A Publication for the
Radio Amateur Worldwide

Especially Covering VHF,
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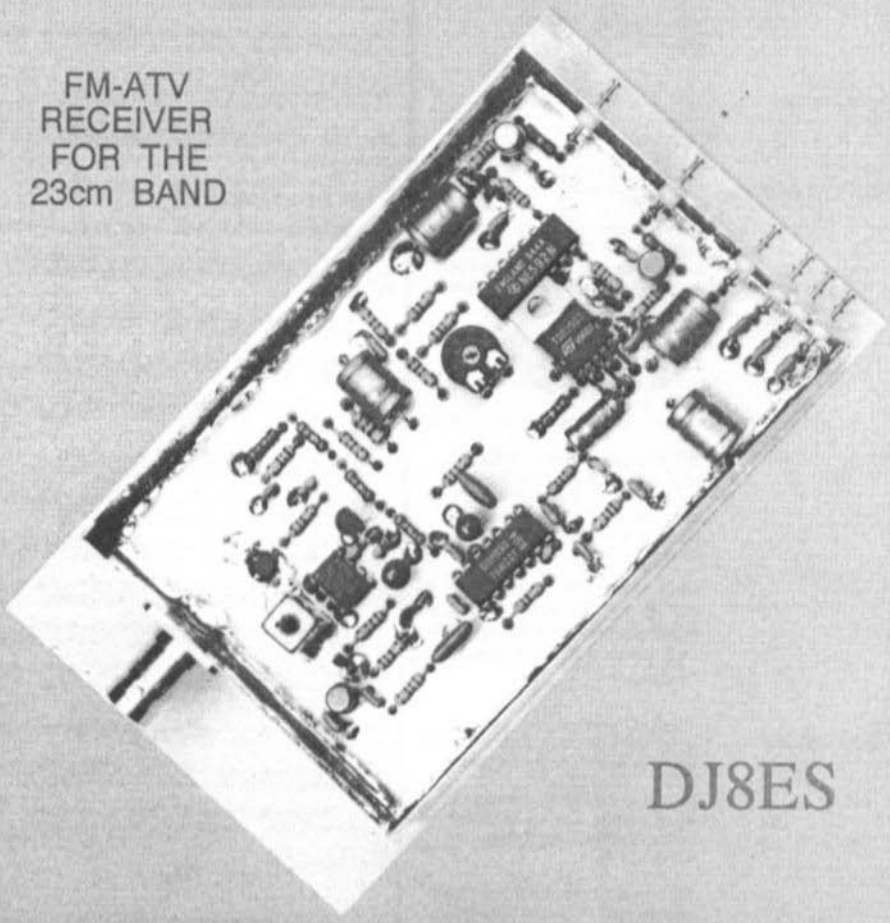
VHF COMMUNICATIONS

Volume No. 23 . Spring . 1/1991 . £3.50

FM-ATV RECEIVER FOR THE 23cm BAND



DJ8ES

VHF COMMUNICATIONS

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Volume No. 23 . Spring . Edition 1/1991

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Mike Wooding, G6IQM

KM Publications a brief introduction

Just a few words by way of introducing ourselves to you. KM Publications consists of myself, Mike Wooding G6IQM and my wife Kim Wooding (hence KM). I have been an active amateur radio enthusiast for several years, my main sphere of activity being amateur television. However, I am now exploring packet radio with some enthusiasm.

Kim Wooding is not an active radio amateur, but is absorbed by all things artistic, thus you may see, from time to time, her influence on the magazine.

KM Publications was set up with the express intention of taking over the publication of VHF Communications, when we heard that UKW-Berichte were going to discontinue it. It has been quite a gamble for us, but we are pleased to say that the majority of the previous subscribers have renewed again, even with the unfortunately necessary price increase.

It is our intention to maintain the very high standard that VHF Communications has enjoyed in the past. In fact, to all intents and purposes, and apart from the occasional difference in published articles between VHF Communications and UKW-Berichte, the magazine will be an almost exact replica of past issues, with a mix of articles from recent issues of the German magazine.

Finally, in this brief introduction, I would invite you to send in your articles to us for consideration of publication. Without the continued support of you, our readership, we cannot maintain this excellent publication.

Also, please write to us with your comments, criticisms, complaints, articles and topics requested, etc. We will be only too pleased to hear from you.

Almost the last word. I would like to express our gratitude and extend our sincere thanks to Corrie Bittan and the staff of UKW-Berichte, for their help, advice and support during the takeover period.

Regards ... Kim and Mike Wooding. KM Publications

Wolfgang Schneider, DJ8ES

FM-ATV Receiver for the 23cm Band

FM is the hot topic among ATVers today. On account of the larger bandwidth required (of the order of 20MHz) FM ATV operation is possible only on 23cm and higher bands such as 13cm 9cm and 3cm. The following contribution describes an FM ATV receiver for the 23cm band.

Previously published designs, particularly early on, leaned either towards narrow-band concepts or were orientated towards satellite techniques (with their own problems). These concepts could only be used without modification for a single purpose.

For universal application - for example direct simplex ATV, ATV repeaters, etc. - a receiver tunable over 50MHz or so is called for.

The receiver described has been conceived specifically for FM-ATV, but could also be used for amplitude modulation. For this the IF on channel 22 (vision carrier 479.5MHz, sound carrier 485.0MHz) is available only at the designated points, that is directly behind the converter or (through a directional coupler) after the IF amplifier.

As fig.1 shows, the FM ATV receiver consists of several modules. First of all is the converter itself, comprised of front end, local oscillator and mixer).

Fig.1 Block circuit diagram of the complete FM ATV receiver for 23cm

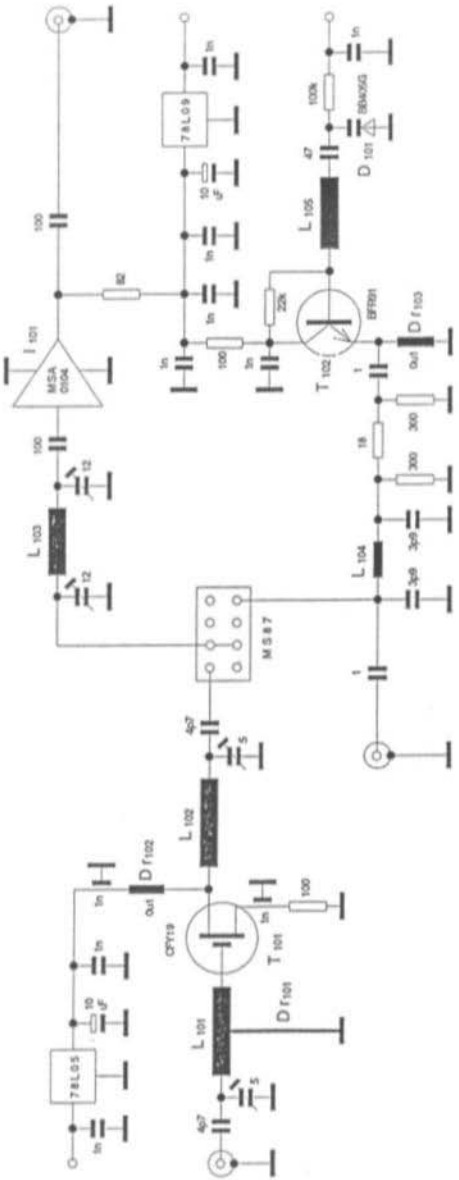


Fig.2 Circuit of the Converter DJ8ES

These stages are all combined in the unit DJ8ES 001.

A digital frequency indication is not absolutely necessary for the operation of the converter, but this would be a pleasant addition to the receiver concept. This module has the designation DJ8ES 002.

The IF amplifier DJ8ES 003 operates at around 480MHz, for which an integrated filter in the unit serves.

The FM demodulator DJ8ES 004 is based on a PLL element, which provides a direct video output. Following de-emphasis a sound demodulator and audio amplifier are also provided in this unit.

The four units are now fully described.

1.

The Receive Converter

Fig. 2 shows the complete circuit of the 23cm converter. Noteworthy elements are the front end, mixer, local oscillator, low pass and IF amplifier. With the indicated dimensioning of the tunable oscillator, the converter covers the whole 23cm amateur band.

1.1. Circuit description

The received signal passes along a quarter-wavelength transformation line to the gate of the GaAsFET T101, is amplified in here and is then led to the ring mixer. Filters were dispensed with between the front end and mixer: conventional quarter-wavelength stripline filters do not provide the necessary bandwidth for 50MHz or more and more specialised designs would not be reproducible on the epoxide PCB material employed. Even so, image rejection of 40dB can be achieved by using the now commonplace high IF of 480MHz.

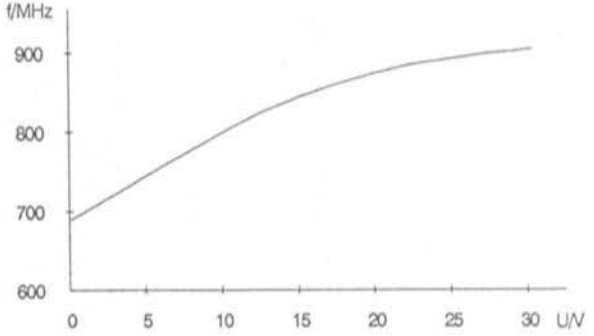


Fig.3 VCO frequency as a function of the tuning voltage

The voltage-controlled oscillator (VCO) was, like the rest of the circuitry, designed from commonly available parts. With this oscillator a frequency variation of about 200MHz can be achieved in the 700-900MHz region. The transistor (T2) employed in this stage provides an output of 20mW. The VCO frequency as a function of the tuning voltage is illustrated in fig. 3.

For decoupling and at the same time power matching between oscillator and mixer a 3dB attenuator element is used. The low pass reduces the harmonic content by 20dB.

A small part of the oscillator power is tapped off through a 1pF capacitor for the frequency display.

1.2. Construction hints

The printed circuit board DJ8ES 001 for the unit described is made of double-sided epoxide material 1.5mm thick; it has the measurements 53.5mm x 108mm (fig. 4), so as to fit into a commercially available tinplate enclosure (55.5 x 111 x 30).

After trimming the board is treated with rub-on silver plating and then drilled. The transistors and the broadband amplifier lie in

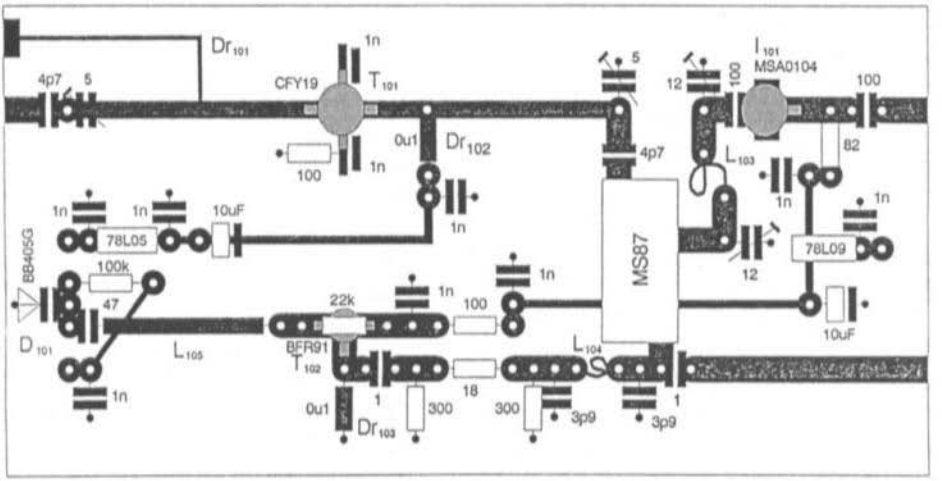


Fig.4 Position of components in the converter

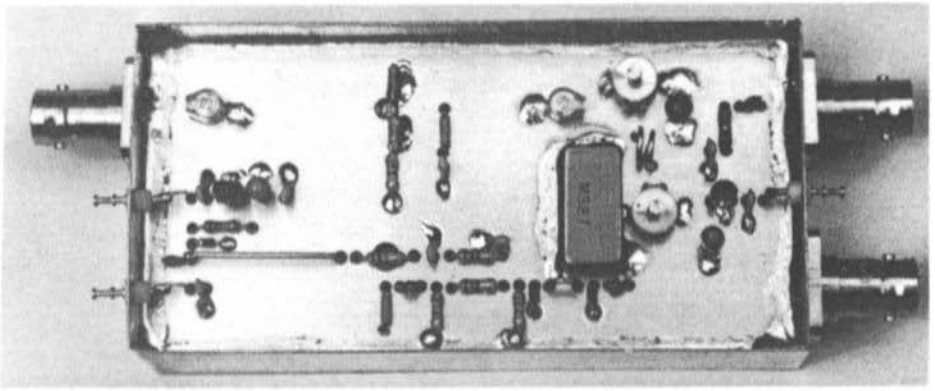


Fig.5 A completed converter unit

the PCB, so you need to drill suitable holes for these and then de-burr them. A recess of approx. 1mm x 8mm should be sawn away at the appropriate edges of the PCB for the BNC connectors.

Following drilling, the holes for the resistors, capacitors, crystal, etc. should be milled with a 3.5mm drill on the groundplane (fully coppered) side so that the surrounding copper surface is removed.

Once these preparations are complete, the PCB can be sprayed with soldering varnish.

After drilling the through-hole ground connections for the transistors and broadband amplifier can be made.

After soldering the PCB into the tinplate enclosure the components can be placed in position.

The BNC flange connectors need to lie with their flanges on the edge of the case. The PCB is set in the case so that the pins of the BNC connectors should just touch the PCB tracks.

Note: the Teflon collar will need to be removed with a sharp knife.

After soldering the PCB to the side surfaces of the housing the components can be inserted.

