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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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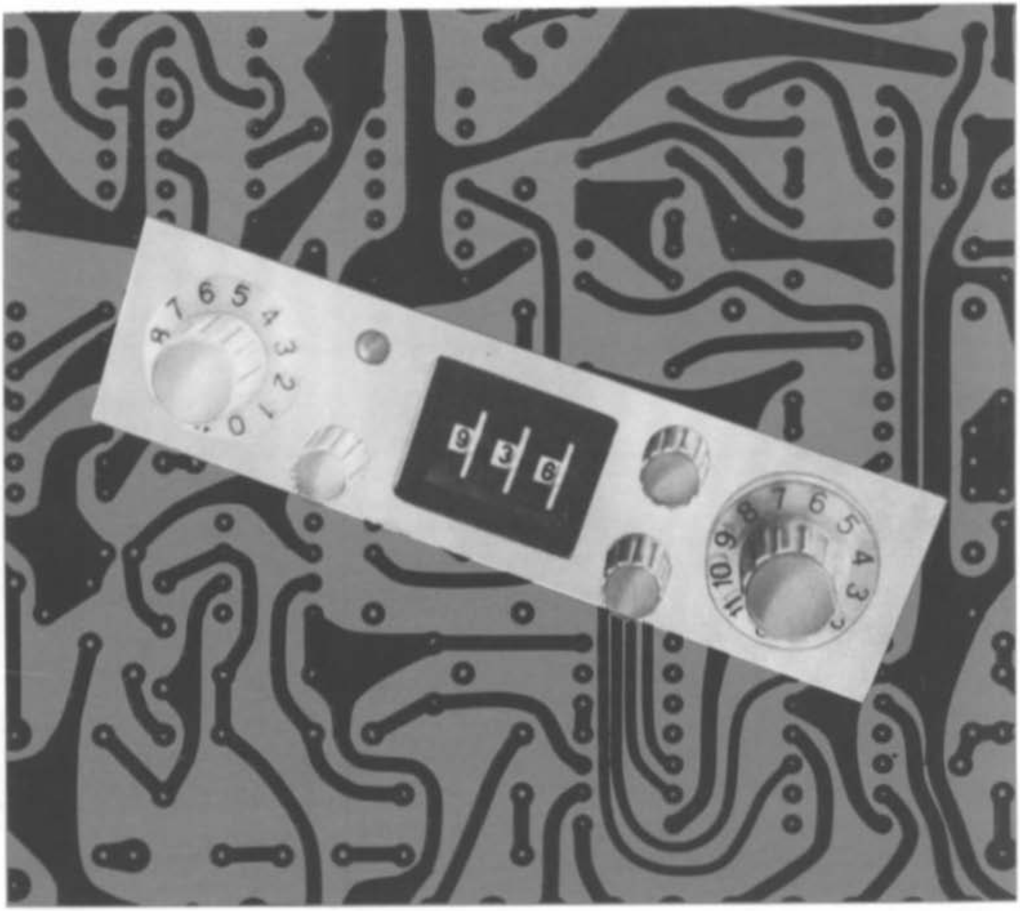

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As you know, VHF COMMUNICATIONS has always attempted to bring its readers really state-of-the-art designs. Sometimes we have branched off from the purely telecommunications equipment to bring modules of general interest such as frequency counters, standard frequency systems etc. This edition of VHF COMMUNICATIONS is to bring two such descriptions: Firstly, a VHF/FM stereo receiver with frequency synthesizer and containing two crystal filters. The other module is a frequency standard with an accuracy of 10^{-8} which is to be made available as a ready-to-operate module. Maybe you would like to mention these modules to non-VHF/UHF amateurs that would probably be interested in this technology.

We have now published our third French omnibus edition, which has been very popular in France. We will probably have to make a reprint of this third edition soon.

Only a few of the 1969 editions are still available. We decided not to reprint them anymore since they are somewhat obsolete now.

A STEREO VHF/FM RECEIVER WITH FREQUENCY SYNTHESIZER
PART I: CIRCUIT DESCRIPTION

by J. Kestler, DK 1 OF

The advances made in the semiconductor technology have had a considerable effect even in the consumer electronic's field. Pre-programmed selector buttons, squelch and automatic frequency-control circuits are to be found in most modern Hi-Fi receivers. Some of the most advanced receivers (such as the Revox-A 720) are even provided with a frequency synthesizer for digital frequency selection. However, the prices of such receivers are extremely high (more than DM 2000. --).

Since most VHF/UHF amateurs will also be interested in receiving VHF/FM broadcast transmissions, it was thought that there will be a considerable interest for a really advanced stereo VHF/FM broadcast receiver with digital frequency selection. Such a receiver should be suitable as both a car radio and for home use.

The described stereo tuner combines many advanced features from both the professional and consumer electronic technologies. It satisfies even the highest demands with respect to quality-of-reproduction, selectivity and ease of operation. Special attention has been paid to the ease of construction so that it can also be built up by less experienced constructors.

1. INTRODUCTION

The frequency range of 87.5 to 104 MHz is allocated to VHF/FM broadcasting in Europe. All transmitters are frequency-modulated with a maximum deviation of $\pm$75 kHz, and virtually all stations transmit at a multiple of 100 kHz, e.g. 87.6; 87.7; 87.8 MHz etc. In the case of stereo transmissions, it is necessary for two audio channels to be transmitted: The left-hand (L) and right-hand (R) channels. Each channel has a bandwidth of 15 Hz to 15 kHz. The sum signal (L + R) is directly fed to the modulator of the FM transmitter so that it is combatible with mono-receivers. The difference signal (L - R) amplitude-modulates a 38 kHz sub-carrier, and the resulting sidebands (23 to 53 kHz) are also fed to the FM modulator. Since the transmission of the full 38 kHz sub-carrier would require too much frequency deviation, a lower-level signal having half the subcarrier frequency is used: this is the 19 kHz pilot carrier. This allows the original 38 kHz subcarrier to be regenerated in the receiver, and is also used for the automatic mono-stereo switching. VHF/FM transmitters in Germany equipped for automatic switching of traffic news transmissions also radiate a 57 kHz subcarrier which is amplitude-modulated with frequencies in the order of 23 to 54 Hz according to the area. An additional tone of 125 Hz is also transmitted for the duration of the information. This ensures that the receiver is muted during normal, music transmissions and is only opened when actual traffic news is being transmitted. Figure 1 shows a diagram of the audio spectrum transmitted, which is in the order of 57 kHz. The required IF bandwidth of the receiver is thus:

$$\Delta f = 2 \times (f_{dev.\,max.} + f_{AF\,max.}) = 2 \times (75 + 57) = 264 \text{ kHz.}$$

Since this value is greater than the channel spacing, a certain overlapping of the frequency spectrum of neighbouring channels will take place which will be audible as unpleasant birdies if the AM suppression of the receiver is poor.

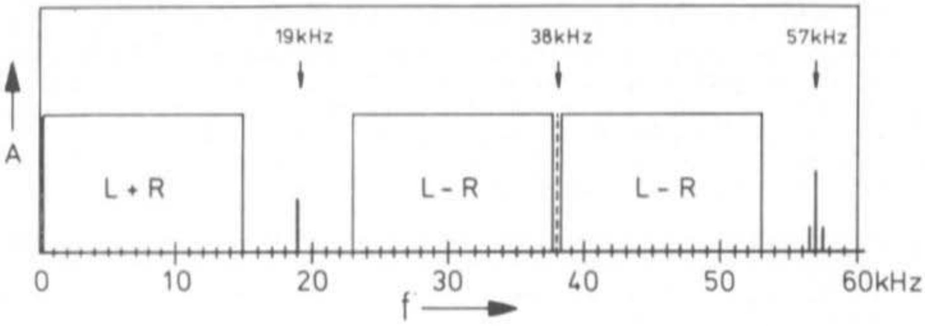

Fig.1: AF spectrum of a stereo transmission operating according to the pilote tone system

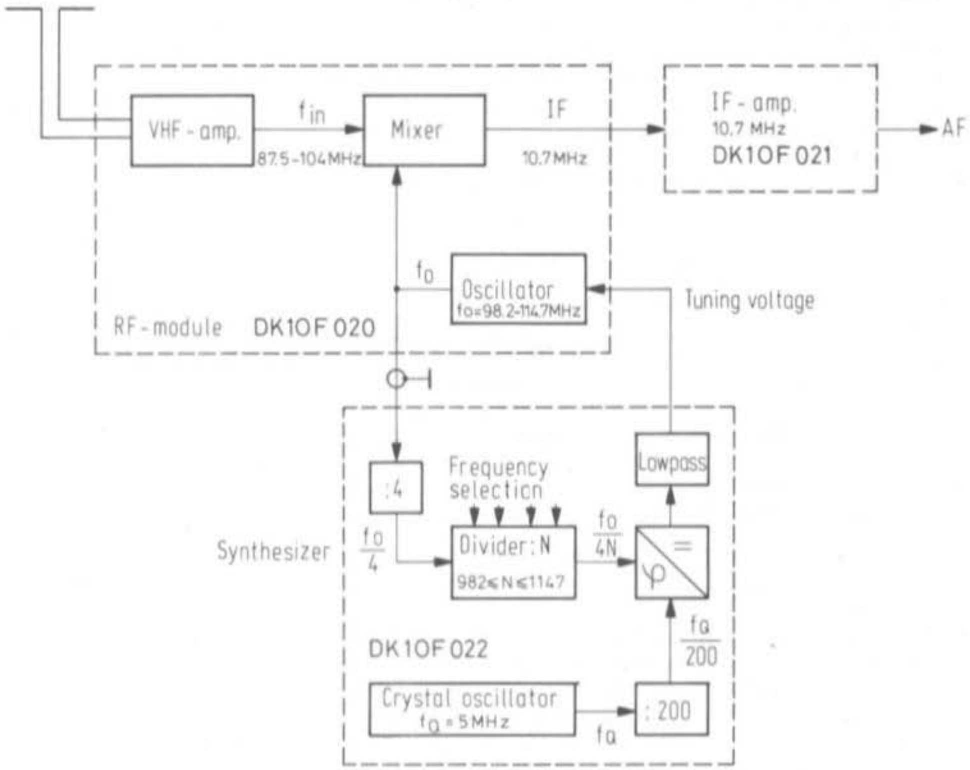

Fig.2: Block diagram of the VHF tuner with synthesizer

2. CONCEPT AND FREQUENCY PLAN OF THE RECEIVER

In Germany, the postal regulations require that the local oscillator of FM broadcast receivers oscillate at a frequency higher than the receive band so that no interference is made to the police and other services operating in the 4 m band between 75 and 87.5 MHz. When using an intermediate frequency of 10.7 MHz and an input frequency range of 87.5 to 104 MHz, it is necessary for the oscillator to operate in the range of 98.2 to 114.7 MHz in steps of 100 kHz.

The block diagram of the tuner with synthesizer is given in Figure 2. The tuner comprises three modules: the RF, IF and synthesizer modules. The RF and IF modules are built up in a conventional manner so that it would be possible for already available modules to be used as long as the actual tuner uses diode tuning.

The frequency synthesizer of the tuner is in the form of a phase-locked loop (PLL) (2), (3). The oscillator frequency f_o is firstly divided by four in a frequency divider before being fed to the programmable frequency divider. The latter divides this frequency further by factor N (dependent on the required channel); the output frequency of this circuit is therefore $f_o/4N$ and is fed to the "actual value" input of the phase comparator. A signal of exactly 25 kHz (5 MHz crystal frequency divided by 200) is present at the "reference" input. The phase comparator generates a DC-voltage which is dependent on the frequency and phase position of these two signals. This DC-voltage is fed via a lowpass filter to the oscillator and ensures that the nominal and actual frequencies are exactly identical. This is virtually the same principle as used for the horizontal synchronization in TV-receivers.

Under locked conditions, the following will be valid:

$$f_o/4\,N = f_Q/200$$

$$f_{in} = f_o - \text{IF remains valid.}$$

After inserting the values for f_Q and IF, the following will result:

$$f_{in} = N \times 100\text{ kHz} - 10.7\text{ MHz.}$$

This means that f_{in} is adjustable in steps of 100 kHz by selection of N.

The output voltage of the phase detector can also be used for tuning the VHF input circuits. The 5 MHz crystal oscillator is entirely responsible for the stability of the receive signal since the VCO is continuously locked to its nominal frequency with the aid of the phase-locked loop. This means that no automatic frequency control circuit is required.

3. VHF INPUT CIRCUIT

Figure 3 shows the circuit diagram of the actual VHF tuner. The input circuit comprising L 201 is used both for selectivity and matching the antenna feeder (50 - 60 Ω, Pt 201) to the input transistor T 201. Due to its low noise figure and low tendency to oscillation, a dual-gate MOSFET has been selected. An AGC voltage can be fed to point Pt 202; a control range of approximately 50 dB is provided by varying this control voltage from +6 V to -3 V. Of course, no gain-control is necessary in the case of an FM receiver, since the sooner limitation occurs, the better. However, if a fieldstrength meter is to be used, it will be necessary to have a logarithmic relationship between the input voltage and the

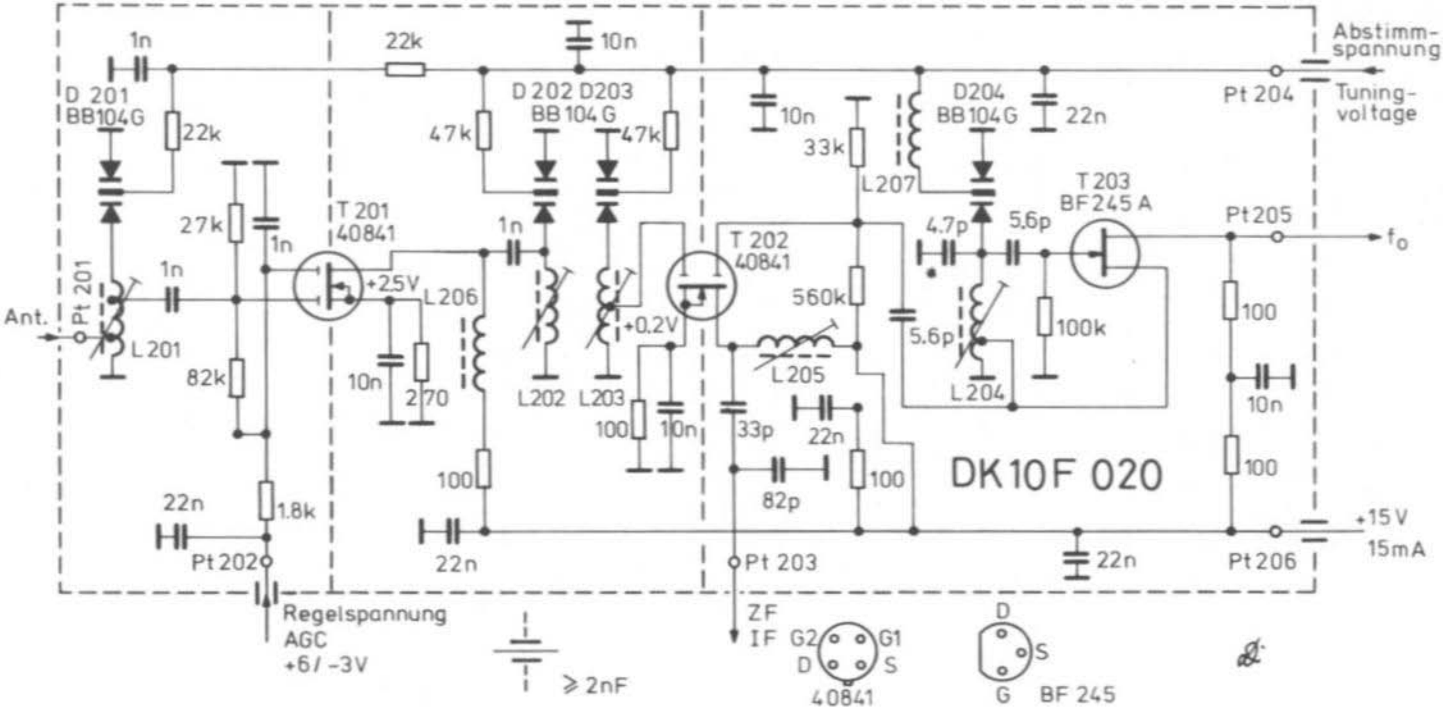

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Fig. 3: Circuit diagram of the VHF tuner

rectified IF voltage, and the control characteristics of a MOSFET are very suitable for this.

The VHF input stage is followed by an inductively-coupled, two-stage bandpass filter comprising L 202 and L 203. This stage is capacitively tuned, as was also the case with the input stage, with the aid of varactor diodes D 201 to D 203. Twin diodes are advantageous since their capacitance is hardly dependent on the RF voltage across the resonant circuit, which is especially important at high signal levels. The mixer stage comprising transistor T 202 is also equipped with a dual-gate MOSFET. The advantage here is that the input and local oscillator frequency can be fed to two, decoupled electrodes so that strong input signals have no reaction on the local oscillator, and that no oscillator signal is fed to the antenna. In addition to this, such a mixer is far less critical with respect to operating point and oscillator injection voltage fluctuations than an additive mixer.

The resulting intermediate frequency is tapped off at the drain circuit (L 205) and is fed via a capacitive voltage divider (33 pF/82 pF) to the IF output of the tuner (Pt 203). The local oscillator (T 203) operates in a common-drain circuit with inductive feedback between gate and source. The tuning is also made capacitively with the aid of diode D 204. A portion of the oscillator voltage can be taken from the drain circuit with relatively low reaction and fed via connection point Pt 205 to the synthesizer.

4. CIRCUIT OF THE IF MODULE

The characteristics of the IF amplifier have the greatest effect on the overall quality of the whole Hi-Fi receiver. As is known, the spectrum of a frequency-modulated signal consists of a wide band of spectral lines which must not only be transmitted as a whole but should also appear at the demodulator simultaneously. This means that the IF module must not only provide a sufficient bandwidth but also a constant group delay within the passband range. This means that the relationship between the frequency and phase of the input and output signal must be as linear as possible. This is rather in contrast to the demands for the highest possible selectivity. Advanced receivers use a large number of LC-filters, which can only be aligned using special measuring equipment such as swept-frequency, and group delay measuring systems. Such filters cannot be used for home construction. Lower priced receivers mainly use ceramic filters; extensive experiments made by the author using such components did not lead to favourable results due to the large manufacturing tolerances and spurious resonances of the ceramic resonators.

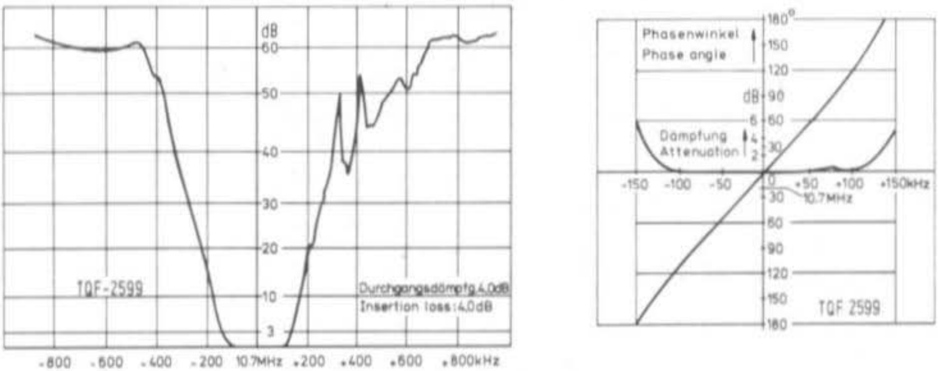

Fig.4: Selectivity and phase characteristics of the crystal filters used

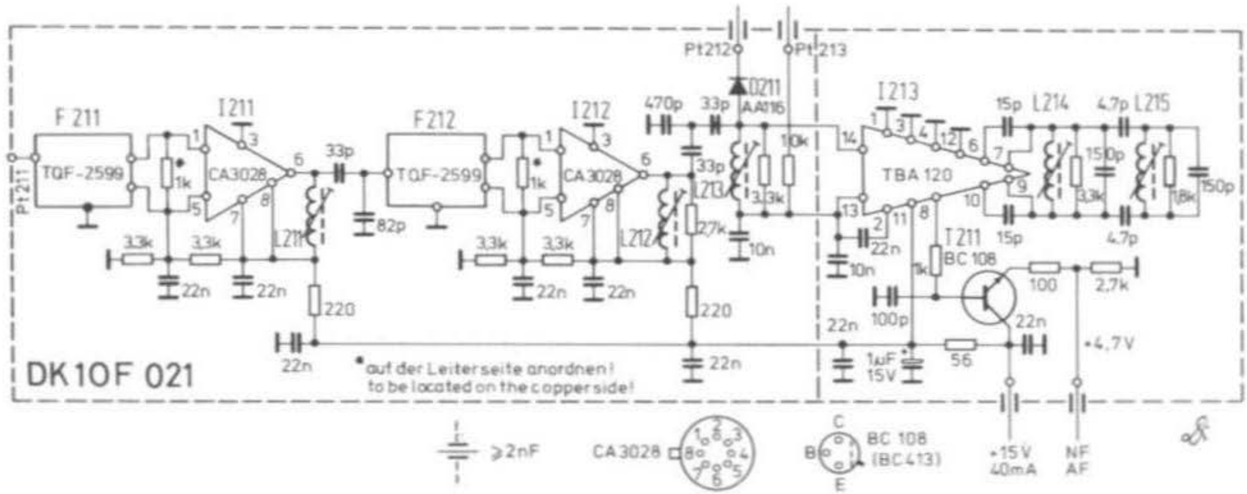

Fig.5: Circuit diagram of the IF module

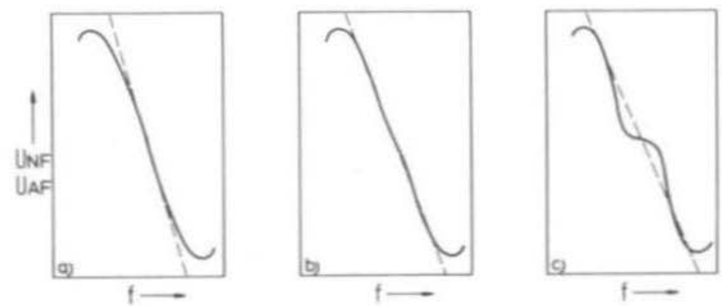

Fig.6:

Demodulation characteristics of:
Simple resonant circuit (a)
optimumly dimensioned bandpass filter (b)
overcoupled bandpass filter (c)

For some time now, special phase-linear crystal filters having a relatively steep slope, have appeared on the market at reasonable prices (Fig. 4). If two such filters type TQF-2599 are used together with several wideband LC-circuits, it is possible for good results to be obtained with respect to the passband and stopband selectivity. The advantage is that the selectivity curve of the crystal filter is only slightly dependent on the adjustment of the resonant circuits required for matching and transformation. This means that no extensive measuring equipment is required for the alignment.

