f) = \frac{1}{2} S_\varphi(f)$$

Usually, the logarithmic value 10 log £ (f) (dB) is used instead of £ (f), and this is thus 3 dB lower than the logarithmic value 10 log S_φ (f).

If one measures in a bandwidth (analyzer bandwidth) of b Hz, the measured power will then be approximately 10 log b (dB) greater than at 1 Hz.

2.1.2. Measurements in the Time Domain

In the first part of this article we have described how the short-term stability can be characterized in a spectral display. The measurement of these magnitudes can be carried out with the aid of a spectrum analyzer.

It is, however, often easier to determine the stability with the aid of a large number of frequency measurements, or period duration measurements with the aid of a frequency counter. Measurements in the frequency and time domain are mathematically equivalent.

The most common value for characterizing in the time domain is the so-called Allan-variance. Using a frequency counter, one reads off a large number of frequency values f_i at predetermined intervals. The interval between two measurements is τ. For evaluation, however, only the relative individual differences

$$y_i = \frac{f_{i+1} - f_i}{f_i} \quad \text{(in ppm)}$$

between two subsequent values are formed. If one assumes that the gate time of the counter is very small with respect to the actual measuring period, the Allan-variance can be defined as follows with m measurements:

$$\sigma_y^2(\tau, m) = \frac{1}{2 \times (m-1)} \sum_{i=1}^{m^{-1}} (\bar{y}_{i+1} - \bar{y}_i)^2 \qquad (8)$$

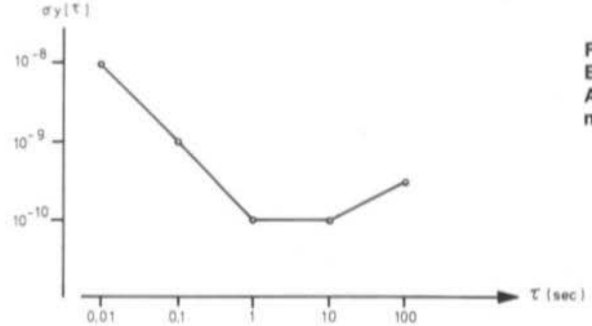

Fig. 6: Example of an Allan-variance measurement

The y-values are the relative frequency variations in ppm (10^{-6}).

This means that the frequency difference between measurement 2 and measurement 1 is formed, the difference (in ppm) is squared, and this value is stored. During the next cycle, the frequency difference between measurement 3 to 2 is formed and the difference square is summed up according to the previous result, which is followed by the third cycle, where the difference 4 to 3 is squared and summed up. After carrying out a sufficient number of measurements, the mean value of this sum of the difference squares is formed and divided by the duration of the measuring interval. This measurement is repeated for various measuring periods, e.g. 1/100 second, 1/10 second, 1 second, 10seconds, etc. and σ_y (τ) is drawn in double-logarithmic scale. This curve is a measure of the dependence of frequency on the measuring (integration) time from one measurement to the other.

In order to compare the characteristics of several oscillators to another, it is necessary for this type of measurement to be carried out at different measuring interval lengths.

Figure 6 shows the typical result of an Allan-variance measurement as a function of the counting time when drawn in a double-logarithmic scale.

2.2. Measuring Methods

A very good reference with respect to this section is given in (2): The NBS Technical Note 632, published by the National Bureau of Standards. This contains a detailed technical introduction with mathematical appendix, a detailed description of typical measuring systems for measuring the short-term stability with exact information for designing these, including circuit examples.

2.2.1. Frequency Domain (Spectrum Analysis)

The fundamental measuring principle is given in **Figure 7**.

A reference oscillator and the oscillator to be examined are converted to frequency zero in a balanced mixer. The conversion product is fed via a lowpass filter to a low-noise DC-amplifier, to which a phase-locked loop is connected. It is in turn loosely coupled to the reference oscillator so that it remains phase-locked to the oscillator to be examined. The phase-locked loop has a relatively large time constant, which means that only very slow frequency variations will be controlled. More rapid frequency variations will be seen as

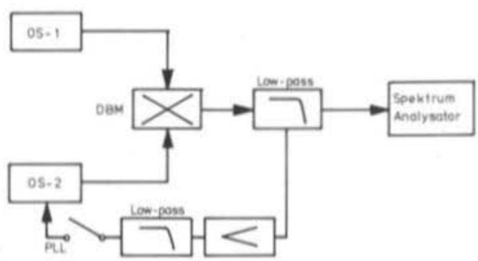

Fig. 7: Measuring principle sideband – phase noise

phase fluctuations, and will generate a noisy DC-signal at the output of the mixer. This can be taken from the output of the low-noise amplifier, and be examined using a spectrum analyzer suitable for low frequencies.

A measuring system was developed according to this measuring principle for examining the oscillators discussed in section 3; this system is constructed using conventional components. If only a coarse evaluation is required, it is sufficient for the noisy signal to be measured on an oscilloscope instead of using an expensive spectrum analyzer. The peak-to-peak amplitude of the noisy signal is a relative value, which can be used to carry out better/worse comparisons. The circuit diagram is given in **Figure 8**.

A peak rectifier or effective value meter can be connected in order to obtain a quantitative evaluation. It is possible to use a number of switchable, active bandpass filters and to measure their output signal instead of using the expensive spectrum analyzer. It is thus possible to evaluate the quantitative relationship between the various spectral components.

However, this method only allows two oscillators to be compared to another, whereas an absolute measurement on an oscillator is not possible. If oscillator 1 is considerably better (derived from a frequency standard), one will obtain a quasi-absolute measurement. If, on the other hand, virtually similar oscillators are compared to another, it will be necessary to deduct 3 dB from the measuring result, in order to obtain the phase noise of a **single** oscillator.

It is possible in this manner to construct various oscillator prototypes and to compare them to another.

The sensitivity of the system can be increased by carrying out the frequency conversion not at the frequency of the oscillator, but after

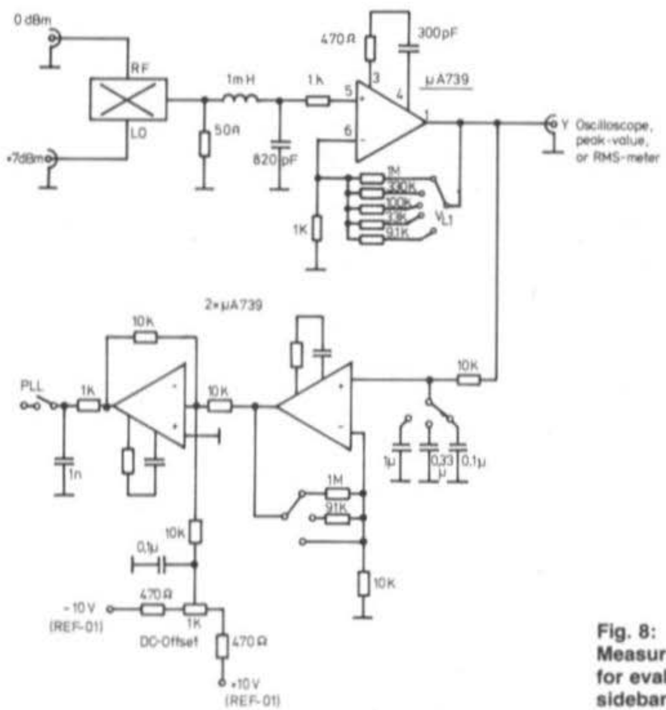

Fig. 8: Measuring circuit for evaluating sideband noise

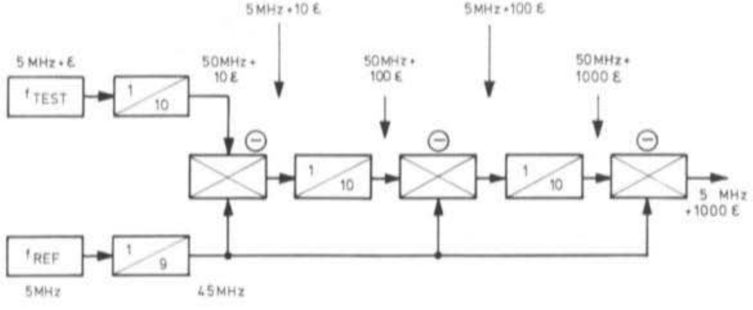

Fig. 9: Successive 10/9 multiplication principle

multiplying the output frequency. A very elegant method is to use »successive 10/9 multiplication« (see **Figure 9**). This has the advantage that the frequency to be measured has the same frequency of the reference signal, whereas the noise sidebands are 10^n wider.

The quantitative measurement of the single-sideband noise £ (f) is made in the following manner:

a) The two oscillators are slightly detuned from another and the amplitude of the difference signal is measured, or adjusted to a reference value (0 dB). The peak-to-peak amplitude corresponds to the sensitivity of the measuring system in V/rad.

b) Both oscillators are locked-in opposite to another and the **effective** noise output voltage is measured at a spacing f from the carrier frequency. The voltage ratio is now converted to dB, and the bandwidth factor 10 log b is added to this. The result is the magnitude £ (f) in »dB/Hz«.

2.2.2. Measurements in Time Domain

Principally speaking, the time domain measurement observes the frequency variations of the oscillator directly using a frequency counter with the fastest possible count, and evaluating this statistically.

The time base of the frequency counter must, of course, have a considerably better short-term stability than the oscillator to be measured. Since a higher frequency resolution, – when using a directly measuring frequency counter – results in long measuring periods, it is usually necessary to make frequency measurements with a high resolution in short cycles, using a reciprocal counter and making period duration measurements.

a) Direct Measurement:
In the case of the direct measurement, the oscillator to be examined is directly connected to the frequency counter; the measuring results of the frequency counter are, for instance, printed out, or are directly evaluated using a fast processor. The most usual evaluation method is to use the Allan-variance.

b) Indirect Measurement:
In the case of the indirect measurement, the oscillator to be examined is mixed with a reference oscillator that is locked-in with the aid of a phase-locked loop (see frequency domain measurement); however, the control is so tight that the phase fluctuations are completely controlled. The correspondingly noisy output signal of the loop is amplified and fed to a voltage-frequency converter that converts the noisy DC-voltage into a rapidly-changing frequency fluctuation. This converted frequency can be measured with a frequency counter according to the method described in a), and evaluated.

One of the next editions of VHF COMMUNICATIONS will bring the final article.

Coupled Microstriplines as Filter

by F. Schmehr, DC 8 EC

This article is based on a lecture given at the Munich VHF-UHF Convention in 1980. The task of this article is to give UHF/SHF amateurs an insight into the exact calculation and construction of microstripline filters. It will also introduce tables that allow a simple and exact determination of the filter dimensions. The tables also offer valuable assistance for further applications in microstripline technology. The complete calculation process with examples would be too extensive to be described here, however, it is possible to obtain a reprint of the lecture, and to obtain completed 13 cm filters from the author. Further details regarding this are given at the end of the article.

FUNDAMENTAL CONSIDERATIONS

Many of our readers with experience in directional coupler technology will know that coupled lines are not defined with the system-dependent impedance Z_0, but with the Even-Mode impedance Z_{oE}, and the Odd-Mode impedance Z_{oo}. Examples for balanced configuration in a homogene medium (air) can be found in many standard works such as in (1).

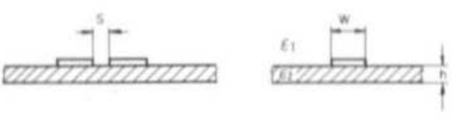

Fig. 1: Striplines in inhomogene medium (microstriplines). Left: Single line; Right: Coupled lines with $\varepsilon_{r\,eff} = K_{EFF} = \sqrt{K_{EFFE} \times K_{EFFO}}$

If, however, the coupled lines **(Figure 1)** are to be found in a homogene medium (dielectric $\varepsilon_r = 2 - 10$), the calculation will become very complicated and approximation formulas will only show a limited accuracy, which means that these are unsuitable for the construction of filters. The same is valid for an individual stripline in an inhomogene medium. The exact calculation of the mechanical dimensions w/h and s/h with respect to Z_0 or Z_{oE} and Z_{oo}, with an error of < 1 %, would be virtually impossible for most radio amateurs. An example of a calculation of w/h using too simple a formula is to be found in (7) where the error is in the order of approximately 15 %.

The »exact« calculation of microstriplines was firstly possible in 1969 with the aid of the »BRYANT-WEISS« method (3). The mean error only amounts to 0.7 % ! In order to take the problem of the inhomogene medium into consideration, the »Dielectric Green's Function« is used. This allows the discontinuities of the fields at the edges of the medium to be considered. The calculations according to (3) can only be made using large computers such as CYBER 176, and these have been carried out. The resulting tables allow anyone working with microstriplines to read off the mechanical dimensions at any given Z.

The given accuracy is especially necessary during the calculation of filters. An extract from this list of tables for coupled lines on a RT/duroid 6010 dielectric with $\varepsilon_r = 10.25$ is given in **Figure 2**.

If this list of tables is available, the dimensions can be read off directly for such applications as matching circuits, lines, inductances, capacitances, open and short-circuited stubs, coupled lines for directional couplers, power dividers, matching networks for antenna arrays, filters with coupled lines, and much

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   RFOPT VERSION 5.1 + 49 - IMPLEMENTED 04/24/80
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 FILE NAME, 'TRXINFO', 'TRLINES', 'LANGE', 'NEWS' OR 'QUIT' ? TRLINES
 BRYANT-WEISS OR APPROXIMATE SOLUTION (BW/A)? BW
 MICROSTRIP, SHIELDED MICROSTRIP OR QUIT (MS/SH/QU) ? MS
 ENTER SINGLE OR COUPLED LINES (S/C)? C
 NO. SUBSTRIPS PER STRIP (25 MAXIMUM) ? 10
 ENTER RELATIVE DIELECTRIC CONSTANT ? 10.25
 UNITS: ENGLISH, METRIC OR RATIOS (E/M/R) ? R
 FOR W/H: ENTER START, STOP, AND STEPSIZE ? 0.5 2 0.5
 ENTER S/H RATIO? 1

    COUPLED STRIPS     S/H =  1.000          K = 10.25

    W/H         Z0E     Z0O         VE       VO         KEFFE   KEFFO
                    OHMS            E+08 M/SEC
    .500       75.332  56.401      1.130    1.235       7.045   5.894
   1.000       55.211  42.545      1.100    1.213       7.431   6.111
   1.500       44.115  34.951      1.079    1.192       7.718   6.329
   2.000       36.874  29.882      1.064    1.173       7.945   6.533


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Fig. 2: Printout of a calculation table

more. It is possible, for instance, for a 13 cm converter to be calculated which comprises two preamplifier stages, a travelling-wave directional filter, or alternately a filter with parallel-coupled lines for 20 dB suppression of the image frequency, a printed subharmonic mixer with $f_{osc.}$ = 1080 MHz, and an IF-preamplifier at 144 MHz, all accommodated on a board of approximately 70 mm x 50 mm. This would represent the end of the considerable mechanical requirements of the chamber technology.

APPLICATIONAL EXAMPLE

The »theory« is now to be shown with the aid of a practical example.

Firstly it is necessary to briefly discuss the calculation fundamentals for filters using coupled lines, after which the mechanical construction and measured values are to be discussed.

This filter is constructed in the manner shown in **Figure 3**:

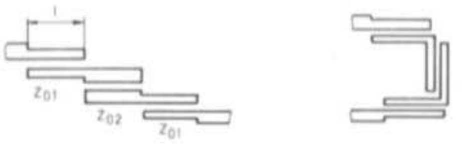

Fig. 3: Filter with coupled lines of stage 2 showing two possible arrangements; $L = \lambda/4 \times 1/\sqrt{\varepsilon_{r\,eff}}$

The following is required for calculation:

- Tables for determining the Tschebyscheff-lowpass coefficients and the degree of filtering n, which can be found in any handbook covering filter fundamentals, such as (1), (2), (5); the complete filter calculation is given in (1) and (8).
- The mentioned tables for determining the values of w and s of the coupled microstriplines.

The calculation process for coupled lines was carried out according to (1).

The values Z_{oE} and Z_{oo} for the individual coupled line sections are obtained with the aid of simple, short calculations. The mechanical dimensions are determined with the aid of the microstripline tables, after which the filter can be constructed.

The filter has the following specifications:

Characteristics	calculated	measured
Bandwidth	200 MHz	160 MHz
Insertion loss	< 1 dB	0.8 dB
Center frequency	2300 MHz	2280 MHz
Stopband loss at 2000 MHz (f_{image})	20 dB	20 dB

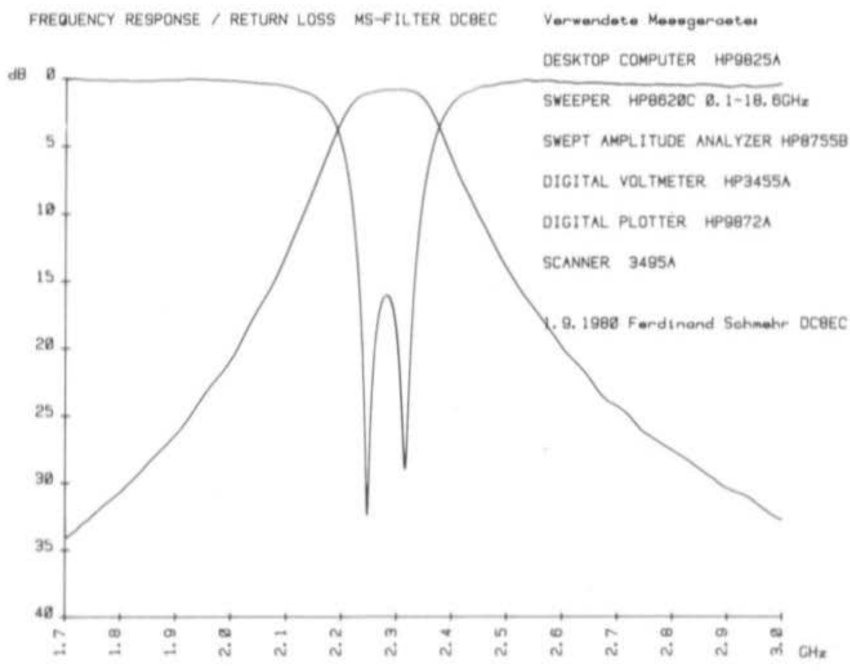

Fig. 4: Frequency response and return loss of the filter shown in Fig. 3

Figure 4 shows the measured frequency response and return loss curves of this filter.

The measurements show that the required filter specifications can be obtained without problems and alignment. The tabular values were calculated using the large computer CYBER 176. The program used was the HF-program COMPACT/RFOPT.

The calculations were carried out for:

- Epoxy material with $\varepsilon_r = 4.8$
- PTFE glas fibre material with $\varepsilon_r = 2.55$
- RT/duroid 5870 with $\varepsilon_r = 2.35$
- RT/duroid 6006 with $\varepsilon_r = 6$
- RT/duroid 6010 with $\varepsilon_r = 10.25$

The following tables are available:

1. According to »BRYANT-WEISS«:

- Single-strip transmission lines for w/h from 0.4 to 15 with 0.2 steps; available for DM 10.—
- Coupled-microstrip transmission lines: 7 tables for w/h from 0.5 to 10 with steps of 0.5, and s/h from 0.3 to 2.1 with steps of 0.3; available for DM 40.—.

2. As under 1., however, calculated with the aid of an approximation (mean error approx. 3 %):

- Single-strip transmission lines (DM 7.—)
- Coupled-microstrip transmission lines (DM 22.—).

Other ranges and steps of w/h and s/h of microstrip and triplate lines are available at short notice from the author.