1.3. List of components

T101:	CFY19 (Siemens)
T102:	BFR91 (Valvo/Mullard)
IC101:	MSA0104 (Avantek)
D101:	BB405G
L101, L102:	etched
L103:	1.5 turns, 5mm diameter, silver-plated wire 1mm thick
L104:	half turn, 5mm diameter, silver-plated wire 1mm thick
L105:	40mm long bent out 5mm, silver-plated wire 1mm thick
Dr101:	etched
Dr102, Dr103:	0.1uH choke (Neosid)

1 Ring Mixer MS87 (or similar 2GHz type)
1 Voltage Regulator 78L05
1 Voltage Regulator 78L09
1 Resistor 82R, 0.1 watt
2 Tantalum bead electrolytics 10uF/25V
2 Chip capacitors 4.7pF
2 Chip capacitors 100pF
2 Trapezoidal capacitors 1nF
2 Foil trimmers 12pF (yellow)
2 Sky trimmers 5pf (green)
1 Tinplate enclosure 111 x 55.5 x 30mm
3 BNC flange sockets (UG-290A/U)
3 Teflon feed-throughs

Ceramic disc capacitors (2.5mm lead spacing)

2 off	1pF
2 off	3.9pF
1 off	47pF
1 off	100pF
8 off	1nF

Resistors (1/8W, 10mm lead spacing)

1 off	18R
2 off	100R
2 off	300R
1 off	22k
1 off	100k

1.4. Commissioning the converter unit

The following measuring devices are required:

- multimeter
- frequency counter (to 1GHz)
- power meter (to 1GHz)
- ATV signal, reaching the converter via an antenna

First only the 12V power supply is connected to the module DJ8ES 001; a current of around 95mA should be drawn, 12mA of which is for the preamplifier. The unit should work straightaway, producing at the oscillator output a signal of around 700MHz with a level of approx. 2mW. If the tuning voltage (max. 30V) is now connected, the frequency variation of the VCO should match that in fig. 3.

Now you should carry out the first trial with a local ATV signal (for instance from a nearby repeater). A TV receiver should be connected to the output socket and set on channel 22. For guaranteeing the best possible tuning of the trimmers in the individual stages an amplitude modulated ATV signal is ideal.

If channel 22 is occupied locally it is also possible to use a different IF. With due consideration to the total concept of the

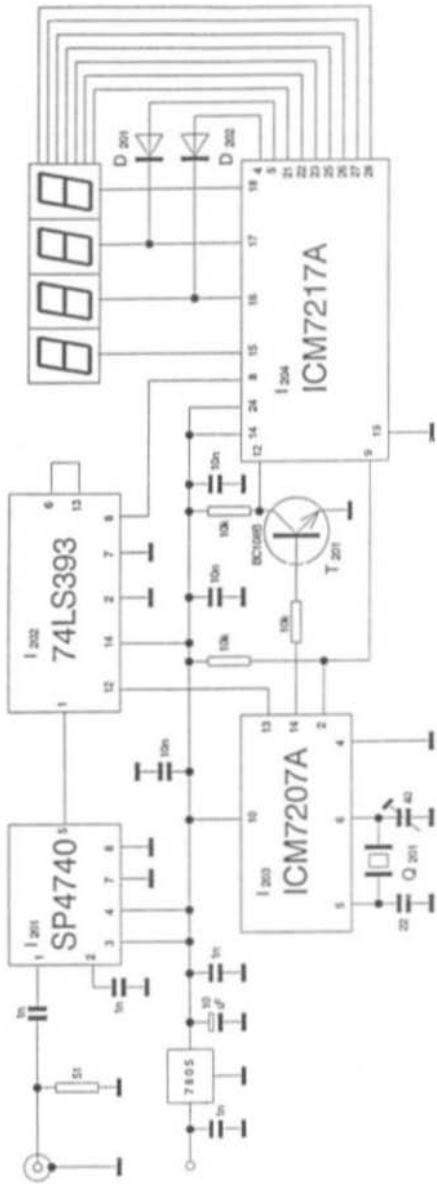


Fig.6 Circuit of the Digital Frequency Indicator DJ8ES 002

converter unit, in particular the tunable oscillator and the incorporated ring mixer, the top frequency is 600MHz, which corresponds to channel 37. In this way the converter covers at IF level virtually the whole of UHF Band IV. naturally this entails corresponding modifications to the following IF and demodulator modules.

2. DIGITAL FREQUENCY INDICATOR

The digital frequency indicator is designed as a separate unit DJ8ES 002, and represents a welcome addition to the FM ATV receiver. The frequency counter evaluates the tunable oscillator of the receive converter and carries forward this frequency, offset by the amount of the IF at 480MHz to the display.

2.1. Circuit description

The circuit of the frequency counter is illustrated in fig. 6. At the input of the unit the oscillator signal to be evaluated is divided by a factor of 256 in IC201 (SP4740). This is an integrated frequency divider from Plessey and this version provides a TTL-level output signal ready for further processing. The following 74LS393 binary counter (IC202) provides, beside the necessary further division by 256, the gate-timing element of the frequency counter.

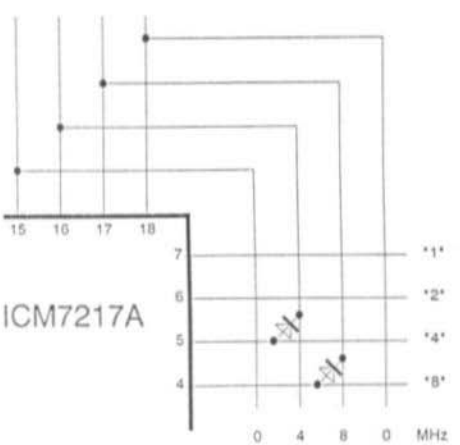


Fig.7 Programming the Counter unit

The counter module serves an ICM7217AIPI from Intersil (IC204). This four digit version has a resolution of 1MHz, which is very favourable for ATV applications. The control signal needed by this unit is provided by IC203, an ICM7207AIPD. This controller provides the driver outputs for use as a frequency counter timebase together with a crystal oscillator and attached divider. The total division factor of the counter calls for a crystal frequency of 8MHz.

The frequency counter must take into account not only the oscillator frequency but also the IF offset. For that reason the counter unit is programmed with the amount of the IF (480MHz). The presetting is done with a diode matrix and can easily be arranged for other frequencies (fig. 7 shows how).

2.2. Constructional hints

The PCB for the digital frequency indicator is again made of 1.5mm thick double-sided epoxide material and has the measurements 110mm x 80mm.

It is unnecessary to build the module in a housing so the PCB has been designed so that it can be cut in half to suit building into other applications. Depending on this construction it may be necessary to fit a heatsink to the fixed voltage regulator: this can be a commercial item or home-made.

Drilling and other work is as for the converter module, although the silver plating is not required.

The ICs should be fitted in sockets, with the exception of the first pre-scaler IC201, which on account of its high operating frequency

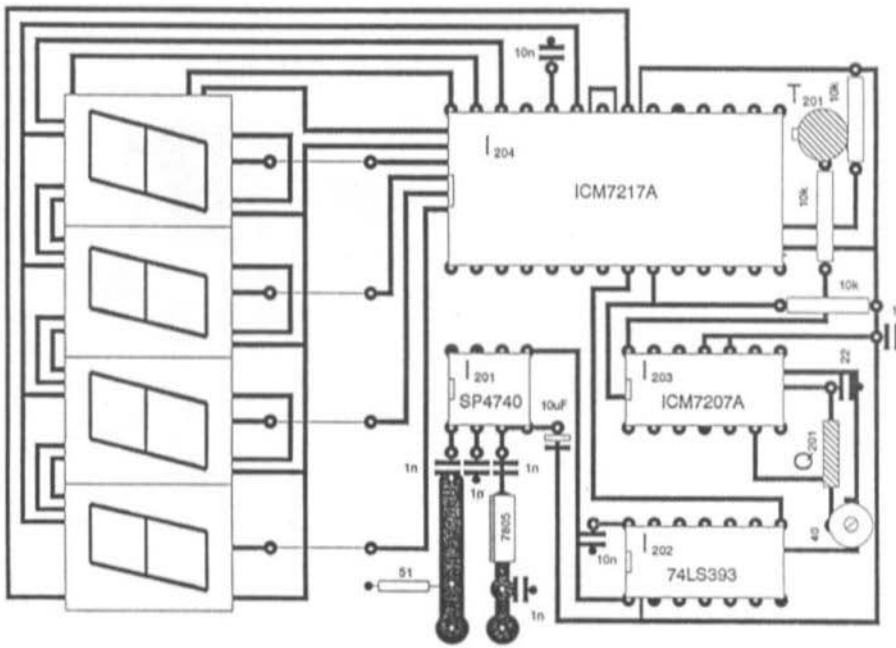


Fig.8 Component plan of the Digital Frequency Indicator

should be soldered direct to the PCB. The IC sockets should be precision-turned ones.

Apart from good contact precision-turned sockets have another advantage: they permit diodes for programming the frequency counter to be laid out beneath the counter (IC204). Fig. 8 is the component layout for the counter PCB DJ8ES 002.

2.3. Component list

IC201:	SP4740 (Plessey)
IC202:	74LS393
IC203:	ICM7207AIPD (Intersil)
IC204:	ICM7217AIPI (Intersil)
Q201:	8MHz Crystal, HC18-U
D201, D202:	1N4148 or equivalent
T201:	BC108B or equivalent

4 7-segment displays SL1110K or equivalent
1 Fixed voltage regulator 7805
1 Tantalum bead electrolytic 10uF/25V
1 Foil trimmer 40pF (grey)
1 Resistor 51R (1/8W, 10mm lead spacing)
1 Resistor 10k (1/8W, 10mm lead spacing)
4 Ceramic disc capacitors 1nF (2.5mm lead spacing)
3 Ceramic disc capacitors 10nF (2.5mm lead spacing)
1 Ceramic disc capacitor 22pF (2.5mm lead spacing)

2.4. Commissioning

Following construction and a visual check (see fig. 9) the frequency counter can be put straight into service.

The unit is connected to the oscillator output of the converter and as soon as power is connected it should display the receive frequency of the converter.

If a known reference frequency is available

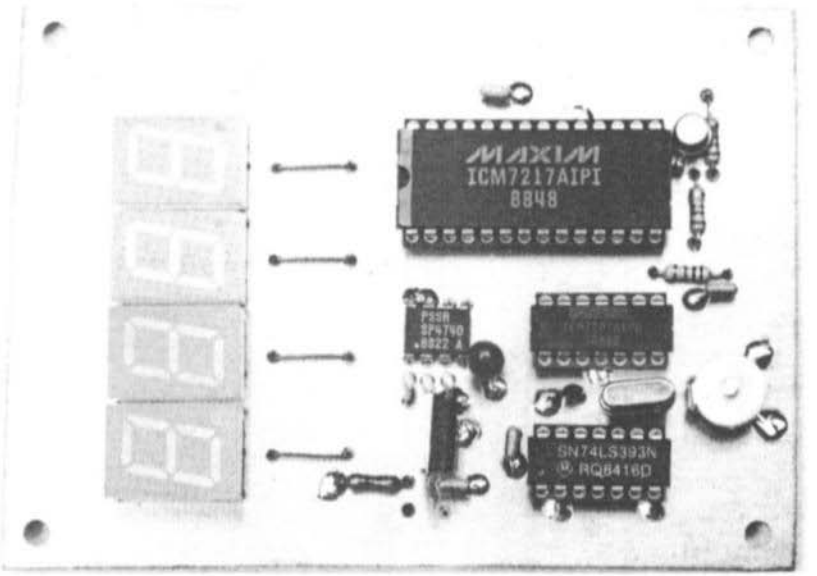


Fig.9 Laboratory sample of the Frequency Counter module

(e.g ATV repeater) fine tuning of the display frequency can be attempted with the trimmer capacitor.

3. THE IF AMPLIFIER

The total concept of the FM ATV receiver is for the IF amplifier to have a through amplification of 45dB at 480MHz.

3.1. Circuit description

Fig. 10 shows the complete circuit of the IF amplifier. The required amplification of 45dB is guaranteed by the three-stage chain of integrated broadband amplifiers of the type MSA0104 (IC301-303). To secure the necessary bandwidth for FM ATV a helical filter is fitted after the first stage.

The integrated amplifiers employ bipolar silicon transistor broadband amplifiers in a Darlington circuit arrangement. These MMICs (MMIC = monolithic microwave integrated circuit) are the simplest way of achieving high levels of reliable broadband amplification. Particularly convenient is that no input or output matching is necessary:

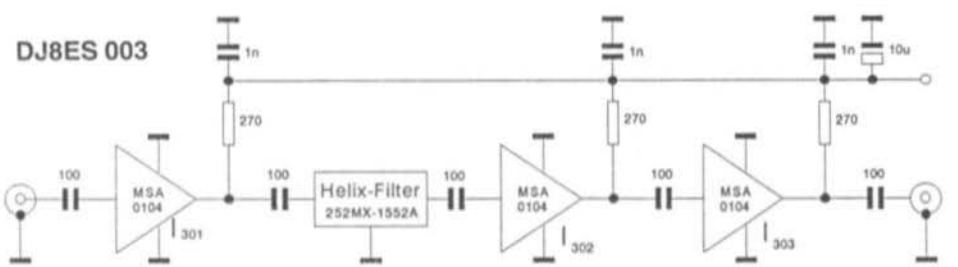


Fig.10 Circuit of the IF Amplifier DJ8ES 003

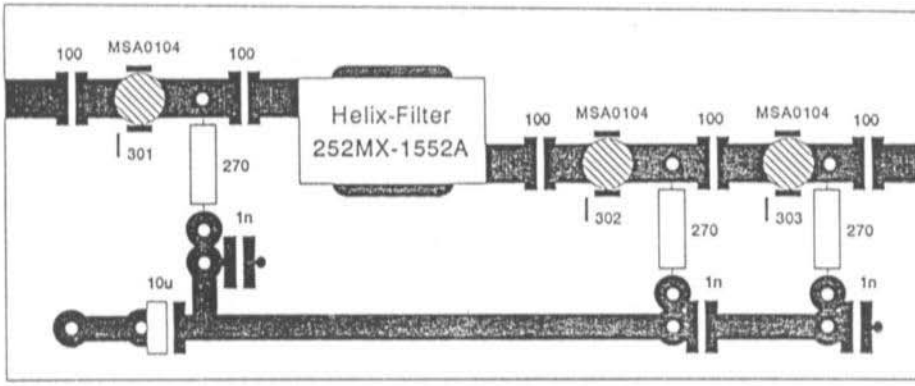


Fig.11 Component overlay for the IF Amplifier DJ8ES 003

impedance in both cases is 50 ohms. The only external components needed for operation of the MMICs are a series resistor for the supply and input and output coupling capacitors.

3.2. Constructional hints

The PCB for the module DJ8ES described consists of double-sided 1.5mm epoxide material and has the measurements 34 x 72mm (fig. 11); it fits conveniently into a commercially available tinplate box (37 x 74 x 30mm). The necessary preparations for the PCB and housing are as for the converter module: this also applies for installing the PCB and mounting the components.