The circuit diagram of the IF module is given in Figure 5. The IF output of the mixer (Pt 203) is connected to the input connector Pt 211. The first crystal filter F 211 is followed by the buffer stage I 211 which is equipped with the integrated differential amplifier CA 3028; the signal is then fed to the second crystal filter via the matching circuit L 211. The second IF stage is identical to the first; a capacitively-coupled bandpass filter comprising L 212/L 213 is provided at the output of this stage, which has been additionally dampened using resistors. This bandpass filter is used to improve the ultimate attenuation. On their own, the two crystal filters would only provide a stopband attenuation of approximately 70 dB (Fig. 4). Diode D 211 rectifies a portion of the IF voltage (Pt 212), and connection Pt 213 provides the reference potential. An AGC and fieldstrength indication circuit which is to be described later is connected here.

The subsequent integrated circuit I 213 represents a six-stage limiting IF amplifier with coincidence demodulator. Similar circuits have been described previously so that the operation need not be described here. It should be noted, that a double-tuned circuit is used (L 214, L 215) instead of the usual phase circuit. This allows a far better approximation to the ideal characteristic, which results in a lower distortion factor, especially at high-frequency deviation levels. Figure 6 shows these relationships in the form of a diagram. In the case of the author's prototype, the non-linearity within the range ±100 kHz from the centre frequency was less than 1% (Fig. 7). The distortion factor resulting from this is therefore in the order of 0.2%.

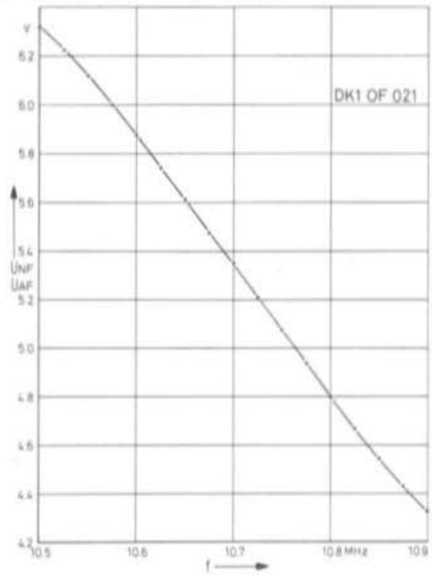

Fig. 7:

Measured demodulator characteristic

The AF output (pin 8 of I 213) cannot be taken directly via a feedthrough capacitor since the resulting lowpass filter (R_{out} = 2.5 kΩ, C = 2 nF) would cause too great an attenuation of the higher audio frequencies. For this reason, an additional stage has been provided comprising T 211 as impedance converter. A capacitor of 100 pF at the base of T 211 is sufficient for RF-bypassing. If the IF section is only to be used for mono-reception, it is possible for this capacitor to be increased to 15 nF and used for de-emphasis.

Experiments made using the integrated circuit CA 3089 (RCA) instead of the TBA 120 were not successful. It is true that the former circuit provides more extra features such as AGC voltage generation, AFC, and squelch, however, it is rather difficult to use this IC due to the large tendency to self-oscillation. Reliable operation was only possible when using a double-coated PC-board with continuous ground surface. In spite of this, it was not possible to obtain a demodulation characteristic that was independent of the IF input voltage. The integrated circuit TBA 120 is completely uncritical in this respect since it is fully balanced with respect to the supply and ground connections. This means that all RF currents are cancelled out. The integrated circuit TBA 120 will still operate reliably even when the operating voltages were not bypassed. Of course, the squelch and control voltage circuits are somewhat more complicated.

5. CIRCUIT OF THE SYNTHESIZER MODULE

5.1. DIVIDER 4 : 1

Since programmable frequency dividers for input frequencies up to 120 MHz can still not be easily realized with components available at present on the market, it is necessary for the frequency of the local oscillator (f_o) to be divided firstly by four. This is obtained in the circuit given in Figure 8. The oscillator signal is injected to connection point Pt 221 which is connected to Pt 205; it is then amplified in transistor T 221 and feeds the phase-reversal stage comprising transistor T 222. This stage provides two antiphase signals for driving the actual frequency divider comprising I 221. This stage is equipped with an ECL-dual flip-flop which can process input frequencies up to typically 180 MHz. The subsequent circuit comprising transistors T 223 and T 224 converts the ECL-level to a signal that is TTL compatible. Since transistor T 224 operates as a saturated switch and since the output frequency can amount to 30 MHz, it is necessary for a very fast switching transistor to be used.

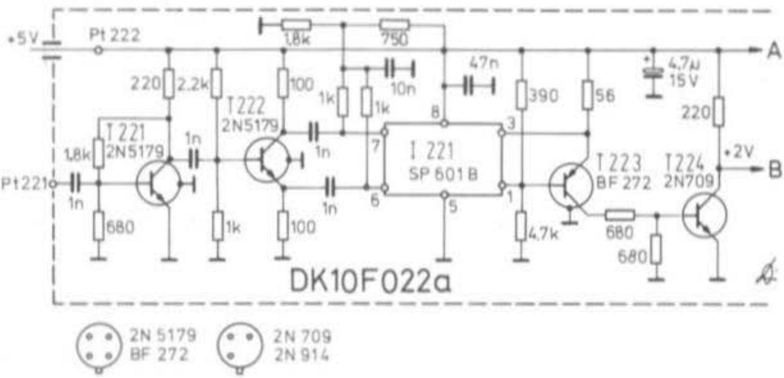

Fig.8: 4:1 divider with preamplifier and level converter

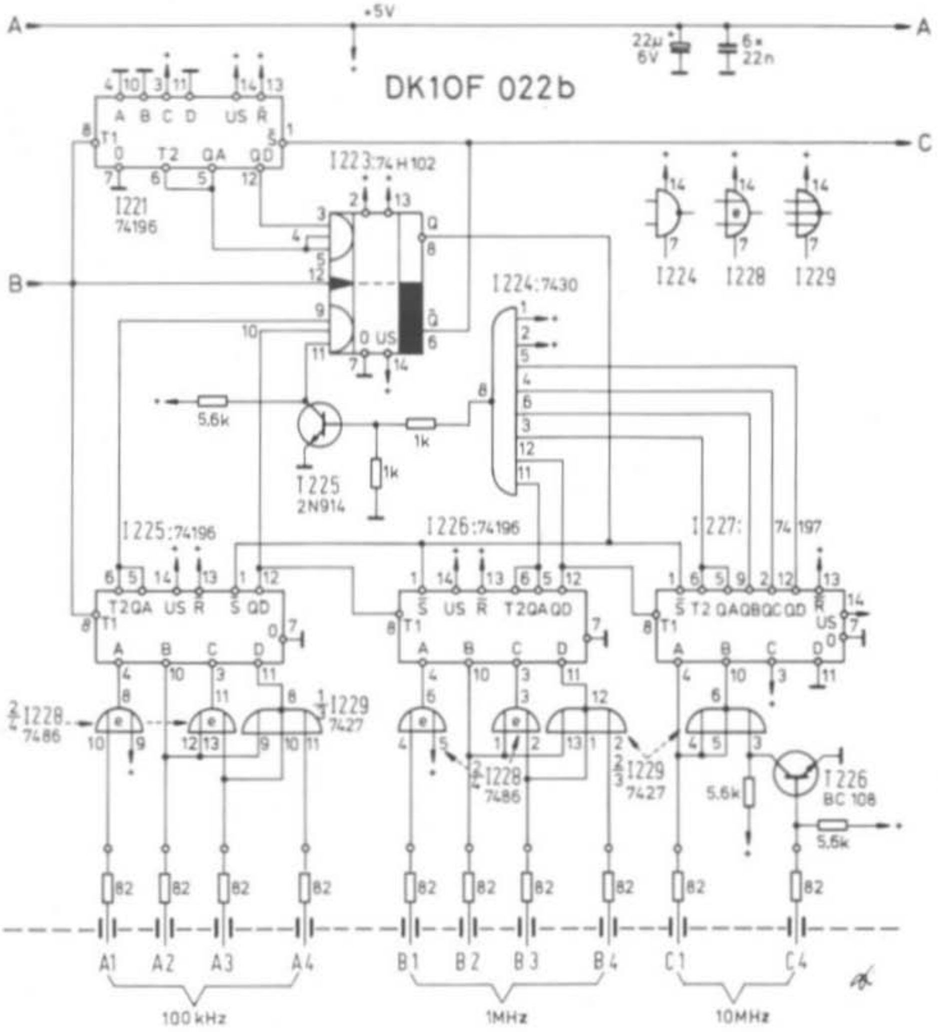

Fig.9: Programmable frequency divider

5.2. PROGRAMMABLE FREQUENCY DIVIDER

The variable divider shown in Figure 9 originates from information given in (4). This frequency divider must divide the output signal of the 4 : 1 divider by N ($982 \leq N \leq 1147$), and is to be programmed directly with the frequency of the required signal (875 to 1040). The signal from the 4 : 1 divider is fed to the clock inputs of the integrated circuits I 222, I 223, and I 225. The operation is now to be explained assuming that the number 881 (corresponding to 88.1 MHz) has been selected at the programming inputs A 1 to C 4. The gate circuits previous to the preset inputs of the decade counters I 225 and I 226 ensure that they are set to the ninth complement of the selected number (the ninth complement of a number n is 9 - n). Integrated circuit I 227 is a 16-counter. It is fed with the fourteenth complement of the number programmed on C 1/C 4. If a zero is given, the preset inputs A to D will receive the fourteenth complement of the number 10. This means that only three rows of numbers are required, since the number zero in the 10-MHz position corresponds to 100 MHz. After 981 clock pulses, the decimal counters I 225 and I 226 are at 9, whereas the 16-counter I 227 is at 15. The output of the NAND-gate I 224 then drops to L (low) level and the output of the inverter T 225 thus to H (high). Since Q_A and Q_D of I 225 are also at H-level, the K-input of the flip-flop I 223 will be activated. This means that it will be switched to its on-state (Q = L) together with the 982th clock pulse. The counters I 225 to I 227 are now preset to their commencement value, and I 222 will switch to the counting mode due to $\overline{Q}$ = H. Since it has been set to the number 4 by the data inputs A to D, it requires five further pulses until the number 9 is obtained. The 987th clock impulse activates the J-input of the flip-flop, and the 988th input pulse switches it back to the rest position. This completes the cycle, the input frequency will have been divided by factor 988, the output frequency can then be taken from the output Q of flip-flop I 223 and fed to the phase comparator circuit. At N = 988, the oscillator frequency of the tuner will amount to 98.8 MHz, which corresponds to an input frequency of 88.1 MHz when an IF of 10.7 MHz is used.

The adjustment of the required input frequency is therefore made on the programming inputs A 1 to C 4. The following table indicates the type of coding:

10 MHz position			1 MHz position					0.1 MHz position				
	C_4	C_1		B_4	B_3	B_2	B_1		A_4	A_3	A_2	A_1
80	H	L	0	L	L	L	L	.0	L	L	L	L
90	H	H	1	L	L	L	H	.1	L	L	L	H
100	L	L	2	L	L	H	L	.2	L	L	H	L
			3	L	L	H	H	.3	L	L	H	H
			4	L	H	L	L	.4	L	H	L	L
			5	L	H	L	H	.5	L	H	L	H
			6	L	H	H	L	.6	L	H	H	L
			7	L	H	H	H	.7	L	H	H	H
H = + 5 V			8	H	L	L	L	.8	H	L	L	L
L = 0 V (ground)			9	H	L	L	H	.9	H	L	L	H

Examples:	C_4	C_1	B_4	B_3	B_2	B_1	A_4	A_3	A_2	A_1
88.2 MHz	H	L	H	L	L	L	L	L	H	L
96.3 MHz	H	H	L	H	H	L	L	L	H	H
102.8 MHz	L	L	L	L	H	L	H	L	L	L

A drawing showing the connections to the selector switches as well as a recommendation for obtaining a program memory are to be given in Part II of this article.

5.3. PHASE COMPARATOR AND REFERENCE OSCILLATOR

A 5 MHz crystal-controlled oscillator comprising transistor T 228 will be seen in the circuit diagram given in Figure 10. The trimmer capacitor in series with the crystal allows the oscillator frequency to be varied by ± 50 kHz. This allows any tolerances of the centre frequency of the crystal filter to be compensated for. It is not necessary for a fine tuning to be provided for the receiver since the IF bandwidth is more than twice as wide as the channel spacing. This means that transmitters not operating exactly on a multiple of 100 kHz will still be received without distortion.

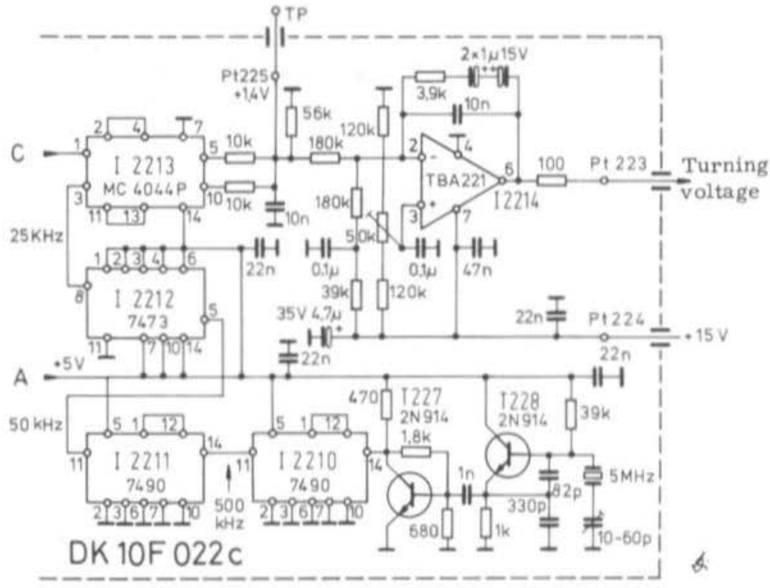

Fig. 10: Phase comparator and reference oscillator

The 5 MHz signal from the crystal oscillator is converted to a TTL-logic signal which is divided twice by factor 10 in the decade counters I 2210 and I 2211, and once again by two in the double flip-flop I 2212. The 25 kHz squarewave signal generated in this manner is fed to the "reference" input (pin 3) of the integrated frequency/phase comparator I 2213 (MC 4044 P); the "actual value" originating from the variable frequency divider is fed to pin 1. At large deviations between the nominal and actual value, I 2213 will operate as frequency discriminator and will then switch to the phase-comparator mode when the control deviation is sufficiently small so that it is able to tune out the remaining frequency difference. The output signal of the module (pins 5 and 10) represents a pulse sequence from which the integrator comprising the operational amplifier I 2214 (TBA 221) forms the mean value of the voltage, which is then fed via Pt 223 to the RF module (Pt 204) and used as tuning voltage.

It will be possible to use other VHF tuners requiring a tuning voltage of up to 30 V if the operating voltage input of the integrator Pt 224 is fed with approximately 32 to 35 V. The DC operating point of the integrator can be adjusted with the aid of the 50 kΩ trimmer resistor. The time constant (180 kΩ/0.5 μF) is in the order of 0.1 s, which means that a 10 MHz jump in frequency requires approximately 1 s.

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A SIMPLE 70 cm POWER AMPLIFIER EQUIPPED WITH THE 2 C 39

by K. Weiner, DJ 9 HO

Used tube types 2 C 39 or YD 1050 are often available cheaply or even free-of-charge. This because a large number of commercial operators using these tubes in their microwave equipment exchange the tubes at regular intervals. This has made them very popular with amateurs for constructing power amplifiers for the UHF bands. The author is to describe how such an amplifier can be constructed more simply than was given in (1). It is to be constructed from lightweight PC-board material or from thin metal plate; even the tube is soldered into place to save provision of a socket. This means that the amplifier can be easily constructed and is therefore not only suitable for use as power amplifier stage but also as driver, etc.

1. THE CIRCUIT DESCRIPTION

The circuit diagram of the power amplifier is given in Figure 1. It will be seen that a common, grounded grid configuration is used. For those readers that have not used such disc triodes before, it should be mentioned that the tube also operates with a DC grounded grid. The required negative grid bias voltage is obtained as positive cathode bias across resistor R 1. This voltage is drive dependent and amounts to 6 V at 60 Ω and cathode current of 100 mA. A zener diode is connected in parallel with this resistor in order to ensure that the bias voltage cannot increase beyond a certain level at higher drive levels. The most favourable values of resistor and zener diode depend on the individual tube, tube condition, anode voltage and operating mode which means that they must be determined individually. At 400 to 600 V and a quiescent current of 20 to 50 mA, a zener diode of 5.6 V and a resistor of 60 Ω should be suitable. The power amplifier can be used for AM, FM, SSB and CW operation, as well as for amateur television. A potentiometer can be used as resistor R 1 (250 Ω) so that a quiescent plate current of 20 mA can be selected.

Since the cathode of the 2 C 39 is connected to one side of the heater, this will mean that the bias voltage is virtually at the same potential as the heater. This requires that one side of the heater winding of the transformer may not be grounded as usual since this would short-circuit the bias voltage. Neither is it possible for several such tubes to be operated from the same heater winding if different bias voltages are required.

The heater voltage should never exceed 6 V. Exact values as a function of the operating frequency and drive level were given in (1). This article also gave the specifications of the various types in this family of tubes.

Since the UHF drive is also fed to the cathode, it is necessary for the heater leads to be provided with chokes for the operating frequency. Chokes L 4 and L 5 ensure that no UHF drive is lost. The capacitive voltage divider comprising C 1/C 2 and inductance L 1 match the impedance of the cable to the complex input impedance of the tube.

The anode circuit L 2 of the power amplifier is in the form of a shortened $\lambda/4$ Lecher line. Since a balanced type of line is used with only a single tube, trimmer C 3 represents the tube capacitance at the open end of the line. This trimmer is also used for tuning the line to resonance which can shift the ba-

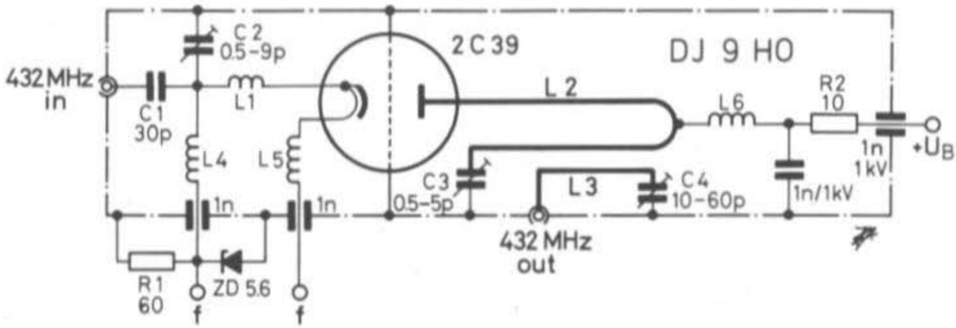

Fig.1: Circuit diagram of the 70 cm power amplifier

lance so that choke L 6 can be required at the normal "cold" end of the line. Resistor R 2 is provided as type of fuse. Its value is not critical (9 to 15 Ω), however, it should not be rated at more than 1/4 W.

The output power is coupled out via L 3. The inductance of the coupling link is tuned to series resonance with the aid of trimmer C 4.

2. OPERATING VALUES

This type of tube operates reliably even with a plate voltage of 300 V. The following table gives some typical specifications for various plate voltages and drive powers which were determined using a calibrated directional coupler (Siemens) and a 50 Ω load (Spinner):

Plate Voltage	Drive Power	Plate Current	Output Power
300 V	0.4 W	28 mA	4 W
300 V	4 - 5 W	72 mA	12 W
420 V	0.4 W	35 mA	7 W
420 V	4 - 5 W	95 mA	22 W
600 V	0.4 W	42 mA	9 W
600 V	1 W	55 mA	15 W
600 V	4 - 5 W	115 mA	42 W

The tube requires a warm-up time of approximately one minute and should be forced-air-cooled even at the lowest plate voltage.

The above values show that the linearity range of the tube is well exceeded when the tube is driven to a plate current of 72 mA at a low plate voltage of 300 V. This can be seen due to the fact that the power amplification is far less than 10 dB. Higher plate voltages are far more favourable.

3. CONSTRUCTION

The 432 MHz power amplifier can be constructed from copper-coated PC-material or thin brass plate. The coated side should be on the inside as well as on the upper side of the intermediate panel (anode side). If copper or brass plate is used for the intermediate panel, this will provide a better heat dissipation for the tube. Figure 2 shows the construction of the power amplifier in the form of a drawing from above, side and below.