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Close-In DF-Receiver for the 144 MHz Band

by H. W. Storbeck, DL 2 DE

Anyone who has taken part in a 144 MHz foxhunt or other DF-meeting, will know that most 144 MHz receivers will not work satisfactorily in the direct vicinity of the transmitter, and will thus no longer allow the direction to be determined. »Professional foxhunters« usually have a special system, which allows them to operate even in the direct vicinity of the transmitter. A large number of amateurs, however, use modified FM-transceivers, and they are very often satisfied when they are able to get within 100 m of the fox transmitter. After the event, the transceiver is modified back to its normal application.

These considerations and the author's own practical experience led to the described close-in DF-receiver for the 144 MHz band. This receiver is also suitable for use as a sensitive field strength meter. It is not designed to be a sensitive or selective receiver, however, it allows the direction to be determined at a distance of several thousand meters right up to the transmit antenna.

The receiver operates as a wideband receiver within the 144 MHz band and is very easy to operate. The circuit is built up on a double-coated PC-board whose dimensions are 106 mm x 42 mm. The component side of the board is in the form of a ground surface. The PC-board can be accommodated in any screened case (maybe made from PC-board material) together with a moving coil meter, a potentiometer with On-Off switch, two 9 V batteries and an antenna connector. No drawings have been given for the case and installation, since most radio amateurs will have the components in their junk box.

CIRCUIT DESCRIPTION

The receiver is equipped with two VHF-preamplifier stages equipped with dual-gate MOSFETs, type BF 900 **(Figure 1)**. The amplified VHF-signal is then rectified, and the resulting DC-voltage is fed at low impedance to an operational amplifier type 741 (I 1), which is connected as impedance converter. It is then amplified in I 2 (741). The field strength is then indicated on a meter.

The gain of the receiver can be varied within wide ranges with the aid of a potentiometer. This is made simultaneously in two ways:

The voltage at the wiper of the potentiometer can be varied continuously from + 9 V to − 9 V. The G 2-voltages of the MOSFETs are obtained with the aid of the voltage divider R 3/R 1 and R 9/R 6. The value of these voltages will thus be in the order of + 2.5 V for maximum sensitivity and −2.5 V for minimum sensitivity. This allows the gain to be varied in the order of 70 dB.

However, this will not be sufficient for all applications, and for this reason the gain of the DC-voltage amplifier I 2 is also reduced as soon as the gain of the MOSFETs has been reduced virtually to a minimum. A junction FET T 3 (BF 245 C) is used for this. As long as the voltage at Pt 3 is ≧ 0 V, diode D 3 will be blocked and will ensure that the gate is not provided with positive voltages. However, as soon as the voltage at Pt 3 becomes negative, diode D 3 will conduct, and T 3 will be gradually blocked after approximately − 4 V.

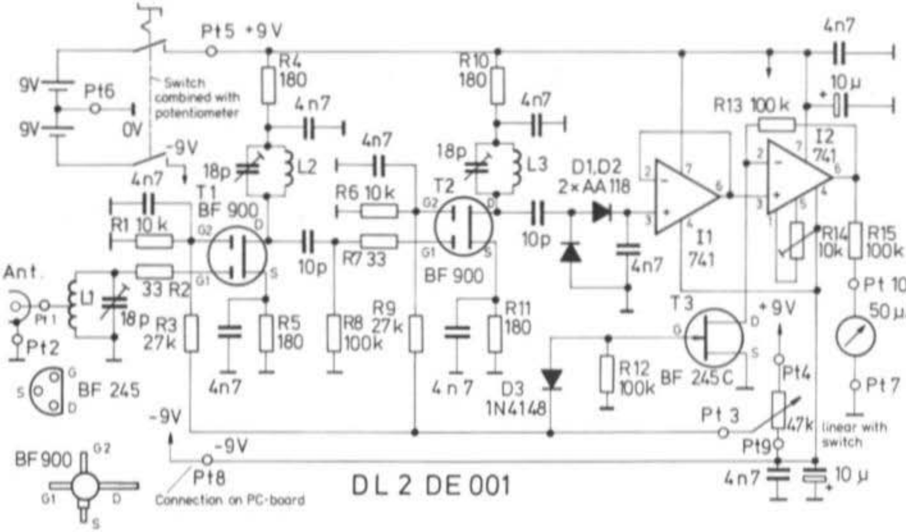


Fig. 1: A straight-through receiver for close-in DF-applications in the 144 MHz band

This means that the drain-source path will be of higher impedance. Since the gain of I 2 is dependent on the ratio of R 13 to R_{ds}, this will reduce the voltage gain of I 2 the more T 3 is blocked.

It should be noted that it is advisable for the given transistor BF 245 C to be used. This transistor has firstly a lower drain-source resistance than the BF 245 A or B and thus ensures a higher gain of I 2; secondly, the control of the DC-amplifier will only commence when T1 and T2 are already virtually blocked. The slope of transistors BF 245 A and B are less steep, and will commence control of the DC-amplifier even at lower negative gate voltages.

The overall adjustment range of the gain is in the order of 85 dB, and input voltages in the order of 1 to 2 V at the antenna will not cause any problems.

The zero adjustment of the meter is made with the aid of trimmer potentiometer R 14. Resistors R 2 and R 7 are provided to neutralize any tendency to UHF-oscillation. If a greater sensitivity is required, it is possible for R 2 and/or R 7 to be bridged, if no tendency to self-oscillation is exhibited. Furthermore, it is possible to increase the value of R 13 to 220 kΩ or 330 kΩ. However, this will cause a somewhat larger zero drift of the meter, especially when large temperature fluctuations occur.

It may seem a disadvantage that the receiver requires two 9 V batteries. However, this circuit does bring some advantages such as a symmetrical (stable) operation of both integrated amplifiers, which means that the zero-point of the meter is not affected by the operating voltage, a low temperature drift, and good control behaviour of all FETs, which can only be correctly blocked with the aid of a negative voltage.

MEASURED VALUES

A prototype was constructed in the publishers' laboratories. The following values were measured on this prototype:

Sensitivity (for a clear indication on the meter):	50 μV
Sensitivity (for FSD):	250 μV
Adjustment range of the gain:	85 dB
3 dB bandwidth:	5 MHz
Current drain at min. (max.) sensitivity:	
+ 9 V battery:	3 mA (10 mA)
− 9 V battery:	4 mA (3 mA)

The receiver will operate at ± 6 V. The life of a minus battery is approximately twice that of a plus battery.

COMPONENTS

T 1, T 2:	BF 900 (TI)
T 3:	BF 245 C (TI)
I 1, I 2:	741 (various manufacturers) DIP 8-pin
D 1, D 2:	AA 118 (various manufacturers)
D 3:	1 N 4148, 1 N 4151 or similar
L 1 - L 3:	6 turns of 1 mm dia. silver-plated copper wire, wound on a 4 mm former, pulled out to 10 mm in length, self-supporting. Coil tap on L 1: 2 turns from the cold end.

3 plastic foil trimmers 22 pF, 7.5 mm dia. (Philips: green)

All ceramic capacitors for 2.5 mm spacing.

All resistors for 10 mm spacing.

1 trimmer potentiometer 10 kΩ, for horizontal mounting, spacing 10/5 mm

2 tantalum electrolytics, drop type, spacing 2.5 mm

The value of R 15 must be found experimentally to suit the meter. The actual meter used (up to 1 mA) should just indicate FSD with 5 V at pin 6 of I 2.

The gain potentiometer R 16 should be equipped with two break contacts for switching off the receiver.

CONSTRUCTION

The circuit of the close-in DF-receiver can be accommodated on the PC-board shown in **Figure 2**. The dimensions of this board are 105 mm x 42 mm; it has been designated DL 2 DE 001. The component side possesses a continuous ground surface. In the case of the PC-board available from the publishers, the ground surface has been removed at positions where interconnections are made through the board to the conductor lanes.

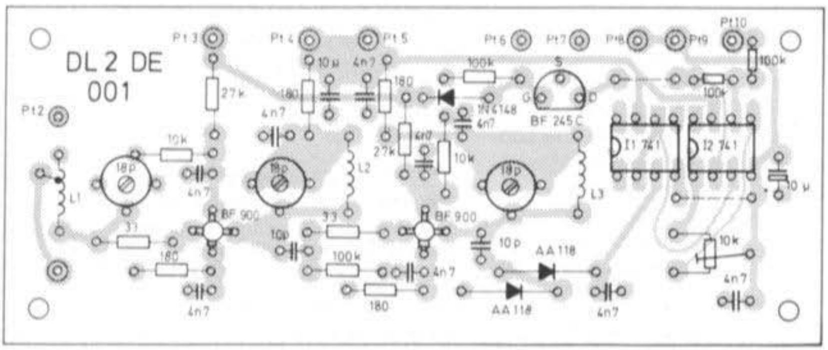


Fig. 2: The circuit given in Fig.1 can be accommodated on this double-coated PC-board (DL 2 DE 001)

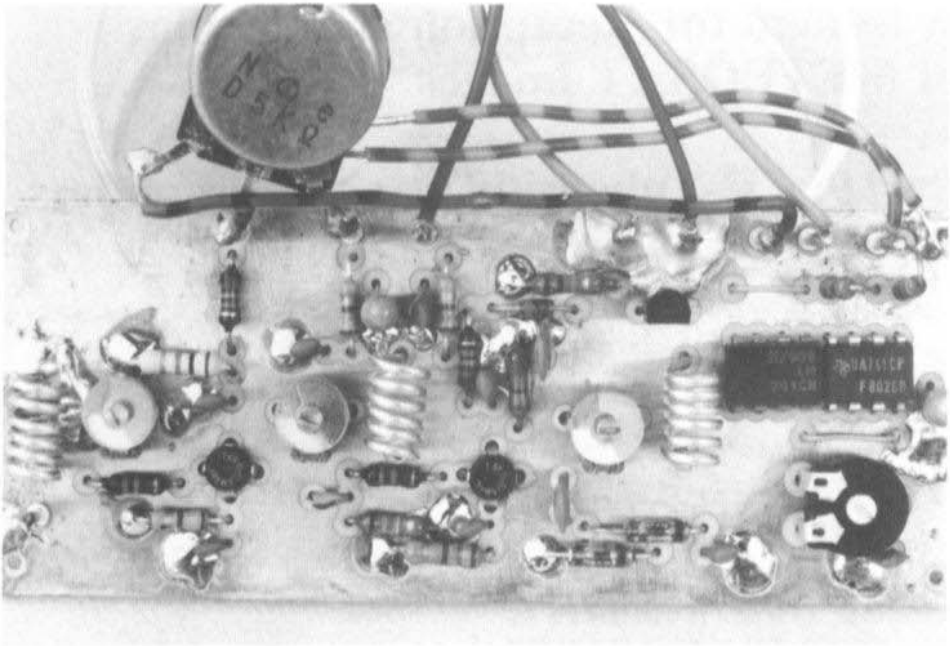


Fig. 3: The complete receiver consists of this PC-board, two 9 V batteries, one potentiometer with on-off switch, 1 meter and an antenna connector

A photograph of the prototype constructed by the publishers, is given in **Figure 3**; it will be seen that two short wire bridges are necessary in the vicinity of the operational amplifiers (see Figure 2).

It is necessary for the board to be enclosed in a RF-tight case for practical operation. It can, for instance, be constructed from PC-board material. In order to assist the installation of the completed, aligned board, it is advisable for the connection pins to be mounted on the conductor side, and not as shown in the photograph.

ALIGNMENT

Firstly set potentiometer R 16 to its fully clockwise position, in other words for full positive voltage on the wiper of the potentiometer. Trimmer R 14 is now aligned without input signal so that the meter indicates approximately 15 % of FSD. This will only be possible when the circuit does not break into oscillation. This adjustment allows a slight drift of the needle without correction and also allows an operational check to be made. If, for instance, R 16 has been turned down, the meter will return to zero after approximately 2/3 of its range. It should be noted that intensive light onto diodes D 1 and D 2 will also affect the zero-point adjustment.

The rest of the alignment only consists of aligning the three VHF resonant circuits for maximum reading at 145 MHz with the aid of the trimmer capacitors.

A System for Reception and Display of METEOSAT Images

Part 9: CR-Tube with X and Y Amplifiers, and EHT-Supply

by R. Tellert, DC 3 NT

The last part of the weather satellite reception system is now to describe the so-called monitor, that consists of a picture tube and three PC-boards:

- Deflection board DC 3 NT 014
- EHT module DC 3 NT 015
- Picture tube connection board DC 3 NT 016

After completing this module, it is possible to receive images from the geostationary satellites GOES, METEOSAT and GMS that are transmitted in an APT format on 1691.0 and 1694.5 MHz, as well as images from the orbiting satellites in the 137 MHz band (TIROS, NOAA, METEOR). These images are then traced line by line on the CRT and photographed with the aid of a camera. Of course, it is also possible for longwave and shortwave FAX transmissions to be recorded if a suitable receiver is provided.

9. SELECTION OF THE CRT

The first consideration was to use an oscilloscope tube, but this possessed a number of disadvantages:

- They are more expensive than mass-produced TV-picture tubes due to the lower quantities manufactured.
- Their resolution only amounts to 200 to 400 lines.
- The image is too small.
- Due to their small deflection angle, an unwanted modulation of the electron beam could easily be possible due to magnetic fields (AC-hum).
- The galvanically coupled drive of X, Y, and Z is very difficult since a potential difference of 1000 to 2000 V is present between deflection plates and cathode.

For this reason, it was decided to use a monochrome TV-tube. These are usually made in two types having either a thin or thick neck. Those with a diameter of 29 mm operate with an anode voltage in the order of 15 kV, which results in a brighter, sharper electron beam than when using the lower anode voltage of only 11 kV in the case of TV-tubes with a thinner neck. For this reason, the author finally selected a medium size TV-tube type A 44-120W with a thick neck and a diagonal screen size of 44 cm.

Of course, a smaller TV-tube can be used, if required.

The TV-tube A 44-120 W used in the author's prototype, requires the following voltages:

- EHT: + 15 kV
- Accelleration voltage A_2: + 500 V
- Focussing voltage: 0 to + 500 V
- Cathode voltage: approx.60 V, since the modulator cylinder is to be operated with a video voltage of 5 to 10 V.
- Heater: 6.3 V / 0.3 A

9.1. Modification of the Deflection System

In its original state, the horizontal and vertical deflection coils comprise two partial windings, each. These are connected in parallel for the horizontal deflection, and in series for the vertical deflection. This must be exactly opposite when using the deflection system as a monitor in the weather satellite system.

This means that the horizontal coil, which is directly on the picture tube, should be connected in series, after which it is still used for the X-deflection.

The coil wound on the ferrite core, on the other hand, is modified for parallel connection, and any thermistor removed.

The modified deflection system can now be placed on the neck of the CRT and be checked with the aid of a variable DC-source after the EHT-module is complete.

Before connecting voltage to the deflection system, a dark point should be aligned on the screen; a square sector should now be marked on the screen with this point as its center. The whole surface of the CRT should not be used due to the distortion at the corners, on the other hand, the surface used should not be too small.

This is followed by connecting a stabilized power supply via an A-meter to one of the deflection coils and increasing the voltage slowly until the dot touches one of the previously-marked lines. The current should now amount to approximately 350 mA ± 50 %. This is followed by repeating this measurement using the other deflection coil.

Finally, the voltage drop of the coils is measured: It should not exceed 3.5 V in the case of the X-coil, and 8 V with the Y-coil. When the measured values are within these limits, the deflection system will be suitable for use in the described monitor.

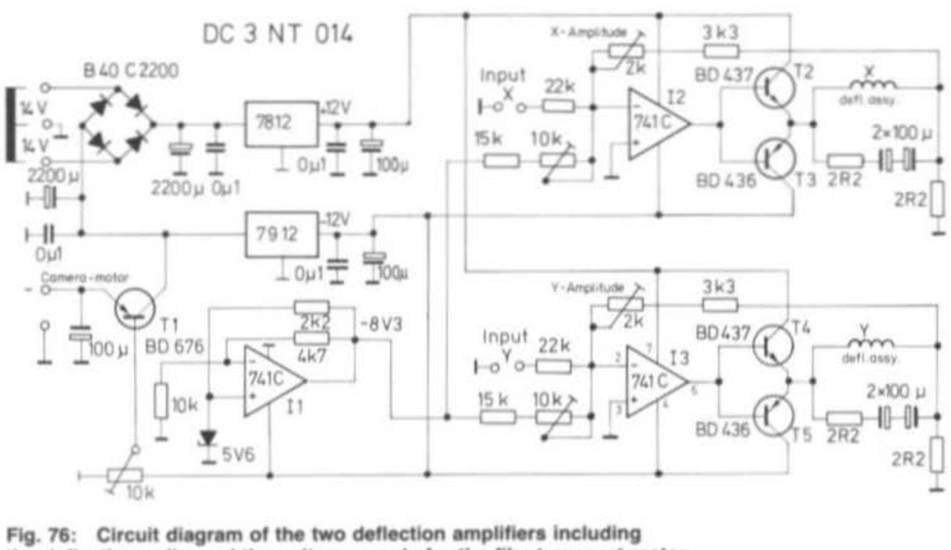


Fig. 76: Circuit diagram of the two deflection amplifiers including the deflection coils, and the voltage supply for the film-transport motor

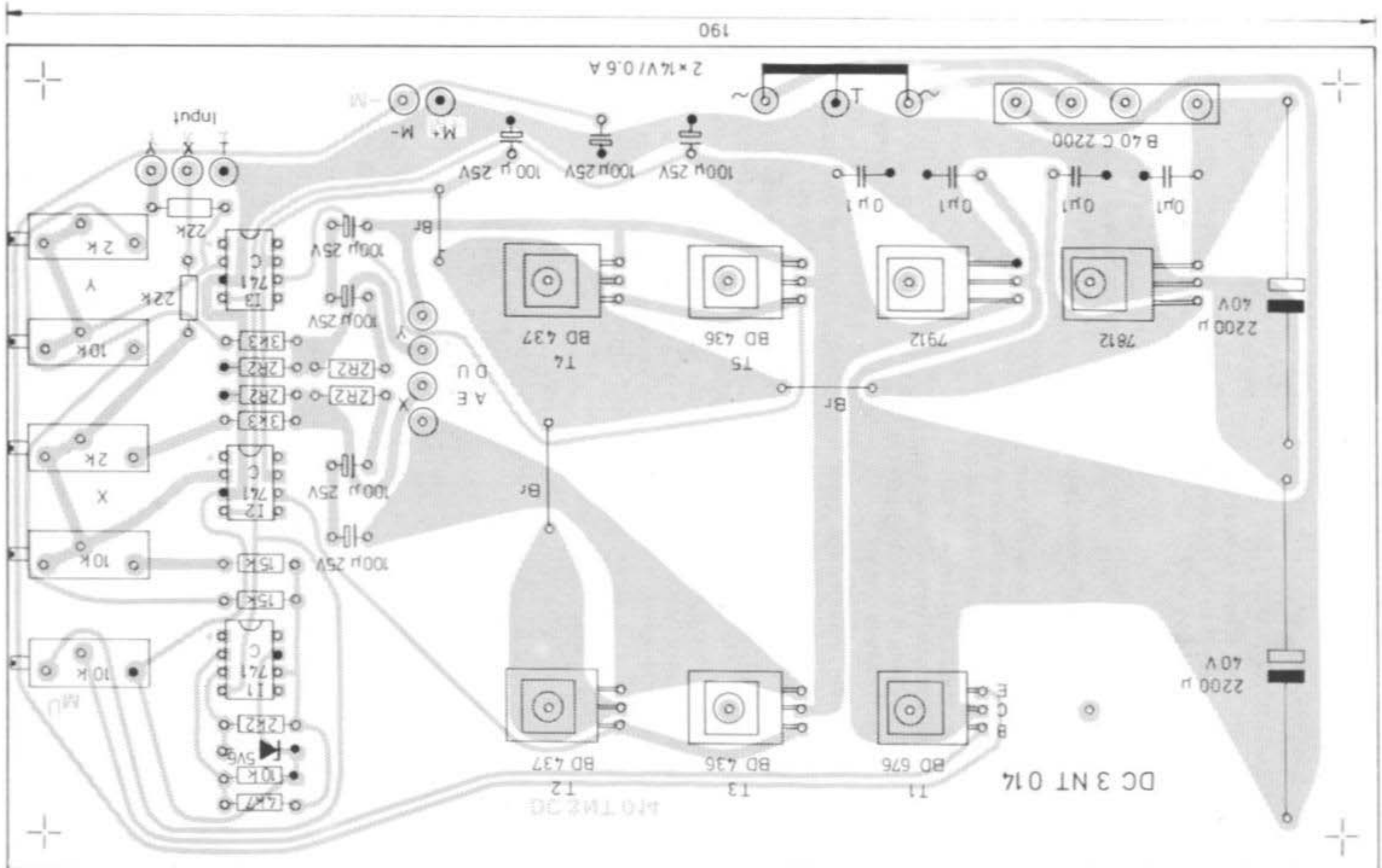


Fig. 77: PC-board DC 3 NT 014 accomodates the circuits shown in Fig.76

10. THE DEFLECTION AMPLIFIER MODULE DC 3 NT 014

The deflection amplifiers convert the X and Y-sawtooth voltages from module DC 3 NT 009 into proportional deflection currents. Since these currents must be in the negative and positive range, two power supply voltages of + 12 V and − 12 V are required. The circuit diagram given in **Figure 76** does not possess any special features. The input deflection voltages have a mean potential of + 7.5 V, which is compensated for using a negative reference voltage. The position of the electron beam can also be adjusted at this point (10 kΩ spindle trimmer potentiometers).