3.3. Component list for DJ8ES 003

IC301,2 & 3: MSA0104 (Avantek)

1 Helical filter 252MX-1552A
5 Chip capacitors 100pF
3 Resistors 270R (1/2W, 10mm lead spacing)
3 Ceramic disc capacitors 1nF (2.5mm lead spacing)
1 Tantalum bead electrolytic 4.7uF/25V
2 BNC flange sockets (UG-290A/U)
1 Teflon feed-through
1 Tinplate enclosure 37 x 74 x 30mm.

3.4. Commissioning

After complete construction of the IF amplifier (fig. 12) it can be put into operation on a

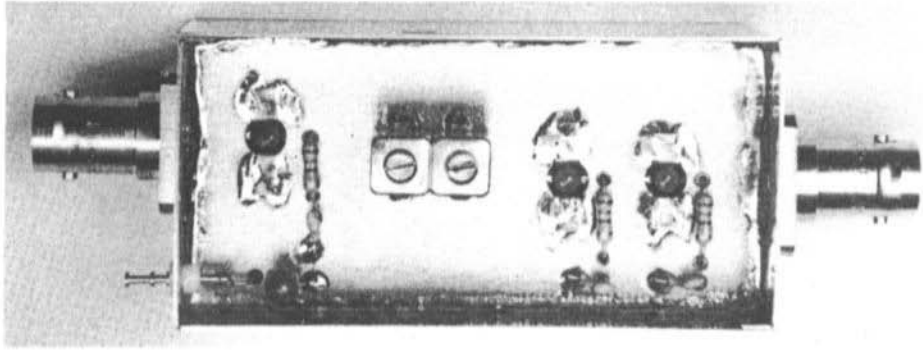


Fig.12 Laboratory sample of the IF Amplifier

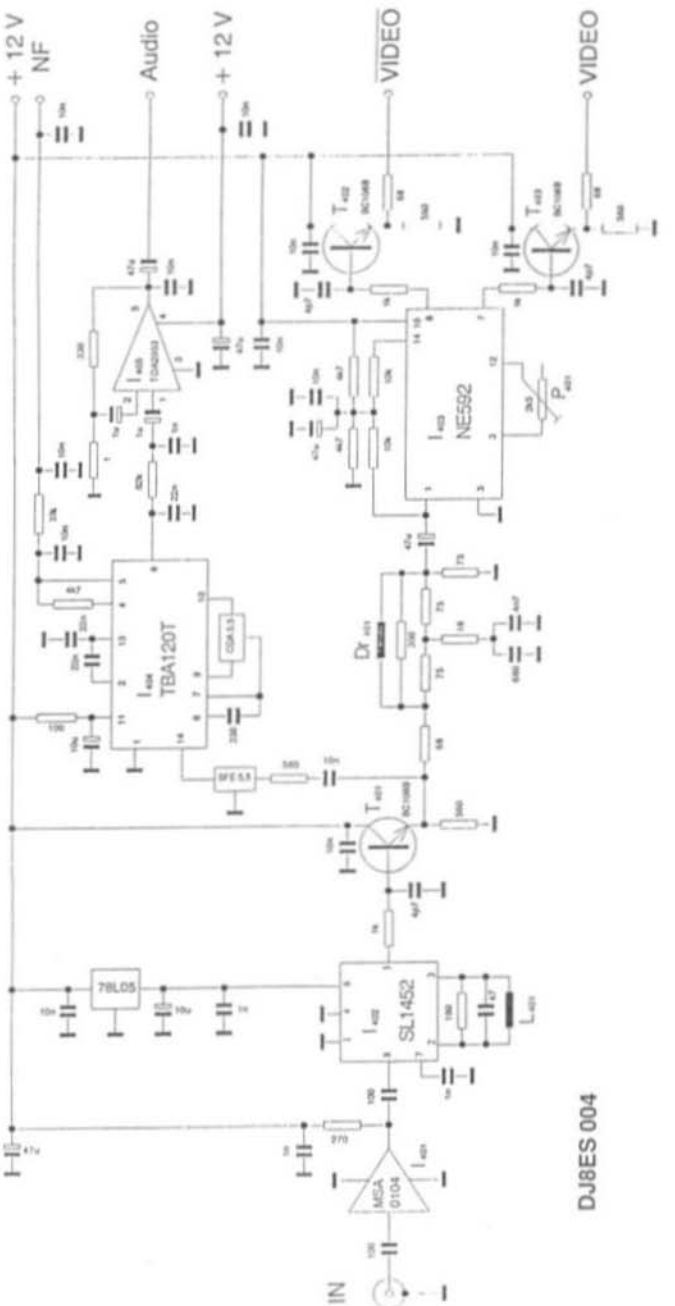


Fig.13 Circuit of the FM Demodulator unit DJ8ES 004 (NF =AF)

trial basis. Apply the +12V supply and look for a current drawn of around 84mA. This implies a current requirement of 28mA for each stage, which can be proved simply by checking the voltage drop of 7.56V across each 270R resistor.

The only frequency-determining component is the helical filter fitted between the first and second amplifier stages. This should be adjusted for the flattest through-pass curve (minimum ripple): this is best checked with a sweep-frequency generator (wobbulator). With a centre frequency of 480MHz a 3dB bandwidth of 22MHz should be achieved.

4. THE FM DEMODULATOR

With regard to the widely used IF of 480MHz which is now commonplace in satellite television receivers a demodulator board was developed to match the rest of the project. This module delivers a normalised video signal together with an audio signal for direct connection to a loudspeaker.

4.1. Circuit description

The heart of the demodulator unit (fig. 13) is the quadrature demodulator SL1452 from Plessey. It is adjusted onto the 480MHz IF to be demodulated with inductance L401. In this frequency-determining resonant circuit the parallel resistor of 180 ohms determines the demodulator bandwidth corresponding to the FM ATV signals being received.

The demodulated video signal passes through a buffer stage (T401) to the de-emphasis filter and on to the video amplifier NE592 (IC403). The amplification factor of this IC is preset by potentiometer P401. On its outputs the NE592 delivers two video signals of the same amplitude but opposed polarity. An impedance transformer (T402, T403) is connected here.

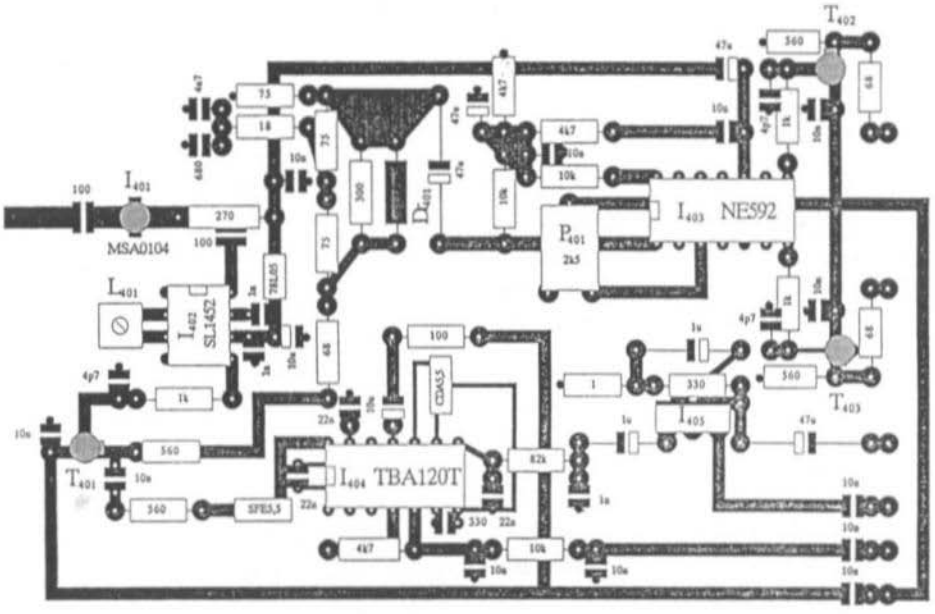


Fig.14 Component plan of the FM Demodulator DJ8ES 004

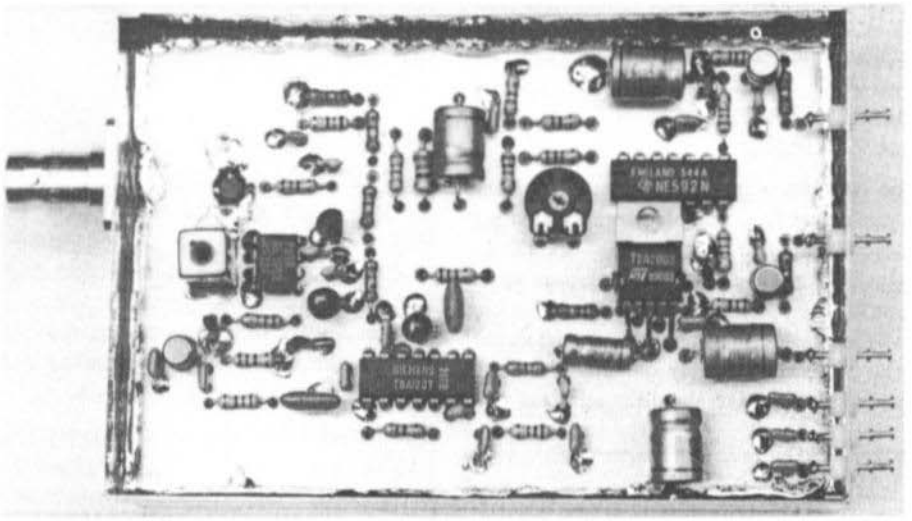


Fig.15 Laboratory sample of a DJ8ES 004 module

The audio subcarrier (5.5MHz) is led via a ceramic filter (SFE5.5) to the amplifier and demodulator TBA120T (IC404). At the same time the volume of the loudspeaker output is preset by a DC voltage: only a potentiometer (5k ohms to ground) is required. In this manner screened audio leads can be kept away from the front panel.

A TDA2003 serves as audio amplifier, with an output power of around 5 watts. A loudspeaker of appropriate rating and 4 to 8 ohms impedance is required. The power supply for this amplifier is separate from the rest of the board: this is to prevent low frequency components from the audio signal interfering with the video signal. The alternative would have been much greater filtering.

4.2. Constructional hints

The PCB DJ8ES 004 for the module described consists of a double-sided board 1.5mm thick made of epoxide material. It has the measurements 72 x 108mm, fitting in a commercially available tinplate enclosure (74 x 111 x 30mm). The track and component layout are to be seen in fig. 14.

After trimming the board is first treated with rub-on silver plating and then drilled. Further work for preparation and construction of the demodulator module is as for the previous two units. The TDA2003 needs cooling: a suitable heatsink should be fitted.

4.3. Component list for DJ8ES 004

IC401:	MSA0104 (Avantek)
IC402:	SL1452 (Plessey)
IC403:	NE592 (Signetics)
IC404:	TBA120T (Siemens)
IC405:	TDA2003 (Siemens)
T401, 2 & 3:	BC108B or equivalent

P401: Trimpot 2.5k (10mm lead spacing)
L401: Neosid coil BV5034 (0.04uH)
Dr401: Choke 33uH

3 Tantalum bead electrolytics 10uF/25V
2 Electrolytics 1uF/25V, axial
4 Electrolytics 47uF/25V, axial
1 Chip capacitor 47pF
2 Chip capacitors 100pF
1 SMD resistor 180R
1 Voltage regulator 78L05

1 Ceramic discriminator CDA5.5
1 Filter SFE5.5 1 BNC flange connector (UG-290A/U)
6 Teflon feed-throughs
1 Tinplate enclosure 74 x 111 x 30mm

Resistors (1/8W, 10mm lead spacing)

1 off	1R
1 off	330R
1 off	18R
4 off	560R
3 off	68R
3 off	1k
3 off	75R
3 off	4k7
1 off	100R
3 off	10k
1 off	270R
1 off	82k
1 off	3R

Ceramic disc capacitors (2.5mm lead spacing)

3 off	4.7pF
1 off	4.7nF
1 off	330pF
12 off	10nF
1 off	680pF
3 off	22nF
3 off	1nF

4.4. Commissioning

The following are required:

- multimeter
- frequency counter to 1GHz
- signal generator to 500MHz
- oscilloscope - FM ATV signal, reaching the converter via the antenna

To begin only the 12V supply is connected: check for a current flow of around 400mW, of which most is taken by the audio amplifier (IC404). The module should work straight-away and the volume of noise of the connected loudspeaker should vary with the potentiometer setting.

After adjusting the video amplifier for maximum amplification (pot P401 clockwise to end stop) only the demodulator coil needs aligning. This can be done by feeding in 480MHz from a signal generator to the BNC connector: the level should be about -40dBm (100uW). Now the core of coil L401 should be adjusted until a DC voltage of 2.3V is measured on pin 7 of the quadrature demodulator (IC402). If correctly aligned the core should be screwed in approximately 80 per cent.

Now the first test with a real local 23cm ATV signal can be commenced. A monitor should be connected to the video output. If necessary carefully draw out the core of the coil L401. In this way the IF can be aligned exactly on 480MHz. The best signal for checking this is a test pattern such as a chessboard: the squares around the border should display no distortion.

The video level at the output should be adjusted to the nominal value of 1V (peak to peak). The most logical way of doing this is with an oscilloscope, remembering to connect the terminating resistance of 75 ohms.

5. LITERATURE REFERENCES

(1) Grimm J., DJ6PI: ATV with Frequency Modulation. VHF COMMUNICATIONS 26 (1986) issue 2, pp 91-102.

(2) Polz R., DC2CS: IF Amplifier and Demodulator for Broadband FM. VHF COMMUNICATIONS 26 (1986) issue 1, pp 2-7.

(3) Hoefer F-R., DF4PN: FM ATV Receive Unit for 23cm. cq-DL 11/89.

(4) Plessey Semiconductors Satellite & Cable TV IC Handbook 1988.

Peter Vogl, DL1RQ

A 6cm Transverter using Stripline Technology.

As presented at the 1990 VHF-UHF conference in Munich

This article describes five building block modules for a 6cm transverter with an IF of 435MHz.

The use of stripline techniques and surface-mount devices (SMDs) in quantity enables a very compact unit to be built with a high degree of reproducibility and reliability.

The achievable noise figure of 1.5dB and power output of 1.5W represent the state of the art in amateur technology.