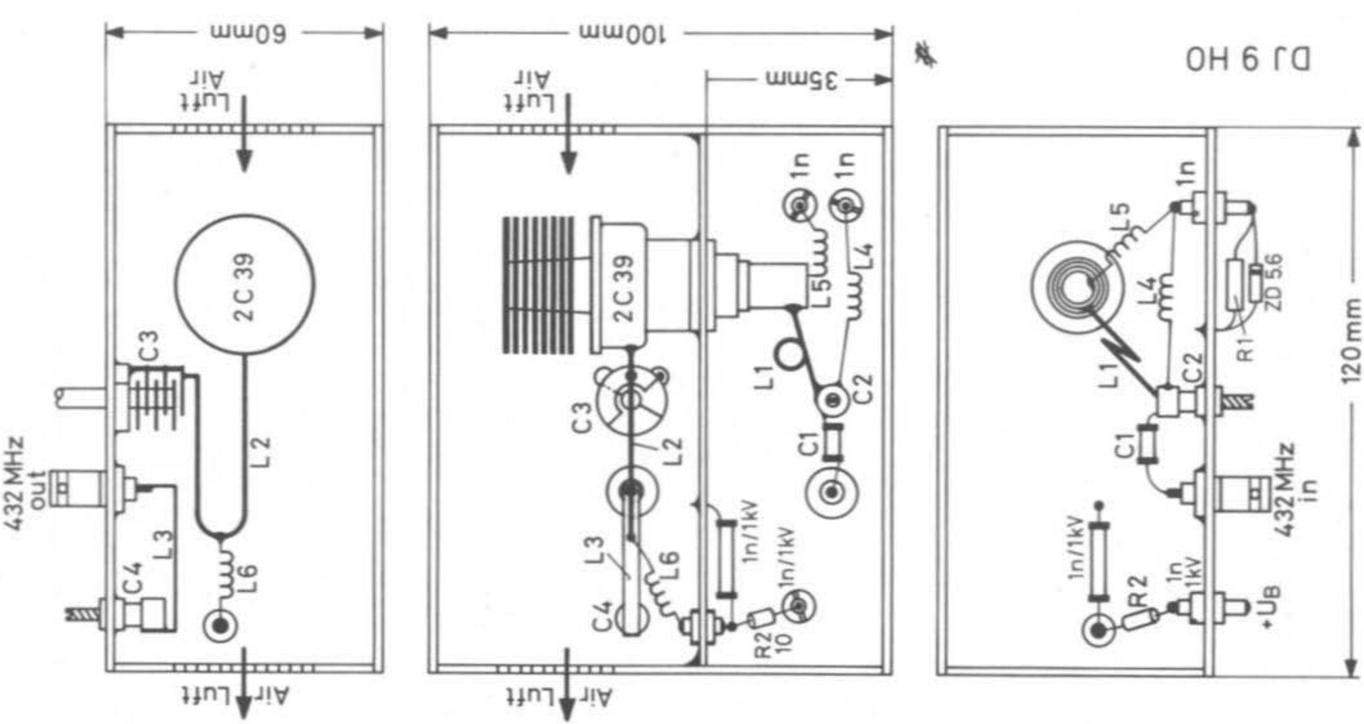

Fig.2: Construction of the power amplifier for 432 MHz

The intermediate panel is firstly provided with a hole for the grid ring of the tube as well as for the high voltage feedthrough for the plate voltage. The front panel is provided with holes for the BNC sockets, feedthrough capacitors and trimmers (e.g. ceramic 0.5 - 9 pF manufactured by Philips or air-spaced trimmers with a spacing of 1.5 mm). This is followed by soldering the framework of the case and the intermediate panel together. The cover and base plates are mounted into place after completing the power amplifier.

The inductances should be constructed as shown in Figure 3. The three chokes L 4 - L 6 comprise ten turns of 0.5 mm diameter (24 AWG) enamelled copper wire wound on a 6 mm former, self-supporting. The tube is soldered to the grid ring on the intermediate panel with the aid of a 100 to 150 W soldering iron. The solder joint should be in the anode chamber. This soldering process should take place relatively quickly in order not to damage the tube. Figure 4 indicates the positions at which the tubes should be soldered and where the anode circuit L 2 should be connected. Figure 5 shows a photograph of a double amplifier stage constructed in the described manner.

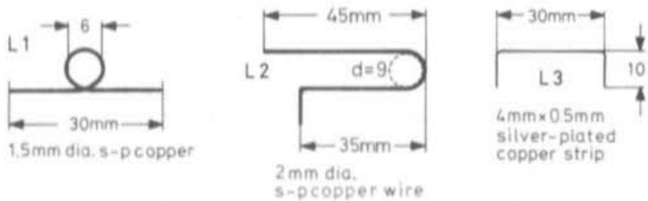

Fig.3 : Shape and dimensions of the inductances

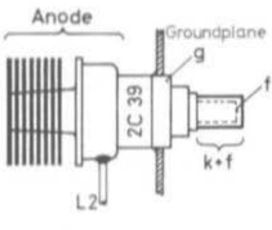

Fig.4 :
Electrodes of the
2 C 39 tubes

4. TUNING AND OPERATION

A standing wave meter (e.g. DK 2 VF 002) should be connected both in the input and output leads. The alignment is commenced at a low anode voltage and a drive power of 1 W; the cathode circuit is aligned for the most favourable compromise between maximum drive and minimum standing wave ratio, as well as the anode side for maximum output power. After this coarse alignment, the power amplifier should be provided with its normal operating voltage after which the fine alignment can be made. With the given values, the power amplifier should operate immediately.

Since the heat dissipation of this type of construction is very low, it is necessary for the tube to be cooled by forced air even when operating at a low plate voltage of 300 V. This can be achieved by drilling a large number of 3 mm diameter holes or sawing a larger window in the side panels. This window can then be screened with a wire grid. These cooling cutouts can be provided either on the longer or shorter sides. It is also favourable for some of the air to be passed through the cathode chamber.

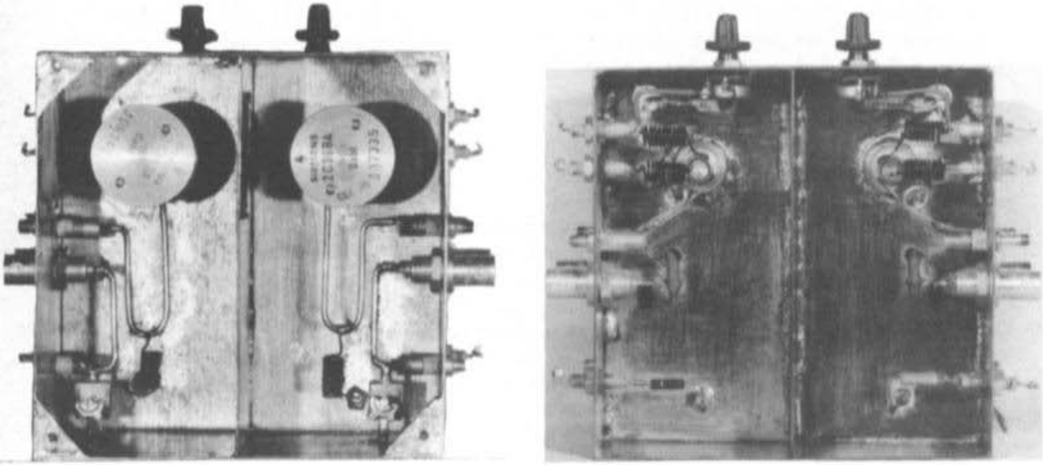

Fig.5: Two-stage amplifier using the described principle

It is possible to use the same type of construction for the 2 m band, which has also been tested in practice. However, it is thought that such a design for the 23 cm band would be of more interest. For this reason, such a power amplifier is to be described in a later edition of this magazine. Since this power amplifier is extremely easy to construct in one evening, it is felt that it can do much to forward operation on the UHF bands.

5. REFERENCES

(1) A. Tautrim: A Stripline Power Amplifier for 70 cm Using a 2 C 39 Tube
VHF COMMUNICATIONS 4 (1972), Edition 3, Pages 144-157

A VERSATILE 70 cm CONVERTER WITH SCHOTTKY-DIODE MIXER

by B. Lübbe, DJ 5 XA

Due to the danger of losing the 70 cm band, it is necessary for the activity on this band to be increased considerably. This description of a 70 cm converter is designed to assist this since it is simple, and can be constructed by less experienced amateurs. Since the UHF circuits of the preamplifier stages are in the form of microstriplines, it is not necessary to carry out the large amount of metalwork required for the resonant chambers etc. The converter can be enclosed in a TEKO box 3 A or 3 B and represents a compact, and reliable converter. It is suitable for use in the 70 cm communications band from 432 - 434 MHz using an IF of 28 - 30 MHz or as an ATV converter for the frequency range of 434.25 MHz to 439.75 MHz when using an IF of 48.5 MHz (CCIR TV channel 2). It can also be used as receive converter in 70 cm transverters or in ATV transceivers.

1. CHARACTERISTICS OF THE CONVERTER

Operating voltage:		12 V (11 - 14 V)
Current requirements:		approx. 20 mA
Overall gain:		approx. 36 dB
3 dB bandwidth of the UHF circuits:		approx. 30 MHz
Of the IF amp.	at 30 MHz:	3 MHz
	at 50 MHz:	7 MHz
Noise figure:		4 dB
Image rejection at an IF of 30 MHz:		>50 dB

2. CIRCUIT DESCRIPTION

The circuit diagram of the converter is given in Figure 1. It will be seen that the converter possesses two preamplifier stages using the well-proved Germanium PNP transistor AF 239. It may seem to some readers that these transistors are not too modern, however, extensive experiments made by the author have shown that Silicon NPN UHF transistors did not bring hardly any better results, even though they were far more expensive. The transistor AF 239 is available virtually everywhere at a very low price. The converter exhibits a very good noise figure that is only approximately 0.2 dB inferior to the best converters available on the market. Experiments made with the transistor AF 279 in the UHF preamplifier stages also brought good results, but the input stages possessed a greater tendency to self-oscillation.

Basically speaking, the UHF preamplifier stages used in this converter are similar to that described in (1). This means that it is not necessary to discuss them in great detail here.

The mixer stage is equipped with two Schottky-diodes type HP 2800 that operate in push-pull. They are fed at low impedance both from the signal and local oscillator input. The oscillator voltage is fed to the mixer via a 50 Ω stripline and the two 27 pF capacitors.

The low-impedance, local oscillator injection offers the advantage that other oscillators can be used with the converter (e.g. local oscillator signal from a transmit converter). In this case, a 50 Ω coaxial cable can be directly con-

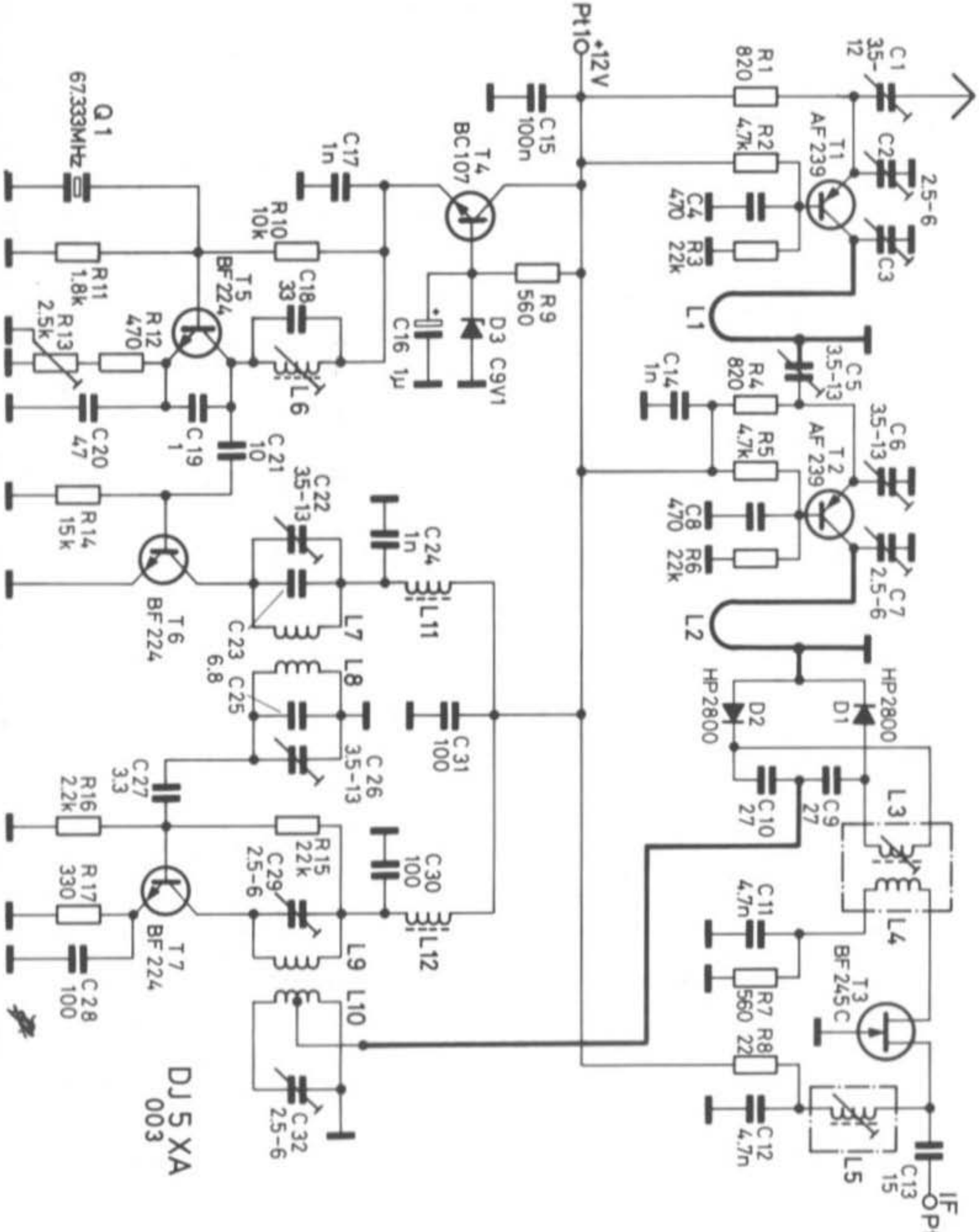

Fig.1: Circuit diagram of the 70 cm micro-strip converter with Schottky-diode mixer

nected to the stripline, and the components for the converter oscillator deleted, or the interconnections broken the positions marked on the PC-board. The UHF preamplifier stages and the mixer then form a common unit which can be used in transceivers and transverters as was recommended in (2).

The mixer is followed by an IF preamplifier equipped with a field-effect transistor operating in a common-gate circuit. It is provided to firstly compensate for the conversion loss, and, secondly to provide sufficient drive for less sensitive shortwave receivers. If, however, the gain of the converter is too high, it is possible for the output circuit of the IF amplifier to be damped with the aid of a resistor. This also ensures an increase of the IF bandwidth. The local oscillator comprises three stages using the inexpensive transistor type BF 224. This circuit is very similar to the oscillator DJ 4 LB 003, with the exception that the last stage has been deleted. The same comments that were made in (2) are also valid for this oscillator which means that it need not be discussed in detail here. There is, however, one special feature which should be mentioned: The zener diode for stabilization of the oscillator voltage must be operated with a current of at least 2 to 5 mA, since its noise level will increase the overall noise of the converter at lower current levels. Special attention should be made to this when the module DJ 4 LB 003 is used in a transverter.

3. CONSTRUCTION

The most favourable means of complementing the PC-board DJ 4 XA 003 is as follows: Firstly mount all resistors, then all capacitors and trimmer capacitors; this is followed by the semiconductors and finally the inductances, bandpass filters, chokes and the crystal. Resistor R 8 (22 Ω) should have a spacing of 5 mm from the PC-board in order to ensure that the field of the oscillator stripline is not affected. The side of the board which is provided with the stripline circuits is the component (upper) side. The components are soldered to the lower side having the larger ground surface. As was previously mentioned, the construction of the converter is extremely simple if the 95 mm by 67.5 mm large, double-coated PC-board of 1.5 mm in thickness, and if the components given in section 4. are used. Special attention must be paid to the base bypass capacitors of the two UHF preamplifier stages and the coupling capacitors of the two UHF preamplifier stages and the coupling capacitors from the oscillator stripline to the mixer diodes. They pay a very major part in the reliability of the overall converter. Capacitors without connection leads, or chip-capacitors should be used. The use of other types is a waste of time. These capacitors are mounted onto the PC-board by sawing suitable slots at the required positions with the aid of a fretsaw. The bypass capacitors for the preamplifier stages are mounted with the base connection and the copper layer on the upper side, and the ground connection at the lower side of the board. This must be made with considerable care since the capacitors are somewhat brittle. The same is valid for the two coupling capacitors between oscillator and mixer. They are soldered to the stripline and diodes on the other side of the board. Only the connections to the IF resonant circuits are made on the lower side of the board.

The IF inductances are made up using special coil sets that are available from the publishers. The coilformers possess six pins of which the two centre pins should be deleted. Since the location is exactly given, it is necessary to pay attention that the connections are correct.

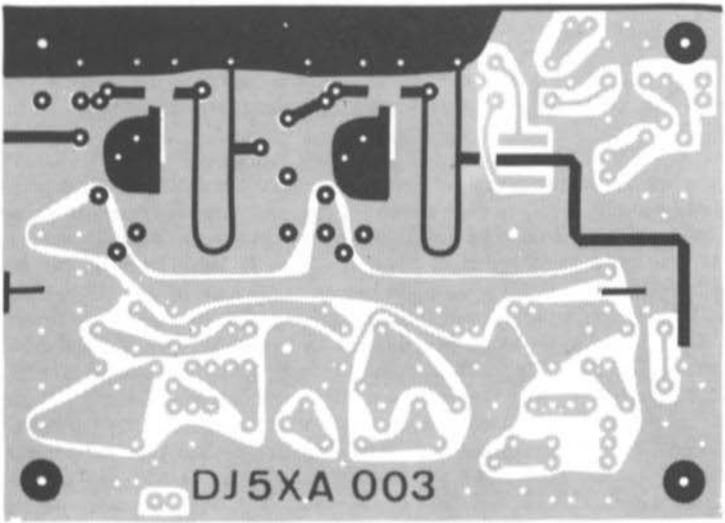

Fig.2: Double-coated PC-board DJ 5 XA 003 for the converter

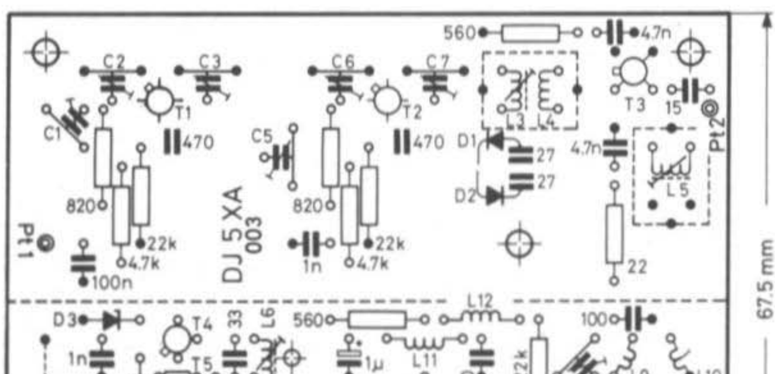

Fig.3: Component locations on PC-board DJ 5 XA 003

The oscillator coil L 6 should be wound so that the cold end faces upward. This ensures that adjacent effects are kept to a minimum. The inductances for the 202 MHz bandpass filter are spaced 5 mm from the PC-board and the spacing is 3 mm in the case of the 404 MHz bandpass filter.

3.1. OPERATION AS AN ATV CONVERTER

An intermediate frequency of 48.25 MHz has been selected for this ATV converter so that it can be connected to virtually any domestic television receiver that is able to tune CCIR TV channel 3 or 4 in TV band I. Crystals oscillating in the order of 60 MHz should not be used since they are liable to cause interference to the subsequent TV receiver. Further details regarding the values for the IF amplifier for this mode are given in Section 4. The oscillator operates at a frequency lower than the input frequency in order to ensure that the correct sideband position is provided. If the ATV signal is to be directly converted to the standard TV IF of 38.9 MHz, and if the signal is only to be processed in the IF amplifier of the TV receiver, it will be necessary for the oscillator to operate above the signal frequency. Further details regarding this are given in (2) so that they need not be dealt with here.

3.2. OPERATION IN A TRANSVERTER OR TRANSCEIVER

If the converter is to be used in a transverter or transceiver in which a local oscillator is already available, it is possible for the board to be sawn off at the dashed line shown in Figure 8 and for only the two preamplifier stages and the mixer to be used. The local oscillator signal is then injected into the converter by connecting a coaxial cable to the 50 Ω stripline; the inner conductor of the cable should be connected to the stripline and the screening to the lower side of the board. The required local oscillator power is in the order of 1 to 5 mW. If oscillator module DJ 4 LB 003 is used, which provides a separate output for receive mixers, sufficient power will be available for this application.

4. COMPONENTS

T 1, T 2: AF 239, AF 239 S or AF 139
T 3: BF 245 C (TI), E 300 (Siliconix) or similar FET
T 4: BC 107, BC 172, BC 182, BC 413, BCY 58 or similar NPN transistor
T 5 - T 7: BF 224 (TI) or BF 173 or BF 199 (Siemens)

D 1, D 2: HP 2800, HP 2811, HP 2817 or similar Schottky diodes
D 3: BZY 83/C9V1 or other 9.1 V zener diode

Q 1:67.333 MHz (HC-6U) for IF 28-30 MHz or 64.333 MHz for IF 48.25 MHz

L 1, L 2: Striplines as described in (1)
L 3 - L 5: Special coilsets. The number of turns are valid for an IF of 28-30 MHz, values in brackets for an IF of 48.25 MHz
L 3: 14 turns (10 turns) of 0.3 mm (29 AWG) enamelled copper wire
L 4: 7 turns (5 turns) of 0.3 mm dia. (29 AWG)
L 5: 22 turns (18 turns) of 0.3 mm dia. (29 AWG)
L 6: 4.75 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire on a coilformer of 5 mm outer diameter with VHF core (brown)
L 7, L 8: 2 turns of 0.8 mm (20 AWG) silver-plated copper wire wound on a 5 mm former, self-supporting, spaced 5 mm from the PC-board
L 9, L 10: 2 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 4 mm former, self-supporting, spaced 3 mm from the PC-board, coil tap on L 10: 1 turn from the cold end
L 11, L 12: 3.5 turns of 0.4 mm (26 AWG) enamelled copper wire placed through a ferrite bead.