The deflection currents generate a voltage drop across each 2.2 Ω resistor, which serves for feedback of the deflection amplifiers that comprise an integrated amplifier, and a complementary pair of transistors, each. The gain can be adjusted with the aid of the 2 kΩ trimmer potentiometers, thus allowing the height and width of the image to be set.

The operating voltage for the film transport motor of the camera is also provided by the power supply of this module. The pass transistor T 1 has a variable base voltage, and is supplied from the negative voltage. This is because the motor will only run when X and Y are at the commencement of the image, in other words, when only the positive power supply is loaded.

The two deflection coils have been included in the circuit diagram for clarity; of course, they are not mounted on the PC-board.

10.1. Construction of the Deflection Amplifier

A single-coated PC-board of 190 mm x 115 mm was designed for accommodating module DC 3 NT 014. **Figure 77** shows the component locations on this board. The seven power transistors are arranged so that their dissipation can be transferred to four heatsinks, which should be constructed according to **Figure 78**. The construction can be seen in the photograph of the author's prototype given in **Figure 79**. This module is designed so that an insulated mounting is not necessary, and one should only pay attention that the heatsinks do not touch another or any other parts of the system.

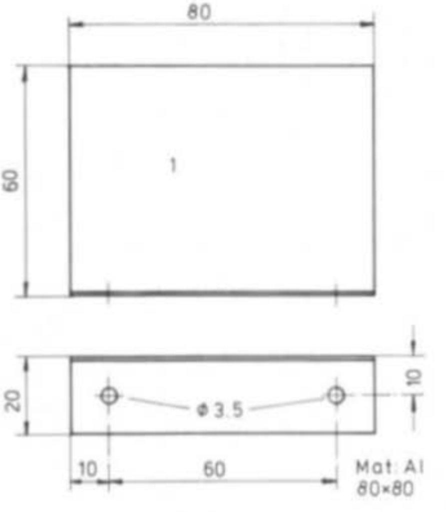


Fig. 78:
Four such heatsinks made from 1 mm alu.plate are required for module DC 3 NT 014

The five trimmer potentiometers are to be found on the edge of the board so that they are readily accessible for adjustment. If module DC 3 NT 014 is mounted suitably in the monitor, it is then possible for the motor voltage, and the position and amplitude of the image to be adjusted externally.

10.2. Components for DC 3 NT 014

T 1:	BD 676 PNP Darlington power transistor (Siemens)
T 2, T 4:	BD 437 NPN power transistors (Telefunken, Siemens)
T 3, T 5:	BD 436 PNP power transistors (Telefunken, Siemens)
I 1 - I 3:	741 C or TBA 221 B

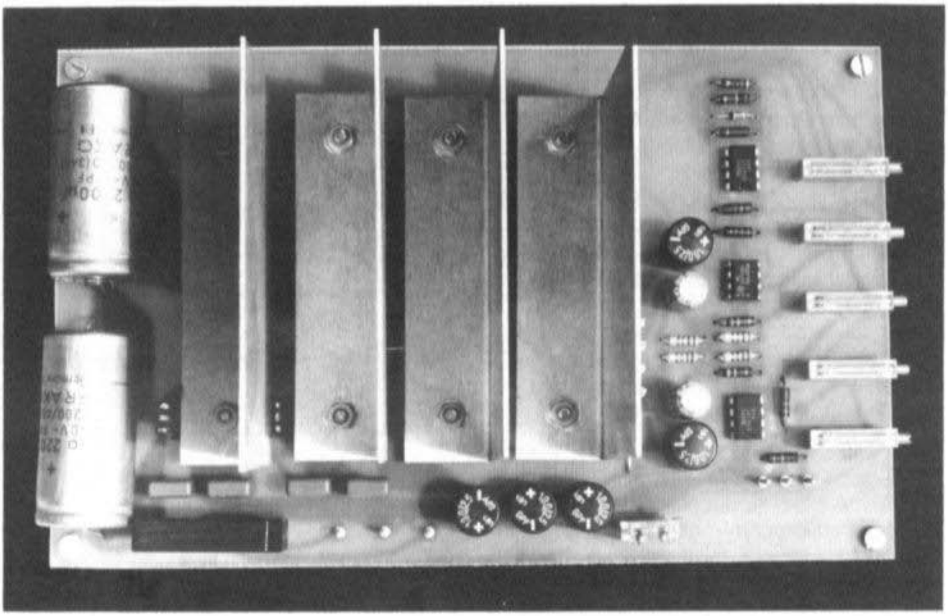


Fig. 79: Photograph of the author's prototype module DC 3 NT 014

1 pc. 7812 (12 V stabilizer, positive)

1 pc. 7912 (12 V stabilizer, negative)

1 bridge rectifier B 40 C 2200

1 zener diode C 5 V 6

2 aluminium electrolytics 2200 µF/40 V, max. 45 mm long

7 aluminium electrolytics 100 µF/25 V, for 5 mm spacing

4 plastic foil caps., 0.1 µF, for 7.5 mm spacing

2 pcs. 10-turn spindle trimmers 2 kΩ

3 pcs. 10-turn spindle trimmers 10 kΩ

All resistors for 10 mm spacing.

11. EHT-MODULE DC 3 NT 015

11.1. Selection and Modification of the Line Output Transformer

During the design and construction of the monitor, it was always attempted to select conventional TV-components. However, since the application differs considerably from that of a TV-transmission, it is necessary for various components such as the line output transformer and the deflection system to be modified.

In the case of an APT-image, the screen is scanned once every 200 seconds, whereas this process must take place in 1/50 s with television. This difference of 1 : 10000 corresponds approximately to the loading of the EHT circuit; it is so low in the case of our monitor that the current of the EHT-module is virtually not measurable (it is in the order of nA!)

A line output transformer originally designed for single-phase rectification, is driven at such a low level that the required HT-voltage can be generated with one cascade. However, the EHT-windings are completely sealed, which means that one has to accept their number of turns.

The line output transformers used in modern, battery-powered portable TV-receivers, are – unfortunately – not suitable for the cascade principle, since rectifier diodes are already sealed within the HT-windings. Finally, a line output transformer was found that was used in TV-receivers equipped with tubes in the line-output stage, which are still available. **Figure 80** shows the original circuit of this line output transformer ZTR 230, as well as its connections 1 to 9. Pin 1 must be isolated from ground and also connected to the board.

The higher HT-voltages are generated in the windings provided for this purpose; an additional three windings must be provided for the converter. This is not difficult since only 2, 7, or 25 turns are required respectively. These additional windings are designated with A/B (seven turns of 0.9 to 1 mm enamelled copper wire), B/C (25 turns of 0.3 mm enamelled copper wire), and D/E (2 turns of insulated wire) in the circuit diagram of DC 3 NT 015/016 (see **Figure 81**). These designations are also to be found as connection points in the component location plan. The direction of winding is given in **Figure 80a**.

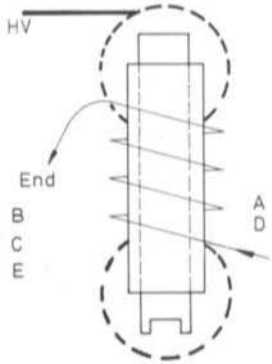
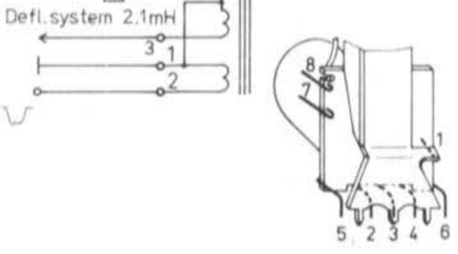


Fig. 80a:
Direction of winding the additional windings on a vacant side of the line transformer



ZTR 230

Fig. 80: A suitable line-output transformer, originally for 110° TV-tubes, 2.1 mH horizontal inductance, 18 kV rectifier, as used for tubes PL 500, PY 88

11.2. Circuit of the EHT-Module

The EHT-voltage is generated in a self-oscillating, blocking oscillator with subsequent rectifier cascade. Since the converter is mainly used to compensate for the transformer losses and to provide the drive power of the transistor, it is possible for the stabilization to be made with the aid of the base current. A parallel circuit of three resistors in the base circuit of the transistor (MJE 13007) is provided for adjustment of the EHT-voltage.

In order to ensure that no holes are burnt in the fluorescent surface of the picture tube, the voltage at the modulator electrode is limited to + 15 V (− 33 V with respect to the cathode!). A 15 V zener diode is provided on PC-board DC 3 NT 016 for this purpose – in other words directly on the socket of the CRT.

If the cathode voltage now falls below + 48 V, which would then cause a higher beam current to flow, the voltage U_{A2} will be controlled down. This is the only possibility of protecting the CRT, since the EHT-voltage cannot be discharged so quickly. Even though, one must be extremely careful when switching on the monitor for the first time, or when making alignments on it.

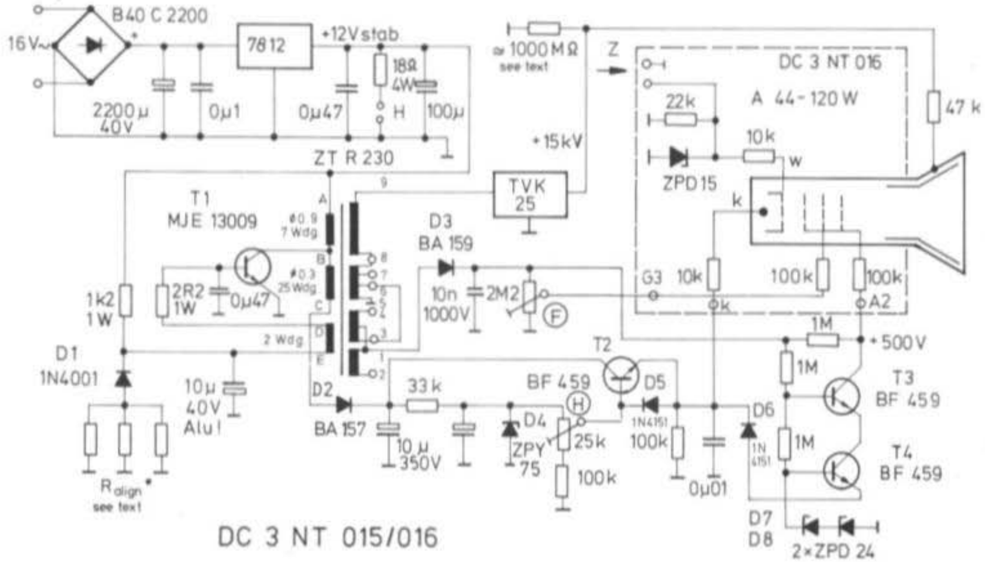


Fig. 81: The EHT-module generates 15 kV using a self-excited blocking oscillator and a voltage tripler cascade. The components contained within the dashed lines are mounted on board DC 3 NT 016

The EHT-voltage is now loaded with the aid of a resistor and discharged after switching off. This resistor must have a breakdown voltage of 18 kV ! The most simple way of doing this is to use a resistor from an EHT-probe. It may be possible to obtain a suitable resistor in the order of 220 MΩ from an old TV-receiver.

In order to avoid corona, and dangerous flash-over of the EHT-voltage, it is necessary for the »hot« connection of the loading resistor to be provided with a plastic cap, or even better for it to be completely sealed. If such a resistor cannot be obtained, it is possible for it to be made from 10 series-connected 22 MΩ resistors which are then sealed, or covered with a heat-shrinking tube. In any case, this resistor must be installed so that it is far enough away from other components.

11.3. Construction of the EHT-Module DC 3 NT 015

With the exception of the previously described load resistor, the rest of the EHT-module is accommodated on a 170 mm x 145 mm single-coated PC-board. **Figure 82** shows the component locations on this board, which is designated DC 3 NT 015. The oscillator transistor and the voltage stabilizer are mounted on a 70 mm high aluminium heatsink, (the MJE 13009 must be insulated !), as shown in **Figure 83**. The five connections in the top right-hand corner of Figure 82 are interconnected via a suitable length of wire to the identical designations on PC-board DC 3 NT 016. The heater voltage for the CRT (designated: H) is also provided here; this is made with DC in order to avoid any hum. The heater current is limited to 300 mA with the aid of an 18 Ω resistor.

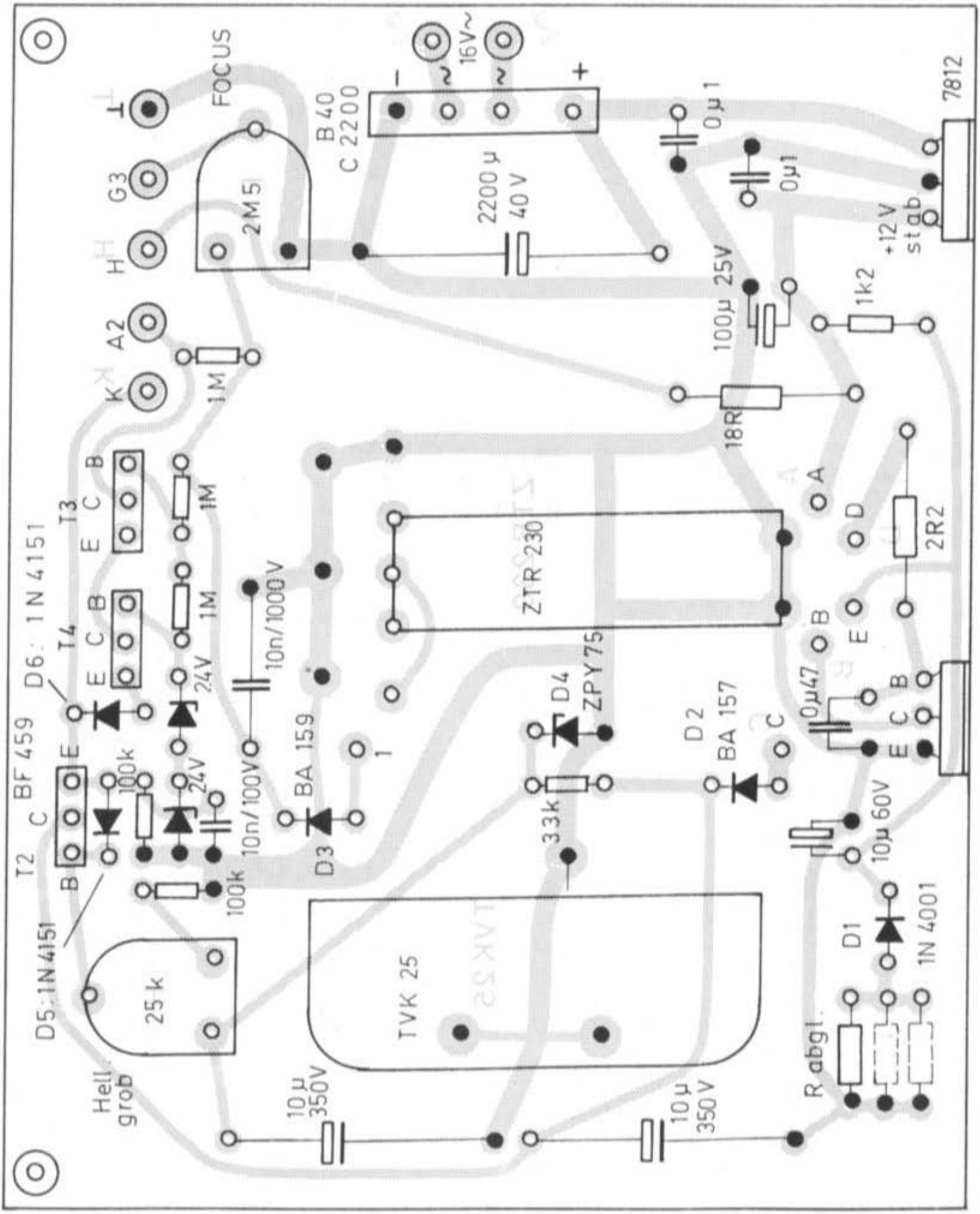


Fig. 82: Component locations on the single-coated EHT-board DC 3 NT 015. Dimensions: 170 mm x 145 mm

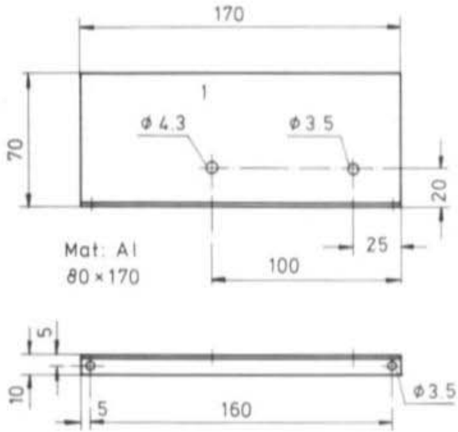


Fig. 83: The heatsink for the MJE 13009 is made from 1 mm thick aluminium (mount the transistor in an insulated manner !). Also required for the 7812

11.4. Components for DC 3 NT 015

T 1:	MJE 13009 NPN power transistor (Motorola), maybe MJE 13007
T 2 - T 4:	BF 459 NPN power transistor with U_{CBo} = 300 V (Siemens)
D 1:	1 N 4001 or similar
D 2:	BA 157 (or 159) fast silicon rectifier for 400 V (ITT)
D 3:	BA 159 (or 3 x BA 157) for 1000 V (ITT)
D 4:	75 V zener diode ZPY 75 (ITT) or BZY 97 C 75 (Siemens). It can also be replaced by a suitable series-circuit
D 5, D 6:	1 N 4151 or similar
D 7, D 8:	24 V zener diode ZPD 24 (ITT) or BZX 97 C 24 (Siemens)

1 silicon bridge rectifier B 40 C 2200 (Siemens)

1 pc. 7812 (12 V stabilizer, positive)

1 line output transformer; modified, see text

1 EHT-cascade TVK 25

1 aluminium electrolytic 2200 μF/40 V (length max. 40 mm)

2 aluminium electrolytics 10 μF/350 V (length max. 37.5 mm)

1 aluminium electrolytic 100 μF/25 V (can type, spacing 5 mm)

1 aluminium electrolytic 10 μF/40 V (can type, spacing 5 mm)

Do not use tantalum electrolytics !