1. POWER SUPPLY AND CONTROL MODULE DL1RQ NT6-1

1.1. Description

The front panel (fig. 1.1) is constructed together with the power supply and control PCB as a compact module, in order to save as much room as possible.

Fig.1.1 Front panel for DL1RQ NT6-1: 145 x 40mm
Material: Epoxy PCB 1.5mm thick, copper-clad both sides

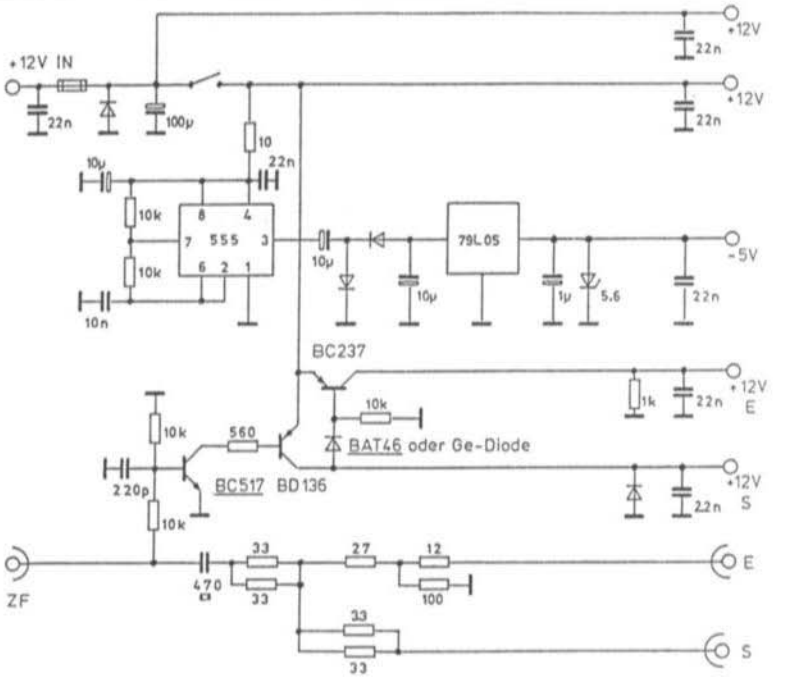


Fig.1.2 Circuit of DL1RQ NT6-1: (BAT46 may be any Germanium Diode)
E = RX; S=TX; ZF = IF

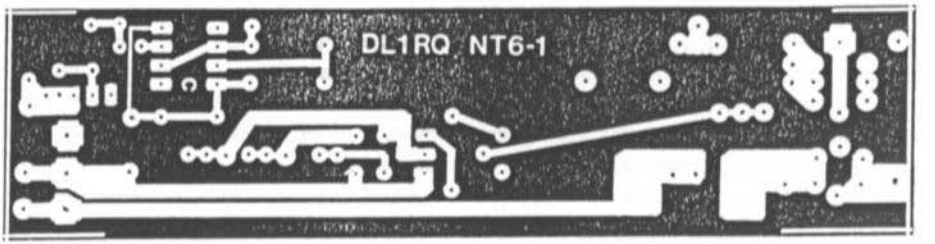


Fig.1.3 Power Supply and Control Module DL1RQ NT6-1
122 x 30mm; Material: Epoxy PCB 1.5mm thick, copper-clad both sides

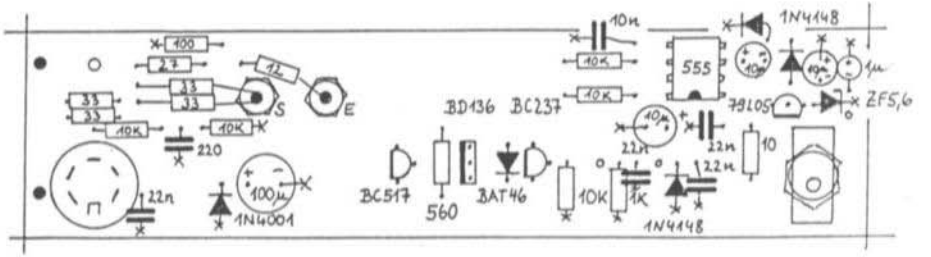


Fig1.4 Position of components on DL1RQ NT6-1

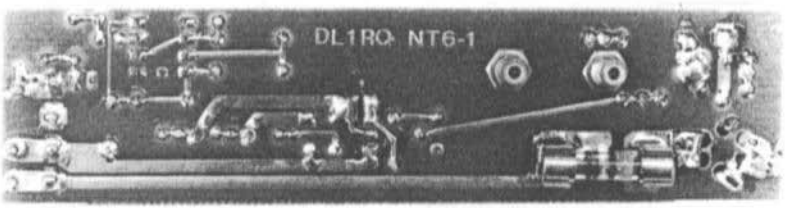


Fig.1.5 Components on conductor side of DL1RQ NT6-1
The SMC connectors are mounted from the conductor side. The fuse holder and the SMD capacitor 470pF are soldered onto the conductor side

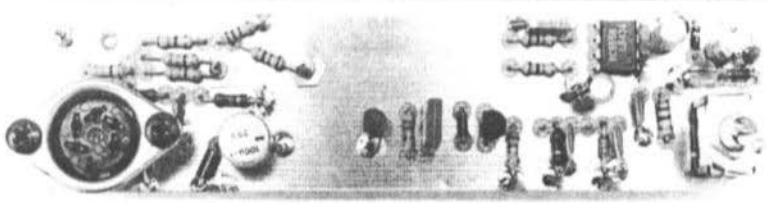


Fig.1.6 View of the component side of DL1RQ NT6-1

The circuit (fig.1.2) is protected against short circuit and reverse polarity and provides a DC-DC converter for the negative supply -5V, a transistor-controlled switch for the 12V supply voltage as well as a resistive divider with 6dB reduction in power load for the send channel and 10dB reduction for the receive side. The measurements of the front panel were chosen to correspond with the width of the FT-790 driver used by the author.

The height of the panel is laid out to match commercially-available side profiles (1). Figs. 1.3 to 1.6 show the printed circuit board DL1RQ NT6-1 and the component overlay in drawing and photographs.

2. OSCILLATOR MODULE FOR 1331.25MHz - DL1RQ 012

A special oscillator module was not designed for the 6cm transverter, instead the respected and much copied oscillator module DL1RQ 012 was put into service again. This unit has already been published in (2) and (3). For the sake of completeness of this article it will be described again, in a slightly improved version. You should be aware that it is built in a tinplate box with lid, 37 x 111 x 30mm, which makes it of larger profile than the other 6cm modules.

2.1. Description

The oscillator module is built out of four stages (fig. 2.1). Beginning with the accepted and low-noise crystal oscillator with a P8002 (or J300) transistor, the frequency is tripled in the following stage using a BFY90.

Another tripler stage with a BFR96 follows and this drives a BFR96 doubler via a two-element filter. The output frequency is coupled through a three-element filter.

Output power is in excess of 40mW, more than enough to drive the module DL1RQ 023.

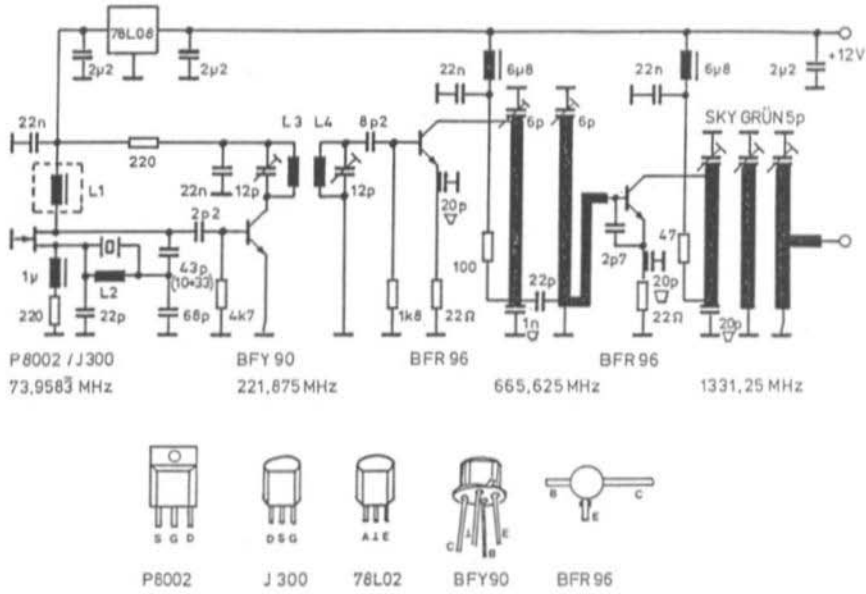


Fig.2.1 Circuit of DL1RQ 012

L1: BV5061 Neosid 115nH brown/blue
L2: 13.5cm varnished copper 0.15mm diameter on 10k resistor
L3: 2.5 turns varnished copper 1.00mm diameter air-wound on 5mm former
L4: as L3
Sky trimmers - 5pF green

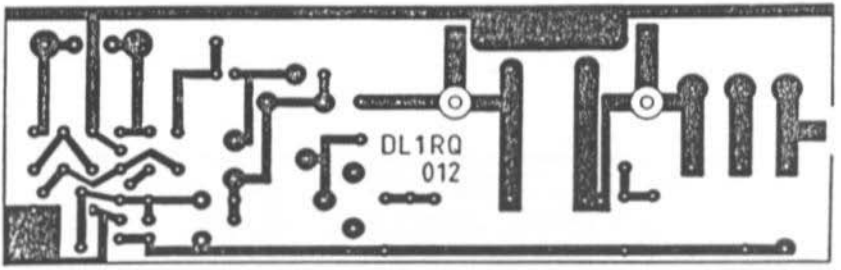


Fig.2.2 PCB for oscillator module DL1RQ 012
108 x 34mm; Material: Epoxy PCB 1.5mm thick, copper-clad both sides

2.2. Construction of DL1RQ 012

PCB, component layout and a sample construction can be seen in figs. 2.2 to 2.4.

Through contacts are indicated on the component overlay by an X in a circle and are formed either by component leads or by small pieces of 1.5mm copper wire. Components to be grounded, where no drilled holes have been provided, are soldered direct to the ground plane. This is indicated by the symbol * on the overlay plan.

The height at which the PCB is mounted in the tinplate enclosure depends on how tall the crystal used is and on the height of the crystal thermostat's insulation.

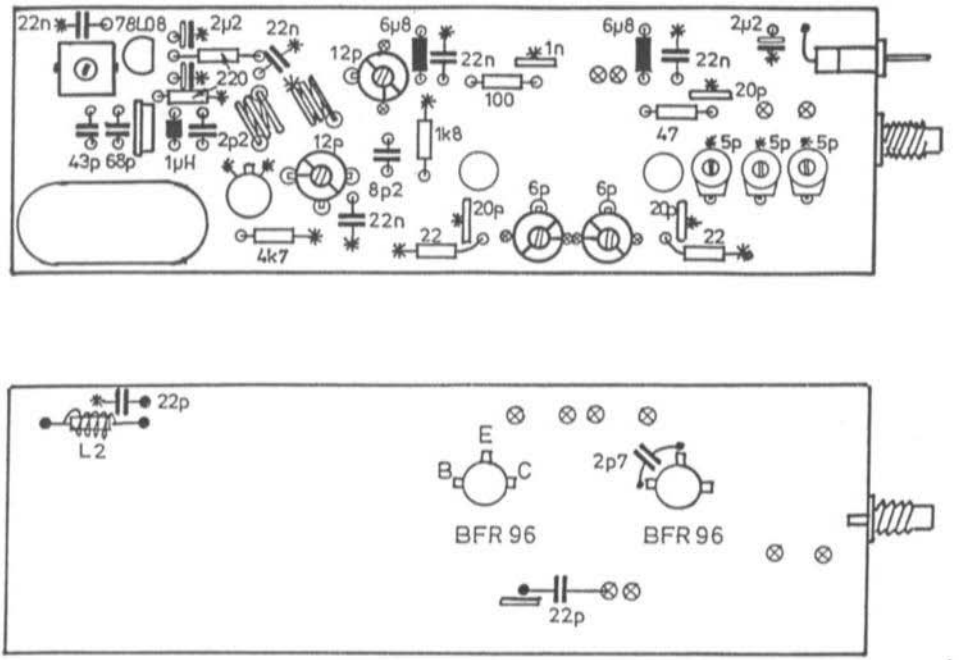


Fig.2.3 Component Layout of DL1RQ 012

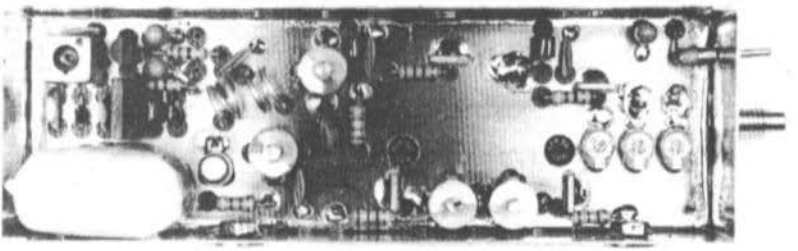


Fig.2.4 View of the completed module in tinplate enclosure.
Lid measurements: 37 x 111 x 30mm

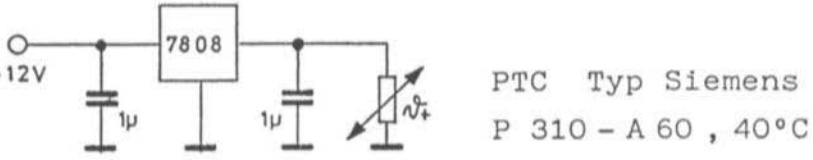


Fig.2.5 Circuit of a simple 60 degrees C thermostat
Hint: The equilibrium temperature lies *above* the indicated doubling temperature of the resistance value of the PTC resistor.

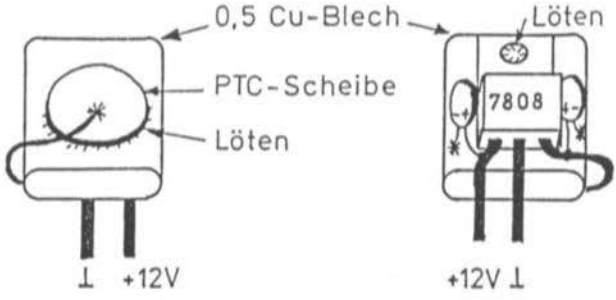


Fig.2.6 Construction details of crystal thermostat.
Cu-Blech = copper sheet material; Löten = solder; Scheibe = disc

2.3. Advice on simple crystal thermostats

Circuit and construction are indicated in figs. 2.5 and 2.6. To avoid thermal instability you need a particularly "intimate" contact between the PTC disc and the copper sheet.