C 1, C 5, C 6, C 22, C 26: 3.5 - 13 pF 7 mm dia. ceramic trimmer or plastic-foil type
C 2, C 3, C 7, C 29, C 32: 2.5 - 6 pF trimmer as above
C 4, C 8: 470 to 1000 pF (uncritical) ceramic disc or chip capacitor without connection leads
C 9, C 10: 20 - 30 pF (value uncritical) as C 4
C 13: 15 pF for IF 28-30 MHz (10 pF for IF 48.25 MHz)
C 16: 1 μF drop-type tantalium electrolytic

All other capacitors are ceramic-disc types for 5 mm spacing.

A spacing of 12.5 mm is available for all resistors.

R 13: 2.5 kΩ trimmer potentiometer for vertical mounting, spacing 5/2.5 mm.

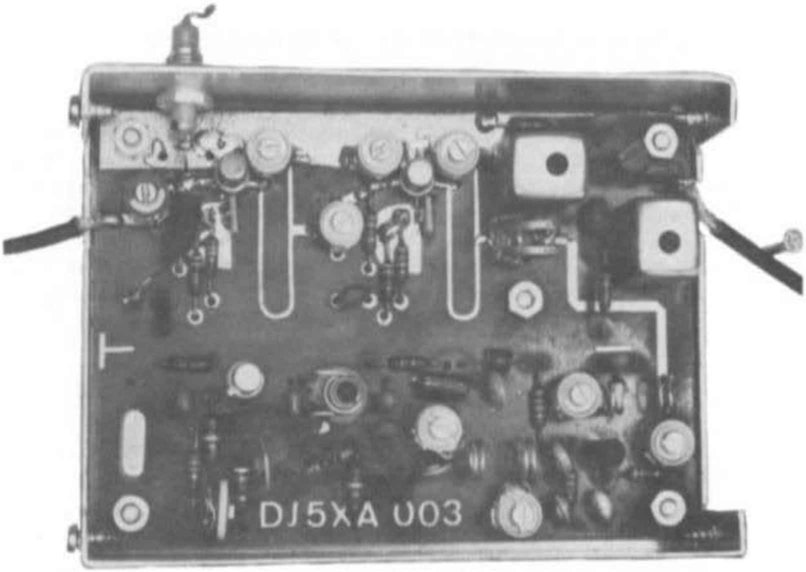

Fig.4: Author's prototype of the 70 cm converter

5. ALIGNMENT

After completing the converter, an operating voltage of +12 V to ground is connected to connection Pt 1. The oscillator and multiplier are tuned to the required frequencies with the aid of a dip-meter. In the standard version of the converter, these frequencies are 67.333 MHz (oscillator), 202 MHz (multiplier) 404 MHz (doubler). With the ATV converter these frequencies are 64.333 MHz (oscillator), 193 MHz (tripler), 386 MHz (doubler). The oscillator injection potentiometer R 13 should be in a central position during this alignment. After ensuring that the oscillator is operating correctly, the IF output should be connected to the shortwave receiver, or TV receiver, and the converter input connected to a 70 cm antenna or signal generator. The two IF circuits are then roughly aligned for maximum noise in the shortwave receiver which is followed by roughly aligning the UHF stripline and matching circuits. The fine alignment is made of a relatively weak signal at the centre of the band where the circuits are aligned for maximum gain and constant gain over the whole band.

In the case of the ATV converter, it should be aligned for best picture quality, e.g. for minimum overshoot at the bright-dark contours as well as for a constant resolution. If no measuring equipment is available and no 70 cm ATV signal can be received, the third harmonic of a 2 m transmitter can be used for alignment. Unwanted cross modulation can occur in the immediate vicinity of strong UHF TV transmitters adjacent to the 70 cm band. In such cases, a 70 cm bandpass filter should be placed in the antenna line.

The alignment is completed with the alignment of potentiometer R 13 (oscillator injection). This is made when receiving a strong (50 μV) signal at the centre of the band, and aligning R 13 until the signal just stops increasing.

6. REFERENCES

(1) W. Schumacher: Dimensioning of Micro Stripline Circuits
VHF COMMUNICATIONS 4, Edition 3/1972, Pages 130-143

(2) G. Sattler: A Modular ATV Transmitter
VHF COMMUNICATIONS 5, Edition 1/1973, Pages 2-15
VHF COMMUNICATIONS 5, Edition 2/1973, Pages 66-80.

AN SHF-WAVEMETER

by K. Hupfer, DJ 1 EE

The following SHF wavemeter is to be described in order to assist radio amateurs using the frequency range in excess of 500 MHz. This wavemeter is relatively easy to manufacture. With the dimensions given, the wavemeter operates in the frequency range of 500 to 2500 MHz, which makes it suitable for the 24 cm and 12 cm amateur bands and for frequencies used during the conversion process. The wavemeter allows measurements to be made down to a minimum of 2 mW, which is very useful for measurements on local oscillators etc. This allows a large number of problems to be solved that are encountered when constructing oscillators and frequency multipliers for these two GHz bands.

1. PRINCIPLE OF OPERATION

The measuring circuit exists of a coaxial arrangement, where the length of the inner conductor is continuously variable. The signal to be measured is fed through the coaxial circuit and a portion of this power is consumed in the indicating circuit. Figure 1 shows the equivalent diagram of the measuring arrangement. The parallel resonant circuit shown corresponds to a minimum coaxial length of $\lambda/4$ which is shortcircuited at the end. The distribution of voltage and current in such a coaxial circuit is indicated in Figure 2.

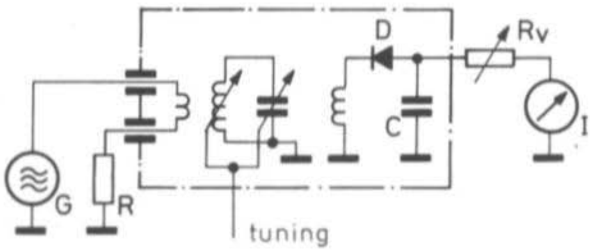

Fig. 1: Equivalent diagram of the wavemeter

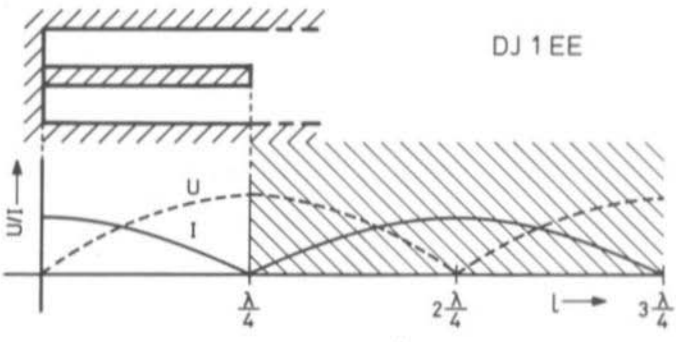

Fig. 2: Distribution of voltage and current in a quarter-wave circuit

When energizing the circuit in the $\lambda/4$ mode, a current maximum will appear at the shorted end and a voltage maximum at the open end:

$$U_1 = I_s \times Z_L \times \text{sine} \frac{2\pi 1}{\lambda} \quad (1) \qquad I_1 = I_s \times \text{cosine} \frac{2\pi 1}{\lambda} \quad (2)$$

Where:

U_1 = voltage at position 1 (spaced 1 cm from the shorted end)
I_1 = current at position 1
I_s = shortcircuit current
Z_L = impedance of the coaxial line
λ = wavelength

With the described wavemeter (Fig. 3), the position of maximum current is determined using a measuring link at the low impedance end of the line. The injection of the signal is also made at this end using a further link having an impedance of approximately 100 Ω. If these links are loosely coupled, the mechanical length of the inner conductor will correspond to the electrical length so that the length of the inner conductor within the circuit will directly coincide with the quarter wave length under resonance conditions. This means that the calibration of the wavelength can be made with the aid of a scale. Practically speaking, there is a difference between the electrical and mechanical length due to the stray field existing between the end of the (thick) inner conductor and the continuing outer conductor. This means that the scale must be corrected by two or three calibration points. In the case of the author's prototype, a shortening of approximately 3 mm takes place at 1.3 GHz and approximately 4 mm at 2.3 GHz.

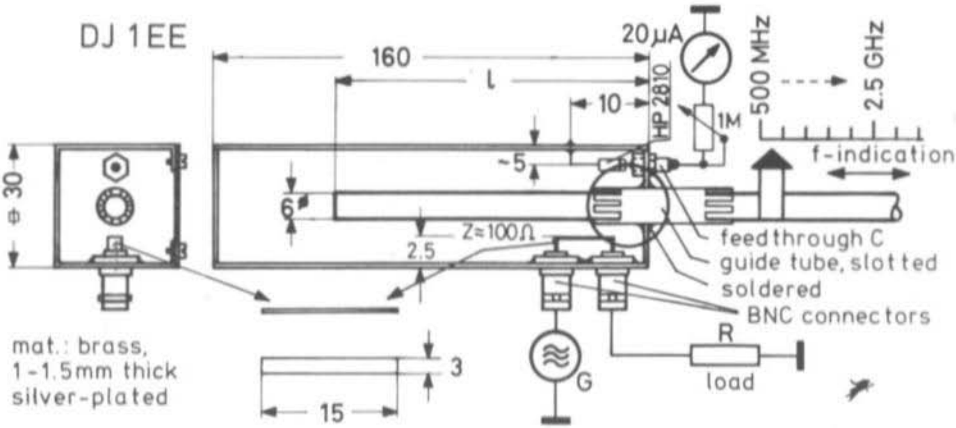

Fig.3: Construction details of a wavemeter for 500 to 2500 MHz

According to equations (1) and (2), the coaxial circuit will also be energized at odd multiples of $\lambda/4$. This means that a frequency which causes an indication of $\lambda/4$ will also cause a second indication when the length of the inner conductor corresponds to $3\lambda/4$. However, this should not cause any confusion when the measurement is commenced with the inner conductor completely extracted.

2. CONSTRUCTION

Since the impedance of the coaxial line is of no great importance, the dimensions and construction details given in Figure 3 are not critical. The outer conductor can have either a round or square cross-section. The coupling link between the input and output is designed for an impedance of approximately 100 Ω so that the frequency meter only provides an SWR of 2 : 1. It is important that a good contact is made to the inner conductor and that it is accurately guided in the centre of the circuit. Any poor contacts in the inside of the circuit will cause the indication to jump so that the measurements will not be reliable. The longer the ends of the slotted tube holding the inner conductor, the greater the accuracy of the inner conductor within the circuit. The complete circuit is mounted on a chassis that includes a meter, which should be as sensitive as possible, and a scale indicating the wavelength. The scale can be directly calibrated with the frequency if the length is transposed in the following manner:

$$f = \frac{300\ 000}{4 \times l} \quad \text{with } l \text{ in mm and } f \text{ in MHz.}$$

Figure 4 shows two examples for applications using the wavemeter.

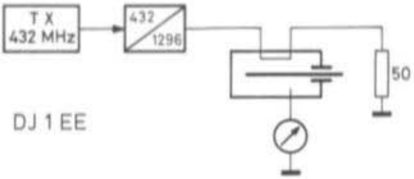

Fig.4a:
Frequency measurement

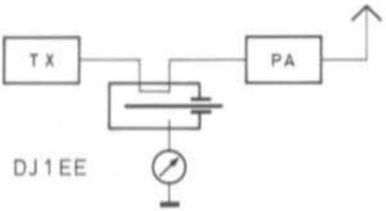

Fig.4b:
Drive measurement

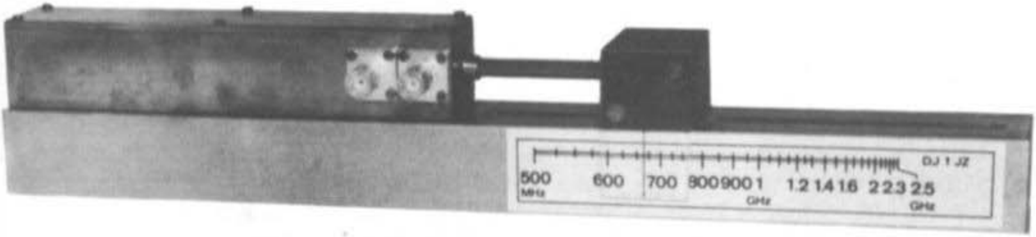

Fig.5: Photograph of author's prototype

ACTIVE RC-BANDPASS FILTERS
PART II: PRACTICAL CONSTRUCTION

by D.E.Schmitzer, DJ 4 BG

As a refresher and to further explain the formulas given in Part I, several examples are to be calculated and the measured results of the circuit are to be compared with the nominal values.

8. EXAMPLES OF THE EQUATIONS GIVEN IN SECTION 3, 4 AND 5 OF PART I

In order to show the difference between the methods of calculation, the calculations are to be based on the same demands: A bandpass filter is required with a centre frequency of 800 Hz and a bandwidth of 200 Hz. This corresponds to a Q of 4. Only the gain is varied from case to case so that the equations of Sections 3, 4 or 5 are valid.

8.1. APPLICATION OF THE EQUATIONS GIVEN IN SECTION 3 (Example 1)

In addition to the previously given values of f_o = 800 Hz and Q = 4, it is assumed that a gain of 2 is required. This means that the demand is fulfilled that Q must be greater than the root of half the gain G. The value of G = 2 can be selected freely; at a Q of 4, the demand $Q = \sqrt{G/2}$ is fulfilled for all values of G that are equal to or less than 32. Capacitors C 1 and C 2 are assumed to be of the same value of 22 nF each and thus to be 22×10^{-9} F. This means that equations 7, 9, 10 and 11 are valid.

$$R\,1 = \frac{Q}{G \times \omega_o \times C} = \frac{4}{2 \times 2\pi \times 800 \times 22 \times 10^{-9}} = 1.8086 \times 10^4 = 18.086\ k\Omega$$

$$R_p = \frac{1}{2 \times Q \times C \times \omega_o} = \frac{1}{2 \times 4 \times 22 \times 10^{-9} \times 2\pi \times 800} = 1130 \times 1.13\ k\Omega$$

$$R\,3 = 2 \times G \times R\,1 = 2 \times 2 \times 18.086\ k\Omega = 72.34\ k\Omega$$

$$R\,2 = \frac{R\,1 \times Rp}{R\,1 - Rp} = \frac{18.086 \times 1.13}{18.086 - 1.13} = 1.205\ k\Omega$$

8.2. APPLICATION OF THE EQUATIONS GIVEN IN SECTION 4 (Example 2)

The same bandpass filter with f_o = 800 Hz and Q = 4 is now to be designed for a gain G of 50. This means that $\sqrt{G/2}$ is greater than Q which means that the equations given in Section 4 are valid. Resistors R 1 and R 2 are assumed to be of 10 kΩ each. The required components are given in equations 14 to 16 where a value of Rp of 5 kΩ results (equation 1).

$$R\,3 = \frac{G \times R\,1}{1 - \frac{Q^2 \times Rp}{G \times R\,1}} = \frac{50 \times 10^4}{1 - \frac{16 \times 5 \times 10^3}{50 \times 10^4}} = 5.95 \times 10^5\,\Omega = 595\ \mathrm{k}\Omega$$

$$C\,2 = \frac{G \times R\,1}{Q \times Rp \times R\,3 \times \omega_o} = \frac{50 \times 10^4}{4 \times 5 \times 10^3 \times 595 \times 10^3 \times 2\pi \times 800} = 8.35 \times 10^{-9}\,\mathrm{F}$$

$$C\,1 = \frac{Q}{G \times R\,1 \times \omega_o} = \frac{4}{50 \times 10^4 \times 2\pi \times 800} = 1.59 \times 10^{-9}\mathrm{F} = 1.59\ \mathrm{nF}$$

8.3. DELETING R 2 (Example 3)

As was explained in Part I, it is possible for the values of R 1 and R 2 to be selected freely if Q is less than $\sqrt{G/2}$. This means that it is possible for R 2 to be selected so that it is infinit, in other words, deleted. In this case, equations 16 to 18 are valid if the gain is to be G = 50 as in the previous example.

There is no change with respect to the equation for C 1, which means that a value of 1590 pF which was found in the previous section also remains valid here.

$$R\,3 = \frac{G \times R\,1}{1 - \frac{Q^2}{G}} = \frac{50 \times 10^4}{1 - \frac{16}{50}} = 7.35 \times 10^5\,\Omega = 735\ \mathrm{k}\Omega$$

$$C\,2 = \frac{G}{Q \times R\,3 \times \omega_o} = \frac{50}{4 \times 7.35 \times 10^5 \times 2\pi \times 800} = 3.38 \times 10^{-9}\,\mathrm{F} = 3.38\ \mathrm{nF}$$

8.4. TRANSITIONAL RANGE

Both methods of calculation are possible in the range of $\sqrt{G/2} < Q < \sqrt{G}$. At the required Q of 4, this is within the range of G = 16 to 32. If a gain G of, for instance, 20 is selected, this will be within this range so that it is possible for the equations of Selection 3, and those of Section 4 to be used.In addition to this, it is possible to delete R 2. The following examples explain this more clearly:

Example 4: According to Section 3, the following are valid at C 1 = C 2 = 22 nF: R 1 = 1.809 kΩ, R 2 = 3.010 kΩ and R 3 = 72.36 kΩ (equations 7, 9, 10 and 11).

Example 5: According to Section 4, the following is valid at R 1 = R 2 = 10 kΩ (as in example 2): R 3 = 333.3 kΩ, C 1 = 3.98 nF and C 2 = 5.97 nF (equations 14, 15 and 16).

Example 6: When R 2 is deleted, and thus when equations 17, 18 and 16 are used, the following will result: R 3 = 1 MΩ, C 1 = 3.979 nF and C 2 = 995 pF as long as R 1 is assumed to be 10 kΩ. The three different groups of components calculated in this manner are considerably different, but the electrical characteristics (f_o = 800 Hz, Q = 4, G = 20) remains valid in all cases (see the measured results in Section 9). This means that several different possibilities present themselves in order to achieve the same results. However, one will usually calculate the filters according to Section 3, since it is possible for the values of the two capacitors to be given.

8.5. IMPEDANCE TRANSFORMATION

Example 7: The calculation in Section 8.4. with C 1 = C 2 = 22 nF resulted in very low-impedance values. In order to ensure that the previous circuit is not loaded excessively, it is advisable for R 1 to be greater than 10 kΩ. This means that the capacitance values must be reduced and the resistors increased. If the values of C 1 = C 2 = 3.3 nF is selected, a conversion factor of 22 to 3.3 = 6.6667 will result, which means that the following resistance values result: R 1 = 12.06 kΩ, R 2 = 20.066 kΩ and R 3 = 482.4 kΩ. It is sufficiently accurate for a value of 12.0 kΩ to be selected for R 1, 20 kΩ for R 2, as well as providing a series connection of 470 kΩ and 12 kΩ for R 3.

Example 8: The values of example 5 in Section 8.4. are now to be used for a further example: R 1 = R 2 = 10 kΩ is valid. The determined values of C 1 and C 2 are rather unsuitable, but have a ratio of 1 : 1.5 to another. It is thus possible for more favourable values for these capacitors to be found by impedance transformation. These can be standard values of 1 nF and 1.5 nF or 2.2 nF and 3.3 nF.

Since the value of 333.3 kΩ is rather high for R 3, the impedance transformation is to be calculated with C 1 = 2.2 nF and C 3 = 3.3 nF. This means that the conversion factor is 3.98/2.2, or 5.97/3.3 which results in 1.8091 in both cases. This means that R 1 = R 2 = 18 kΩ is sufficiently accurate, whereas a series connection of 470 kΩ and 130 Ω provides sufficient accuracy for the required R 3 = 603 kΩ.

9. SUMMARY AND MEASURED VALUES

The following table summarizes the values determined in the previous Sections and shows the values measured in a practical circuit.