1 plastic foil capacitor 10 nF/1000 V (spacing 22.5 mm)

1 plastic foil capacitor 10 nF/100 V (spacing 7.5 mm)

1 plastic foil capacitor 0.47 μF (spacing 7.5 mm)

2 plastic foil capacitors 0.1 μF (spacing 7.5 mm)

1 trimmer potentiometer 2.5 MΩ (horizontal mounting, spacing 17.5/10 mm)

1 trimmer potentiometer 25 kΩ (horizontal mounting, spacing 17.5/10 mm)

1 resistor 2.2 Ω/4 W (25 mm spacing)
1 resistor 18 Ω/2 W (25 mm spacing)
1 resistor 1.2 kΩ (12.5 mm spacing)

R_{align}:	3 resistors of 22 Ω/1 W each (spacing 15 mm), if required, correct as described in the text.

1 EHT resistor (18 kV) 220-1000 MΩ (see text).

Figure 84 shows a photograph of the author's prototype module DC 3 NT 015.

11.5. Module DC 3 NT 016

This small PC-board, whose dimensions are 40 mm x 40 mm, does not have its own circuit diagram, and is only used for accommodating the components around the CRT-socket, which were already given in Figure 81. The video signal, which has not been mentioned up to now, is fed to the input designated with »Z« on PC-board DC 3 NT 016.

The conductor lanes are arranged so that the voltage peaks are grounded if any flashover should occur in the picture tube. For this reason, the ground surface and all metal parts of the picture tube should be grounded to this board. The component locations are given in **Figure 85**, and **Figure 86** shows a photograph of a prototype board.

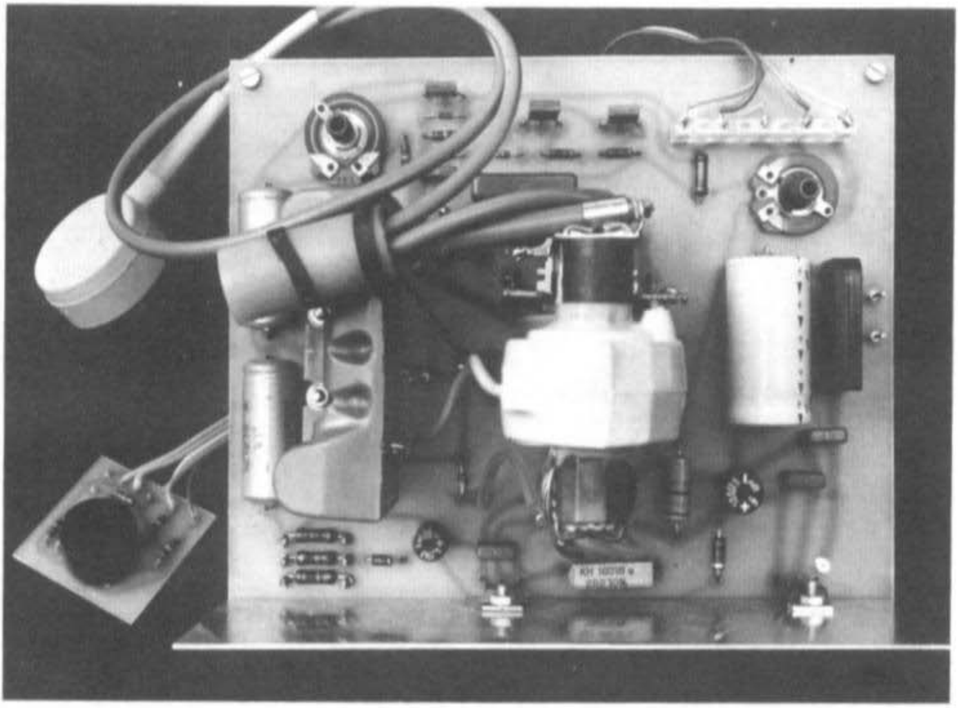


Fig. 84: Photograph of an prototype of the EHT-module DC 3 NT 015; the EHT-resistor is between the cascade (left) and the line transformer (center)

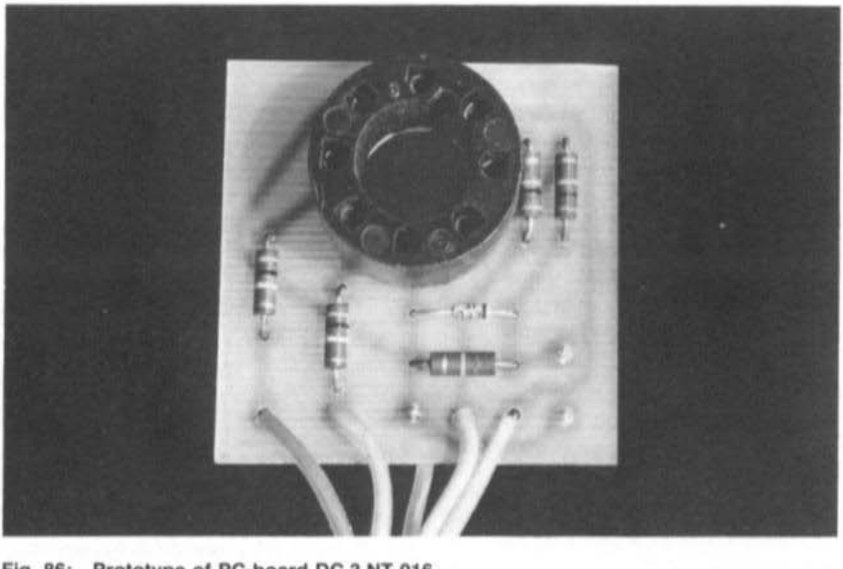


Fig. 86: Prototype of PC-board DC 3 NT 016

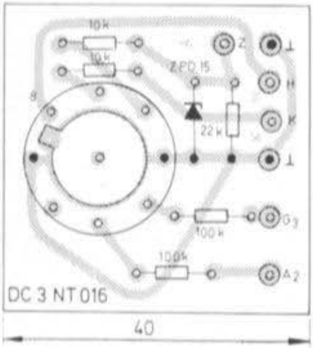


Fig. 85: This small board accommodates the components at the CRT-socket

Components for DC 3 NT 016

1 octal socket for PC-board mouting
1 zener diode, 15 V: ZPD 15 (ITT) or BZX 97 C 15 (Siemens)
5 resistors for 10 mm spacing.

11.6. Interconnection and Alignment of the EHT-Module

The EHT-module should be firstly checked without connecting the CRT. The EHT-voltage is firstly set by replacing one of the base resistors – given as »R_{align}« in Figure 81 or 82 – by a 50 Ω trimmer. The alignment is commenced with the highest resistance value and it is reduced until an EHT-voltage of 14.5 to 15 kV is measured. The actual resistance value of the trimmer is then measured and replaced by a fixed resistor of the same value. The power loading of the resistor combination (approx. 7 Ω) in the base circuit is in the order of 2 W.

This is followed by checking voltage U_{A2}, which should be in the order of 500 V. Attention should be paid during the measurement that the impedance of this voltage source amounts to approximately 1 MΩ.

The CRT can be connected after checking the other operating voltages.

In order to ensure that no dot is burnt into the fluorescent surface, the beam is deflected by connecting a 50 Hz AC-voltage to one of the deflection windings and allowing a current of approximately 200 mA to flow. The result should be a straight line on the screen of the CRT, whose brightness can be varied by altering the DC-voltage at the modulator electrode (input »Z« on PC-board DC 3 NT 016).

If the EHT-module is working correctly, it is now possible for the deflection amplifiers to be checked together with the deflection system until one is able to measure the deflected point on the screen of the CRT as already described in section 9.1.

12. ASSEMBLY OF THE MONITOR

It was already mentioned in Part 8 of this description that the CRT could be mounted in a wooden case with the screen facing upwards **(Figure 87)**. The wooden case must be light-tight, and be large enough to accept the selected CRT together with its mounting plates.

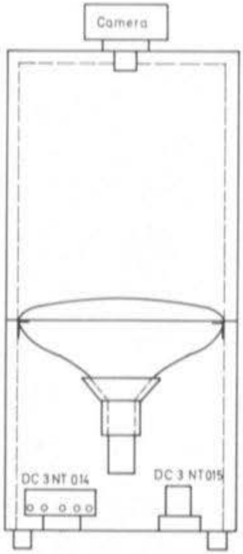


Fig. 87: Possible construction of the monitor as used by the author

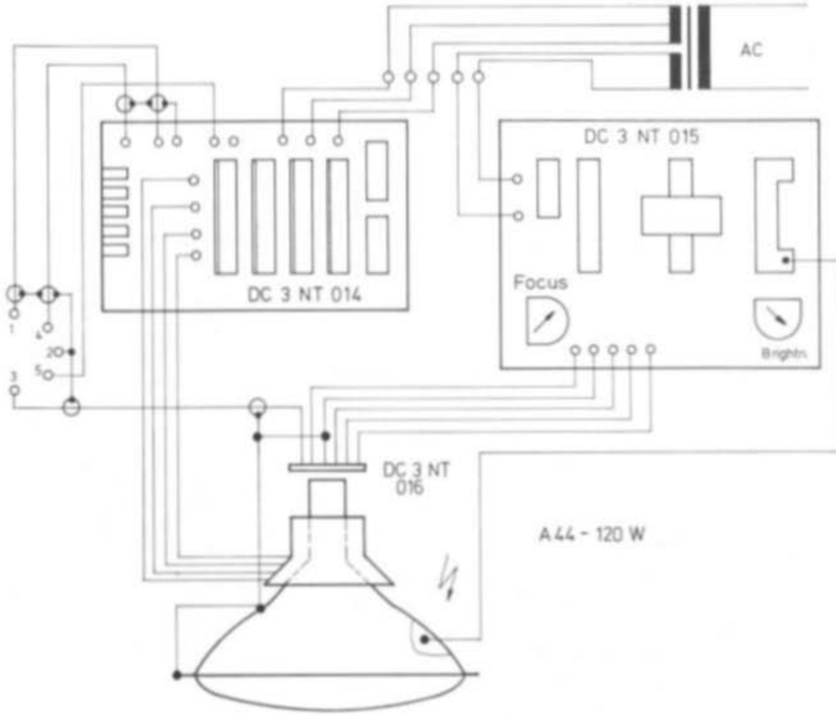


Fig. 88: Wiring of the monitor

Four brackets constructed from 2 mm thick aluminium plate should be prepared and screwed into the corners for mounting the CRT. The upper edge of the case must be sawn carefully and remain free so that it can accept the light-tight, matt-black camera adapter, which can also be made from wood. The height of this adapter is dependent on the focal length of the camera lens. In the case of reflex cameras, the most favorable distance can be determined with the aid of the square marked on the CRT screen.

Modules DC 3 NT 014 and 015 are screwed into position in the bottom of the monitor box. PC-board 016 is plugged onto the CRT-socket and wired as shown in **Figure 88**.

The power transformer should not be installed in the monitor itself, since magnetic fields in the order of the video frequency (0 to 2 kHz) would be visible as a patterning of the image.

Fig. 89: This METEOSAT test image shows the fine resolution, clean grey steps, and the slight distortion at the corners caused by the CRT

For this reason, the transformer should be accommodated in a separate case, and be mounted at least 1 m from the picture tube. The following voltages are fed to the monitor using a five-core cable:

28 V/0.8 A with center tap for DC 3 NT 014
16 V/1 A for DC 3 NT 015/016.

The lower part of the monitor case is provided with a large cutout in the vicinity of the CRT-socket. An aluminium plate is mounted (light-tight) over this cutout and is provided with one connector each for interconnection to the electronic module (DC 3 NT 009), and to the previously mentioned power supply. In addition to this, one can provide holes for adjusting the spindle trimmers on board DC 3 NT 014. These should also be covered (light-tight) after completing the alignment.

The following images are now to demonstrate the quality that can be obtained using this system. They were directly photographed using the described monitor with a 6 x 6 cm, or a 35 mm reflex camera of the prototype system constructed by the publishers.

This completes the description, however, T. Bittan, DJ Ø BQ is to describe suitable antennas for use in conjunction with this system.

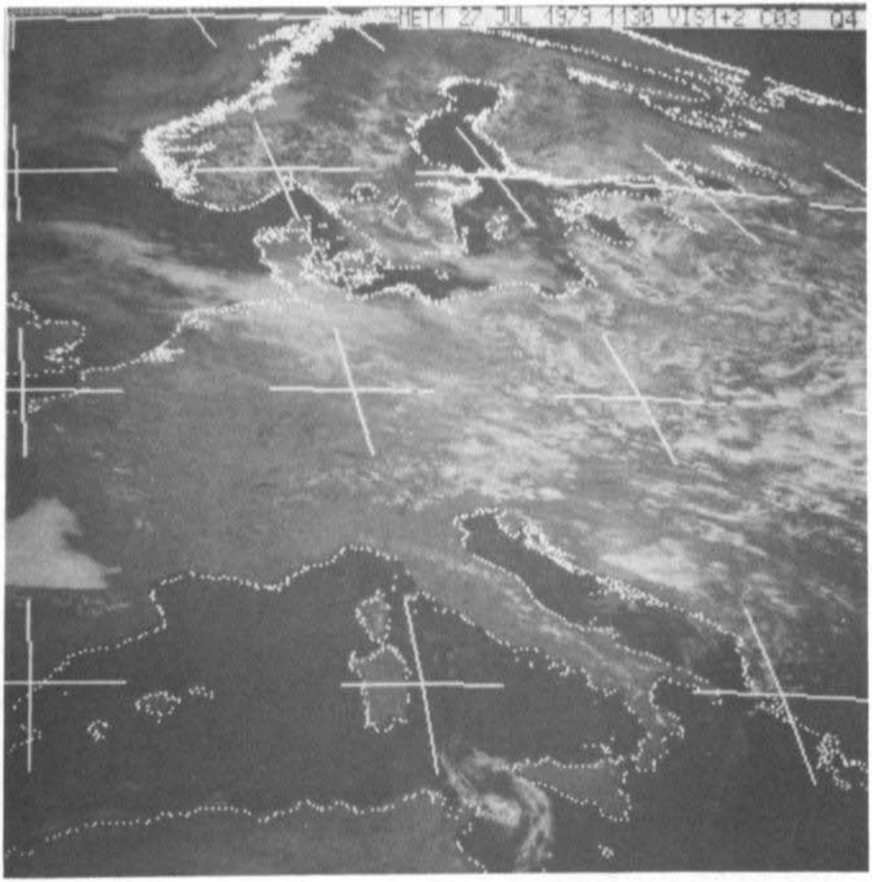


Fig. 90: This is how METEOSAT saw the summer weather over Europe at 11.30 on the 27.7.79; the white pointed lines and the data line originate from the computer of the ESOC in Darmstadt, West Germany

Fig. 91: An infrared image from TIROS-N (137.620 MHz). This shows that it is just as hot (dark) in Spain and Sicily as in Algeria and Tunesia. This satellite ceased operation on 1.11.80, and has been replaced by NOAA 7 which is operating on the same frequency

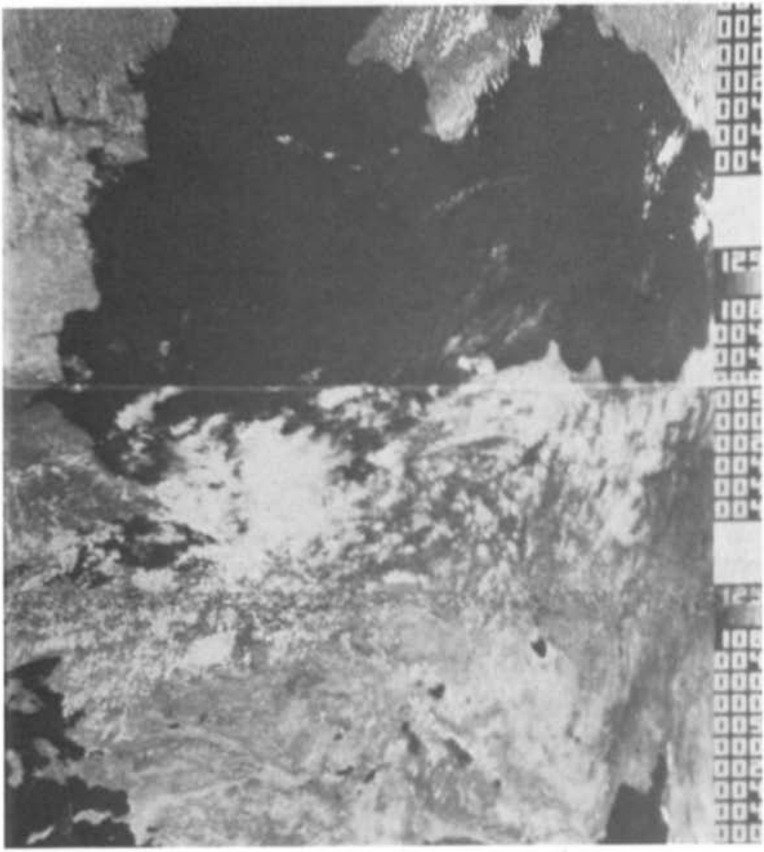


Fig. 92: Image from a METEOR satellite, (137.150 MHz, 240 lines per minute) – showing the Black Sea and Turkey virtually free of clouds. You will notice a considerable distortion due to the curvature of the earth

Fig. 93:
Such large-area images were seldom sent via METEOSAT:
It will be seen that Europe was completely free of clouds on 1.9.1979 with the exception of Greece, Scandinavia, and England/Ireland

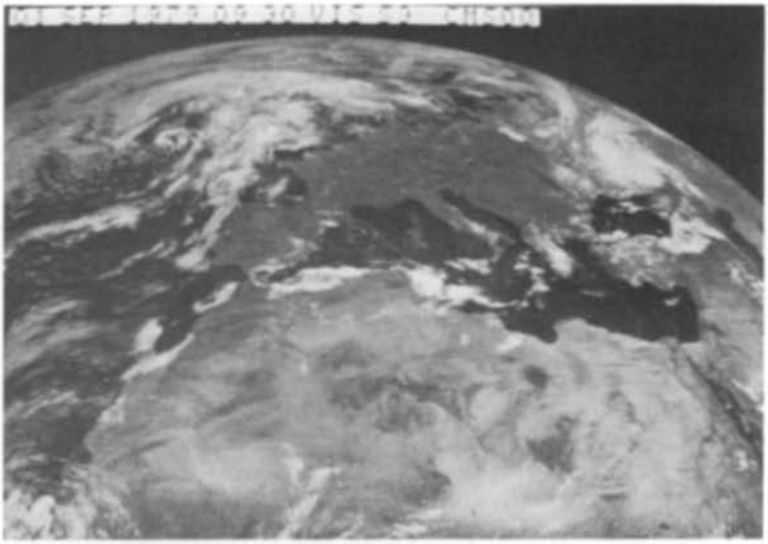


Fig. 94:
Image from one of the METEOR satellites (137.3 MHz/ 120 lines per minute) showing an impressive image of a low-pressure area with its fronts

A Ring Mixer Module for the DJ 4 LB ATV-Transmitter

by B. Roessle, DJ 1 JZ

It is well known to all those readers actively involved with TV-transmitters that intermodulation will occur when sound and video carriers are processed or amplified together. These interference signals are spaced 5.5 MHz from the video or sound carrier, as well as from another. These intermodulation products cannot be completely avoided when using the IF-principle used by DJ 4 LB (1); however, they can be suppressed to more than 30 dB down on the video carrier by using a high-level mixer stage and a really linear »linear amplifier«. This means that it is possible for them to be more than 60 dB down when using a filter such as described by DL 4 FA in (2).