The constructor soldered the PTC disc flat onto the copper sheet. Unfortunately the discs are prone to developing hair-line cracks when heated rapidly, and this leads to significant instability. The use of good flux and preheating of the PTC disc are recommended.

The thermostat is placed up against the crystal using plenty of heat-sink paste and surrounded by as much expanded polystyrene as possible.

3. TRANSISTOR QUADRUPLER 1331.25/5325MHz DL1RQ 023

3.1. Description

The quadrupler consists of two doubler circuits connected in cascade. The first doubler, using a BFR91, displays very good input matching. It operates with a fixed-tuned stripline circuit at 2662.5MHz. The second doubler is equipped with an MGF1302 and after a stripline filter delivers around 40mW output on 5325MHz. By means of a 10dB stripline coupler an oscillator level of approx. 4mW is fed out to the receiver module.

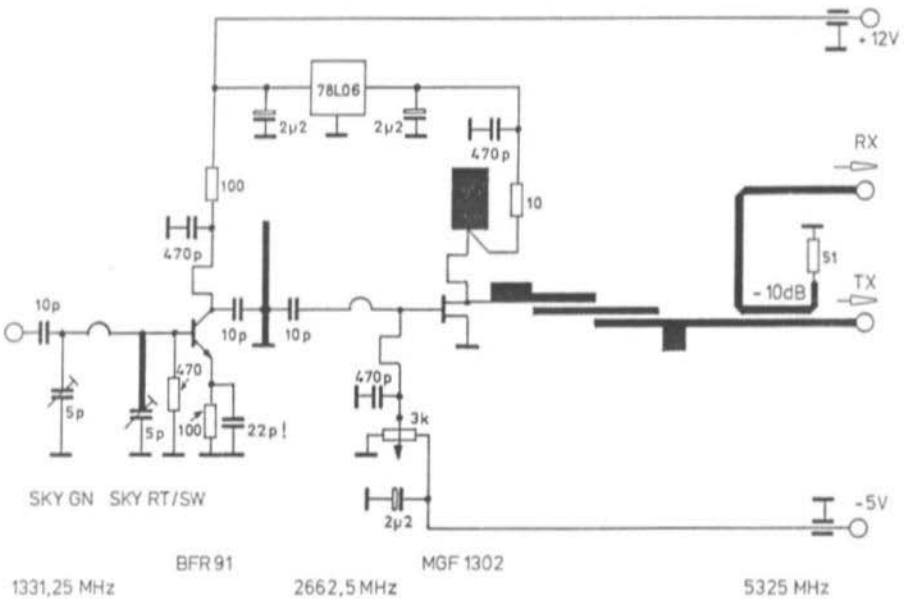


Fig.3.1 Circuit of DL1RQ 023
GN = green; RT/SW = red/black

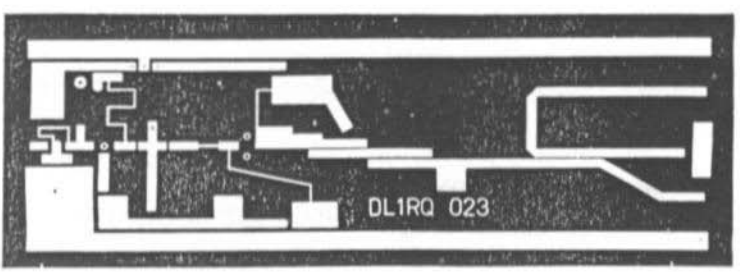


Fig.3.2 PCB for Quadrupler DL1RQ 023
108 x 34mm; MateriaL: RT-Duroid d-5870; 0.439mm

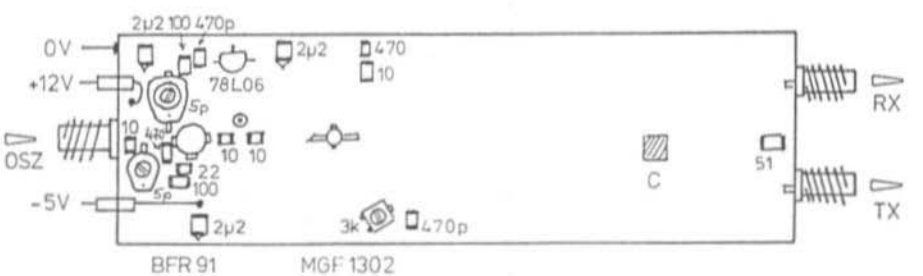


Fig.3.3 Component layout of DL1RQ 023
OSZ = oscillator

Fig.3.4 The completed module in tinplate enclosure.
Lid measurements 37 x 111 x 13

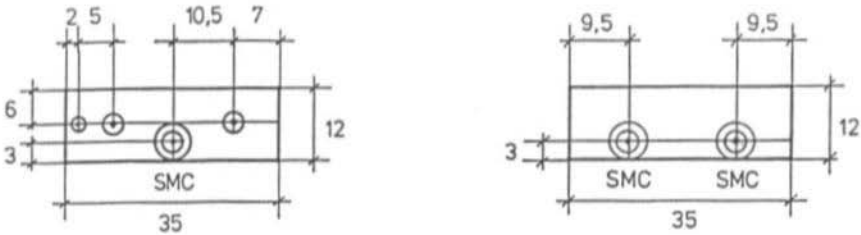


Fig.3.5 Dimensioned sketch of the end panels of DL1RQ 023

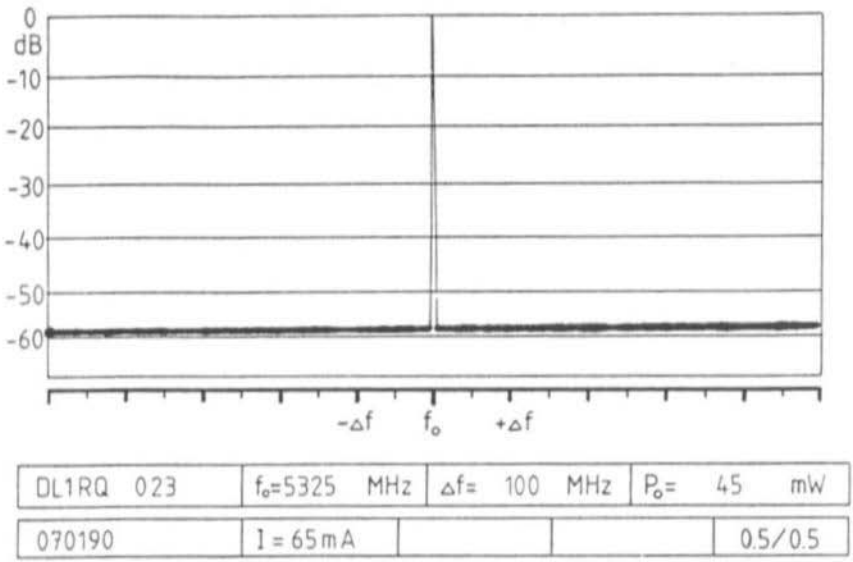


Fig.3.6 Output spectrum of the quadrupler DL1RQ 023

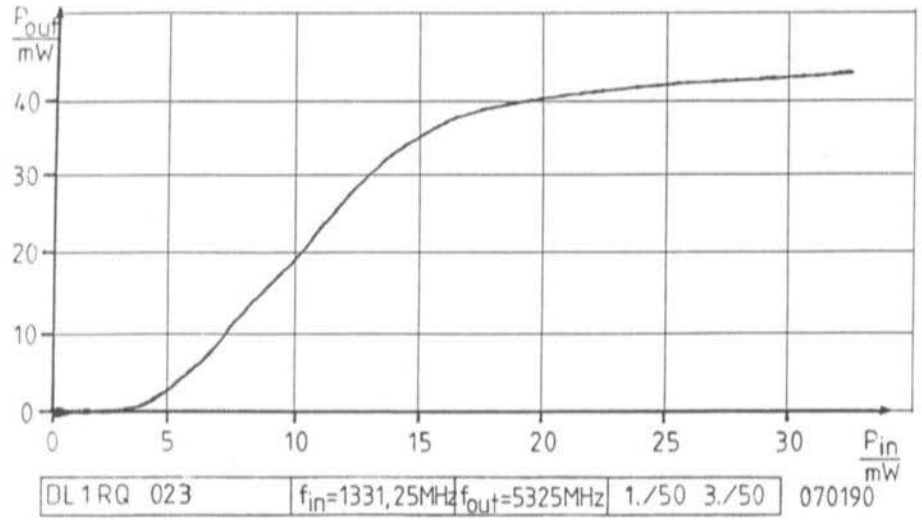


Fig.3.7 Relationship between input power at 1331.25MHz and output power at 5325MHz

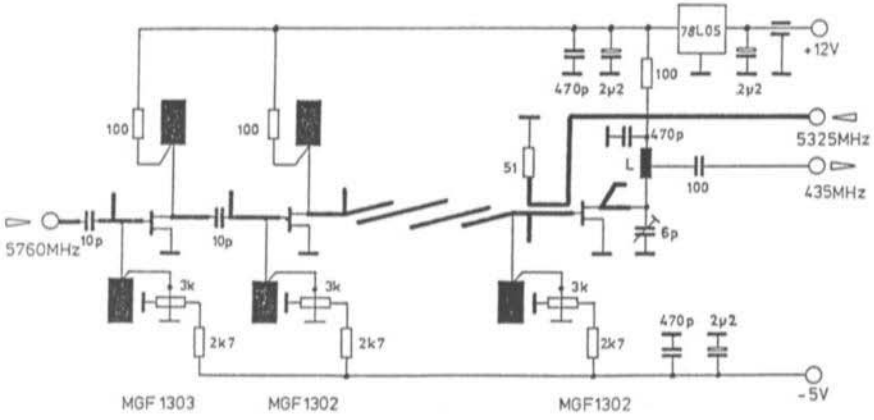


Fig.4.1 Circuit diagram DL1RQ 670
L: 2 turns of 0.5mm silver-plated copper wire on 3mm former.
All components SMD pattern except voltage regulator 78L05 and 6pF trimmer (Valvo/Mullard grey).
10pF coupling cap's: Valvo 2222 851/853 (D104.520) 2 x1.25 x 0.51mm

3.2 Construction of DL1RQ 023

PCB, component layout and a sample construction can be seen in figs. 3.2 to 3.4.

C is a 4 x 4mm piece of single-sided RT-Duroid PCB material 0.439mm thick and sets the coupling attenuation of 10dB. It is fixed with a small dab of universal glue (UHU, etc.)

Transistor BFR91 is mounted in the fashion of an SMD component. The cold end of the stripline resonator at the collector of the BFR91 must be through-contacted to the

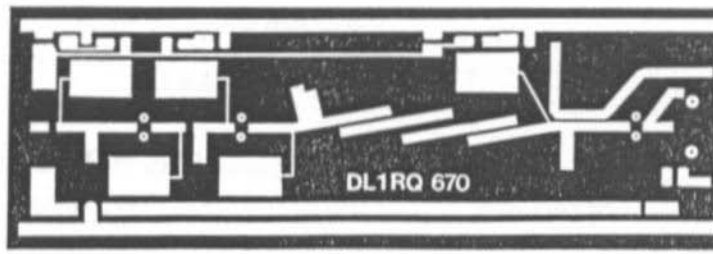


Fig.4.2 PCB DL1RQ 670
108 x 34; Material: RT-Duroid D-5870; 0.439mm

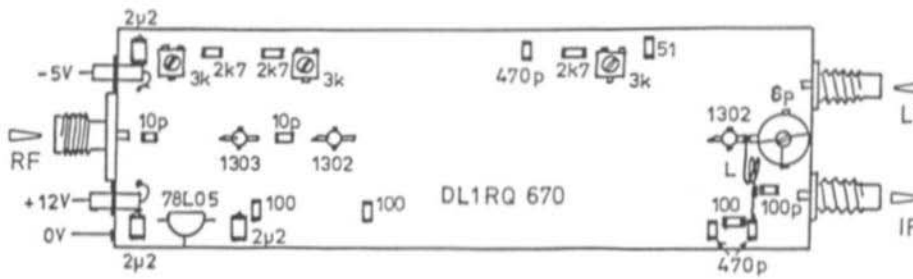


Fig.4.3 Component layout of DL1RQ 670

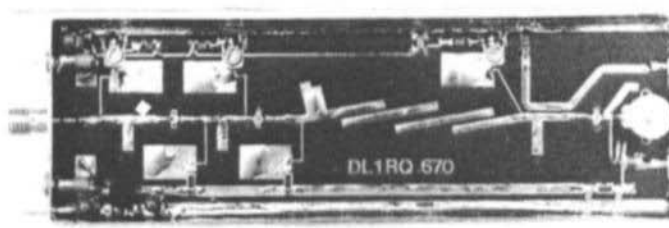


Fig.4.4 The completed module in tinplate enclosure.
Lid measurements 37 x 111 x 13

groundplane side. The source connections of the MGF1302 are bent downwards, passed through holes and soldered to the groundplane.

A tinplate enclosure with the (lid!) dimensions of 37 x 111 x 13 is not commercially available to the knowledge of the writer. Accordingly just the lid of an available box with the dimensions 37 x 111 x 30 was used [type wg3113 from (4)]. The side walls 12mm tall were home-made to match the originals.

The SMC connectors are soldered from outside the end panels (fig. 3.5). The Teflon insulation which projects inside should be removed right back to the wall of the enclosure and the contact pin shortened somewhat.

The Teflon PCB should be trimmed so that the tracks line up with the pins of the connectors, then soldered to them. After this the edges of the PCB can be made "watertight" by soldering all round to the walls of the

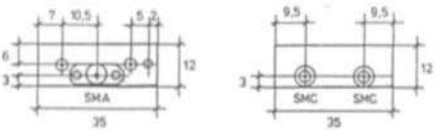


Fig.4.5 Dimensioned sketch of the end panels of DL1RQ 670

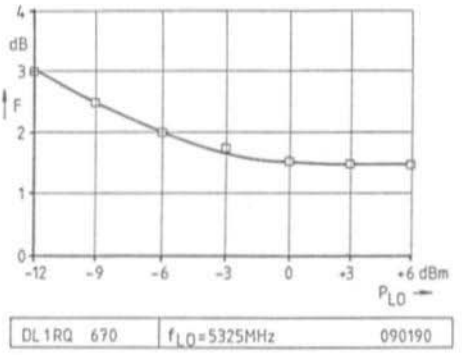


Fig.4.6 Noise measurements for the receiver module DL1RQ 670 in relationship to the oscillator perfomance

enclosure; this applies not only to the ground-plane underside but also to the side strips or islands on the track side.