Example	1	2	3	4	5	6	7	8
R 1	18.1 kΩ	10 kΩ	10 kΩ	1.81 kΩ	10 kΩ	10 kΩ	12.1 kΩ	18 kΩ
R 2	1.2 kΩ	10 kΩ	-	3.01 kΩ	10 kΩ	-	20.1 kΩ	18 kΩ
R 3	72.3 kΩ	595 kΩ	735 kΩ	72.4 kΩ	333 kΩ	1 MΩ	482 kΩ	603 kΩ
C 1	22.0 nF	1.59 nF	1.59 nF	22.0 nF	3.98 nF	3.98 nF	3.3 nF	2.2 nF
C 2	22.0 nF	8.35 nF	3.37 nF	22.0 nF	5.97 nF	995 pF	3.3 nF	3.3 nF
Q nominal	4	4	4	4	4	4	4	4
Q actual	3.9	4.02	3.84	3.96	4.02	3.94	4.04	3.94
G nominal	2	50	50	20	20	20	20	20
G actual	1.94	49.5	49.7	19.8	19.9	19.8	19.9	19.8

As can be ssen from these values, the deviations between the required nominal value and the actual measured value is always in the order of 1 % or less. The exception to this is the Q of example 3; however, even this value possesses only a deviation of 4%. Resistors and capacitors with a tolerance of ± 1% were used during all measurements, as well as an operational amplifier type 741 C, which was known to possess the values given in the data sheet. It is permissible for many applications that larger tolerances exist with respect to Q, centre frequency and gain so that larger component tolerances are acceptable. However, the tolerances should be kept as low as possible. If considerable differences exist between the calculated and measured values, it is advisable

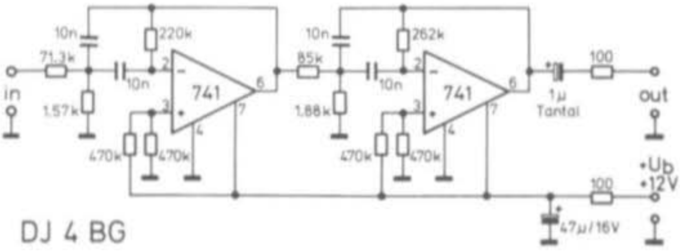

Fig.2a: Circuit diagram of a three-stage filter

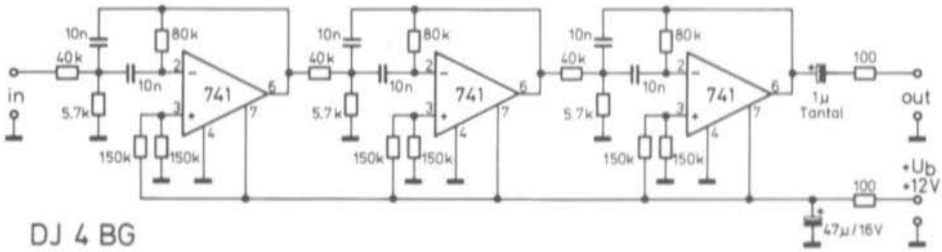

Fig.2b: Two-stage filter with staggered centre frequencies

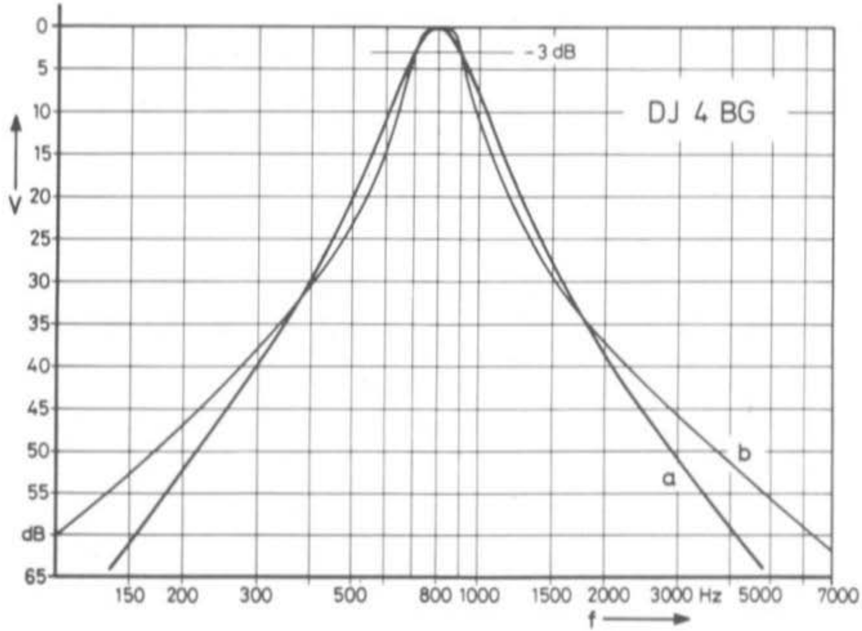

Fig.3: Selectivity curves of the filters shown in Fig.2a or 2b

to check the actual tolerances of the components used in the circuit. The weakest link in the chain is always the operational amplifier, whose input impedance is possibly too low, or whose non-load gain is too low at the operating frequency.

10. SERIES CONNECTION OF SEVERAL FILTERS

The selectivity of a single-stage filter will probably not be sufficient for a large number of applications since the roll-off of gain only increases by 6 dB per octave at a certain spacing from the centre frequency. This results in the typical triangular characteristic of a single resonant circuit. The ultimate attenuation is increased by connecting two or more such filters in series. These filters can be designed for the same centre frequency, or be staggered so that the required filter curve results. The following should, however, be observed.

In the case where several filters having the same centre frequency are connected in series, the individual circuit must possess a wider bandwidth than the required -3 dB bandwidth. When two such filter circuits are connected in series, the common -3 dB bandwidth will coincide with the -1.5 dB bandwidth of the individual filter; and to the -1 dB bandwidth when three filters are used. The characteristic of this filter in the passband range is always rounded so that a lower degree of ringing is present with pulse-type signals than when using filters with staggered centre frequencies. It is possible using this type of circuit to obtain a selectivity characteristic similar to a rectangle (flat top); however, the calculation is far more complicated, and the ringing in the case of pulse-type signals such as interference pulses is greater. However, the extent of circuitry is less for a given selectivity.

Informative details were given in (6) for experiments with filters where the individual circuits possess a staggered centre frequency. Two filters were constructed in order to compare the information given above. Each filter possesses roughly the same ultimate attenuation, centre frequency and -3 dB bandwidth.

Figures 2a and 2b show the circuits used, and the selectivity curves are given in Figures 3a and 3b. It will be seen that the two-stage filter is only inferior at attenuation values of over 30 dB when using the selected values. The transient behaviour of both filters is shown in Figures 4a and 4b (drive: 1.7 V, squarewave, 25 Hz; scale: 0.1 V/cm and 50 ms/cm). It will be seen that the filter with the staggered centre frequencies is somewhat less favourable with respect to pulse-handling ability, but only requires two stages to obtain nearly identical characteristics. On the other hand, the other filter with identical centre frequencies requires three stages, but possesses a better pulse behaviour.

11. UNIVERSAL BOARD FOR THE CONSTRUCTION OF ACTIVE BANDPASS FILTERS

A universal PC-board was designed to allow experimentation and the design of active RC-filters for various applications. It is also used as part of the system board DJ 4 BG 016 which was described in (7). This board allows up to three individual filter circuits to be connected in series. A drawing of PC-board DJ 4 BG 015 and the component locations are given in Figure 5.

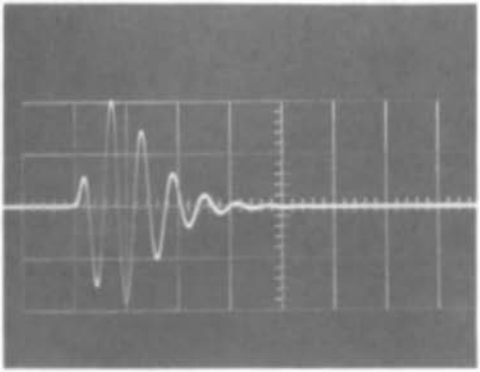

Fig.4a: Transient behaviour of the three-stage filter

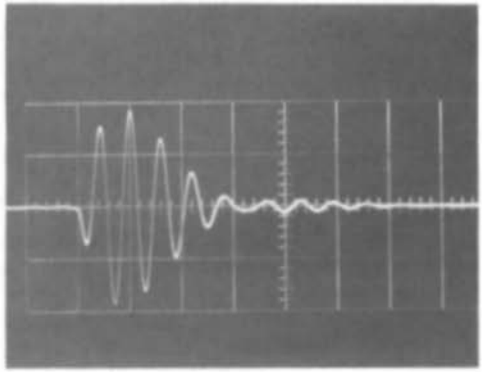

Fig.4b: Transient behaviour of the two-stage filter with staggered centre frequencies

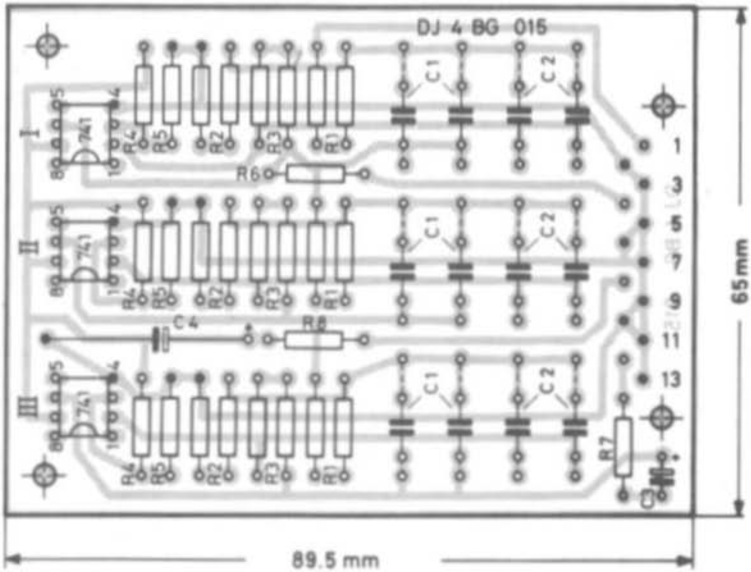

Fig.5: Component locations on PC-board DJ 4 BG 015

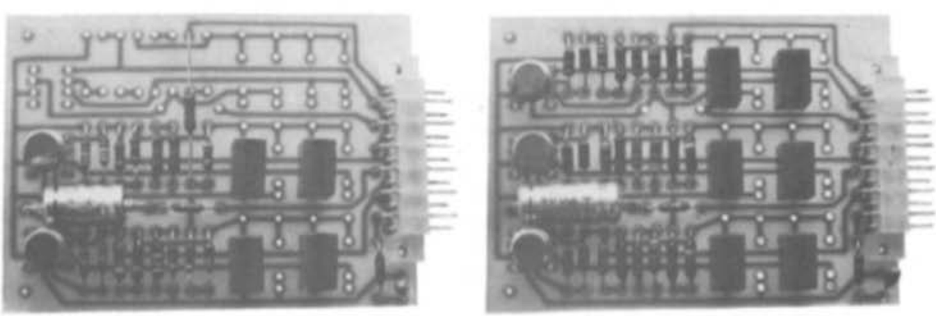

Fig.6: Author's prototype of a two-stage and three-stage filter

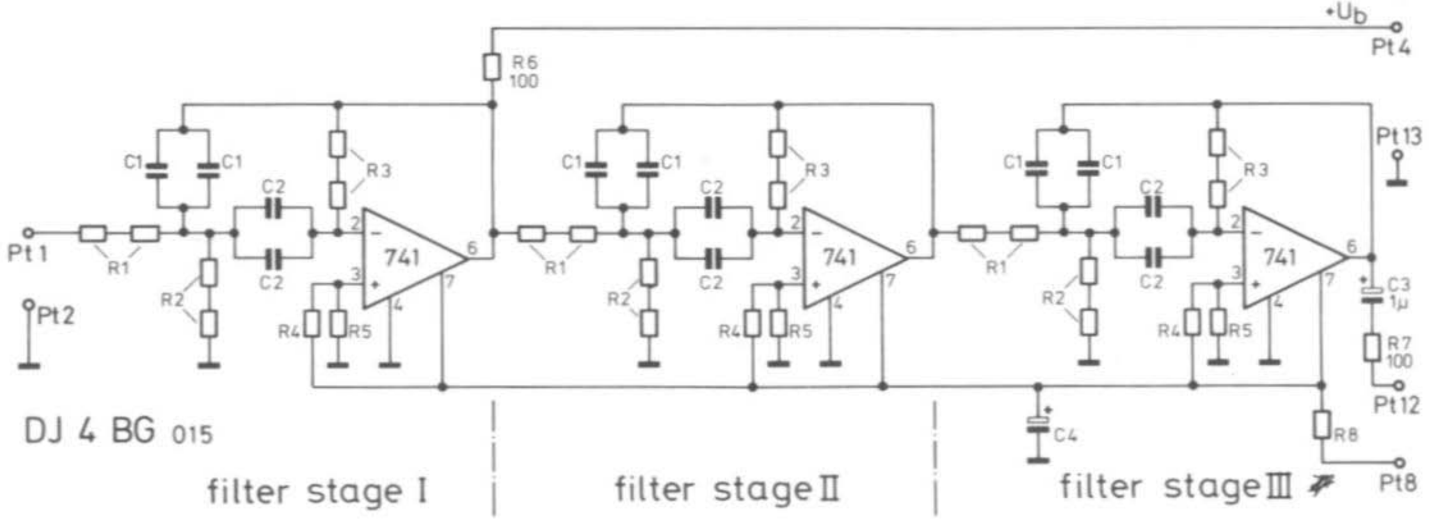

Fig.7: Circuit diagram of a filter using the maximum complement on PC-board DJ 4 BG 015

In order to allow non-standard values to be obtained, two places have been provided for each frequency-dependent capacitor and resistor so that capacitors can be varied by parallel connection, and resistors by series connection so that the required values can be obtained. A spacing of 12.5 mm has been provided for the resistors, whereas capacitors of 7.5, 10, 12.5 and 15 mm spacing can be used.

In order to avoid confusion due to differing component numbering, the same designations have been used in the circuit diagram (Fig. 6) and component location plan (Fig. 5) that were used in the theoretical part. This means that they are not numbered from front to back, as is usually the case, but commence again for each filter stage (I, II and III) as R 1, R 2, R 3, C 1 and C 2. This is not important when constructing multi-stage filters having the same centre frequency since each stage will possess the same component values in the same place. It is only important that care is taken when constructing filters with staggered centre frequencies (see components in Fig. 2b). Resistors R 4 and R 5 are used for adjusting the operating point, since the operational amplifier is not fed with two operating voltages as is usually the case (e.g. ±15 V). For this reason, it is necessary for half the operating voltage to be connected to connection point Pt 3. The values of resistors R 4 and R 5 are therefore equal and approximately twice the value of R 3. This is especially important at larger values of R 3. If the value of R 3 is relatively low, its effect on the DC operating point of the operational amplifier will be correspondingly low so that the values of R 4 and R 5 need not be lower than approximately 10 kΩ. Resistors R 6 and R 7 are used for decoupling and to avoid any tendency to self-oscillation of the operational amplifier when loading the outputs capacitively, e.g. by long screened cables. Values between 50 Ω and 200 Ω are usually sufficient. Resistor R 6 is only required when the output after the first filter stage (Pt 4) is used.

The components R 8 and C 4 are used for filtering the operating voltages in order to reduce the coupling to any other stages. A capacitance of at least 47 μF should be chosen for C 4, but 100 μF would be better. The value of R 8 is dependent on the operating voltage which should not be less than 12 V. Resistor R 8 should not be much higher than 100 Ω so that the voltage drop across it remains sufficiently low.

In the modular receive system using the system board described in (7), module DJ 4 BG 015 is mainly used as CW-AF filter so that it is possible to use a SSB receiver also for CW reception. It is, of course, also possible to construct a special receiver for CW-reception with a fixed bandwidth of, for instance,500 Hz (crystal filter XF-9M) and to switch in an audio bandpass filter under interference conditions. As an extension to the application described in (7) it was decided to provide the possibility of tapping off the signal after the first filter. The reason for this is that when a receiver is used where the whole selectivity is made in a crystal filter at the input of the IF amplifier, it is possible for wideband noise to be generated in one of the subsequent stages which can reduce the signal-to-noise ratio. Module DJ 4 BG 015 can be used for the first filter stage in order to reduce this wideband noise. In the case of a special CW receiver, one will provide somewhat more than the bandwidth of the IF amplifier, e.g. 500 Hz to 1 kHz in the case of a 500 Hz bandwidth, whereas corner frequencies of approximately 300 Hz and 3 kHz will be selected as corner frequencies in the case of a combined receiver.

The PC-board for the active audio filter is built-up in a similar manner to the TEKO-box modules and has been designated DJ 4 BG 015. The filter described in Section 10 has been constructed using this board; Figure 7 shows photographs of the author's prototypes.

12. TABLES

The following tables are provided for a centre frequency of 800 or 1000 Hz when G = 1 and C 1 = C 2 = 10 nF. This is provided to simplify calculation for the user of PC-board DJ 4 BG 015. The components are given for various Q-values and give the appropriate bandwidths. The designation B (1) is the bandwidth of a single filter stage, whereas B (2) or B (3) gives the bandwidth of a series connection of two or three identical filters.

Table 1: f_o = 800 Hz, G = 1, C 1 = C 2 = 10 nF

Q	0.707	0.8	1	1.35	1.5	1.65	2	2.5	3	4
R 1 (kΩ)	14.065	15.9	20	27	30	33	40	50	60	80
R 2 (kΩ)	∞	57	20	10.2	8.55	7.4	5.7	4.35	3.53	2.58
R 3 (kΩ)	28.13	31.8	40	54	60	66	80	100	120	160
B(1) (Hz)	1135	1000	800	590	530	485	400	320	267	200
B(2) (Hz)	725	640	512	378	339	310	256	205	171	128
B(3) (Hz)	580	510	408	301	271	247	204	163	136	102

Table 2: f_o = 1000 Hz, G = 1, C 1 = C 2 = 10 nF

Q	0.707	0.8	1	1.35	1.5	1.65	2	2.5	3	4
R 1 (kΩ)	11.3	12.8	16	21.6	24	26.4	32	40	48	64
R 2 (kΩ)	∞	45.6	16	8.15	6.83	5.92	4.56	3.48	2.82	2.06
R 3 (kΩ)	22.6	25.6	32	43.2	48	52.8	64	80	96	128
B(1) (Hz)	1420	1250	1000	738	662	606	500	400	334	250
B(2) (Hz)	910	800	640	472	424	388	320	256	214	160
B(3) (Hz)	725	638	510	376	339	309	255	204	170	128

Example: required are the components for a three-stage filter with a centre frequency of 1000 Hz and a bandwidth of 300 Hz. This is obtained by quoting the 1000 Hz table in row B (3) to find the nearest value to 300 Hz. This is 309 Hz in the column Q = 1.65. This means that all component values are known, the resistance values are given in the table, and the capacitance values were given. This example was made using Table 2. Table 1 gives the components for the three-stage filter described in Section 10.

13. ADDITIONS TO SECTION 4

For clarity, it should be noted that an error was made in the determining of the transition range between the two methods of calculation given in Section 3 and 4: Equation 16 was referred to in the fourth line of page 18 of Edition 1/1975 of VHF COMMUNICATIONS. This was a printing error since reference should have been made to equation 17. However, the limit derived in this manner is only valid for the special case of R 2 = ∞. Equation 14 is more general, which shows that the limit is at Q^2 x Rp/G x R 1 or when transposed, at Q = G x R 1/Rp.

This means that the range of validity for the equations given in Section 4 is greater to the value of $\sqrt{R\,1/Rp}$. This means that they are valid at

$$Q \leq \sqrt{G \frac{R\,1}{Rp}}$$

In the special case of $R\,2 = \infty$ and thus $R\,1 = Rp$, the result will be the previously mentioned limit of $\sqrt{G}$.

As was mentioned in the introduction to this article (Section 1), the equations given in the references were classed as given. They were given in the clearest possible manner and methods of simplification were also given. The examples calculated according to these equations and the comparison to the practical circuits have shown that they can be used with success.

One disadvantage does exist in that when calculating according to Section 4 ($Q \leq \sqrt{G/2}$): it is not possible to give the capacitors. When considering equation 16, it will be seen that this can be transposed easily so that:

$$R\,1 = \frac{Q}{G \times C\,1 \times \omega_o} \text{ is obtained from } C\,1 = \frac{Q}{G \times R\,1 \times \omega_o}$$

This is, of course, equation 5 from Section 3. This means that at least one of the two capacitors can be given and an available, standard type used.

The calculation can then be made according to equations 5, 14 and 15a.

As has been mentioned in the previous sections, simplifications can also be made here, e.g. by selection of $G = 1$, or $R\,2 = \infty$, or both.

14. REFERENCES

(6) K. Schmidt: Berechnungsbeispiele für aktive NF-Bandpässe
Funkschau 1974, Edition 6, Pages 187-190

(7) E. D. Schmitzer: A System Board for the TEKO-Modules
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 220-229.

STACKED TUBULAR SLOT ANTENNA FOR THE 23 cm BAND

by Gerd Körner, DK 2 LR

A 23 cm antenna is to be described that can be manufactured from a copper or brass tube of 90 cm in length and approximately 7 cm (uncritical) in diameter. Four vertical slots are sawn into the tube that operate as four dipoles and reflectors (Fig. 1). The roughly $\lambda/2$ slots in the metal tube represent "negative" dipoles; they generate horizontal polarization. The stacking of four such slots provides a gain of approximately 6 dB over a dipole due to the reduction of the vertical beamwidth. The horizontal diagram can be altered by tuning the reflector slots. The variation range is from a virtually omnidirectional characteristic (A) to a somewhat reduced radiation to the rear as shown in (B) in Figure 2. The characteristics of this antenna make it extremely suitable for beacons and for local contacts.

1. CONSTRUCTION

It is possible that one may be able to purchase a suitable tube from a local building source. The diameter is not too critical and can be between 70 and 80 mm. If the antenna is to be constructed from metal plate, it is advisable to cut the slots before bending. The dipole and reflector slots are mounted 180° from another and are opposite another on the circumference of the tube. Their dimensions are identical. The phase difference between the two elements is obtained afterwards with the aid of the feed of the dipole and the capacitive tuning of the reflector.

The feeder and transformer lines are made from 1 mm thick copper or brass plate, and are 3 mm and 6 mm wide respectively. They are soldered to one edge at the centre of each of the four dipole slots, after which they are bent up and fed parallel to the inner surface of the tube with a spacing of approx. 9 mm.

The reflector slots are capacitively tuned with the aid of metal discs whose spacing from the centre of the slot can be altered with the aid of adjustment screws. Figure 3 shows a suitable arrangement: The four 30 mm x 30 mm square plates are glued onto a piece of plastic with rectangular crossection. The spacing of this plastic piece to the reflector slot can be adjusted with the aid of the spring-loaded screws.

After completion, the antenna should be sealed at each end with the aid of a plastic disc. It is advisable for the antenna to be completely enclosed in a plastic cover for weather protection.