It has often been suggested that a Schottky-diode ring mixer should be used in a module DJ 4 LB 004 instead of the transistor push-pull mixer. Unfortunately, no detailed article resulted from this. This article is to describe a ring mixer/linear amplifier module which can be constructed and used to replace module DJ 4 LB 004 in the ATV-transmitter.

When driven with 0.5 mW at 38.9/33.4 MHz, the output power will be 40 mW at 434.25/439.75 MHz. As can be seen in **Figure 1**, the strongest intermodulation products (video carrier − 5.5 MHz and sound carrier + 5.5 MHz) are suppressed by appproximately 47 dB ! The oscillator frequency (473.15 MHz) is suppressed by approximately 50 dB. This module is used in the ready-to-operate transmitter ATV-7010, in which it drives a two-stage linear amplifier to an output of 10 W.

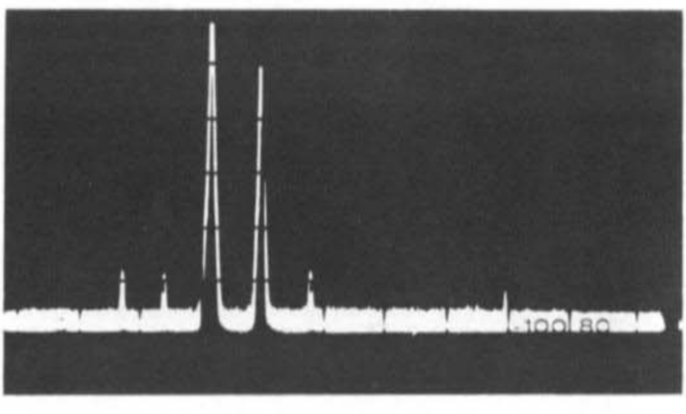


Fig. 1: Spectrum of the module when aligned to 36 mW video, and 4 mW sound carrier. The intermodulation products are appr. 47 dB down on the video carrier; the oscillator signal (4 sections to the right of SC) is appr. 50 dB down.
(Photo: VHF COMMUNICATIONS)

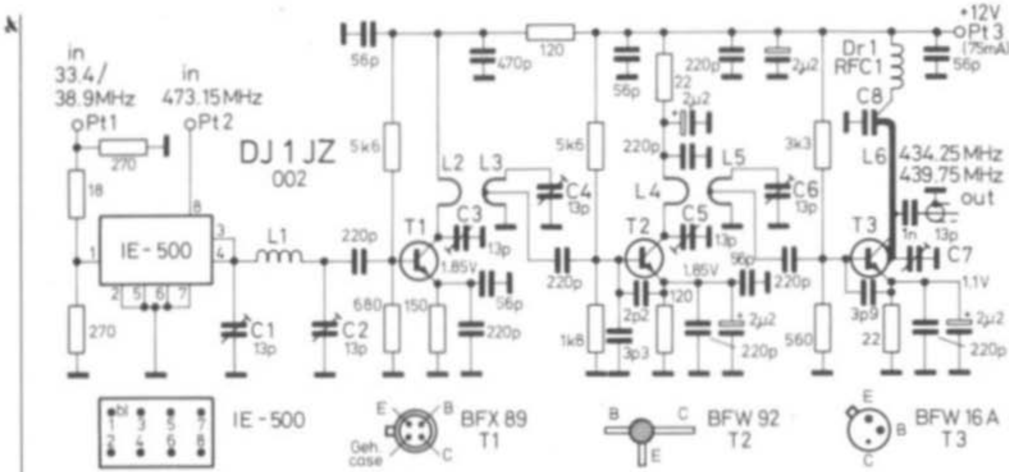


Fig. 2: The bypass and »neutralizing« capacitors are very important. These have different – however not very critical – values

CIRCUIT DESCRIPTION

Figure 2 shows a ring mixer with subsequent three-stage linear amplifier. The local oscillator input Pt 2 is provided with approximately 5 mW at 473.150 MHz at 50 Ω from the unmodified module DJ 4 LB 003.

Input Pt 1 is fed with the video and sound signal at IF-level. A 3 dB attenuator with an impedance of 50 Ω ensures the correct matching between ring mixer and residual sideband filter (DJ 6 PI 004). The IF video carrier level should not exceed 2 mW at Pt 1, otherwise the specifications of a standard level mixer such as IE-500, SRA-1, or MD-108 will not be obtained. Approximately 0.1 mW at 435 MHz will be available at the output of the ring mixer (pins 3 and 4). This signal is then passed via a Pi low-pass filter to the first amplifier stage.

A two-stage bandpass filter coupling is provided between the amplifier stages to ensure a good suppression of the image and oscillator frequencies. A capacitively shortened λ/4 circuit is used at the output of the low-level stage equipped with T 3. This allows the output coupling at 50 Ω to be made easily.

Finally, it should not be forgotten that the described module is, of course, just as suitable for construction of transmitters or transverters for other modes, especially for SSB.

CONSTRUCTION

The double-coated PC-board DJ 1 JZ 002 (see **Figure 3**) has dimensions of 135 mm x 50 mm, and is suitable for accommodation in a tin-plate case of the same dimensions (30 mm high). The continuous, copper surface represents the component side, and is only provided with countersunk holes where the component connections are passed through the board to the conductor side. The PC-boards offered by the publishers are already drilled.

After drilling the board, it is soldered into the case with a spacing of approximately 5 mm between the conductor lanes and the lower cover. The solder joint should be made all around the board on the component side. Finally, the screening panels should be soldered across the module, and in the vicinity of Pt 2 as shown by the dashed lines given in Figure 3. These panels are approximately 23 mm high, and are very important in achieving a clean output spectrum.

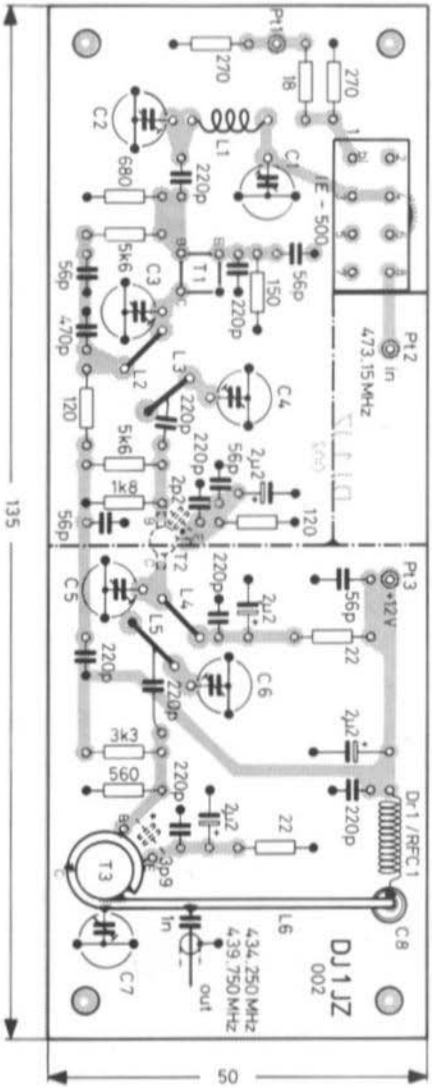


Fig. 3:
One screening panel is placed over T 2, and the other between this and the ring mixer. Both panels should be as high as the case and must be completely soldered

After this, it is possible for the other components to be installed. All ground connections are soldered to the upper side. Three components (marked with * in Figure 3) should be soldered to the conductor lanes; a neutralized capacitor of 2.2 pF from the base to emitter of T 2, T 2 (BFW 92) itself, and a similar capacitor of 3.9 pF from the base to emitter of T 3 (BFW 16 A). Special attention must be paid that these components are soldered into place with the very shortest connection length.

Pin 1 (RF-input) of all known ring mixers is marked in some way. After the ring mixer has been inserted into the board correctly, and pins 1, 3, 4, and 8 have been soldered to the conductor lanes, the case should be soldered to the ground surface on one side. The soldering process should be made quickly ! Pins 2, 5, 6, and 7 are internally connected to the case in type IE-500.

The line inductance L 6 is bent around transistor T 3 so that it fits tightly and soldered to the case of this transistor. The line is thus also used as heat sink. The collector connection of this transistor can be removed or considerably shortened and soldered to line L 6. This means that only base and emitter of T 3 must be connected to the PC-board, and should be so long that L 6 possesses a spacing of 4 mm from the ground surface. The ground connections of the trimmer capacitor of this stage are also only connected to the board, and its hot end is bent up and soldered to T 3/L 6. The cold end of L 6 is bent down and soldered to a chip capacitor that is soldered tothe ground surface as shown in Figure 3. An RF-choke is soldered into place here for filtering the operating voltage.

COMPONENTS

T 1:	BFX 89 (Siemens)
T 2:	BFW 92 (Siemens) or 2 N 6621
T 3:	BFW 16 A (Siemens)
Mixer:	IE-500 (MCL). When using other 500 MHz ring mixers, it may be necessary for the ground pins 2, 5, 6, and 7 to be grounded externally !

L 1:	3 turns of 1 mm dia. silver-plated copper wire wound on a 5 mm former, pulled out to fit the hole spacing on the board, soldered into place so that it has a spacing of 2 mm from the board, self-supporting.
L 2 - L 5:	Hair-pin loop from 1 mm dia. silver-plated copper wire, bent around a 5 mm former, and soldered into place 11 mm above the board; tap on L 3 and L 5: 7 mm from the hot end.
L 6:	Approx. 60 mm of 1.5 mm dia. silver-plated copper wire of which 3 mm is bent down, and approx. 25 mm wound around T 3. Tap: approx. 25 mm from the cold end.
RFC 1:	19 cm of 0.4 mm dia. enamelled copper wire, wound around a 3 mm former. Pull out the first 5 turns somewhat, self-supporting
C 8:	Ceramic chip capacitor, approx. 470 pF

6 x 13 pF, 1 x 6 pF plastic foil trimmers of 7.5 mm dia.

4 drop tantalum electrolytics: 2.2 μF / 25 V

All other capacitors:
Ceramic disk types for 5 mm spacing

All resistors: for 10 mm spacing

1 metal case 50 mm x 135 mm x 30 mm

ALIGNMENT

Provide the module **(Figure 4)** with an operating voltage of 12 V and check the operating points of the three transistors: In the case of T 1, a voltage of approximately 1.9 V should be set across the 150 Ω emitter resistor, a voltage of 1.9 V should also be present across the 120 Ω emitter resistor of T 2, and finally approximately 1.1 V at the 22Ω resistor of T 3.

The alignment can be carried out easily using a (simple) sweep generator: The demodulator of the sweep generator is connected to the output of the module and the generator output signal is fed via a coupling link of two turns, and placed firstly in the vicinity of L 5 in order to align L 6. This is followed by injecting into L 4 and aligning L 5, after which the signal is injected into L 3 and L 4 is aligned. This process is continued back to L 1, and the resonant circuits should be aligned with the aid of frequency markers to obtain a flat passband curve. The Pi-filter of the ring mixer is then finally aligned for maximum output power, or – if possible – for minimum intermodulation products; Figure 1 shows the output spectrum that can be obtained.

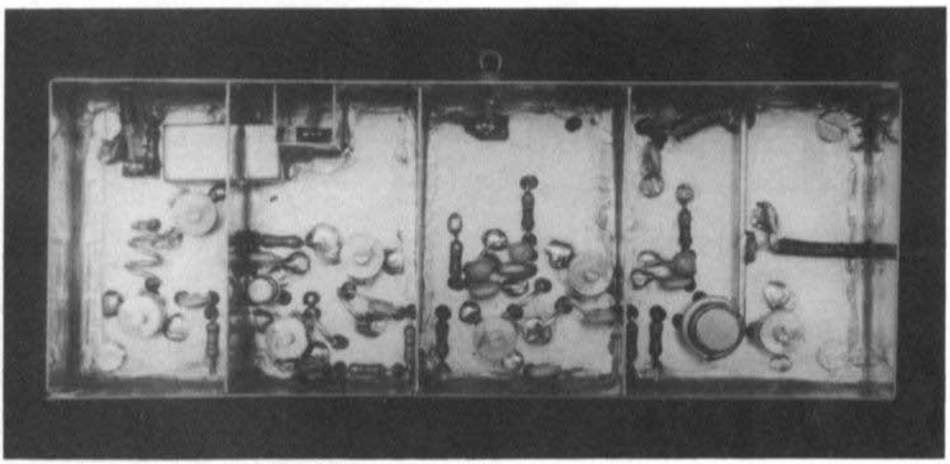


Fig. 4: Photograph of the author's prototype. Not all measures have been made at this stage, however, several additional panels have been installed

EDITORIAL DISCUSSION:

Is it really necessary for the video/sound carrier spacing to be 10 dB ?

When attempting to obtain better intermodulation rejection of ATV-transmitters as described in this article, it will be seen that the intermodulation products decrease rapidly when the sound carrier is more than 10 dB down on the video carrier. The publishers have experimented to find out how far one can go and what can be obtained with these measures.

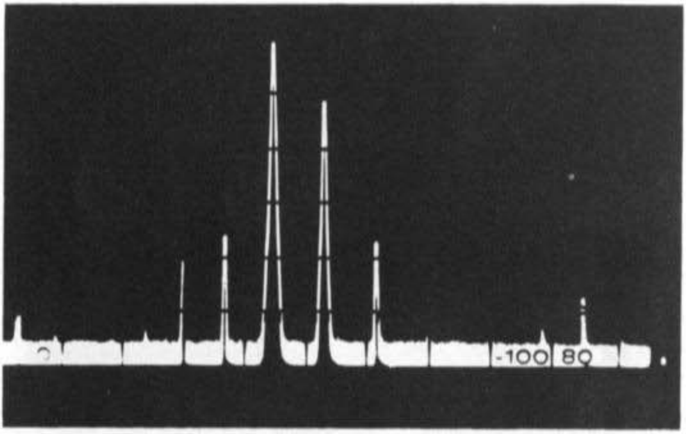


Fig. 5: Output spectrum; the VC-power is 9 W and the SC-power is 1 W. The IM_3-products are approx. 37 dB down on the VC. (Photo: VHF COMMUNICATIONS)

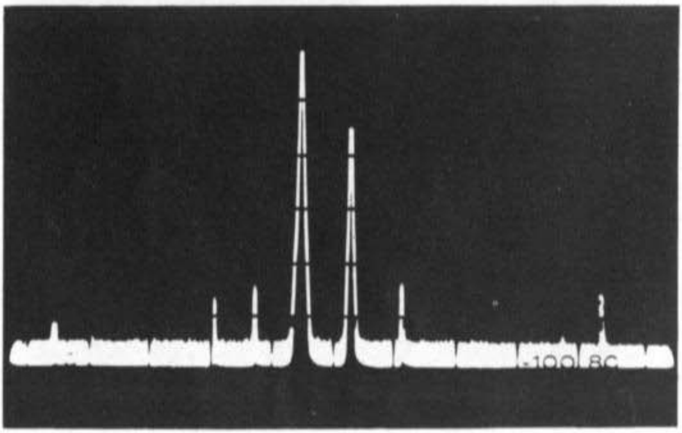


Fig. 6: The VC-power is now approx. 10 W and the SC-power only 250 mW. This allows the IM-rejection to be increased to approx. 45 dB. (Photo: VHF COMMUNICATIONS)

When using an ATV-transmitter as described by DL 4 LB and a conventional monochrome TV-receiver, the attenuation was increased until only a few rough structures were visible in the noise. When the sound carrier is 10 dB down on the video carrier, the voice modulation is still completely readable.

After this, the output level of the sound carrier was reduced until the readability was just R 5. This showed the quality of the AF-amplifier used: it was virtually independent of the microphone and speaker, and voice communication was still possible when the sound carrier was 20 dB down on the video carrier !

In order to ensure that the drive is not quite so critical, and in order to have a certain power reserve, the sound carrier was finally adjusted to be 16 dB down on the video carrier. This reduction of the sound carrier by 6 dB with respect to the standard, provided approximately 8 dB improvement of the third-order intermodulation, which will be seen by comparing **Figure 5 and 6**. These spectra were made at a video carrier output of 10 W, using a linear amplifier equipped with transistors 2 N 5944 and MRF 644 subsequent to module DJ 1 JZ 002 !

As can be seen in Figure 6, the third-order intermodulation amounted to 45 dB, and fifth-order intermodulation to 47 dB. The oscillator and image rejection was approximately 47 dB. This means that the low (IF) concept of ATV-transmitters can still be improved.

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A Microcomputer for Amateur Radio Applications

Part 7: The TV-Interface

by W. Kurz, DK 2 RY

This article represents the last of the series, and is to describe the TV-interface completing the microcomputer system. The TV-interface generates characters on a conventional TV-receiver.

7.1. Principle of Operation

The TV-interface consists of three parts:
Data storage, character generator, and video modulator (see **Figure 48**). The data are passed from the CPU via a bidirectional bus driver to the data storage, and the addresses are passed via an address multiplexer. At the same time, the write-order is activated. If the CPU does not write into the data storage, the addresses from the frequency divider will be present at the address inputs of the data storage. A frequency divider is driven by a clock of 6 MHz, and divides firstly by 384 to 15 625 Hz: the line frequency of the TV-receiver. Secondly, this frequency is divided by 320 to obtain the frame frequency of 48.8 Hz.

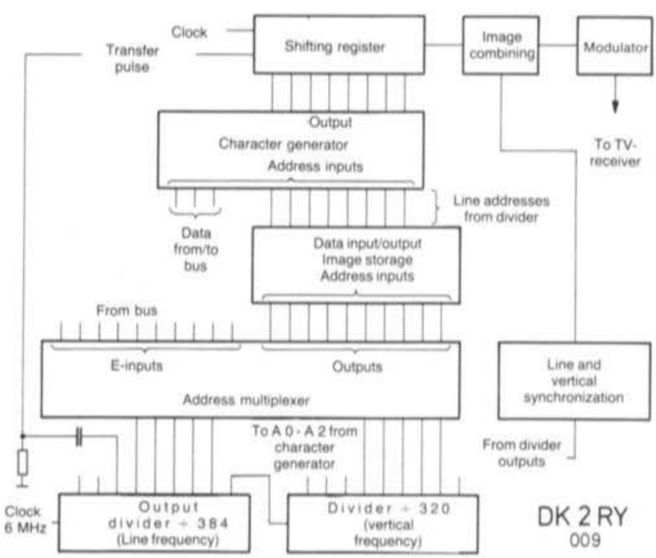


Fig. 48: The TV-interface comprising three parts: Data storage, character generator, video modulator

Fig. 49: Generation of a row of characters

As was mentioned previously, the addresses from the frequency divider are fed to the RAMs as long as the CPU is not writing into the data storage of the TV-interface. These then feed a seven-bit data word to the upper seven address inputs of the character generator. The lower three address inputs are fed with the first three address bits of the second frequency divider. These are used for displaying the 8 lines of the character. The addressed storage contents of the character generator are fed to an 8 bit, parallel-series converter (register) and fed out in series.

The characters are combined with the blanking pulses from a fixed value storage in an OR-gate, and finally provided with synchronizing pulses in a dual inverter. This generates a complete video signal with synchronizing pulses that can be fed into the video circuit of a TV-receiver for displaying the data. In this case, the readability of the characters is somewhat better then when modulating them onto a VHF or UHF carrier and passing them through the VHF or UHF portions of the TV-receiver. When converting the video signal to a VHF or UHF signal, one has the advantages of being able to use a conventional TV-receiver.

The method of generating the characters is now to be described in more detail:

The frequency divider is set to zero at the commencement of an image. This means that the address zero will be present at the address inputs of the RAMs. The data word stored there will be fed to the character generator, which passes on its stored contents to the register. This is read out during eight system pulses.