3.3. Alignment hints for DL1RQ 023

The following should be set up first:

The trimmer in the BFR91's base circuit should be fully turned "out", the input trimmer turned inwards a quarter. The gate-bias trimmer of the MGF1302 should be adjusted to its middle position.

After this the input trimmer should be adjusted for best input matching. Using the base circuit and the gate-bias, tune for maximum output power. In this operation, turn the base circuit in as little as possible and watch the drain current by the voltage drop on the drain resistor!

3.4 Measured values for the frequency quadrupler DL1RQ 023

Figs. 3.6 and 3.7 shown what should be achieved with this module.

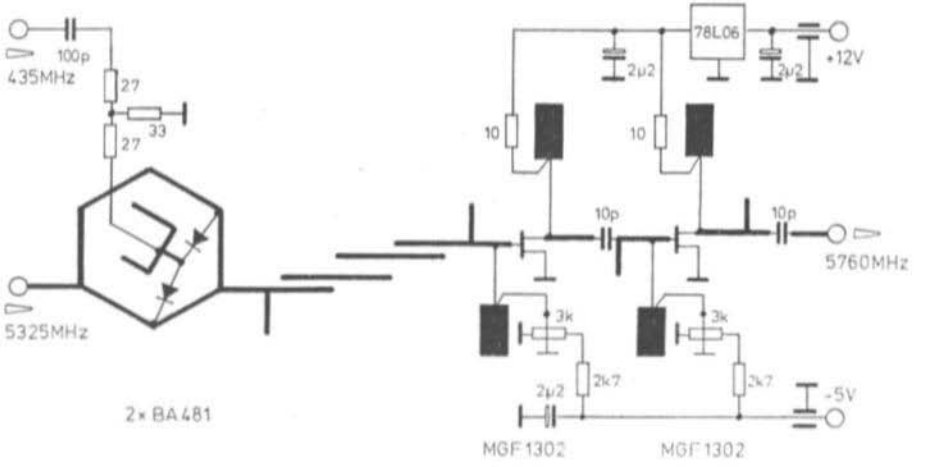


Fig.5.1 Circuit diagram of DL1RQ 706

4.
RECEIVER MODULE DL1RQ 670

4.1. Description

The receiver module (fig. 4.1) contains a two-stage preamplifier with transistors MGF1303 and MGF1302. Following is a two-pole stripline filter for selecting the receive frequency of 5670MHz. The mixer stage is realised with an MGF1302 transistor and in this form is already proven in the writer's 10GHz concept (5). The noise measurement of the receive module remains constant at 1.5dB from 1mW oscillator loading to above 10mW.

4.2. Construction of the receive module DL1RQ670

Figs. 4.2 to 4.5 show all the details.

4.3. Alignment hints and measurements

The trimmer of the 435MHz circuit should be tuned for maximum noise in the 70cm receiver connected (turned about halfway inwards). On the trimmers for the gate voltage the following drain currents are measured:

Preamplifier stage MGF1303: approx. 12mA, corresponds to 1.2V on drain resistor Preamplifier stage MGF1302: approx. 15mA, corresponds to 1.5V on drain resistor Mixer stage MGF1302: approx. 1.5mA, corresponds to 0.15V on drain resistor.

The currents of the first stage and the mixer stage should ideally be optimised on a noise measurement station, but checks on several lab. samples indicated that the techniques described achieved close to ideal noise performance (fig. 4.6).

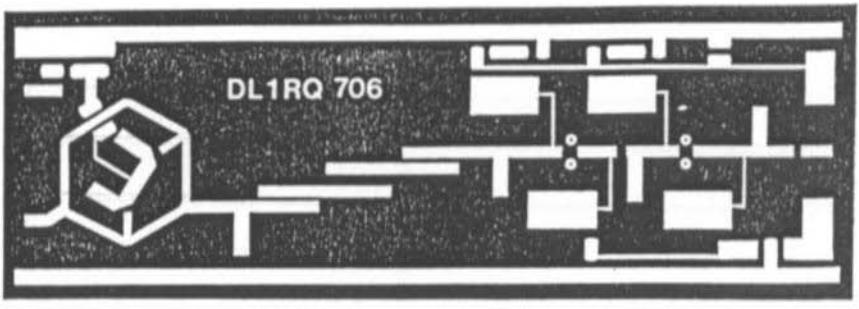


Fig.5.2 PCB DL1RQ 706
108 x 34mm; Material: RT-Duroid D-5870; 0.439mm

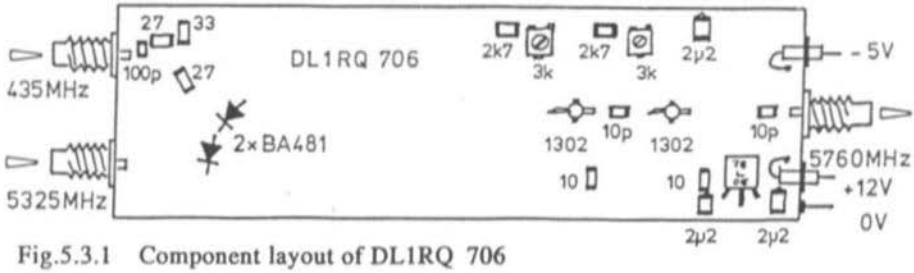


Fig.5.3.1 Component layout of DL1RQ 706

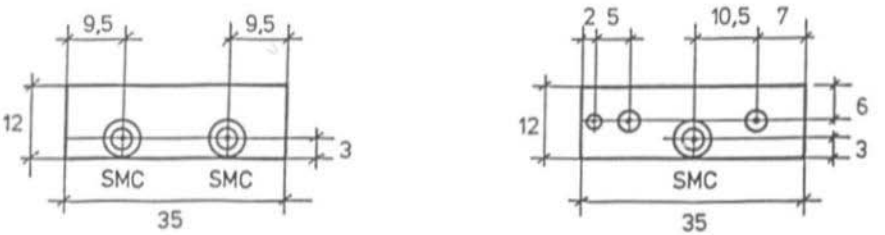


Fig.5.3.2 End plate dimensions for DL1RQ 706

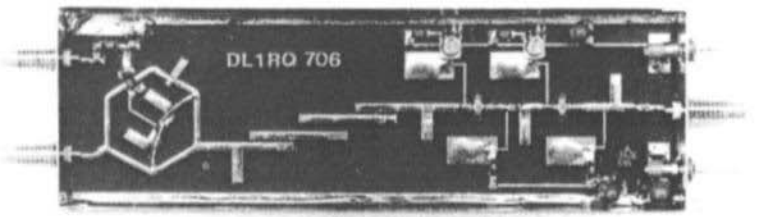
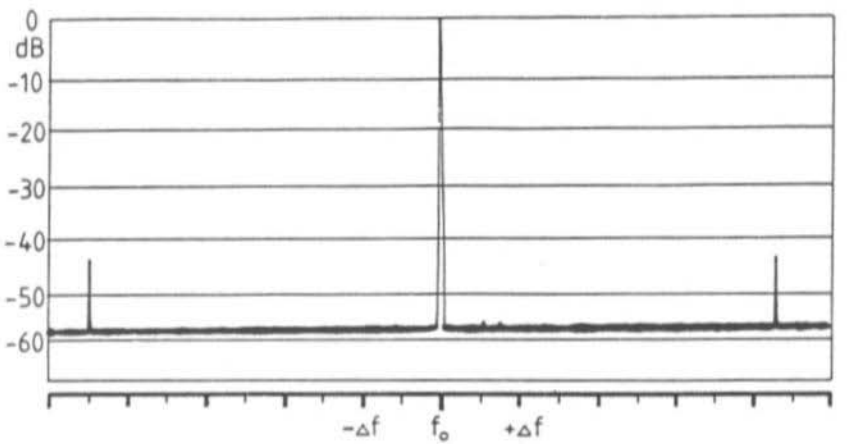


Fig.5.4 The completed module in tinplate enclosure.
Lid measurements 37 x 111 x 13.



DL1RQ 706	f_o= 5760 MHz	Δf= 100 MHz	P_o= 100 mW	
070190	I = 75mA	f_{in}=435 MHz	f_{LO}= 5325MHz	0.5/0.5

Fig.5.5 Output spectrum of the transmit module DL1RQ 706

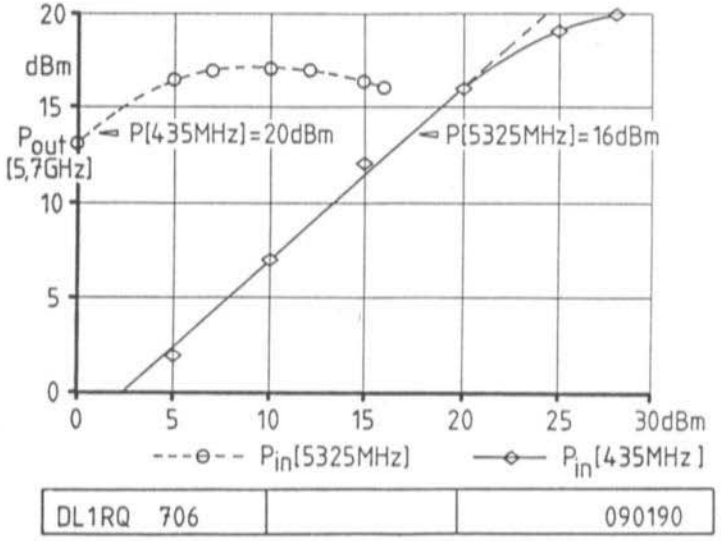


Fig.5.6 Oscillator level and linearity behaviour ofthe transmit module (with 6dB divider from NT6-1)

5. TRANSMITTER MODULE DL1RQ 706

5.1. Description

The transmitter module (fig. 5.1) contains a push-pull ring-hybrid mixer (rat-race balanced mixer) with two diodes of the type BA481. The 70cm signal which has already been attenuated 6dB in the driver module is reduced once more here by 10dB before it reaches the mixer diodes.

The oscillator power is divided in the ring and fed to the diodes in push-pull. Under ideal conditions there should be good suppression of the oscillator on the output of the mixer.

A stripline filter connected behind this suppresses the image frequency and provides additional suppression of the oscillator. Following this is a two-stage amplifier which raises the mixer signal to a maximum of 100mW.

Figs. 5.2 to 5.4 give the details necessary for construction.

5.2. Alignment hints for the transmitter module DL1RQ 706 and measurements

The only alignment is adjusting the quiescent current of the two transistors:

First stage MGF1302: approx. 30-40mW, corresponds to 0.3-0.4V on drain resistor
Second stage MGF1302: approx. 40mA, corresponds to 0.4V on drain resistor.

Figs. 5.5 and 5.6 indicate first-class linearity and adequate suppression of undesired mixer products.

Arising from these measurements it is evident that the mixer presented here is very tolerant of the oscillator levels offered to it.

TO BE CONTINUED

(References and suppliers overleaf.)

7.
LITERATURE AND SOURCES OF SUPPLY

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(2) 10GHz Transverter using Stripline Technology. Peter Vogl DL1RQ. 1986 VHF-UHF Conference notes, pp 58ff. South Bavaria District of DARC.

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(4) Giga-Tech Karl Himmler, Postfach 1160, W-6805 Heddesheim, Germany. Modules and special components for the GHz-Amateur.

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Bernd Neubig, DK1AG

A Modern, Professional Look at the Design of Stable, Crystal Oscillators Working at High Frequencies Part 1

From a lecture given to the 13th Session (10th Feb. 90) of the GHz- convention in Dorsten.

1. CHARACTERISTICS OF OSCILLATING CRYSTALS AT HIGH FREQUENCIES

1.1 Frequency Range and Overtones

Stable, high-frequency Quartz oscillators, almost without exception, use AT-cut crystals. The frequency determining mode is that of the thickness-shear, which means the thickness of the plate determines the frequency of oscillation according to the expression:

$$f(kHz) = N1660/d \text{ where } N = 1, 3, 5, ... [1]$$

The upper frequency limit for the basic Quartz (N = 1) is determined by the technology available in the manufacturing process - the precise working of two thin surfaces, which must be absolutely smooth and plane parallel to each other. Nowadays, these limits are reached, for example, by the firm *Telequarz*, under production thickness tolerances of 40 to 45um - which corresponds to a frequency of 40 to 45MHz. In the laboratory, however, crystal wafers of 22um thick (75MHz fundamental) can be achieved. By etching out a thin membrane of 50um from the centre of a thick Quartz slab by means of an ionic-beam process, a fundamental frequency of 300MHz can be obtained.

The disadvantage of these high-frequency fundamental crystals is that their Q decreases inversely as the frequency increases, such that the product Qf remains constant. A rule-of-thumb would be a figure of 10^{12}Hz. Therefore, a crystal having a fundamental frequency of 10MHz would have a Q around 10^5, whereas the 300MHz crystal would have a Q of only 3000.

As a rule, overtone crystals are employed at frequencies above 30MHz, with factors of N = 3, 5, 7, 9 or 11. The Q is considerably

higher at the 3rd-overtone, Qf = 4-7 * 10^{12}Hz, but from the 5th-overtone on the dynamic capacitance C_1 reduces, and with it the crystal's ability to be pulled.

At the 7th or 9th overtones it is very difficult to pull more than 5 - 10ppm, and this is very important for tuning and modulation purposes.

A reasonable limit for overtone working would be the 9th overtone, which, at 40MHz fundamental, would give an upper final frequency limit of 360MHz. This limit is exploited frequently in professional applications, but 100 - 150MHz seems to be the end of this particular world for amateurs.

In theory, all crystals possess an ability to oscillate at the fundamental frequency and on their overtones, but the design is always optimized for the desired overtone.

1.2 Spurious Resonances

In addition to its fundamental and overtone resonances, every crystal has a number of undesirable spurious modes of oscillation. These are not harmonically related and always lie above the fundamental mode, at distances varying between 10 and 120kHz. These spurious resonances are some 10 - 20dB down on the fundamental (plane-parallel crystals), but overtone crystals are prone to stronger spurs and they are closer to the main resonance (Fig.1).