2. MATCHING

Figure 4 gives a drawing of the feeder and stacking lines: Since the slot dipoles are part of the ground surface, the feeder lines are directly connected asymmetrically as striplines and fed beside the slots along the inner surface of the tube. The slots are connected in groups of two. An impedance of approximately 600 Ω is present at the centre of each slot. This impedance is reduced to approximately 300 Ω by the interconnection of two such slots. The two groups of two dipoles are connected once again in parallel so that an impedance of approximately 150 Ω results. This impedance is then transformed to 50 Ω with the aid of a π-link comprising two trimmers of approx. 5 pF and the line T.

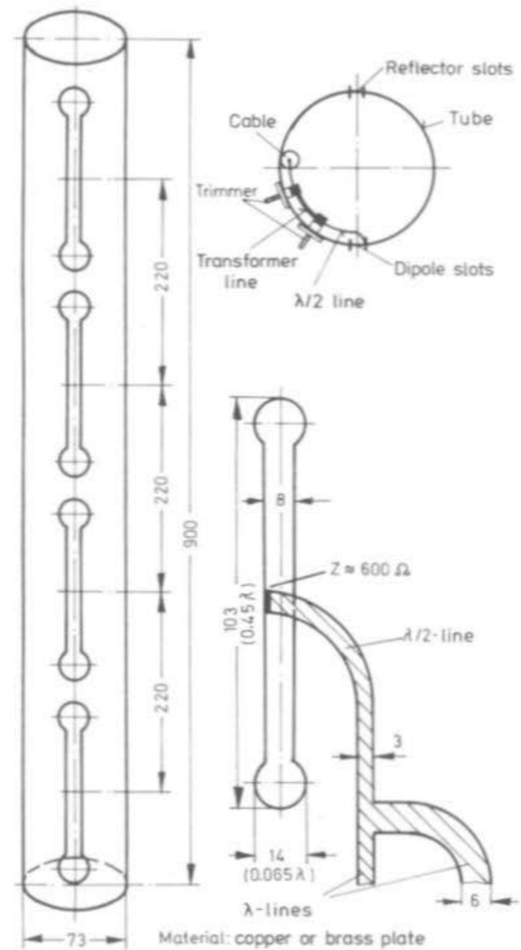

Fig.1: Dimensions of tube and slots

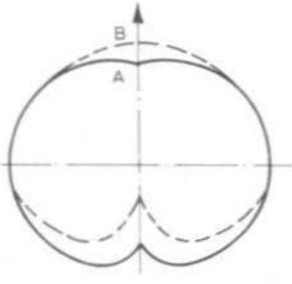

Fig.2: Horizontal polar diagram

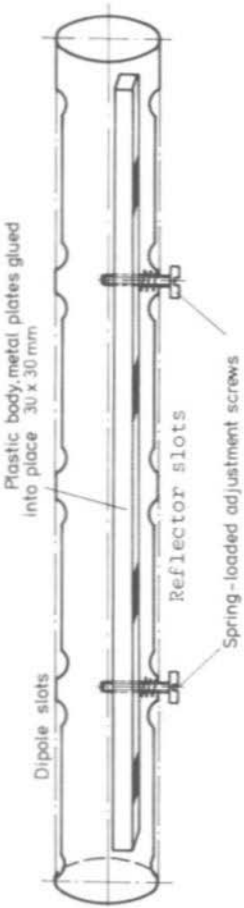

Fig.3: Tuning arrangement for the reflector slots

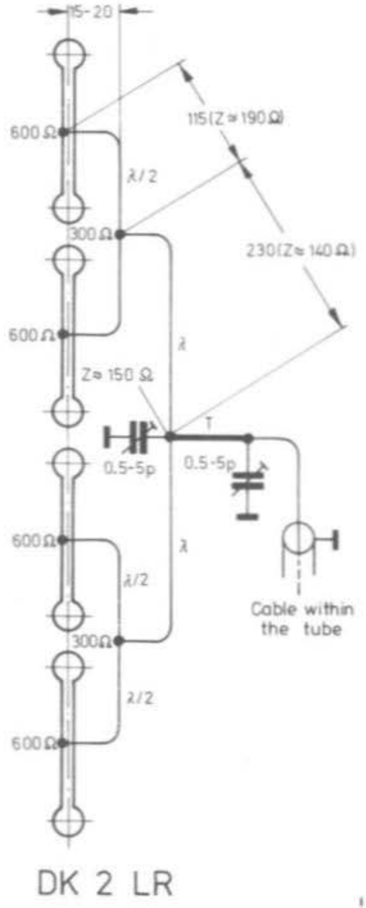

Fig.4: Matching of the dipole slots

The interconnection of the slots and groups of slots is made without transformation using $\lambda/2$ lines. The impedance of these lines should be a mean value between the two values present at each end; however, it does not seem to be critical (in contrast to the length of the lines). The selected stripline width results in an impedance of approximately 190 Ω for the 3 mm width when spaced 9 mm from the inner surface, and approximately 150 Ω for the 6 mm wide line. The transformer line T is 9 mm wide and 35 mm long (0.15 λ) and possess an impedance of approximately 90 Ω.

The whole matching system is only held by the four soldered connections at the centre of the slots and by the two trimmers. No further supports are used.

A virtually omnidirectional horizontal characteristic (A) is obtained by deleting the capacitor plates from the reflector slots. The characteristic B is obtained with a spacing of approximately 3 mm.

3. REFERENCES

(1) Karl Rothammel: Antennenbuch, 4th Edition, Page 416.

A FOUR-ELEMENT YAGI ANTENNA FOR THE 23 cm BAND USING A STRIPLINE BALUN

by B. Lübbe, DJ 5 XA

A four-element yagi antenna is to be described for the frequency band in the order of 1300 MHz. It indicates an application of the stripline balun transformer described in (1). The gain of the antenna is in the order of 6 dB, and is therefore suitable for test, portable and local operation. The author required a simple method of connecting and matching the antenna to 50 Ω coaxial cables. As can be seen in Figure 2, the use of the stripline balun DJ 5 XA 002 a/b (Fig. 1) allows the antenna to be matched extremely well without measuring equipment.

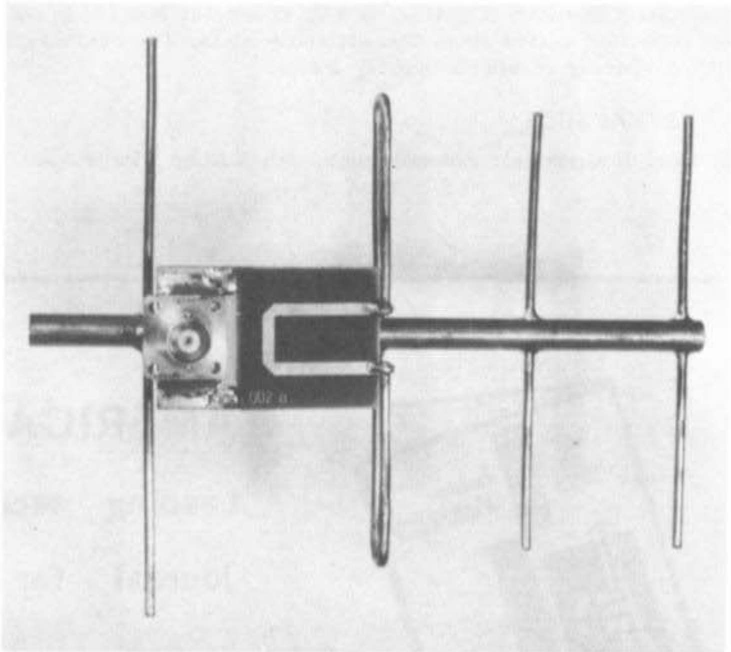

Fig. 1:

Photograph of the author's prototype

The antenna is constructed from a brass boom of 7 mm diameter, thick copper wires. The elements are placed through holes drilled in the boom and soldered into place (Fig. 3). PC-board DJ 5 XA 002 b is provided with a 7 mm diameter hole for a BNC connector at the end of the printed conductor, which is then connected to the inner conductor of the connector with the aid of a short wire. The flange of the connector is soldered to the ground surface of this PC-board. PC board DJ 5 XA 002 a is shortened so that sufficient room is available for the flange of the connector. The parts are joined together as was described in (1) and are soldered to the folded dipole using two short metal strips. A spacer is soldered between the boom and the flange of the connector to provide the required support.

REFERENCES

(1) M. Münich and B. Lübbe: Six-Element Colinear Antenna with Reflector Plate for the 24 cm Band Using a Stripline Balun
VHF COMMUNICATIONS 6, Edition 2/1974, Pages 85-88.

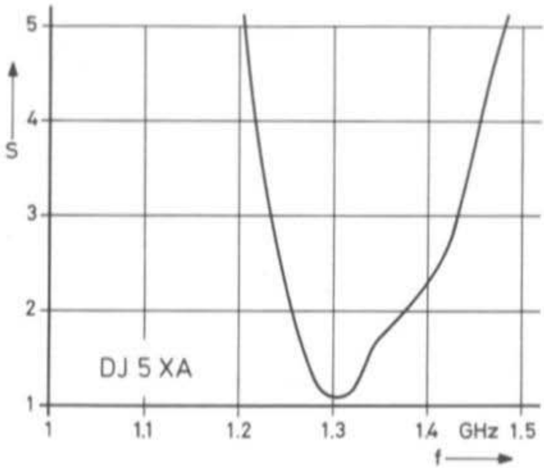

Fig. 2: Standing wave ratio s of the prototype antenna

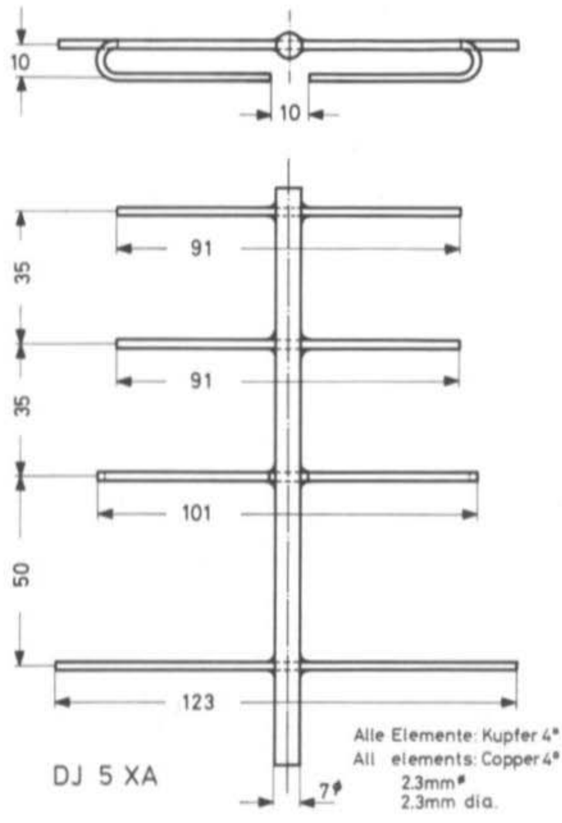

Fig. 3: Dimensions of the four-element yagi antenna constructed by DJ 5 XA

A 40-ELEMENT COLINEAR ANTENNA FOR 23 cm

by G. Körner, DK 2 LR

The described colinear antenna possesses 40 $\lambda/2$ elements mounted in front of a reflector panel. The description is to describe some interesting solutions for some of the problems encountered using such antennas (Fig. 1).

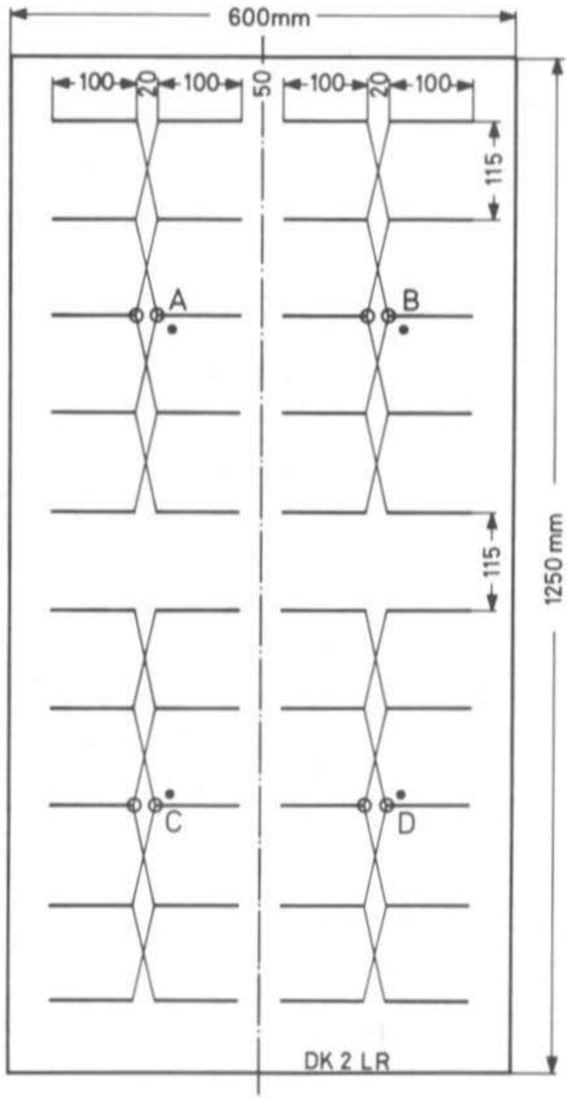

Fig. 1: Dimensions of a 40-element colinear antenna for 23 cm

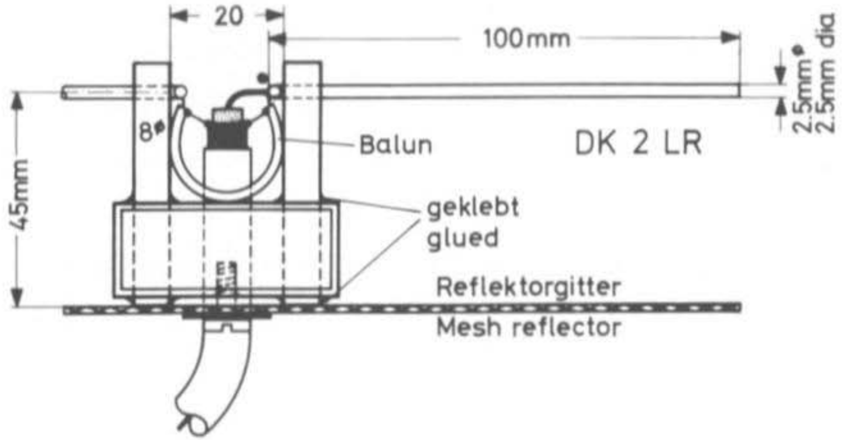

Fig.2: Construction of dipole elements, supports, framework, reflector and balun

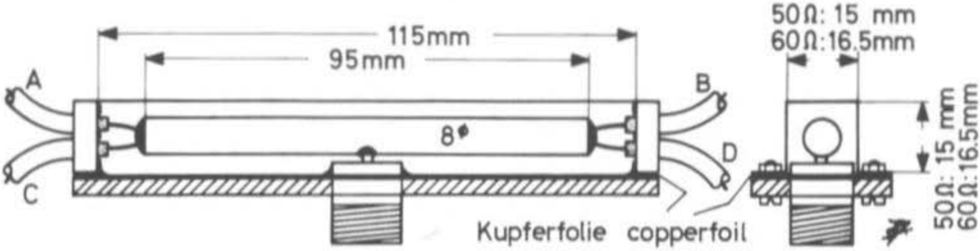

Fig.3: Double transformation line

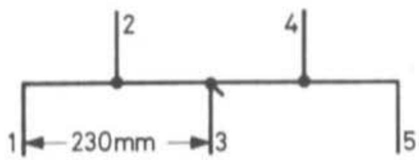

Fig.4: Half of a radiator sub-group

Fig.5: Bending of the interconnection cables

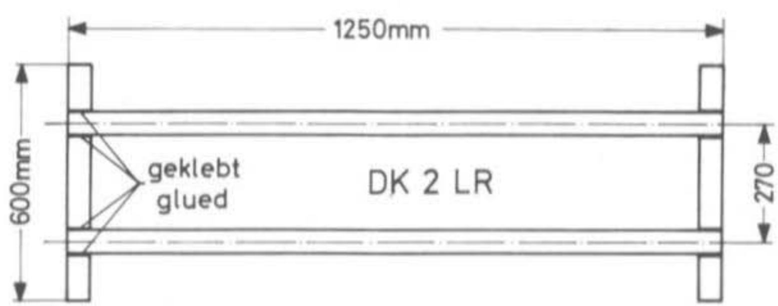

Fig.6: Framework of PVC boxes

The gain of the antenna is in the order of 16 to 18 dB. Electrically speaking, the antenna consists of four sub-groups of 10 elements each. Each sub-group is fed with the aid of a balun transformer located at the centre dipole. The balun transformer is best made using teflon coaxial cable having copper tubing of 4 mm in diameter as outer conductor. It is, of course important that the four baluns are connected identically at the same phase position, e.g. that all inner conductors, for instance, are connected to the right-hand dipoles which are marked with a point in the diagram (Fig. 1 and Fig. 2). The four cables from the sub-groups are electrically 2.5 λ in length (57.5 cm times velocity factor of 0.66 or 0.84). The impedance of these cables should amount to 50 Ω. Since they are not terminated with the correct impedance, it is important, that low-loss cable be used and that an electrical length of a multiple of $\lambda/2$ is accurately maintained.

The resulting impedance of 43 Ω from the four antenna groups is transformed to the required impedance of the feeder cable in a double $\lambda/4$ transformer as shown in Figure 3. The inner conductor is 8 mm in diameter, and the inner diameter of the square outer conductor is 15 mm for 50 Ω feeders and 16.5 mm for 60 Ω cables. Copper foil is used to interconnect the ground connections between the outer conductors of the feeder cables to the sub-groups and the output connector for the feeder.

Each of the four sub-groups consists of two rows of five elements each, together with the required interconnection lines (Fig. 4). These parts are all made from the same material (2.5 mm diameter copper wire). It is constructed by firstly making elements 1 and 5 including the whole interconnection cable. After this, elements 2, 3 and 4 are soldered into place. Element 3 is extended slightly and bent so that a connection is provided for connecting the cable or balun.

Two such groups of elements are now mounted together to form the sub-group and the interconnection cables should be bent in a semi-circle at all cross-over positions (Fig. 5). This bending means that the vertical spacing between the elements is somewhat less than 115 mm. It is therefore advisable for the framework and supports to be constructed after completion of the groups of radiators.

The reflector is made from aluminum mesh. The reflector mesh is mounted onto a frame of PVC boxes of 20 mm x 40 mm x 1 mm (cable ducting). The PVC supports for the dipole elements are also glued to these PVC boxes. The framework, dipoles, four baluns and the four stacking cables then form a unit. The reflector can be screwed to the lower part of the framework using self-tapping screws. The spacing between reflector and the centre of the elements amounts to 45 mm.

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M. Münich, B. Lübbe: A 6-Element Colinear Antenna with Reflector Plate for the 24 cm Band Using a Stripline Balun
VHF COMMUNICATIONS 6, Edition 2/1974, Pages 85-88.

MODIFICATIONS TO THE ATV TRANSMITTER DJ 4 LB

by P. A. Johnson, G 8 EIM / G 6 AFF/T

1. MODIFICATION TO MODULE DJ 4 LB 004 (1) (2)

The author would like to offer some improvements to this very successful ATV transmitter. Many of the ATV constructors in Great Britain have encountered some difficulty in suppressing the 478 MHz local oscillator signal from the side-band video signal at 436 MHz. The following modification solves this problem and provides a further attenuation of the unwanted lower sideband and other spurious signals.

PC-board DJ 4 LB 004 is cut at the position shown in Figure 1. This separates the mixer and linear amplifier sections, and allows a helical filter to be used after the mixer. A tighter coupling to inductance L 401 allows the loss of the filter to be compensated for. The filter which was originally described in (3) is given in Figure 2. The output impedance of the filter is 75 Ω, and a suitable matching to transistor T 405 is provided by tapping inductance L 402 half a turn from the cold end.

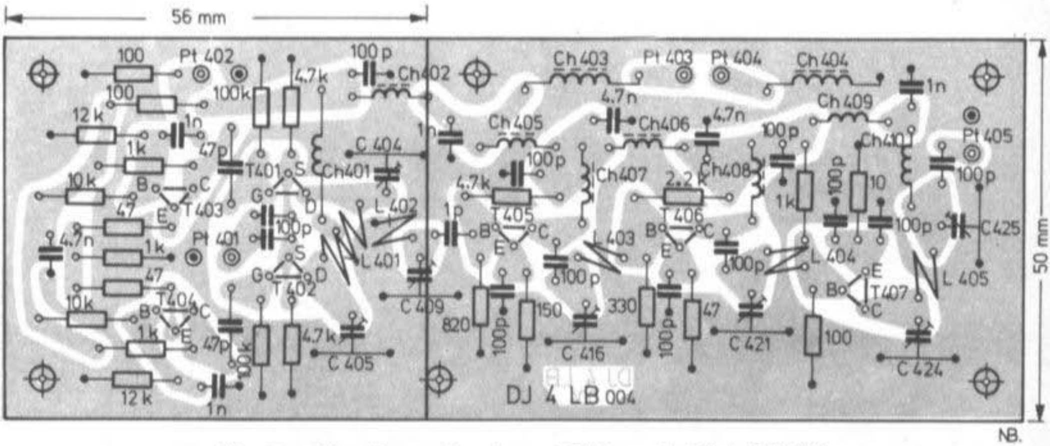


Fig. 1: Position of cut on PC-board DJ 4 LB 004

A further stage of amplification should be provided at this position (Fig. 1). The simplest method of doing this is to duplicate the circuit of transistor T 405 (Fig. 3) and mounting this on a small extra board, which is then soldered to the larger part of PC-board DJ 4 LB 004 (Fig. 5). It is now important that the linear amplifier section is housed in a separate box to the mixer and filter.