After the eight system pulses, the fourth flipflop of the frequency divider will be actuated so that address 1 will be connected to the RAMs. This process is repeated up to 256 clock pulses, in other words until 32 characters have been outputted. Due to the division by 384, blanking finally takes place and the line synchronizing pulse is sent. This represents one line of the image (not one row of characters !). After 384 pulses have passed through the divider, the first flipflop of the second frequency divider will be actuated. This means that H-level is present at the address bit of the character generator.

The above process is repeated, and the second line of the image is completed. The whole process is repeated until the eight image lines of a character sequence are outputted. The next flipflop is then actuated and the second row of characters is commenced, and so on until a complete TV-image is displayed. **Figure 49** shows the way the image is generated.

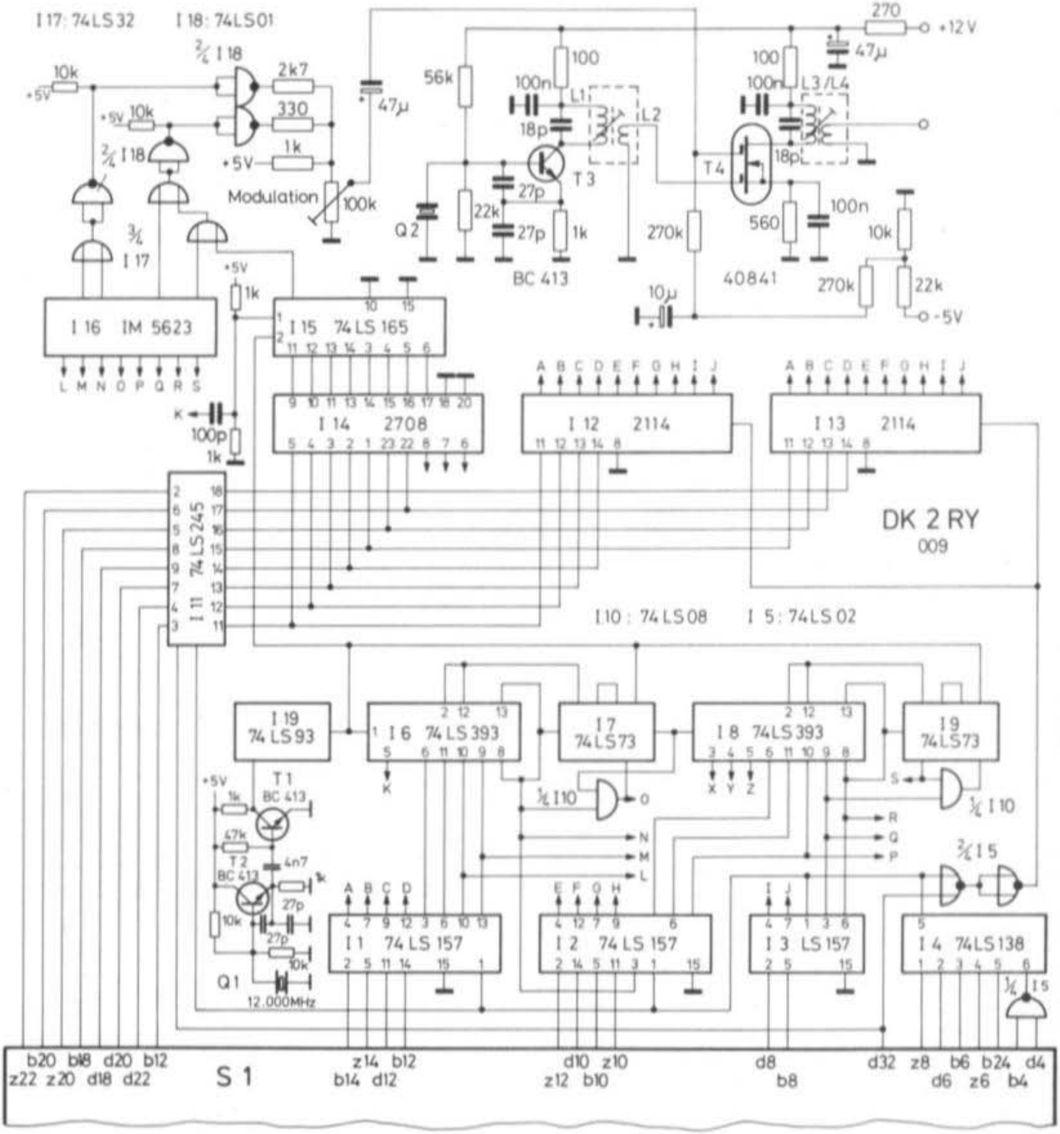


Fig. 50: The TV-interface DK 2 RY 009

7.2. Circuit Description

The complete circuit diagram is shown in **Fig. 50.** The integrated circuits type SN 74 LS 157 are used as address multiplexers. Their B-inputs are connected to the frequency divider, and their A-inputs to the address bus. An SN 74 LS 138 is used as selector switch for the address multiplexer. If an address of the TV-interface is addressed by the CPU, the select-input will be multiplexer L, and the address is fed via this multiplexer into the storage. In the other case, it will be select-input H, and the addresses will be fed from the frequency divider into the storage.

The frequency divider consists of two parts:

The line frequency divider, and the vertical frequency divider. A simple 12 MHz Colpitts oscillator is used as clock. This is fed via an amplifier stage to a frequency divider that divides the 12 MHz signal by two to 6 MHz. This is the system clock of the TV-interface. At the same time, it is converted to TTL-level. The system clock pulse is then fed to the register and to the input of the counter.

An 8-bit divider type SN 74 LS 393 is used as frequency divider together with a part of an SN 74 LS 73. Such a divider would normally divide by 512. However, since a division ratio of 384 is required, it is necessary for the divider to be reset at a counter state of 384, which is achieved using a synchronous reset. This is done by combining the divider output of the SN 74 LS 73 and the D-output of the SN 74 LS 393 via an AND-gate. If 384 is present at the output of the divider, the inputs of the AND-gate will be at H-level, and thus also the output. This is, in turn, connected to the K-input of the master-slave flipflop. Since the J-input also has H-level, Q will be also at H-level and $\bar{Q}$ at L-level. These are connected to the reset input of the frequency divider. Since the clock input of the master-slave flipflop is also connected to the system clock, this will wait until a clock pulse is completed before it accepts the given states. The counting chain is now reset, which means that the frequency divider will divide by 384.

This complicated method was selected in order to obtain a defined reset time. When directly combining the gate with the reset inputs, it would be possible for frequency dividers not to be reset correctly. The output signal of this divider is fed to the input of the vertical frequency divider, which operates according to the same principle.

As was previously mentioned, the storage is addressed. Two type 2114 circuits are used as storage. The data input/outputs are connected to a bi-directional bus driver type SN 74 LS 245. Its connection CE is connected to the SN 74 LS 138, and the DIR-input to the WR. The CPU can thus write in and read out from the storage. Furthermore, the data input/outputs D 0 - D 6 are connected to the 7 higher-valancy address inputs of the character generator type 2708. These data outputs are connected to an 8-bit register, which converts the parallel characters into series. The system clock pulse is present at the clock input, whereas the set-input is connected via an RC-link with bit 2 of the line counter. An impulse is fed to the set-input of the register at the transition from H to L-level. It will then take over the data from the time generator and will write them at the speed of the system clock. The output signal from the register is then fed to an OR-network where it is provided with the vertical and horizontal synchronizing pulses.

A PROM type IM 5623 generates the vertical and line synchronizing pulses. Whereas four addresses, each, are connected to the higher-valency outputs of the line and vertical frequency dividers, output 0 1 and 02 will generate the line or vertical blanking. Output 0 4 and 0 3 provide the synchronizing pulses, which are combined in an OR-gate and finally added to the video signal.

The combined signal is fed to gate 2 of the dual-gate MOSFET T 4 where it is modulated onto the signal from the crystal-controlled oscillator. Gate 2 is set to a fixed bias voltage of − 1.5 V, in order to achieve the most linear modulation possible. The modulation depth can be set with the aid of the 100 kΩ potentiometer. The output signal is fed from Pt 1/Pt 2 via a coaxial cable to the antenna input of the TV-receiver. The output voltage is approximately 30 mV.

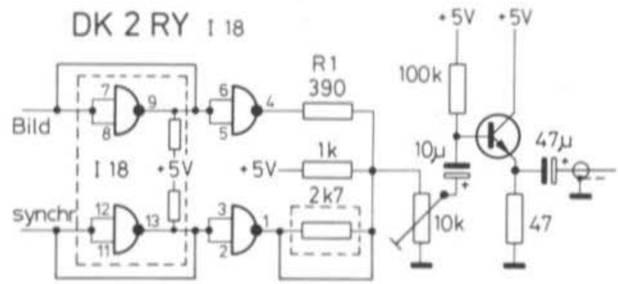


Fig. 51:
Simple drive at video level; modify R 1 from 390 Ω to 150 Ω, delete dashed parts, and replace by bridges

It is also possible for the output signal of the adding network to be fed to the video input; however, it is necessary for the phase to be shifted by 180°. This can be achieved using a simple transistor amplifier. Directly feeding into the video circuits of the TV-receiver has the advantage of a considerably better quality. Oscillator and modulator can then be deleted. **Figure 51** shows a circuit for directly feeding the output signal into a video circuit.

The following modifications will be required:

The interconnection between the OR-gate (I 17) and the inverter (I 18) should be disconnected as shown in Figure 51, as well as between this inverter and the second inverter. Instead of this, two bridges are made between I 17 and I 18. A wire bridge is now connected instead of the 2.7 kΩ resistor. The output signal can now be matched to 50 Ω using an emitter follower. One can use the oscillator transistor T 3 for this. The modulator (T 4) can be deleted. The output signal from the emitter follower is fed via a short coaxial cable to the video input of the TV-receiver.

7.3. Components

I 1 - I 3:	SN 74 LS 157 (TI)
I 4:	SN 74 LS 138 (TI)
I 5:	SN 74 LS 02 (TI)
I 6, I 8:	SN 74 LS 393 (TI)
I 7, I 9:	SN 74 LS 73 (TI)
I 10:	SN 74 LS 08 (TI)
I 11:	SN 74 LS 245 (TI)
I 12, I 13:	2114 (INTEL, NEC)
I 14:	2708
I 15:	SN 74 LS 165 (TI)
I 16:	IM 5623
I 17:	SN 74 LS 32 (TI)
I 18:	SN 74 LS 01 (TI)
I 19:	SN 74 LS 93 (TI)
T 1 - T 3:	BC 108 C or similar NPN transistor (TI)
T 4:	BF 351, 40481 or similar dual-gate MOSFET
Q 1:	Crystal 12.000 MHz HC-25/U
Q 2:	Crystal according to TV-channel (45-70 MHz)
S 1:	Connector strip C 74334-A80-A60 (Siemens)
L 1, L 3:	7 turns of 0.5 mm dia. enamelled copper wire in special coil set
L 2, L 4:	1.5 turns on L 1, or L 3 respectively.

7.4. Construction and Alignment

The TV-interface is constructed on a PC-board of 101.6 mm x 160 mm, and has been designated DK 2 RY 009. After mounting the components **(Figure 52)**, the coil cores are aligned for maximum output voltage. In order to achieve this, it is necessary for the character generator (2708) to be removed from its socket.

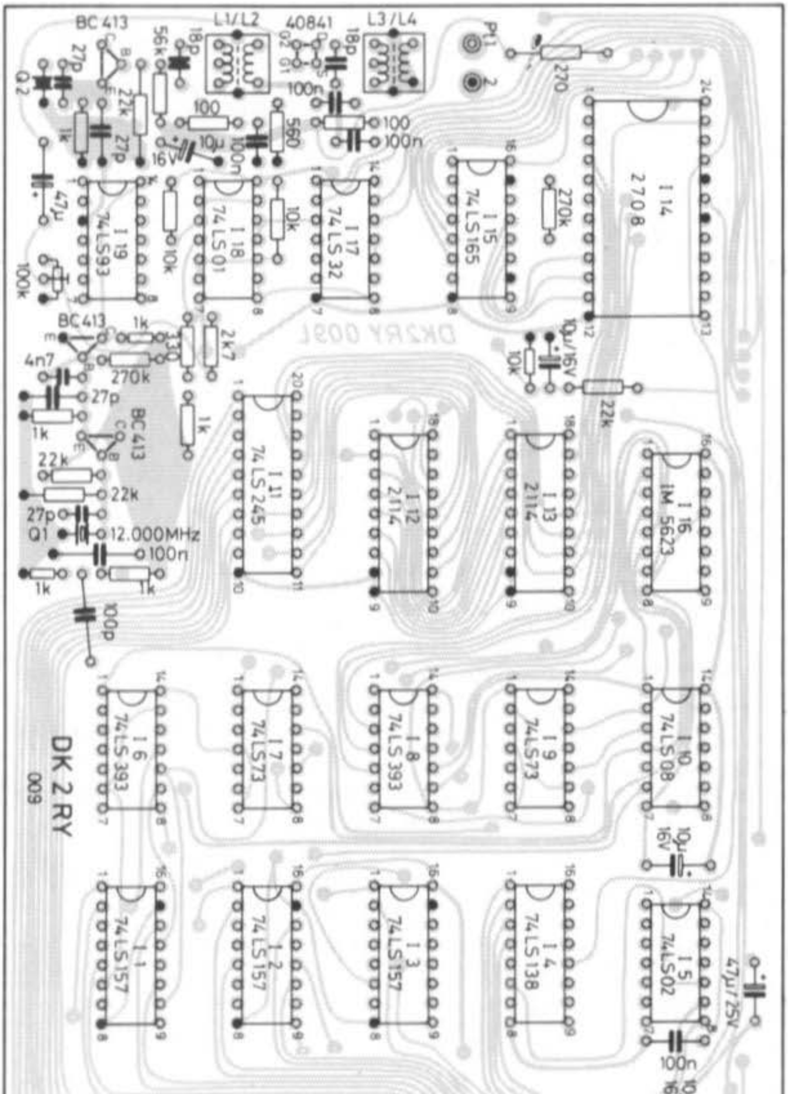


Fig. 52: PC-board DK 2 RY 009 of the TV-interface

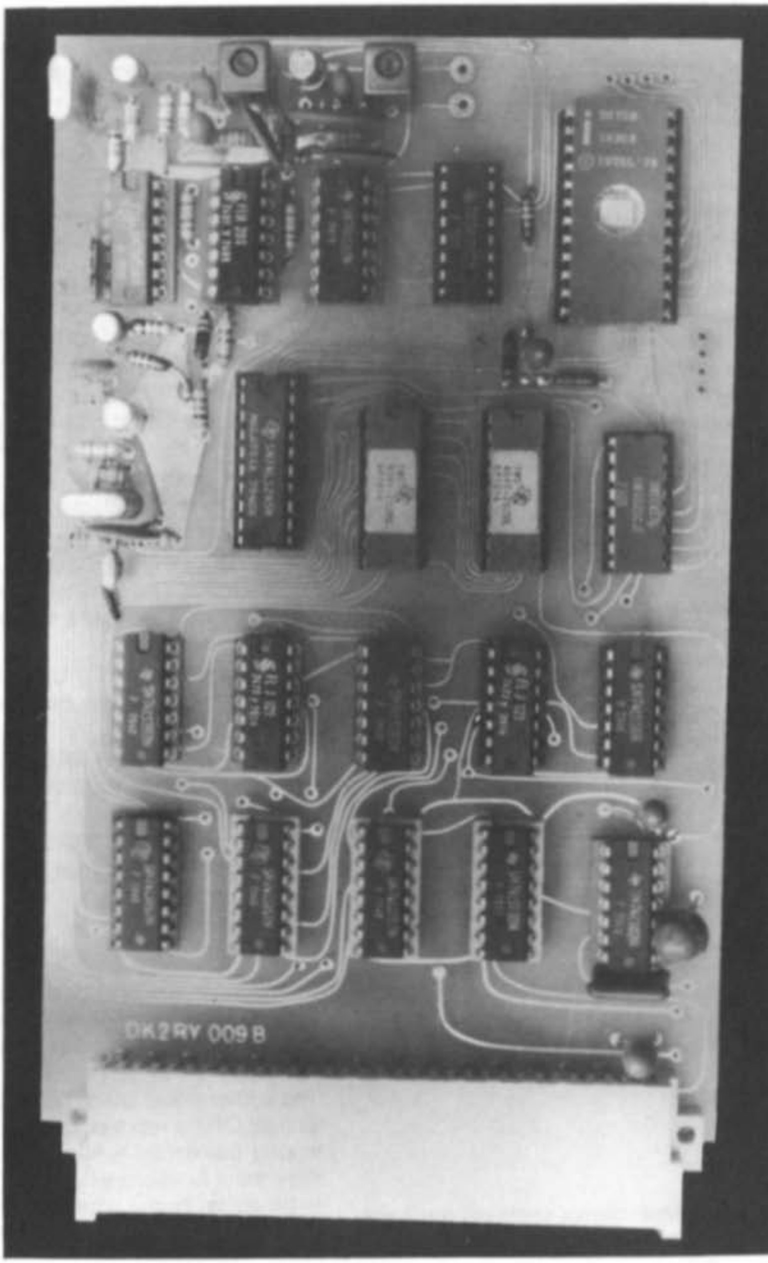


Fig. 53: A complete TV-interface

After this, the character generator is replaced and the modulation potentiometer slowly increased until a random pattern of characters is visible on the TV-screen. The modulation level is now increased until a good contrast is to be seen between black and white.

It is more favorable for the alignment to be made with the aid of an oscilloscope. In this case, it is necessary for the 2708 to be removed and the inputs of the register should be grounded. It will then be possible for the image contents (25 % modulation), the front and back black shoulders (75 % modulation), and the synchronizing pulses (100 % modulation) to be recognized and aligned to the required values.

After completing the alignment, the board is placed in one of the connector strips on the bus board DK 2 RY 002.

The TV-interface is designed to produce black characters on a white background. This is far more favorable when using a VHF or UHF modulation than when displaying white characters on a black background. In the latter case, the TV-receiver will reduce the gain if only a few characters are present and a black bar will appear in the line, which means that the character itself will not be clearly displayed.

This is immaterial when displaying at video level. If one requires white characters on a black background, this can be achieved by disconnecting the connection between pin 7 of the register and pin 9 of the SN 74 LS 08, and providing instead a wire bridge between pin 9 of the register and pin 9 of the SN 74 LS 08. The characters will then be white on a black background.

7.5. Drive

The TV-interface cannot carry out the functions of a monitor, but requires the aid of the CPU for this. It is designed so that the CPU can access it in the same manner as a storage (memory mapped). This means that the CPU can read and modify any storage position (character) at any time. This means that it is advisable to allow the program to take over the function of the »controller«. The software for the driving system will be available later from the publishers. This includes the complete cursor-drive. The »cursor« is a small black rectangle on the screen which marks the position of the next character that can be inputted.

The cursor-drive consists of the following parts:

a) Cursor home position which is the position of the first character on the screen.

b) Carriage return (CR) corresponding to the carriage return of a typewriter, but without line feel.

c) Line feed (LF), corresponds to the line feed, without carriage return.

d) Back space (BS): Back-space of the cursor by one character without cancelling the last character.

e) Horizontal TAB (HT): Forward shift of the cursor by one position, without cancelling the character at the present position of the cursor.

f) Vertical TAB (VT): Upward shift of the cursor by one line (required for graphic displays).

g) Delete or Rubout (DEL): This allows the last character inputted to be cancelled and corrected.

h) End of Text or CTRL C:
Stop of the running program and return to the command state.

i) Form Feed or Page: Cancelling the full screen.