At the 3rd-overtone 6-9dB can be reckoned with and at the 5th-overtone and higher, 3dB is often difficult to achieve. At frequencies near the limit (e.g: 300MHz using 9th-overtone) 2dB is the most that can be expected. Very often, the first spurious resonance lies only some 20 - 30kHz above

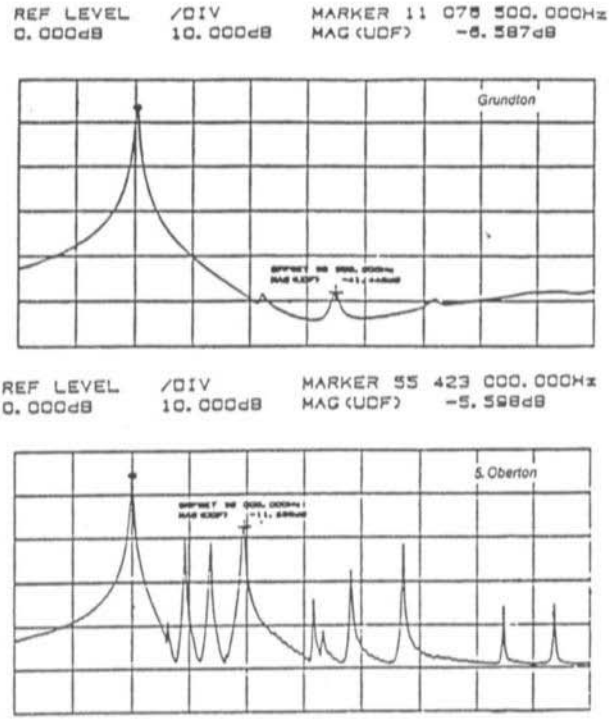

Fig.1 Spurious Resonances present with Fundamental (top) and 5th-Overtone

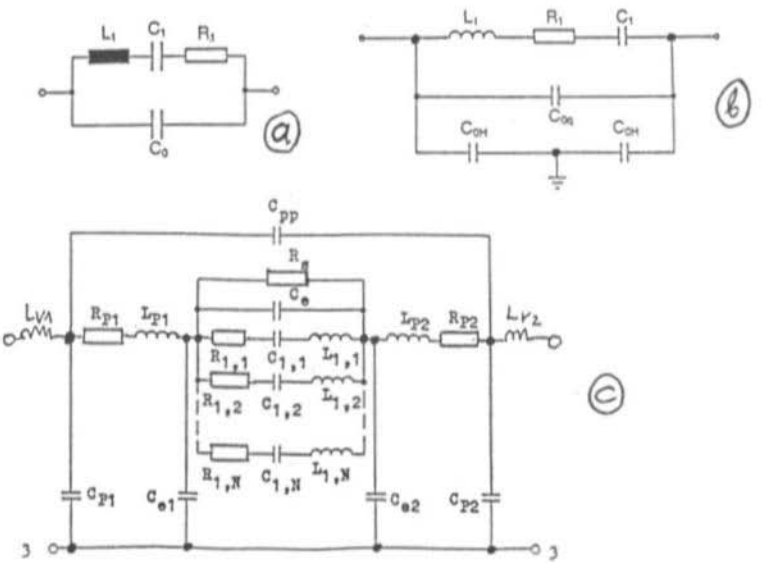

Fig.2 Crystal Equivalent Circuits

the main resonance. It cannot really be predicted with any certainty what oscillation will break out in a crystal using any particular oscillator circuit. The amplitude of the spurious oscillations is controlled by the size of the vibrator electrode. That is, the capacity of C_1 and the conductance of the crystal. The demands for a better suppression of spurious resonances means, very often, a crystal with small pulling ability and a higher resonant resistance R_1.

1.3 Equivalent Circuit

The well-known equivalent circuit of an oscillating crystal shown in Fig.2a is not sufficient to show its behaviour at high frequencies.

The housing of the crystal should always be earthed in applications demanding the highest frequency stability. The capacitance of the glass connector feed-throughs amounts to 1pF, and can no longer be neglected at the higher frequencies. The spurious resonances are represented by additional series-tuned circuits. This leads to the more representative circuit in Fig.2b. Later the influence of the internal and external connection leads, together with mounting pins and sockets for frequencies above 150MHz must be taken into account. All this is shown in Fig.2c.

1.4 Locus-Curve diagram

The high frequency behaviour of the impedance or admittance of electronic components is represented by a locus diagram. The complex conductance of the simple equivalent diagram of Fig.2a is represented, near the resonant frequency, by a circle whose centre at the low frequencies lies just above the real axis. The (right-hand) intersect with the real axis is the resonant frequency f_r (Fig.3a).

At high frequencies the centre point moves upwards owing to the influence of Co and resistance R_1. The resonant frequency f_r, at which the crystal impedance is real, separates from the extreme right-hand point of the circle (the series resonance frequency f_s in the

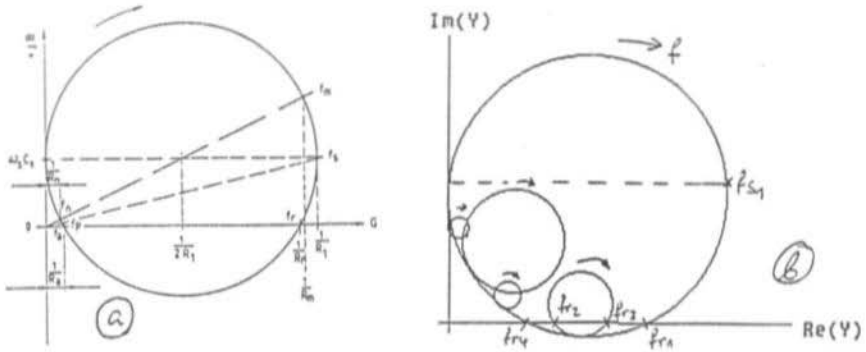

Fig.3 Locus-Curve Diagrams
a) ideal; b) with spurious resonances

equivalent circuit). Also, the frequency f_m, at which the crystal is at its lowest impedance, wanders away from f_s. Above a certain frequency the locus diagram no longer intersects the real axis, that means that there is no real working impedance in the vicinity of the main resonant frequency. Most crystals exhibit this characteristic above the 5th or 7th overtones, 120 to 150MHz.

The static capacity C_0 must be compensated for stable and reliable operation. This is carried out, as a rule, with a parallel inductance, but also - as in the case of crystal filter circuits - with a differential transformer, together with a compensation capacitor in the opposite leg. The compensation inductor has the following value:

$$L_p = 1 / ((2 * PI * f_o)^2 * C_0) \qquad [2]$$

where C_o is the not-to-be-forgotten winding and stray capacitance of the circuit. The inductor must not have a very high Q and does not need to be tuned. It is recommended that the above formula [2] should be used for the calculation of the value and the nearest lower value chosen from the standard series of chip inductors.

At still higher frequencies the locus diagram will become distorted from the ideal owing to the near presence of the spurious resonances, which give rise to pigtails on the main circle. Lead inductance losses force the circle away from the Im-axis (Fig.3b). This deteriorates the impedance relationship between the resonant and anti-resonant frequencies and the crystal becomes unstable.

2. CRYSTAL OSCILLATOR CIRCUITS FOR VHF AND UHF

Many circuits have already been dealt with in earlier editions of VHF Communications. Here a few important aspects and new developments of the intervening years will be considered.

2.1 Clapp-Guriett Oscillator

The Clapp-Guriett crystal oscillator (Fig.4) is a derivative of the ubiquitous Colpitts circuit. In order that reliable overtone operation can be achieved, the normal Colpitts capacitive-divider capacitors C_a and C_b between base, emitter and ground are modified by replacing the latter by a tuned circuit L_eC_e. Its resonant frequency:

$$f_{ok} = 1/(2 * PI * L_e * C_e) \qquad [3]$$

is chosen so that it lies between the desired overtone frequency and the next lower

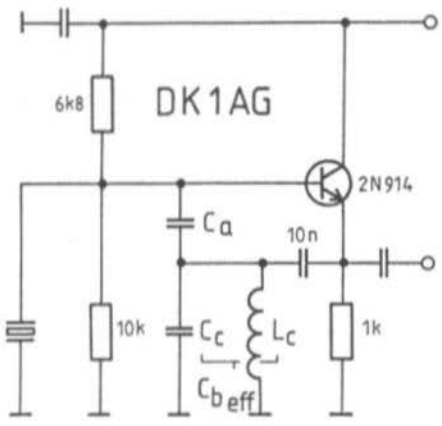

Fig.4 Clapp-Guriett Oscillator

overtone (or fundamental). This makes the tuned circuit capacitive above f_{ok} and the feedback conditions are thereby maintained. Below f_{ok} the tuned circuit is inductive and the combined oscillator circuit cannot support oscillations.

Both values of the capacitors C_a and $C_{b\ eff}$ should be of the same sort of value, but it may be advantageous for the latter to be up to 20% smaller in value - but never greater than C_a. the reactance of C_a should be, at the higher frequencies, from 20 to 50 Ohms - but no more than 100 Ohms. It may be determined by the following expression:

$$C_a = 1/(2 * PI * f_n * X_{ca}) \qquad [4]$$

where X_{ca} = 20 to 50 Ohms and $C_{b\ eff}$ must be equal to or less than C_a. [5]

With the resonant frequency of the tuned circuit f_{ok} between the overtones:

$$f_{N-2} < f_{ok} < f_N \qquad [6]$$

L_c and C_c are calculated as follows:

$$L_c = 1/((4 * PI^2 * C_{b\ eff}) * (1/f_{ok}^2 - 1/f_n^2)) \qquad [7]$$

$$C_c = C_{b\ eff}/(1 - (f_{ok}/f_o)^2) \qquad [8]$$

The crystal then works with a load capacity C_L of about:

$$C_L = (C_a * C_{b\ eff}) / (C_a + C_{b\ eff}) \qquad [9]$$

There is, however, no point in specifying a load capacity when ordering crystals for use above 100MHz as the measurement of this capacity is a very involved process. Instead, the crystals should be ordered as series-resonant, together with a deposition which must be pre-determined experimentally. The crystal manufacturers will most probably be helpful in obtaining a value for the deposition §f/f (§ represents Delta), but it may be calculated as:

$$§f/f = -C_1/2(C_o - C_L) \qquad [10]$$

Values for the dynamic capacitances C_1 and C_o can be obtained from manufacturers' literature (4), C_L can be calculated using formula [9].

The Clapp-Guriett oscillator has the advantage that the crystal sees only a de-tuned circuit and nothing on the working frequency. The tuned circuit does not require any tuning and the oscillator frequency does not react so much to variations of L_c. Its use on higher overtones (after the 7th) is, however, more difficult, because the relative frequency separation of the overtones become smaller, thus necessitating a higher tuned circuit Q, and tuning then becomes axiomatic.

2.2 The Clapp-Oscillator circuit

The grounded base circuit of Fig.5 is well established, and is therefore used very extensively. Its disadvantage is that the oscillator

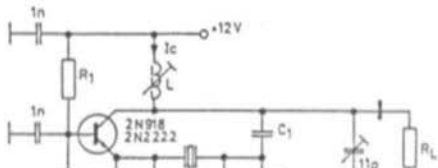

Fig.5 Clapp Oscillator (grounded-base)

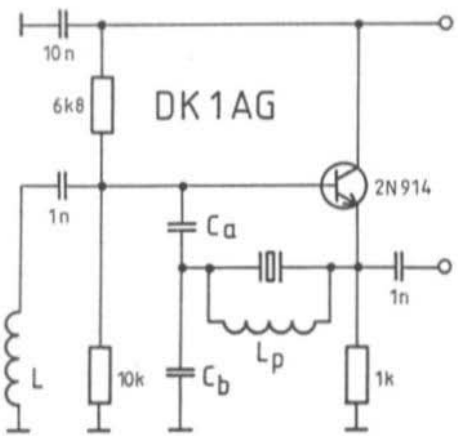

Fig.6 Crystal-stabilized Colpitts Oscillator

stage has a very high voltage gain, and therefore has a tendency both to overload the crystal element and also to develop parasitic oscillations at frequencies between 600 and 1000MHz. In view of this the following points should be observed:

- The transit frequency of the transistor should be at least four times that of the intended frequency of operation at the chosen collector current. For the lowest phase-noise, a low-noise LF transistor, with a just adequate transit frequency, is very often better than a fast microwave transistor.

- Attention to the extensive and effective use of both RF and LF decoupling measures, which avoid the establishment of loops, and also using good RF layout practice, will prevent the setting up of parasitic oscillations. Of particular importance is the provision of a good short ground for the base lead.

- The collector tuned circuit must have a low working Q and be damped with a 1 to 2k resistor.

- The oscillator is not a PA stage! This means that it should not be expected to deliver much power: -10dBm at the most. The supply voltage (or U_{ce}) should also be considerably lower than 12V.

- When it is properly tuned, the crystal will oscillate precisely when it is in the series mode, and above 100MHz the C_o compensation with a parallel inductor is employed. By properly tuned is understood to mean: the crystal should be replaced by a resistor having a value equal to that specified for R_1 of the crystal. The oscillator should then oscillate reliably. Tune the collector circuit for oscillation at the nominal frequency. After refitting the crystal, the oscillator should then function to within a few ppm of the crystal's nominal frequency.

- The capacitive voltage divider in the collector circuit should be so dimensioned that there can be no large excess of gain. The gain of the basic circuit is:

$$V = I_T / U_u R_c \qquad [11]$$

with U_T = 26mV at 25°C. The collector resistor R_c is determined by the Q_L of the tuned circuit inductor L_c and the loading of the following stage R_L, together with the impedance transformed by C_a, C_b of the series circuit of the crystal's R_1, and the emitter resistance r_e, as follows:

$$R_c = R_p \| R_L \| (R_1 * r_e) * \ddot{u}^2 \qquad [12]$$

with $$R_p = Q_L * 2 * PI * f_o * L_c \qquad [13]$$

and $$r_e = U_T / I_c \qquad [14]$$

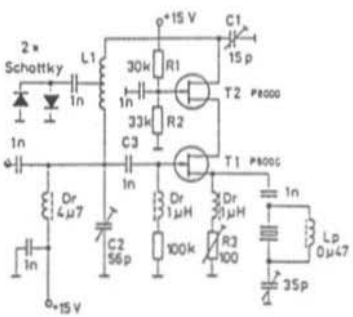

Fig.7 Hartley LC Oscillator with Crystal in Emitter feedback

also $\ddot{u} = (C_a + C_b) / C_a = (C_b / C_a) + 1$ [15]

The capacitive divider C_a, C_b must now be so dimensioned that the transistor amplification V, the attenuation k of the feedback network C_a, C_b and the voltage divider formed from the crystal's R_1 and the emitter resistor r_e, fulfils the well known conditions for feedback:

$$kV > 1$$

by a factor of two, to provide some gain reserve.