No undesirable effects have been noticed since making these two modifications. The suppression of spurious signals is extremely good, and TVI problems were cured, even when using the 10 W linear amplifier DJ 3 SC 001 (4). The use of the helical filter also simplifies the alignment procedure.

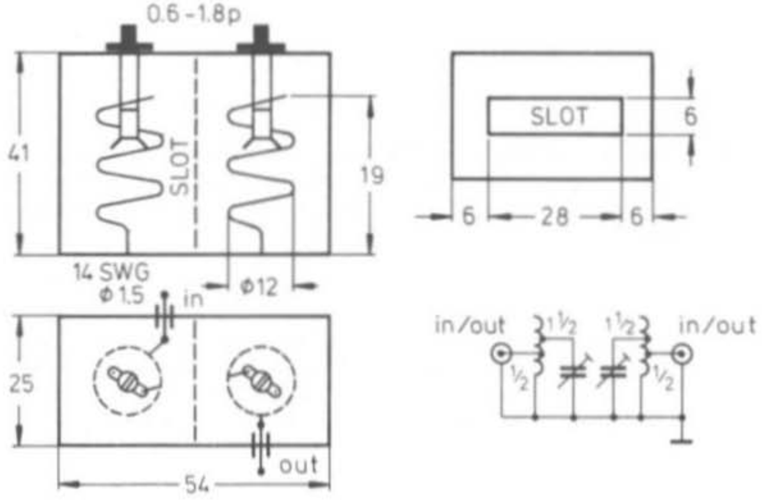


Fig.2: Helical filter for the 70 cm band

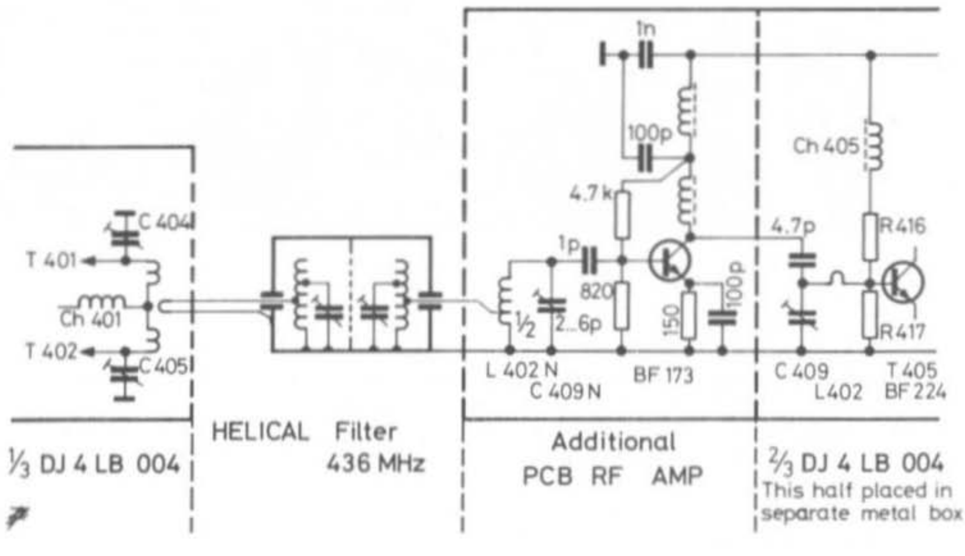


Fig.3: Interconnections between the various stages

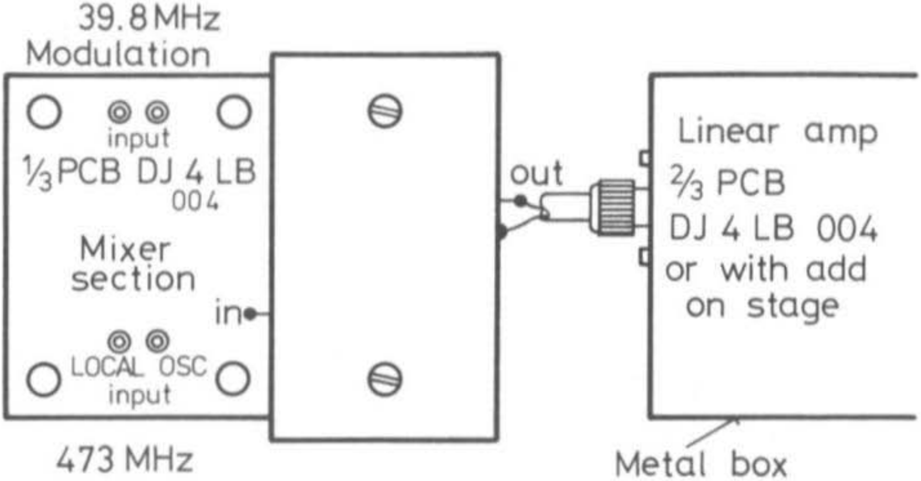


Fig.4: Mixer, helical filter, and linear amplifier sections

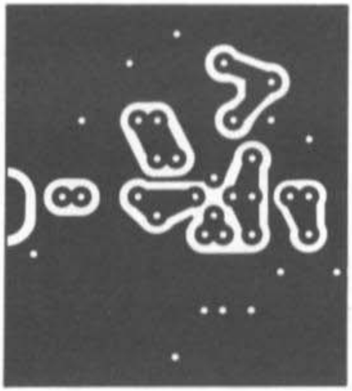
2. MODIFICATIONS FOR UK USE

ATV stations in Great Britain do not usually receive intercarrier FM sound but prefer to receive AM sound on the video carrier frequency. The following description shows how it is possible to obtain an AM sound carrier in the cheapest and easiest manner .

The sound IF module DJ 4 LB 002 is modified by adding a connection point Pt 20a at the junction of Ch 202 and R 201. The signal taken from this point is fed via a trimmer resistor to a transistor BC 108 as emitter-follower. This transistor provides an output impedance of 75 Ω which is suitable for driving the input of the video IF module DJ 4 LB 001 (Pt 101).

The mode selection is made using a four-pole, three-way switch which provides the following modes: AM sound, video only, video plus FM sound. Further details are given in the block diagram in Figure 6.

The author recommends a common UK video frequency of 433. 8 MHz with 6 MHz intercarrier spacing at 439. 8 MHz.



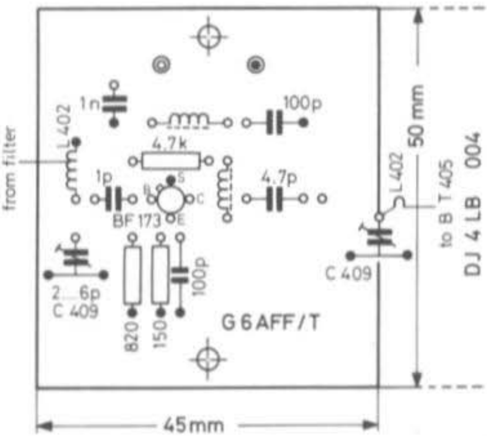


Fig.5: PC-board for the extra amplifier stage

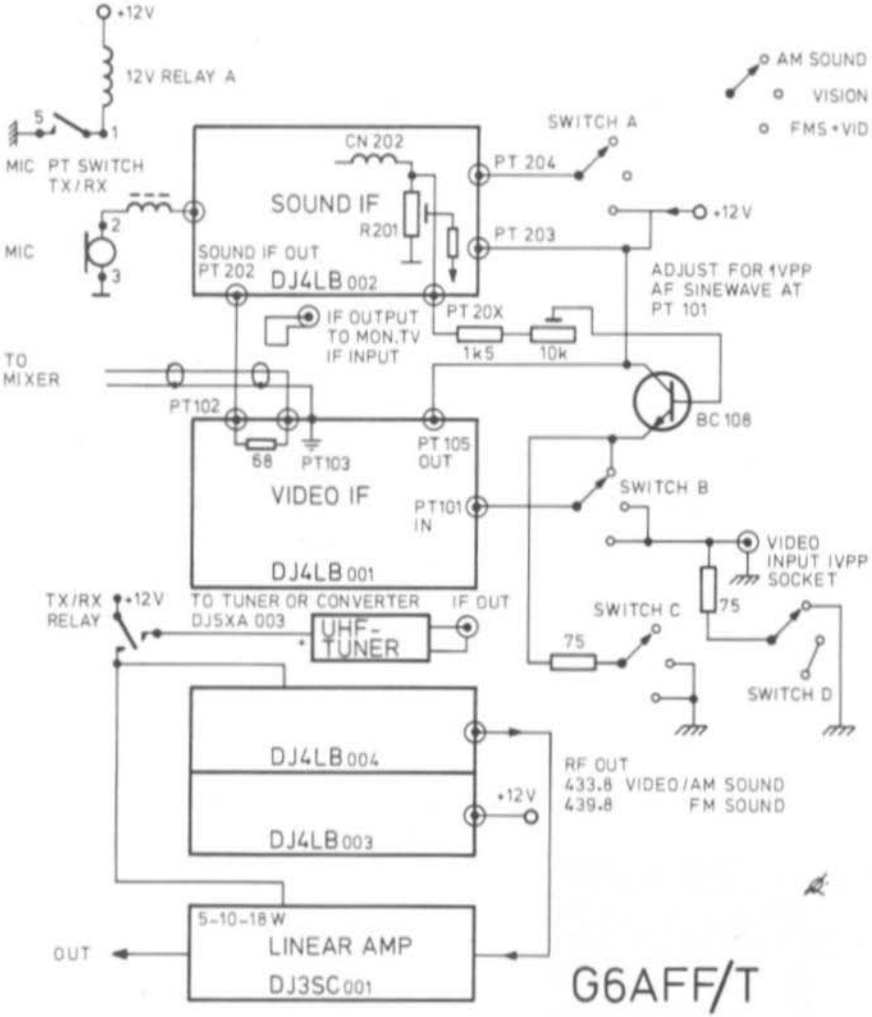


Fig. 6: Switching required to obtain the various modes

3. FURTHER NOTES

3.1. MODULE DJ 4 LB 001

The value of capacitor C 108 can be increased from 2.2 pF to 5.6 pF, as well as increasing the value of resistor R 104 to 1.8 kΩ. This results in a greater signal swing from the video IF-module DJ 4 LB 001. It is important that a high-gain transistor type BF 224 be used (ß 120), however, these are difficult to obtain.

A PAL colour signal was transmitted through the ATV transmitter after completing the modifications and no deterioration of the colour signal was observed. Both multiburst and sawtooth signals have been used to check the linearity.

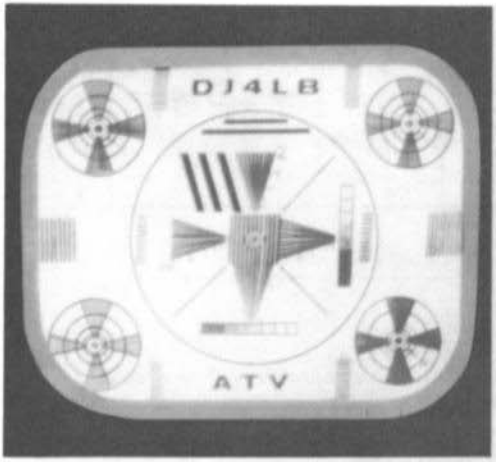
4. REFERENCES

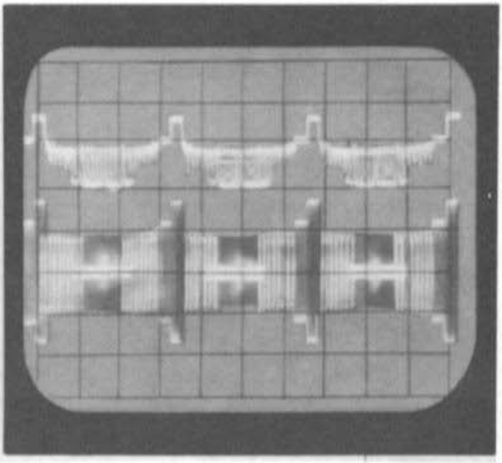
(1) G. Sattler: A Modular ATV Transmitter
VHF COMMUNICATIONS 5, Edition 1/1973, Pages 2-15

(2) G. Sattler: A Modular ATV Transmitter
VHF COMMUNICATIONS 5, Edition 2/1973, Pages 66-80

(3) VHF-UHF Manual
RSGB-Publications

(4) G. Freytag: A Transistorized Linear Amplifier
VHF COMMUNICATIONS 6, Edition 1/1974, Pages 30-37.





NOTES AND MODIFICATIONS

1. PROGRAMMABLE FOX-HUNT RECEIVER DL 9 FX 006/007 (Ed. 2/74)

1.1. ERROR ON PC-BOARD DL 9 FX 007 (Switching circuit):

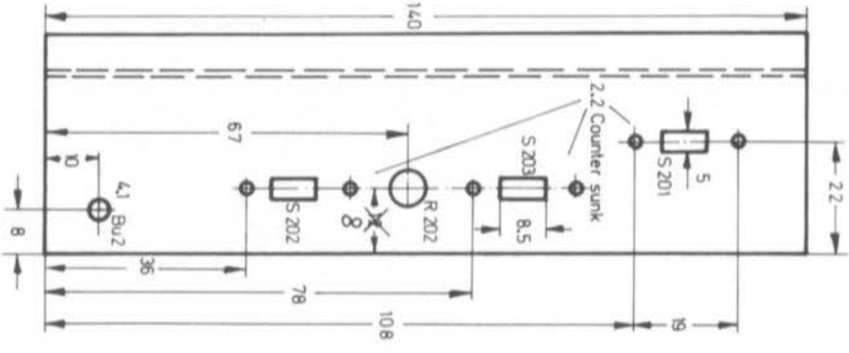
The first series of PC-boards possessed a drawing error which led to a short between connection R 209 E and the adjacent conductor lane. This should be removed. However, it is not to be expected that any of these boards were exported.

1.2. RECOMMENDED NEUTRALIZATION

Should any tendency to self-oscillation be noticed, it can be avoided by placing ferrite beads around the connection wires of transistors T 9 and T 10, and connecting capacitors of 4.7 nF between the emitters and collectors of these transistors. A capacitor of 10 nF can also be connected from pin 3 of the integrated circuit I 1 to ground. The demodulator transistor T 14 should be shunted by a capacitor of 10 pF in order to ensure that the S-meter indication is not affected.

1.3. CASE

The cutout for switch S 202 in Figure 8 should be only 8 mm from the edge of the case and not 12 mm as given in the original article. This is to ensure that there is sufficient room for the battery. The following drawing shows the new dimensions:



2. FM-IF MODULE DC 6 HL 007 (Ed. 3/72)

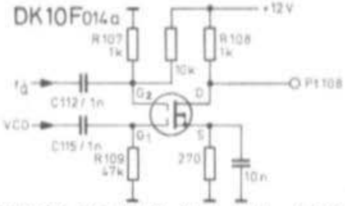
The value of the coupling capacitor C 703 of the first bandpass filter should not exceed a value of 1.8 pF. This is to ensure that the coupling is not too tight that a simple alignment to maximum is no longer possible.

3. PLL-OSCILLATOR DK 1 OF 011/014 (Ed. 2/74)

A number of readers have informed the publishers that the phase-locked loop possesses a tendency to self-oscillation,which indicates an additional integration time constant. On observing the internal circuit of the integrated frequency/ phase comparitor MC 4044 it will be seen that one output (Pin 10) is in the form of an open emitter. Pin 5 is connected to a diode that is in the pass direction for positive voltages. This means that the RF-bypass capacitor C 145 will practically not be able to discharge at voltages of less than approximately 1 V, which represents a very long time constant. A control loop with two or more integral links with identical or similar time constants will always possess a tendency to oscillation. In the case of the PLL-oscillator, this will be observed in the form of unstable spurious signals at both sides of the wanted signal.

This problem can be avoided by connecting a resistor of approximately 10 to 50 kΩ in parallel with C 145 to ensure that this capacitor is discharged sufficiently quickly. Attention should also be paid that the values of R 156, C 124 and C 149 are accurately maintained so that the resulting time constant is not too long.

It has been found in some cases that the output voltage of mixer T 106 is too low to fully drive the pulse shaper T 141. This can be improved by using a dual-gate MOSFET (40673 or similar) for T 106 in the following circuit:



Attention should be paid that only first choice ICs from wellknown manufacturers are used for the operational amplifier I 143. Second and higher choice ICs can produce considerable noise voltages at the output.

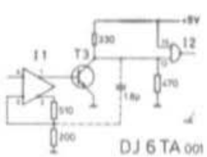
4. HIGH-IMPEDANCE COUNTER PREAMPLIFIER DJ 6 TA 001 (Ed. 3/74)

A few enquiries regarding this module have led to a few tips on the construction and fault finding:

4.1. The value of the current-limiting resistor between connection U_A and ground is too high (15 Ω), which would lead to a current limiting at approximately 45 mA. The value of this resistor should be reduced to approximately 6.8 Ω so that a current of 65 mA can be taken without current limiting.

4.2. Due to the tolerances of the zener diode D 5, the value of +1.8 V at pin 14 of I 1 can only be classed as approximate value. It is only important that transistor T 3 is fully conductive or blocked according to the input signal and threshold adjustment. This can be checked by measuring the collector voltage of T 3. With zener diodes with voltage values of over 5.6 V, the integrated circuit I 1 can become too warm so that the threshold is unstable.

4.3. In a few cases, oscillations have been observed on the switching slope which is indicated when higher frequencies can be counted correctly, but lower frequencies are indicated too high, and the indication is not stable. The cause of this is a somewhat longer switching time of some individual ICs type MC 1035 (more than 10 ns). If transistor T 3 has a very high gain, the combination will cause oscillations to be formed on the switching slope which will be converted to needle pulses of approximately 5 ns in I 2 (will only be observed on an oscilloscope having a sufficiently high bandwidth). The simplest method of removing this fault without exchanging components is for T 3 to be included in the feedback link of the Schmitt trigger for higher frequencies. This is achieved by soldering a disc capacitor of 1.8 to 2.2 pF from the collector of T 3 to pin 3 of I 1 (non-inverting input). The steep slope at T 3 now forces the Schmitt trigger to switch more quickly. This disc capacitor can be soldered to the lower side of the board and bent back. The following drawing indicates this modification.



A STANDARD FREQUENCY OSCILLATOR WITH AN ACCURACY OF 10^{-8}

by R. Görl, DL 1 XX

1. APPLICATION

Demands on the stability and absolute accuracy of frequency determining components and modules have increased considerably in the professional communications and measuring technology over the last 10 years. A similar tendency is also to be observed in amateur radio technology. Complicated modulation modes such as SSB, as well as new communication modes such as EME and MS operation require higher frequency accuracies, especially on the UHF bands. In addition to this, the advances made in the digital electronics technology have made advanced frequency measuring, and frequency comparison methods with numerical indication (frequency counters) also available to the radio amateur.

In most cases, the owners of digital frequency counters usually do not have a suitable, accurate frequency standard without which exact absolute measurements of the frequency are not possible. The continuous re-calibration of an inexpensive oscillator against a standard frequency transmitter is unfavourable in practice since this cannot always be made when required due to the propagation conditions. The increasing tendency to use frequency synthesizers in contrast to the use of a large number of crystals also tends to favour the use of a single crystal oscillator with a short warm-up time of less than 10 minutes and high repeatability which is then used as frequency standard for the complete station.

For the following reason, it is not favourable for a crystal oscillator to be operated continuously, and perhaps only used to drive a digital clock as a secondary mode: Electric clocks using synchronous motors are more accurate with respect to the long-term stability than an inexpensive crystal clock. This is because the AC-line frequency is monitored and controlled by crystal clocks having a higher accuracy. Furthermore, the aging of a crystal that is in continuous use is considerably higher than a well-aged crystal that is not oscillating continuously. Since a radio amateur only uses his frequency standard occasionally, maybe for a few hours, the required absolute accuracy can be maintained over a considerable period when one assumes that the aging rate during this short operating period probably amounts to $10^{-9} \frac{\Delta f}{f}$ per hour.

The described frequency standard was developed with this in mind. The price is extremely low when compared with the excellent characteristics and technical complement. This oscillator is not to be offered in a form of a kit since most amateurs would not be able to adjust and align the oscillator with respect to temperature behaviour, crystal loading and control characteristics due to the accuracy of the measuring equipment required. However, the unit is to be offered as a ready-to-operate module in fully aligned condition. Finally, several terms are to be described in order to understand the operation better. The physical magnitude, such as $\frac{\Delta f}{f}$/day for the aging rate; $\frac{\Delta f}{f}/^{o}C$ for the temperature coefficient, or $\frac{\Delta f}{f}$/V for the stability with respect to supply voltage fluctuations. It is therefore not possible for the accuracy of a frequency standard, to be completely specified by one value. It is only terms such as repeatability, absolute accuracy, etc. that can be given as $\frac{\Delta f}{f}$.

2. THE CRYSTAL

The crystal oven used in the described module is designed to work with crystals having a HC-6U or HC-36U holder. All fundamental crystals in the frequency range of 200 kHz to 20 MHz can be used in the circuit. The use of overtone, standard crystals is not planned since their advantages with respect to stability to variable circuit parameters are only advisable at far higher degrees of accuracy. This, however, requires a far higher degree of circuitry and temperature stability.

Generally speaking, it could be said that the higher the crystal frequency the greater the stability. The actual conditions are so that the stability of a crystal with respect to temperature fluctuations is mainly dependent on the crystal cut used. The individual cuts, however, can only be used for certain frequency ranges for economical reasons. The most stable frequency range with respect to temperature fluctuations when using a AT-cut is the frequency range from 800 kHz upwards. The characteristic temperature-to-frequency corresponds to a third-order parabola whose reversal point is in the order of 27 °C, and whose relatively flat minimum is especially suitable as temperature reversal point for use in conjunction with crystal ovens.