Furthermore, the operating system also contains a so-called page-full software that ensures that the screen address range is not exceeded.

The author would like to point out that an additional CPU is required together with the previously mentioned software in order to fulfill the control functions when the TV-interface is to be driven from a different computer.

It is, however, then possible for the TV-interface to be used for other applications, such as an RTTY-converter that drives the TV-interface so that the text is displayed on the TV-screen. The author can assist readers with the production of such a program, as long as a Z 80-CPU is used.

8. CONNECTION AND TESTING THE COMPUTER

The microcomputer can be interconnected and operated after all described modules have been built up, aligned, and tested.

Firstly, the EPROMs programmed by the publishers are placed into their sockets. A small numeral is present on the upper side, which is identical to the number of the socket in which the EPROM should be inserted. If an EPROM is placed into a different socket, the program and computer will not function.

The card is now placed into one of the connection strips on the bus board, after which the computer is ready for operation. A random mixture of characters will appear on the screen after switching on the operating voltage. After depressing the reset-key, this will disappear and a large white rectangle with black edge will appear on the screen. The first question of the computer: TIME will now appear in the first line. This means that the computer requires to know the time. The local time is then inserted using the keyboard, for example: 09/45 (CR), thus 09.45. The character (CR) indicates depression of the »Carriage Return« key. It is absolutely necessary to provide the slant stroke and no spaces should be present between the numerals. If an error is made, e.g. 09/61 (CR), a nonexistent time of 09.61, the computer will answer: TIME SYNTAX ERROR, and request the time again in the next line

T I M E

It is now possible for the time to be inputted again.

If the time has been given in the correct manner, the computer will verify the time as:

T I M E 09 45 00

This is followed by:

C M D > in the next line

which means that the computer is in its command state and is waiting for the next order.

8.1. Programs

Since the programs that have been developed and are under preparation, are all in the assembler program language, they are too extensive (over 2000 orders) to be described in this magazine. They would also be boring to a large number of our readers.

As was previously mentioned, an operating system was developed that does not only carry out the cursor control, but also has another task. This includes, for instance, the extensive drive of the number cruncher, the display of mathematical calculation results on the screen, the calculation of the time, the data, and order input, as well as the drive of the rotator interface. For amateur radio applications, a program has been developed for calculation of the distance and direction using the QTH-locator, as well as the pointing of the antenna to the calculated direction. Furthermore, the author is preparing a program for calculating the location and direction (azimuth and elevation) of satellites. Further programs are planned for logging (also contest logging), and for RTTY and CW-decoding.

However, it is necessary for an interface to be developed for all planned programs in order to achieve these applications. The development of these depends, of course, on the interest in the computer itself.

If sufficient interest is available, the next part of this description is to describe the orders, as well as information regarding the commencement addresses of various sub-programs, which are required for developing one's own programs. When ordering the programs, the following must be given, since they are stored in the program:

1) Call sign
2) One's own QTH-locator

The program consists of the following:

3 k operating system (3 pcs. 2708)
1 k amateur radio programs (1 pc. 2708)

9. EDITORIAL NOTES

Since the interest in obtaining PC-boards for this computer is very low, it has been decided not to produce any boards for this computer system. However, any readers requiring programmed EPROMs should contact the publishers who will arrange this on their behalf.

A Compact 144 MHz/28 MHz Transverter with Low-Noise Preamplifier, Schottky Ring Mixer, and Clean Transmit Signal

by R. Albert, DK 8 DD

144 MHz transverters have been used for some time now for converting the output frequency of 10 m transceivers to the 144 MHz band. Although this cannot be seen from the designation, such transverters are linear transverters suitable for all operting modes. This article is to describe an extremely compact, reproducible transverter. On the receive side, it is equipped with an extremely low-noise dual-gate MOSFET (BF 981), which is followed by a Schottky ring mixer. On the transmit side, an output of approximately 100 mW is available with a very good harmonic and spurious rejection. This power is sufficient for local communication, but can, of course, be amplified to any required level using linear amplifier stages. The dimensions of the transverter amount to 140 mm x 63 mm x 31 mm (including BNC-connectors). As can be seen in the photograph given in Figure 1, the transmit and receive paths are accessible separately and are provided with separate connectors.

1. CIRCUIT DESCRIPTION

As can be seen in **Figure 2**, the heart of this transverter is a common crystal-controlled oscillator that provides the local oscillator frequency for the transmit and receive conversion. This local oscillator does not possess any frequency-multiplying stages, but oscillates directly at the required frequency of 116 MHz. A low-noise junction FET (T 1) is used, and the 116 MHz resonant circuit comprises L 4 and a capacitive voltage divider. This is followed by a subsequent double-gate MOSFET buffer (T 2) that amplifies the oscillator signal to a value of 7 to 10 dBm (≙ 5 to 10 mW) required for the ring mixer (IE-500 or similar type).

Four PIN-diodes (D 1 to D 4) are provided in the vicinity of the ring mixer which switch on the receive path when + 12 V is present at the »RX« operating voltage pin. The transmit path is switched in, when + 12 V is present at the »TX« pin.

In the receive mode, the input signal is passed via the »IN« connector to the low-noise dual-gate MOSFET T 4, where it is amplified and fed to the ring mixer (pin 1) via the bandpass filter comprising L 10 and L 9. Diodes D 1 and D 3 will conduct, and D 2 and D 4 will be blocked. The intermediate frequency is formed as the difference of the input signal and local oscillator signal, which is then passed from pins 3 and 4 of the ring mixer via D 3 and a series-circuit comprising L 2 to the IF-preamplifier (T 3). This low-noise FET operates in a common gate circuit; a subsequent Pi-filter comprising L 3 transforms its output impedance to 50 Ω, which allows the 10 m receiver to be fed using a 50 Ω coaxial cable that can be connected to connector »RX«.

In the transmit mode, the 28 MHz drive signal is fed at a power level of max. 100 mW to the »TX« connector. Trimmer potentiometer P 1 is used for adjusting this input power level to the input of the transmit mixer that only requires a maximum of 1 mW. The wideband IF-circuit comprising L 1 filters out unwanted signals outside of the passband.

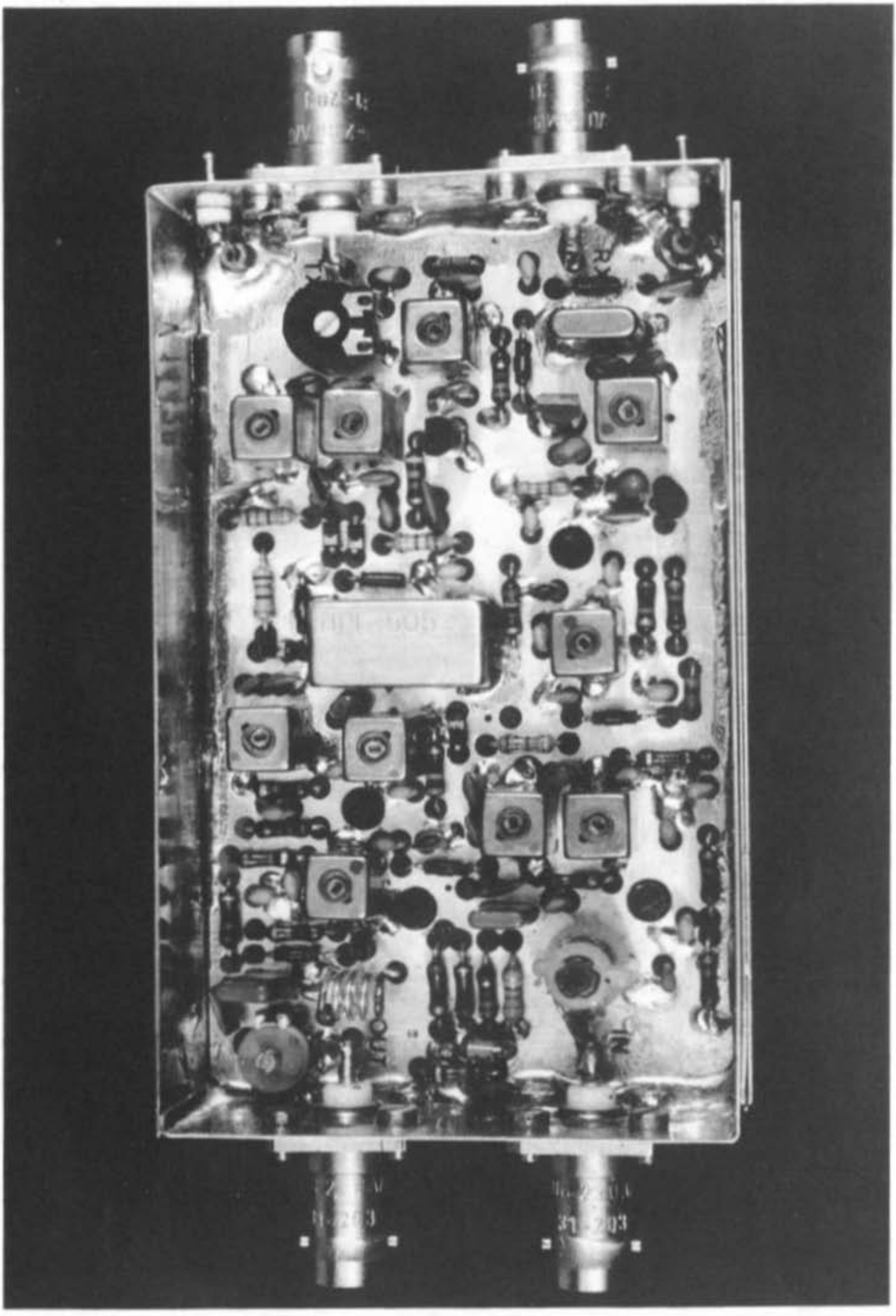


Fig. 1: Photograph of the complete transverter from 28 MHz to 144 MHz

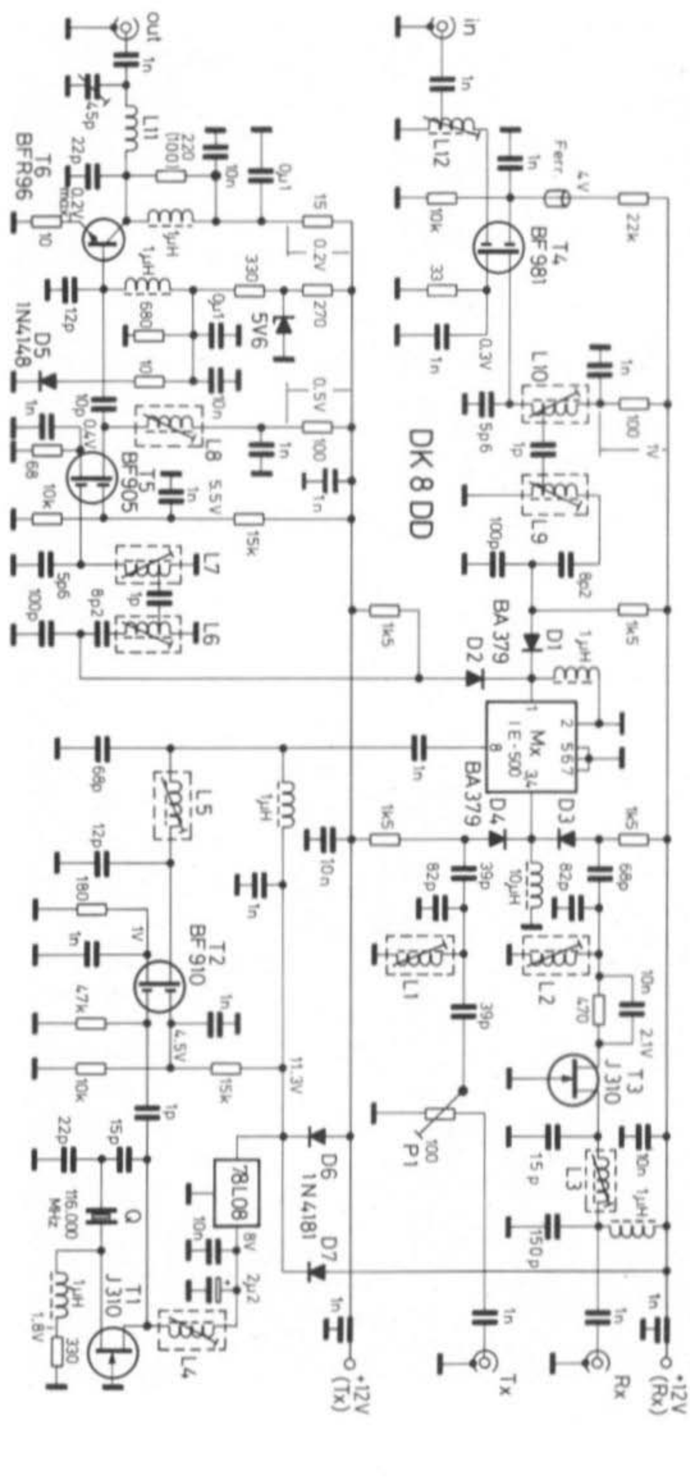


Fig. 2:
The ring mixer is switched using PIN diodes so that it can be used both in the transmit and the receive mode

Diodes D 4 and D 2 will now conduct and diodes D 3 and D 1 will be blocked. This means that the drive signal is converted up to 144 MHz with the aid of the local oscillator signal. This signal is then passed via the bandpass filter comprising L 6 and L 7 to the driver transistor T 5. The drain circuit comprising L 8 increases the suppression of oscillator and image frequency. A Pi-filter matching has been selected in the collector of the output transistor T 6 for improved harmonic suppression. After correct alignment, all spurious signals will be suppressed by more than 50 dB. Due to the high gain of transistor type BFR 96, special chokes and a separate operating-point stabilization is provided for the output stage.

1.1. Components

T 1, T 3:	J 310 (Siliconix)
T 2:	BF 910 (Texas Instruments)
T 4:	BF 981 (Philips)
T 5:	BF 905 (TI) or (new !) BF 907
T 6:	BFR 96 (Siemens, Philips)
Mixer:	IE-500, HPF-505, SRA-1 or similar
D 1 - D 4:	BA 379 (Siemens)
D 5 - D 7:	1 N 4148 or 1 N 4151 or similar

Voltage stabilizer: 78 LO 8 (National Semic.)

1 zener diode: C 5 V 6

L 1, L 2:	Neosid-BV 5049 (yl/wt)
L 3:	Neosid-BV 5056 (gn/bl)
L 4, L 5, L 8:	Neosid-BV 5061 (bl/bn)
L 6, L 7, L 9, L 10:	Neosid-BV 5063 (bl/orange)
L 11:	5 turns of 0.8 mm dia. silver-plated copper wire wound on a 4 mm former, self-supporting, pull out to fit the holes in the board
L 12:	6.5 turns of 0.8 mm dia. silver-plated copper wire wound on a 6 mm ribbed coil former, with VHF-core, coil tap approx. 1 turn from the cold end

5 min. ferrite chokes (spacing 10 mm): 1 μH

1 min. ferrite choke (spacing 10 mm): 10 μH

1 ferrite bead, 3 mm long

1 plastic foil trimmer, 7 mm dia.: 45 pF

All ceramic disk capacitors: spacing 2.5 mm (exception: 2 ceramic flat, tubular capacitors of 100 nF: 5 mm spacing)

2 ceramic feed-through capacitors for solder mounting, short type: approx. 1 nF (value uncritical)

1 trimmer potentiometer 100 Ω, spacing 10/5 mm (Piher)

All resistors:
Carbon resistors for 10 mm spacing

1 tinplate case 102 mm x 60 mm x 28 mm with two covers

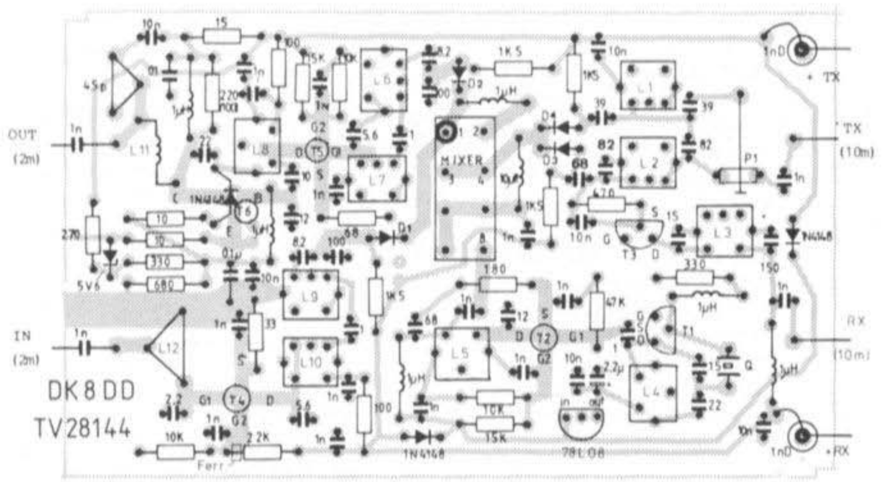
4 BNC flange connectors UG-290 A/U

1 crystal 116.000 MHz, 5th overtone, HC-18/U

2. CONSTRUCTION

The circuit given in Figure 2 can be accommodated on a 101 mm x 60 mm large, double-coated PC-board as shown in **Figure 3**. The component side is provided with a continuous copper surface, which is used as ground surface, and is only etched where the components pass through the board. 5 mm holes are required for the four stripline transistors (T 2, T 4, T 5, T 6), and two 2.6 mm holes for the two feed-through capacitors. All other holes on the board are 0.8 mm in diameter.

The two side pieces of the case are now depressed into one of the covers, after which the two side pieces are soldered together, but not to the cover. The inner conductors of the 4 BNC-connectors should be shortened to 2 mm in length. These connectors are now placed over the mounting holes on the outside of the case and screwed into place with the aid of 2.5 mm screws, which are inserted from the inside of the case. The PTFE feed-throughs for the operating voltages: + 12 V (RX) and + 12 V (TX) are pressed into place from the outside and shortened to a length of 2 mm on the inside. The PC-board can now be placed into the case from the lower side with the component side (ground surface) facing the connectors. It is now fitted into the case so that the position is determined by the



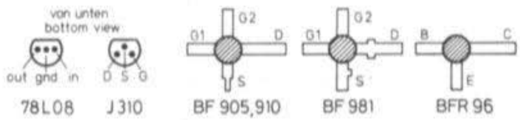


Fig. 3:
The dimensions of the double-coated PC-board are only 101 mm x 60 mm

mounting screws of the connectors. Approximately 3 mm remain free on the conductor side of the board. Attention should be paid that the holes for the two feedthrough capacitors are directly below the PTFE feedthroughs. The photograph shown in Figure 1 can provide valuable information during construction.

The edge of the ground side of the PC-board is now soldered on all sides to the case, after which it is turned around so that the conductor side of the of the board can be soldered twice to the case in the vicinity of L 12. The feedthrough capacitors are placed through the board from the connector side and soldered to the ground surface.

On the conductor side, the connection wires of these two capacitors should be bent, shortened and soldered to the required conductor lanes. On the upper side of the board they are connected to the PTFE-feedthroughs.

It is now possible for the other components to be mounted into place on the boards, and this should be commenced with the ceramic capacitors. The capacitors that are grounded on one side can be connected to soldering points on the inductance cans, or to resistors, which means that they need not be soldered to the ground surface itself. In the case of all others, the connection leads should be kept as short as possible so that no unwanted line inductances result. The inner conductor pins of the »IN« and »OUT« connectors should be connected using one 1 nF capacitor, each, to the required points on the board. The connectors designated »RX« and »TX« respectively are, on the other hand, connected with the aid of short wire bridges.

Solder the plastic foil trimmer and inductance L 12 into place. There is only one inductance to be wound (L 11); this should be made according to section 1.1. and soldered into place. Attention should be paid to the direction of the winding, which is given on the board.

Mount all resistors, chokes, and diodes with the exception of D 5.

Mount all canned inductances, and solder the screening cans additionally to the ground surface. This should be done quickly by using a hot soldering iron (approx. 3 mm wide).

Solder the ring mixer into place, and also solder the case to the ground surface quickly. It is sufficient to solder the two short sides to the board.

This is followed by soldering the crystal and the trimmer potentiometer into place.

After this, mount the transistors and the voltage stabilizer. The soldering iron and PC-board should at least be grounded during this phase, since transistors type BF 905, BF 910, and BF 981 are MOSFETs, and it is thus possible to destroy them by higher voltages inspite of the gate protective diodes.

Finally, diode D 5 is soldered into place on the conductor side of the board and pressed into position directly adjacent to the case of T 6.

3. ALIGNMENT OF THE TRANSVERTER

In order to carry out an exact alignment, the following will be required: A multiplier with RF-probe, or even better an RF-millivoltmeter, or oscilloscope with RF-probe, furthermore an RF-power meter with a 100 mW range, and a signal generator for aligning the receive converter. It is necessary for the lower cover to be fitted during alignment, and in operation.

3.1. Alignment of the Local Oscillator

Connect a 12 V source: + to one of the two feedthroughs, and – to the case. Connect the RF-probe to G 1 of transistor T 2 and align the core of inductance L 4 for maximum reading. Switch the operating voltage temporarily off and on and slightly shift the core of L 4 until the oscillator commences oscillation reliably on switching on.

Place the RF-probe to pin 8 of the ring mixer and align the core of inductance L 5 for maximum reading. If suitable measuring equipment is available, it is possible for the frequency and power to be checked at this position (unsolder the 1 nF). Any deviation of the frequency from 116.000 MHz can be corrected by aligning inductance L 4, however, reliable commencement of oscillation, and sufficient output (at least 5 mW) are more important than a deviation of a few hundred Hz.

3.2. Alignment of the Receive Converter

A 28 MHz receiver is now connected to the »RX« connector, and a 144 MHz antenna or signal generator connected to the »IN« connector using 50 Ω coaxial cable. The preamplifier transistor T 4 can oscillate when the input connector is open, and this should be avoided by providing either an antenna, a 50 Ω terminating resistor, or other 50 Ω load to the »IN« connector at all times. The + U_B (RX) feedthrough can now be provided with + 12 V.

Firstly, inductances L 3 and L 2 are aligned for maximum noise in the connected receiver. This is followed by also aligning inductances L 9, L 10, and L 12 for maximum noise. If a clear maximum results for all inductances, this will mean that the receive converter is operating correctly, and it will then be possible for the fine alignment to be made:

Adjust the receiver to 28 MHz. Align L 3 and L 10 for maximum noise, or maximum reading on the S-meter.

Set the receiver to 30 MHz. Align L 2 and L 9 for maximum. Finally, inductance L 12 should be aligned for this signal-to-noise ratio either at the center of the band (145 MHz/29 MHz) or in the frequency range of interest (for example 144.3 MHz/28.3 MHz). A very weak beacon signal is most suitable for this, of course, where no noise-figure measuring equipment is available.

3.3. Alignment of the Transmit Mixer

Attention should be paid that the drive power from the 10 m transmitter cannot be more than 100 mW in any operating mode ! If only 1 mW is available, the trimmer potentiometer can be rotated to its fully clockwise position. If the power is between 1 and 100 mW, the potentiometer should be adjusted to obtain the required value at the ring mixer.

A milliwattmeter is now connected to the »OUT« connector and a voltage of + 12 V connected to the feedthrough + U_B (TX). Firstly, connect the RF-probe to pins 3 and 4 of the ring mixer and align inductance L 1 for maximum reading.

This is followed by connecting the probe to G 1 of transistor T 5 and aligning L 6 and L 7 to resonance. Usually, the power-meter will indicate a small output power at the output, which means that the resonant circuit comprising L 8, and the output trimmer of the Pi-network can be aligned for maximum output. If this is not the case, the RF-probe is connected to the collector of T 6 and L 8 aligned for maximum reading. Now the 45 pF-trimmer is aligned for maximum output power.

This is followed by the fine alignment:

Tune the exciter to 28 MHz and align L 1 and L 7 for maximum output power.

Set the exciter to 30 MHz and align L 6 and L 8 for maximum output. Set the exciter to 29 MHz and align the 45 pF trimmer carefully for maximum output power.

A harmonic rejection of at least 45 dB will be achieved after a careful alignment (especially of the 45 pF trimmer). The given values in excess of 55 dB can usually only be achieved in conjunction with a spectrum analyzer and other measuring equipment.

If the transverter is only to be used in a narrow portion of the 144 MHz band (e.g. only SSB-band), it is possible for the alignment to be made here. This means that the power gain of the transmit converter will then be more than 20 dB. The lowest possible drive level to achieve the required output power should be used so that the mixer is able to provide the cleanest possible signal.

Since transistor T 6 (BFR 96) has a very high gain at 144 MHz, it is very important for a real termination of 50 Ω to be present at all times at the TX-output socket (OUT). When connected to a subsequent power amplifier, the fluctuating input impedance of this amplifier as a function of drive could cause T 6 to break into oscillation. In this case, the 220 Ω damping resistor in parallel to the collector choke of T 6 should be exchanged for 100 Ω. However, the output power will be reduced to approximately 50 mW due to this.

4. MEASURED VALUES

The following values were measured by the publishers on a prototype:

Oscillator frequency
stable between 8 and 13.5 V DC

Receive Converter

Noise figure: 1.6 dB. This could be improved to 1.2 dB after changing the coil tap on L 12 to 1 turn from the cold end (1 nF capacitor directly soldered from the connector to the coil).

Transmit Converter

Maximum output power: 250 mW
(U_B = 12 V; P_{in} = 1 mW)
Maximum output power with good linearity:
150 mW (P_{in} = 0.5 mW)
Harmonic rejection
(measured up to 500 MHz): min. 55 dB

5. NOTES REGARDING OPERATION

Many shortwave transceivers have built-in connectors for a transverter. If such connectors are not available, a socket should be provided for extracting a small portion of the drive power. In addition, a two-pole change-over switch is required that switches off the RF-output stage during transverter operation, and switches the drive to the transverter socket.

The required drive level is then passed from this socket to connector »TX« of the transverter. The »RX« connector is then connected with the antenna connector of the transceiver. In this manner, it is not necessary for an extra relay to be provided for transmit-receive switching at 28 MHz.

The operating voltage for the transverter must be switched between transmit and receive mode (possibly using a relay). This is carried out by feeding + 12 V to the PTFE feed-through RX in the receive mode, and to TX in the transmit mode.

The two 144 MHz connectors »IN« and »OUT« should be connected to the antenna using a coaxial relay when used without power amplifier, or connected to the required power amplifier and possibly masthead pre-amplifier. The sensitivity of the receive converter is extremely good with its noise figure of 1.2 dB, which means that an additional pre-amplifier will only bring an improvement when mounted directly at the antenna, where it is able to compensate for cable losses.

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 3/1981 of VHF COMMUNICATIONS

DJ 6 PI 011	**Prescaler with Preamplifier for upto 1300 MHz**		**Ed. 3/1981**
Semiconductors	Not available from us, contact nearest Plessey and Philips agency		
PC-board	DJ 6 PI 011	Double-coated, thru-contacts	DM 28.—
Minikit	DJ 6 PI 011	9 cer.caps., 1 tantalum electrolytic, 1 feedthrough cap., 1 ferrite bead, 2 trimmer pots., 4 resistors, 1 metal case, 2 BNC connectors	DM 19.—
Kit	**DJ 6 PI 011**	**with above parts (without Semiconductors)**	**DM 45.—**
DL 2 DE 001	**Close-In DF-Receiver**		**Ed. 3/1981**
PC-board	DL 2 DE 001	Double-coated, drilled	DM 16.—
Semiconductors	DL 2 DE 001	3 transistors, 2 op.amps., 3 diodes	DM 16.50
Minikit	DL 2 DE 001	1 m silver-pl.wire, 3 pl.foil trimmers, 11 cer.caps., 2 tantalum el., 1 trimmer pot., 14 resistors	DM 15.50
Kit	**DL 2 DE 001**	**complete with above parts**	**DM 46.—**
DC 3 NT 014	**Satellite Reception System / Monitor Deflection Module**		**Ed. 3/1981**
PC-board	DC 3 NT 014	Single coated, drilled, with plan	DM 25.—
Semiconductors	DC 3 NT 014	5 trans., 2 voltage stab., 3 op.amps., 1 bridge rectifier, 1 zener diode	DM 36.—
Minikit	DC 3 NT 014	9 alu.el., 4 pl.foil caps., 5 trimmer pot. (10 turn), 13 resistors, 12 solder pins	DM 39.—
Kit	**DC 3 NT 014**	**complete with above parts**	**DM 98.—**
DC 3 NT 015/016	**Satellite Reception System / Monitor EHT-Module**		**Ed. 3/1981**
PC-board	DC 3 NT 015	Single-coated, drilled, with plan	DM 25.—
PC-board	DC 3 NT 016	Single-coated, undrilled, with plan	DM 8.—
Semiconductors	DC 3 NT 015/16	4 transistors, 1 voltage stab., 1 bridge rectifier, 5 diodes, 4 zener diodes, 1 EHT-cascade	DM 124.—
Line-output trans.	DC 3 NT 015	with additional windings	DM 69.—
TV-tube	DC 3 NT 015/16	A 44-120 W with socket and EHT-connector (without freight)	DM 140.—
Deflection unit	for above TV-tube		DM 59.—
Minikit	DC 3 NT 015/016	5 alu.el., 5 pl.foil caps., 2 large trimmer pots., 2 power resistors, 16 resistors for 12.5 mm spacing, 9 resistors for 10 mm spacing, 15 solder pins	DM 27.—
Kit	**DC 3 NT 015/016**	**with all above parts (excl. freight on TV-tube)**	**DM 438.—**
DC 3 NT	**FAX**	**Set of Drawings for FAX-machine**	**DM 8.—**
Complete Package-Price for all kits DC 3 NT 003-016			**DM 1650.—**
All kits for operating only with TV-tube: DC 3 NT 003-009 and 011-016			**DM 1550.—**
All kits for operating only with FAX-machine: DC 3 NT 003-008 and 010-013			**DM 1050.—**

DJ 1 JZ 002	**Ring Mixer / Linear Amplifier Module for ATV-transmitters (Replaces DJ 4 LB 004)**		**Ed. 3/1981**
PC-board	DJ 1 JZ 002	Double-coated, drilled, with plan	DM 19.—
Semiconductors	DJ 1 JZ 002	1 ring mixer IE 500, 3 transistors	DM 68.—
Minikit	DJ 1 JZ 002	7 pl.foil trimmers, 19 cer.caps., 1 feedthrough cap., 4 tantalum el., 14 resistors, 1 metal case, 3 wire sizes	DM 38.50
Kit	**DJ 1 JZ 002**	**complete with above parts**	**DM 122.—**
DK 8 DD 001	**Compact 10 m / 2 m Transverter**		**Ed. 3/1981**
PC-board	DK 8 DD 001	Double-coated, drilled	DM 30.—
Components	DK 8 DD 001	6 transistors, 1 ring mixer, 3 diodes, 1 voltage stab., 11 already wound coils, 6 miniatur chokes, 1 pl.foil trimmer, 1 ferrite bead, 2 feedthrough caps., 47 cer.caps., 1 tantalum cap., 25 resistors, 1 trimmer pot., 1 crystal 116.000 MHz (HC-18/U), 1 metal case (drilled), 4 BNC-conn., 2 PTFE-feedthroughs	DM 169.—
Kit	**DK 8 DD 001**	**complete with above parts**	**DM 198.—**

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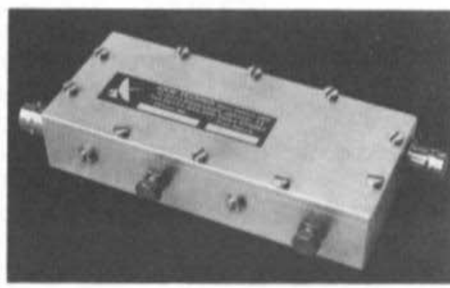
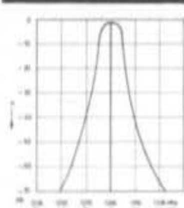
4-stage, sealed bandpass filters for 1152 MHz, 1255 MHz, 1288 MHz or 1297 MHz centre frequencies.

3 dB bandwidth:	12 MHz
Passband insertion loss:	1.5 dB
Attenuation at ± 24 MHz:	40 dB
Attenuation at ± 33 MHz:	60 dB
Return loss:	20 dB
Dimensions (mm):	140 x 70 x 26



Ideal for installation between first and second pre-amplifier or in front of the mixer for suppression of image noise, and interference from UHF-TV transmitters and out-of-band Radar Stations. Also very advisable at the output of a frequency multi-plier chain, or behind a transmit mixer.



Price: DM 168.—

Please list required centre frequency on ordering.

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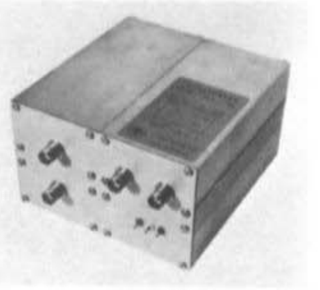
A New Generation of Transverters for 1296 MHz



We would like to introduce our second generation of 23 cm transverters for operation in conjunction with either 10 m or 2 m transceivers.

- Double-conversion both on transmit and receive with the 10 m version to obtain the extremely high image and spurious rejection and clean spectrum.
- Overall noise figure of the receive converter typically 3.9 dB
- Transverters are available in the following versions:

ST 1296/144 A:	2 m IF, output 1 W	**DM 655.—**
ST 1296/144 B:	2 m IF, output 3 W	**DM 825.—**
ST 1296/28:	10 m IF, output 1 W	**DM 798.—**

NEW COAXIAL SPECIALITIES

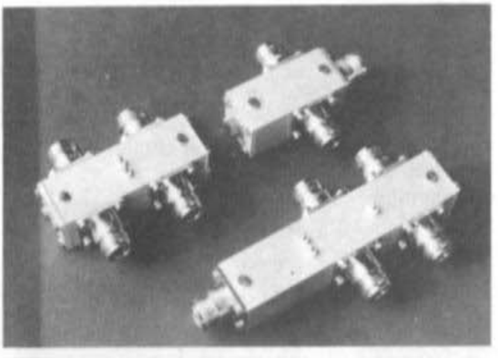


Fig. 1

Fig. 2

Fig. 3

A completely new programme of coaxial products offering some entirely new possibilities in the HF, VHF and UHF-range:

- **Inexpensive multi-port coaxial relays (Fig. 1)** with 2, 3, or 4 input ports (antennas) to 1 output port (feeder). 50 Ω N-Connectors. Low-loss with good crosstalk rejection (isolation).
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- **Multi-port coaxial switch** with 50 Ω cable connections suitable for installation within equipment. RG-58/U.
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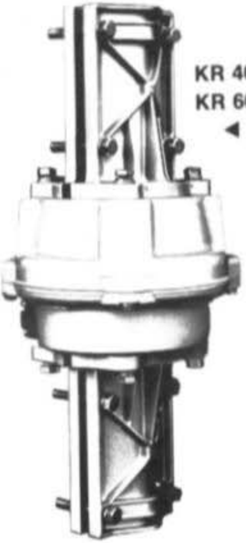
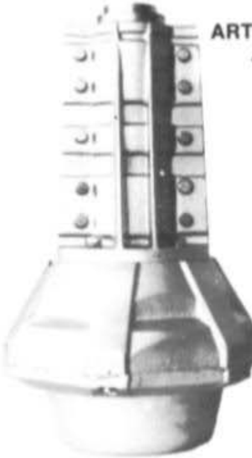
ANTENNA ROTATING SYSTEMS



KR 400
KR 600
◄

KR 2000
◄



ART 8000
◄

SPECIFICATIONS

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

*) 1 kpm ≙ 9.81 Nm

Controllers for above rotators

Our well-known rotators KR 400, KR 600 and KR 2000 are now available with large 360° compass indicators of 105 mm diameter. These models are designated by the suffix »RC«.

KR 400, KR 600, KR 2000

KR 400 RC, KR 600 RC, KR 2000 RC

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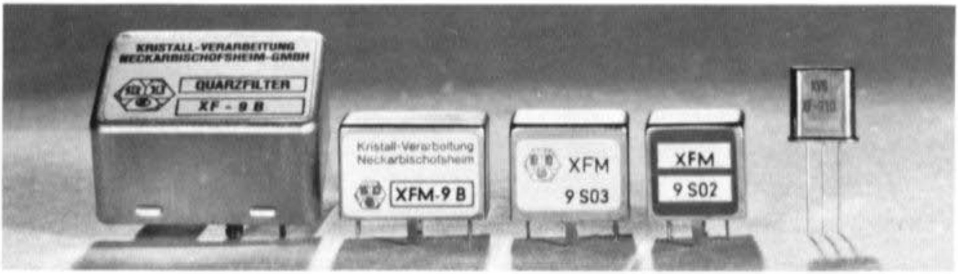

Case: 1 15 14 13 17

DISCRETE CRYSTAL FILTER	Application	MONOLITHIC EQUIVALENT with impedance transformation			MONOLITHIC EQUIVALENT without impedance transformation		
		Type	Termination	Case	Type	Termination	Case
XF-9A	SSB	XFM-9A	500 Ω ‖ 30 pF	15	XFM-9S02	1.8 kΩ ‖ 3 pF	13
XF-9B	SSB	XFM-9B	500 Ω ‖ 30 pF	15	XFM-9S03	1.8 kΩ ‖ 3 pF	14
XF-9C	AM	XFM-9C	500 Ω ‖ 30 pF	15	XFM-9S04	2.7 kΩ ‖ 2 pF	14
XF-9D	AM	XFM-9D	500 Ω ‖ 30 pF	15	XFM-9S01	3.3 kΩ ‖ 2 pF	14
XF-9E	FM	XFM-9E	1.2 kΩ ‖ 30 pF	15	XFM-9S05	8.2 kΩ ‖ 0 pF	14
XF-9B01	LSB	XFM-9B01	500 Ω ‖ 30 pF	15	XFM-9S06	1.8 kΩ ‖ 3 pF	14
XF-9B02	USB	XFM-9B02	500 Ω ‖ 30 pF	15	XFM-9S07	1.8 kΩ ‖ 3 pF	14
XF-9B10*	SSB	—	—	–	XFM-9S08	1.8 kΩ ‖ 3 pF	15

*** New: 10-Pole SSB-filter, shape factor 60 dB : 6 dB 1.5**

Dual (monolithic twopole) **XF-910;** Bandwidth 15 kHz, R_T = 6 kΩ, Case 17
Matched dual pair (four pole) **XF-920;** Bandwidth 15 kHz, R_T = 6 kΩ, Case 2 x 17

DISCRIMINATOR DUALS (see VHF COMMUNICATIONS 1/1979, page 45)
for NBFM **XF-909** Peak separation 28 kHz
for FSK/RTTY **XF-919** Peak separation 2 kHz

CW-Filters – still in discrete technology:

Type	6 dB Bandwidth	Crystals	Shape-Factor	Termination	Case
XF-9M	500 Hz	4	60 dB : 6 dB 4.4	500 Ω ‖ 30 pF	2
XF-9NB	500 Hz	8	60 dB : 6 dB 2.2	500 Ω ‖ 30 pF	1
XF-9P*	250 Hz	8	60 dB : 6 dB 2.2	500 Ω ‖ 30 pF	1

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