All other crystal cuts in this frequency range (200 kHz to 20 MHz) possess a temperature-to-frequency characteristic corresponding to a second order parabola. The temperature coefficient $\frac{\Delta f}{f}$/°C is considerably higher. An exception to this represents the very expensive GT-cut which are hardly manufactured nowadays.

A further and probably most important point is the crystal aging. This is mainly dependent on the care made during the manufacture of the crystal. The aging of the crystals is mainly observed as continuous increase of frequency, which never completely disappears even after months and years of operation. The real reasons of crystal aging are still not known completely; however, it is known that the surface of the crystal is damaged during the cutting process so that particles continue to be lost decreasing the mass of the crystal. The reason why a low crystal loading (low oscillating amplitude) is advisable can thus clearly be seen. Crystals with a low aging rate usually have a high Q (not always the case). This is also favourable since the feedback can be kept at a very low value.

The attainable accuracy as a function of Q has been found experimentally to be:

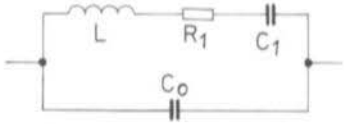
$$\frac{\Delta f}{f} \sim \frac{1}{1000\ Q}$$

This means that crystals with a Q of at least 100 000 must be used if an accuracy of 10^{-8} is to be obtained. When referred to the diameter of the AT-crystal, which is determined by the holder in use, the Q of a crystal increases with frequency. After considering all factors, it will be found that a precisely manufactured crystal in the order of 5 MHz is especially favourable for this application.

3. CRYSTAL OSCILLATOR CIRCUIT

The discussion of the most favourable crystal oscillator circuit is as old as the crystal technology. Opinions differ mainly with respect to the question whether the crystal should operate at series or parallel resonance. However, since any crystal must be "pulled", the actual frequency of oscillation of the crystal is affected by the external circuitry and will not oscillate at either its parallel or series resonance.

Fig. 1:
Equivalent diagram
of a crystal



When considering the equivalent diagram of a crystal as given in Figure 1, it will be seen in a simplified manner that the series resonance of a crystal will formed from L and C_1 when the following is valid:

$$\omega L = \frac{1}{\omega C_1}$$

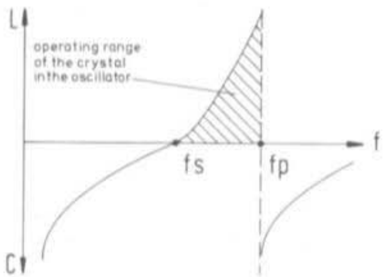
On the other hand, a parallel resonance can bc derived from the series circuit of C_1 and C_0 (holder capacitance), e. g. when the following is valid:

$$\omega L = \frac{1}{\omega \dfrac{C_0 \times C_1}{C_0 + C_1}}$$

This parallel resonance cannot be achieved in practice since the required external circuitry places reactances in parallel with the crystal. This means that the complete oscillator circuit operates at the frequency at which the crystal together with its reactive impedance is real.

Capacitive feedback circuits are mainly used to obtain variation ranges on pulling crystal oscillators, and to obtain the best characteristics with respect to the temperature coefficients. The crystal then oscillates in the inductive part of its reactive impedance characteristics, e. g. between the series and parallel resonance frequency (Fig. 2). This means that it would not be correct to judge the stability of a crystal oscillator alone on the criteria of the crystal. It is far more important to keep the stabilizing factor of the crystal as high as possible in the circuit. This means that fluctuations of the operating parameters such as operating voltage, capacitances, etc. shall have the lowest possible effect on the frequency. Under these conditions, it is sufficient for the crystal and the varactor diode used for alignment to be kept at a stable temperature. The low heat capacitance (mass) of these two components allows a short warm-up period.

Fig. 2:
Reactive impedance characteristic
of the crystal
f_s = series resonance
f_p = parallel resonance



Now to details regarding the crystal loading: If the fact is ignored that an oscillating crystal warms itself via its own impedance, which is valid more or less for every electrical component in a circuit, and which can amount to several degrees centigrade at a crystal loading which is even far below the maximum level, it is of the greatest importance with respect to the aging rate that the crystal loading is as low as possible. Values of 100 μW or less are advisable.

One often sees circuits in which "overloaded" crystals are used in conjunction with a NAND-gate (e. g. SN 7400). However, there are cheaper methods of destroying crystals.

4. THE CRYSTAL OVEN

High demands are placed on the temperature stabilization of crystal oscillators with an accuracy in the order of 10^{-8}. The warm-up period should be as short as possible in order to ensure that the unit is ready for operation in the shortest possible time. The inverse ambient-temperature reduction factor of the crystal oven must be as favourable as possible so that the ambient temperature fluctuations within the crystal oven only result in temperature fluctuations of maximum 0.1 °C when it is assumed that the temperature coefficient of the crystal in the vicinity of its reversal points is approximately $\frac{\Delta f}{f}$ /°C $\approx 1 \times 10^{-7}$.

In this range, it is possible for the temperature shifts of the overall circuit as a result of aging to be considered. In addition to this, the adjustment of the nominal temperature should be provided so that it can be adjusted with a resolution of 0.1 °C in a range of approximately $\pm$ 5 °C. Such demands cannot be fulfilled using bi-metal thermostats: The system-dependent temperature ripple (1) will only be at a minimum when the nominal temperature can just be kept after considering the heat dissipation of the system. This means that an effective heating of the crystal oven must be obtained using an additional, switchable heater winding. It is practically impossible to precisely adjust a small bi-metal thermostat with an accuracy of 0.1 °C. This means that a proportional control of the temperature stabilization must be used.

A thermistor is used as heat-probe which should have a value that is as high as possible in order to ensure that the intrinsic heating due to the inavoidable measuring current is as low as possible. The control amplifier is in the form of a DC-voltage amplifier, whose output stage is mounted on the case of the crystal oven for cooling. This ensures that a separate heater winding is not required.

The heat-probe, case of the crystal oven, crystal, varactor diode and output transistor are combined in such a manner that the best possible heat conduction exists between the various components so that the lowest possible heat capacitance is obtained. The crystal holder which protrudes out of the case of the crystal oven has no noticeable effect on the accuracy of the control circuit since the heat dissipation of the electrical connections to the crystal are very low due to the low cross section of the holder.

The crystal oven does possess one undesirable feature: Rapid fluctuations of the heating due to a defect in the temperature dissipation, or due to large voltage fluctuations cause to a relatively large overshoot of the frequency. This is not only due to the control circuit where an overshoot of the heater current can be observed. The cause is possibly in the inhomogenities of the warm-up process or in the cooling of the crystal element. A temperature gradient will be formed in the crystal temporarily which can be lead to a frequency variation that has nothing to do with the expected characteristic as a result of the cut. This will be observed during the warm-up process. Since the crystal (AT-cut) is operated at its reversal point, the frequency should only reduce to the adjusted frequency during the transient period, and should increase if a temperature over-control condition exists. The measured transient behaviour does not indicate this (Fig. 3: observe the two different ordinates which differ by two orders of ten). This "weakness" of the crystal oven is, however, of no importance since the frequency fluctuations remain below $\frac{\Delta f}{f} = 10^{-8}$ during normal operation and since these transient processes are completely without hysteries, and therefore do not cause any residual control deviation.

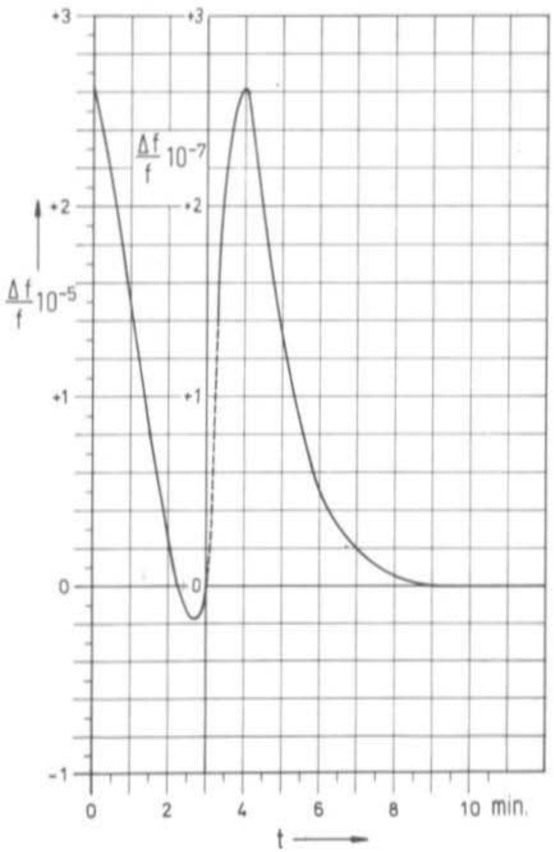


Fig.3:
Transient characteristic of the oscillator

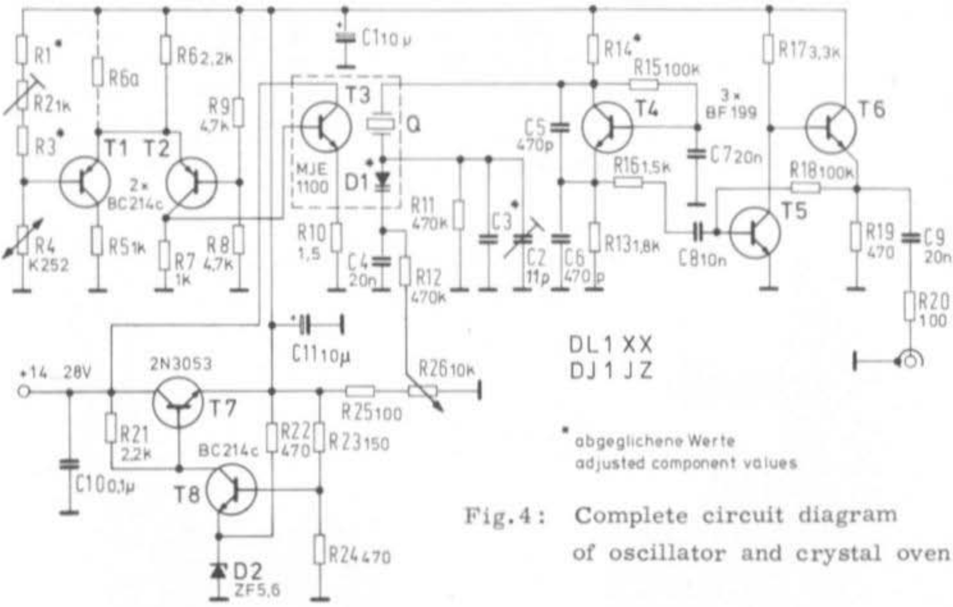


Fig.4: Complete circuit diagram of oscillator and crystal oven

5. CONSTRUCTION

The complete circuit diagram of the oscillator and crystal oven are given in Figure 4. A modified Pierce crystal oscillator circuit is used with transistor T 4 in a common-base circuit. The coarse frequency adjustment is made with the aid of trimmer capacitor C 2, or by exchanging capacitor C 3 if large frequency variations are present. Varactor diode D 1 is provided for fine tuning; the bias voltage of this diode can be adjusted with the aid of a precision potentiometer (R 26).

The values of the feedback capacitors C 5 and C 6 are selected so that oscillation commences readily and the temperature-dependent semiconductor capacitances are sufficiently shunted. The correct value of resistor R 14 is selected for each individual crystal so that the correct crystal loading results.

The output signal is taken from the emitter of transistor T 4. If the operating points have been adjusted correctly, a virtually undistorted (sinusoidal) signal will be available at this point. The output signal is amplified to approximately 1 V (RMS) in a subsequent, two-stage amplifier and coupled out at low impedance.

The comparison between the actual and nominal value for the temperature control is made in a differential amplifier stage comprising transistors T 1 and T 2. The required crystal oven temperature is adjusted with the aid of resistors R 1 to R 3, and the required control gain, or heating current is selected by connecting resistor R 6a in parallel to R 6. The differential amplifier drives a power transistor (T 3) whose power dissipation heats the case of the crystal oven.

With the exception of the heater transistor, all operating voltages are taken from the stabilizer circuit comprising (T 7, T 8, D 2).

The complete circuit including crystal oven is built up on a single-coated PC-board of 115 mm x 50 mm. The completed PC-board is mounted in an aluminum case of 125 mm x 55 mm x 55 mm (Fig. 5). The output signal is taken from a BNC-connector which is mounted at one end of the box; the connections for the operating voltage (feedthrough capacitors) and the precision trimmer potentiometer for frequency calibration are to be found at the other end of the box.

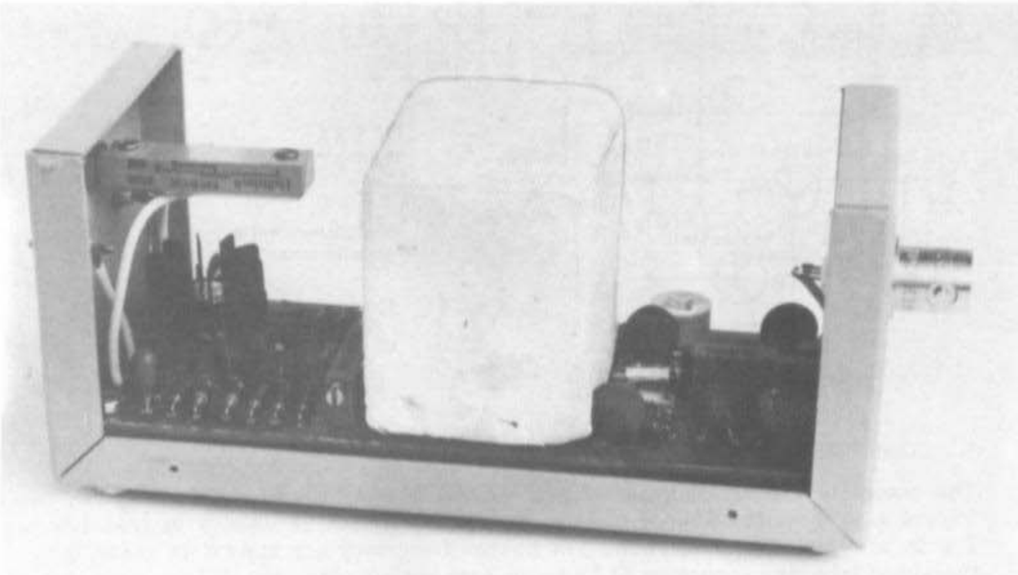


Fig.5: Author's prototype with cover removed

6. ALIGNMENT

Although it will not be possible for most radio amateurs to carry out this calibration process, it is briefly to be described. Due to the difficulty in calibrating this module, no kit is to be offered, only a ready-to-operate module.

The equivalent data f_{series}, R, C_1 and L are now determined in a passive resonator circuit (pi-circuit) and the Q of the crystal is then calculated. If the Q of the crystals are not sufficient, such crystals should not be used in this circuit. This is followed by measuring the temperature-reversal point on increasing and decreasing the temperature, which is made in a circulating water bath. All crystals that have too high or too low a reversal point as well as any crystals that possess a measurable hysteresis in the reversal point between heating and cooling should also not be used.

The crystal oven is now heated to within 1 to 2 °C of the required crystal temperature. The crystal is now inserted and adjusted to a loading of approximately 100 μW by aligning of R 14. The maximum permissible amplitude for the crystal was calculated previously from the equivalent data of the crystal.

It is now possible for the exact crystal oven temperature to be selected whilst observing the frequency variation. This adjustment takes some time since one must wait until the transient effects of the temperature are stabilized after each adjustment. This is followed by adjusting the frequency coarsely with capacitor C 2 and finely with R 26, and allowing the oscillator to run for at least fourteen days continuously. The frequency variations are noted once per day. The module can be classed as ready for operation after the aging rate has dropped to far less than $10^{-8} \frac{\Delta f}{f}$/day. Any oscillators whose aging rate is not less than this value after an operating period of 30 days are classed as not successful and will be dismantled.

7. FREQUENCY ADJUSTMENT

After manufacture, the module is aligned for better than $10^{-8} \frac{\Delta f}{f}$ at an operating voltage of 18 V. However, one must assume that the crystal will age by 10^{-7} to $10^{-6} \frac{\Delta f}{f}$ after a considerable period, which means that the frequency standard should be calibrated occasionally.

The method of adjusting the frequency to that of a standard frequency transmitter has been published on so many occasions that only a few details are to be mentioned here:

The longwave standard frequency-transmissions such as DCF on 77.5 kHz Droitwich on 200 kHz and the Deutschlandfunk on 151 kHz are preferable to the shortwave transmissions. Of these, the transmitter with the nearest location should be selected. Delay variations can occur during the measuring period due to propagation conditions, which could simulate a frequency deviation (of up to $10^{-7} \frac{\Delta f}{f}$ on shortwave). This means that any signals with a high degree of fading should be avoided. On the other hand, longwave signals have the disadvantage that the observation time is relatively long for an accuracy of 10^{-8}. One requires, for instance, a measuring period of at least 1000 seconds in the case of a 100 kHz signal.

Finally, two methods of comparing the frequency to a standard frequency transmission:

A longwave standard frequency signal is received in a TRF-receiver with feedback, and the synchronized beat frequency is multiplied to the highest possible frequency. This frequency is measured with the aid of frequency counter which uses the described frequency standard as time base.

The standard frequency from the longwave receiver is fed to a phase comparator circuit and compared to a harmonic of the oscillator to be adjusted. A meter connected to the phase discriminator allows the beat of the oscillation to be observed and brought to zero.

The manufacturer is prepared to align such oscillators once per year for those amateurs that are not able to make this correction adjustment, it is only necessary to pay the transport costs.

8. SPECIFICATIONS

Frequency:	5 MHz (optimal 1 MHz to 20 MHz)
Crystal:	AT-cut, fundamental
Crystal holder:	HC-36U
Frequency adjustment:	Mechanical: approx. 3×10^{-5} $\frac{\Delta f}{f}$
	Electrical: approx. 1×10^{-6} $\frac{\Delta f}{f}$
Output voltage:	approx. 1 V, sinusoidal
Output impedance:	approx. 200 Ω
Ambient temperature range:	-20 to +40 °C

Frequency deviation:

Due to aging:	less than 1×10^{-8} $\frac{\Delta f}{f}$ /day after 24 h period
Due to temperature:	less than 1×10^{-9} $\frac{\Delta f}{f}$ /°C

Due to supply voltage fluctuations of ±10% at 18 V:
less than 5×10^{-9} $\frac{\Delta f}{f}$

Due to loading (no-load/short): less than 1×10^{-8} $\frac{\Delta f}{f}$

Frequency error after switching on at a temperature of 25 °C and a supply voltage of 18 V:

Deviation $\frac{\Delta f}{f}$ from the required frequency after 3 min.: less than 10^{-6}
after 6 min.: less than 10^{-7}
after 10 min.: less than 10^{-8}

Operating voltage:	+ 14 V to + 28 V to ground
Heating current:	700 to 900 mA
Operating current:	Less than 150 mA at 18 V
Dimensions:	125 mm x 55 mm x 55 mm
Weight:	Approx. 200 g

9. REFERENCES

(1) R. Görl and B. Rössle: A Stable Crystal-controlled Oscillator in the Order of 10^{-7} for Frequency and Time Measurements
VHF COMMUNICATIONS 4, Edition 4/1972, Pages 235-240

(2) D. E. Schmitzer: A 200 kHz Receiver for Synchronizing 1 MHz Crystal Oscillators to the Droitwich Longwave Transmitter
VHF COMMUNICATIONS 4,. Edition 2/1972, Pages 111-118.

MATERIAL PRICE LIST FOR EQUIPMENT
described in Edition 2/1975 of VHF COMMUNICATIONS

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Minikit	DJ 5 XA 003 (2 coilsets, 1 coilformer with core, 2 ferrite beads, 10 trimmer capacitors, 4 chip capacitors, 1 TEKO Box 3A)	DM 23.--
Crystal	67.333 MHz (HC-6/U)	DM 22.--
Kit	DJ 5 XA 003 with above parts	DM 110.--

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13-pole connector	for PC-board mounting, each	DM 4.90
13-pin connector	for above	DM 4.10

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Type TA Type TA-S Type TA 4

Type U 3 Type U 4 Type U 5

Model	Type	Frequency	Gain	Weight	Features
TA	$5/8\ \lambda$	144-175 MHz	3 dB	275 g	Glass-fibre whip
TA-S	$5/8\ \lambda$	144-175 MHz	3 dB	275 g	Glass-fibre with 5 m cable
TA 4	$1/4\ \lambda$	144-175 MHz	0 dB	130 g	Stainless steel (PH 17-7)
U 3	$5/8\ \lambda$	400-470 MHz	3 dB	100 g	Silver-plated, epoxy coated
U 4	Colinear	420-470 MHz	4 dB	150 g	Stacked $\lambda/4$ and $5/8\ \lambda$
U 5	Colinear	420-470 MHz	5 dB	175 g	Stacked $5/8\ \lambda$ and $5/8\ \lambda$

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Application		SSB-Transmit.	SSB	AM	AM	FM	CW
Number of Filter Crystals		5	8	8	8	8	4
Bandwidth (6dB down)		2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz
Passband Ripple		< 1 dB	< 2 dB	< 2 dB	< 2 dB	< 2 dB	< 1 dB
Insertion Loss		< 3 dB	< 3.5 dB	< 3.5 dB	< 3.5 dB	< 3 dB	< 5 dB
Input-Output	Z_t	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω	500 Ω
Termination	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
Shape Factor		(6:50 dB) 1.7	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:40 dB) 2.5
			(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:60 dB) 4.4
Ultimate Attenuation		> 45 dB	> 100 dB	> 100 dB	> 100 dB	>90 dB	> 90 dB

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