Example: Crystal oscillator 100MHz in 5th-overtone, HC-45/U, $R_1 < 100$ Ohms. Transistor $f_T = 800$MHz at $I_c = 5$mA.

Collector tuned circuit: $L_c = 100$uH, $Q_L = 30$, $\ddot{u} = 6$.

$$C_{cg\,Total} = 1 / ((2 * PI * f_o)^2 * L_c) = 25.33pF$$

$$C_a / C_b = \ddot{u} - 1 = 5$$

From this:

$$C_a = 25.33pF * 6 / 5 = 30pF$$

$$C_b = 5 * C_a = 150pF$$

$$R_p = 30 * 2 * PI * 100MHz * 100uH = 1.885k$$

$$r_e = U_T / I_c = 5.2 \text{ Ohms}$$

$$(R_1 + r_e) * \ddot{u}^2 = 105.2 * 36 = 3.79k$$

$$\text{for } R_2 R_p R_p = 1.885k \| 3.79k = 1.26k$$

Therefore, the amplification is:

$$V = I_e / (U_T * R_c) = 242$$

On the other hand, the attenuation of the feedback network (collector to emitter) is:

$$k = r_e / (r_e + R_1) / \ddot{u} = 5.2 / 105.2 / 6 = 0.00824$$

therefore:

$$kV = 242 * 0.00824 = 1.99$$

The conditions for oscillation (kV > 1) are therefore fulfilled by a factor of 2.

2.3 Crystal-Stabilised Colpitts LC oscillator

The circuit of Fig.6 shows a crystal also working in the series-resonant mode.

As opposed to the circuit of Fig.2.b, this circuit has the advantage that the frequency determining portion of the circuit has a very low resistance, and the voltage gain is only about unity.

Also, the RF grounding of the collectors is not so critical, as is the base of the Clapp circuit.

This circuit can work reliably and well up to some 300MHz and, with a correct design, there will be no problems with parasitic oscillations.

The method of arriving at the circuit dimensions and the tuning procedure is very similar to that described above.

2.4. Hartley LC oscillator with the Crystal in the Emitter feedback

A more developed form of this circuit (Fig.7) was described by the author in (6) as a cascode using two high-current FETs. This was successfully constructed by C.H.Schübbe DJ6JJ (7) amongst others. Its chief advantage is that the crystal is remote from the tuned circuit.

The danger is also present here that the crystal can be overloaded by too great a degree of feedback, or too much emitter-collector potential. A correct design will result in very good values for phase-noise.

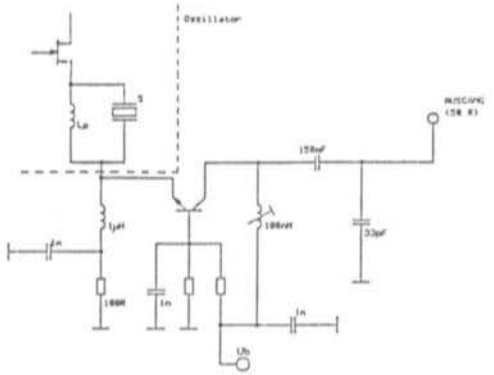

Fig.8 Using the Crystal also as a filter

In order to further improve the phase-noise characteristics of the output signal, the crystal can, at the same time, also be used as a filter. This is done by taking the current direct from the crystal, e.g: via a transformer in the base circuit as shown in Fig.8.

Part-2 of this article will appear in issue 2/91.

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(2) Neubig.B., DK1AG: *Design of Crystal Oscillator Circuits. Part-2.* VHF Communications Vol.11, Ed. 4/1979, pp 223-237.

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(5) Rech.W.H., DF9IC: *47GHz SSB Komponenten und - Baugruppen.* Tagungsband 12. GHz-Tagung VHS-Dorsten (11.2.89) pp 23ff.

(6) Neubig.B., DK1AG: *An Extremely Low-Noise 96MHz Crystal Oscillator for UHF/SHF Applications.* VHF Communications Vol.3, Ed. 3/1981, pp 135-143 and Ed. 4/1981 pp 194-203.

(7) Schübbe.C.H., DJ6JJ: *Anfordderungen an Quarzoszillatoren für Mikrowellenbaugruppen mit Lösungsbeispeil.* Tagungsband 12 GHz-Tagung VHS-Dorsten (11.2.89) pp 69-81.

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10GHz ATV The Easy Way
Part-1

At last year's British Amateur Television Club's convention, along with Dave G8NND, Steve G8JMJ and Bob G8OZP, I put on a display of my simple 10GHz ATV station. A great many people showed a lot of interest in the 10GHz set up because of the quality of the pictures and the simplicity of construction. Basically, all that I have done is to put together a lot of ideas from various publications, including the RSGB VHF/UHF Manual 4th Edition, The Microwave Newsletter and CQ-TV. My thanks to all.

1. THE TRANSMITTER

1.1. The Circuit Diagram and Construction

I use a separate transmitter and receiver utilising Solfan heads without the in-line mixer diodes for both. It is quite permissible to use Solfan heads with in-line mixers, but the results obtained are not as good.

The circuit diagram of the transmit modulator and Gunn diode power supply is shown in Fig.1, the PCB layout in Fig.2 and component overlay in Fig.3.

Once the board has been assembled and checked for short circuits and mis-placed components, **WITHOUT** the Gunn diode connected, apply the 12V supply. Confirm that by adjusting the 10-turn preset potentiometer that the output voltage to the Gunn diode can be varied between 6 and 10 volts.

Re-set the Gunn supply to 7 volts. Turn off the supply and connect the head unit, not forgetting to fit the 470-ohm resistor and the 0.01uF capacitor across the Gunn diode (see Fig.4).

WARNING: Gunn diodes are negative resistance devices, the lower the bias voltage, the lower the resistance and hence the higher the current. If bias below +5.5 volts is applied to the Gunn diode it may be damaged.

Also, NEVER look into a head unit with the power switched on. Similarly, it is inadvisable to look into, or stand in front of, an

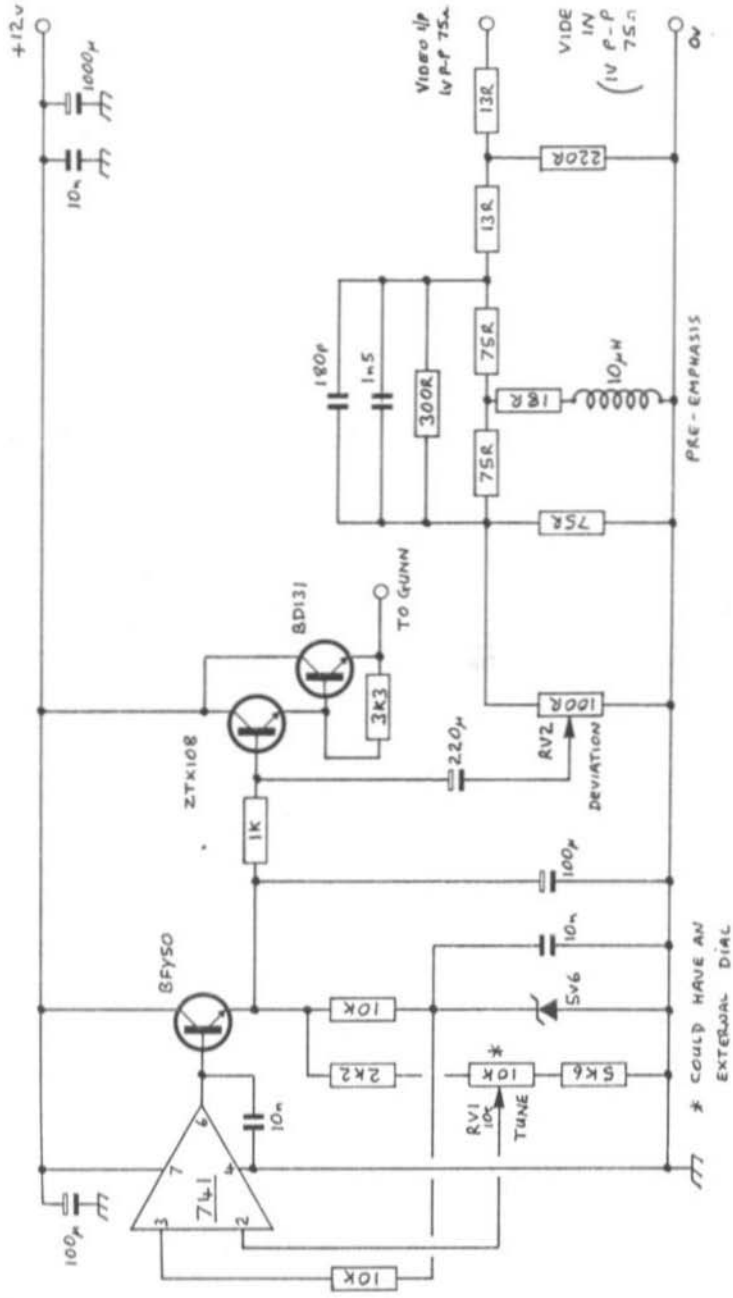

Fig.1 Transmitter Modulator and Gunn Diode supply.

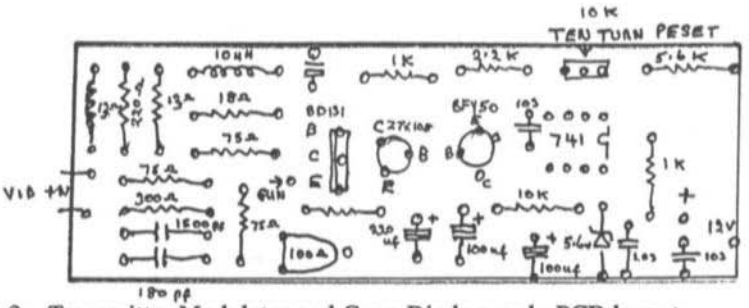

Fig.2 Transmitter Modulator and Gunn Diode supply PCB layout

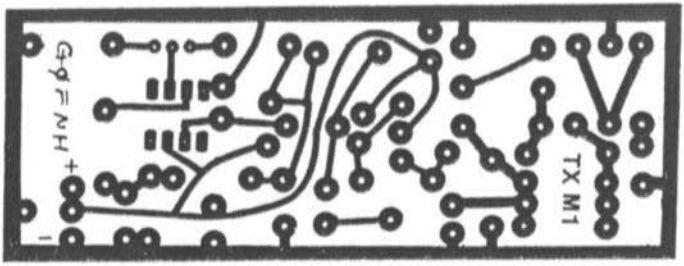

Fig.3 Transmitter Modulator and Gunn Diode supply PCB component overlay

illuminated dish or launching unit. **MICROWAVE RADIATION CAN BE DANGEROUS.**

1.2. Setting up a Gunn Diode Head

The transmit frequency should be set to 10.250GHz, and if an SHF frequency counter is not available this alternative method can be used: Remove any components at the Gunn diode. (Whilst microwave devices of this type are not particularly sensitive to handling, it is advisable to keep any contact with unconnected Gunn diodes to a minimum. Also, the use of isolated soldering equipment is recommended).

If a Solfan head is being used then it is likely that the oscillator will be set to around 11Ghz. Initially undo the brass locknut and screw in the brass screw by 1-turn. DO NOT move the steel screws yet. Fix the head unit firmly to a flat surface and apply about +7.5 volts to the Gunn diode. Connect a multimeter between the mixer diode and ground with the Gunn supply on a current of between 1 and 3mA should be shown on the meter.

Take a flat piece of metal and, keeping it parallel to the mouth of the horn, move it slowly away and watch for a minimum reading on the meter. When this condition occurs mark the position of the plate. Continue moving the plate outwards until another position of minimum current is found and again mark the position of the plate. The distance between the two marks is half the wavelength of the oscillator and from this can be determined the frequency.

The accuracy of this method can be improved by finding several minimum positions and averaging the result. If the frequency is not correct make small adjustments to the BRASS screw until the correct frequency is obtained, turning the screw in reduces the frequency and screwing it out raises it.

The following equation is used to determine

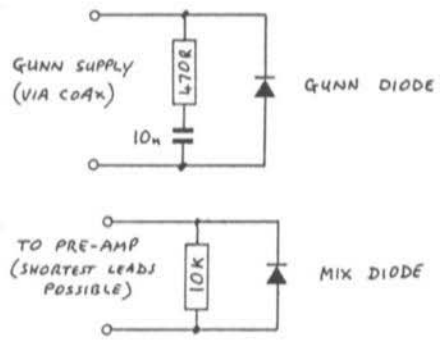

Fig.4 Connections to the Gunn Diodes

the actual frequency from the wavelength measurements:

Frequency = 150/distance mm

If a transmit head unit other than the Solfan type is used, the same method as above can be employed to calculate and thus set the oscillator frequency, but some experimentation with the tuning screws will be required to ascertain the correct procedure.

Finally, fine-tuning of the transmit frequency is obtained by adjusting the 10-turn preset on the modulator PCB.

1.3. Setting up the Modulator

Apply a 1 volt peak-to-peak video signal to the input and, with the aid of an oscilloscope, confirm that approximately 50mV of video is superimposed onto the Gunn diode supply voltage whilst varying the deviation potentiometer. If an oscilloscope is not available, set the potentiometer towards its low end and then adjust as necessary for best pictures on a receive system.

If required an external tuning potentiometer can be used so as to vary the transmit frequency whilst in use. However, I have found, as have most other operators, that it is better to have the transmit frequency fixed and to tune the receiver, as is done on the other bands.

The modulator PCB should be fitted into a small die-cast box, with the Solfan head mounted onto it and a small dish or horn aerial mounted onto the head. Sockets for the video input and DC power supply can then be fitted to the box as required.

2. REFERENCES

1) The British Amateur Television Clu journal CQ-TV, various issues.

2) The Radio Society of Great Britain VHF/UHF Manual 4th Edition.

Bring your index up-to-date !

The index of the preceeding year is always inserted in the middle of the Spring (this) edition of the magazine. It can be easily removed to update a comprehensive index. You will find the index for Volume 22 1990 on pages 31 to 34 of this issue.

A. Vilaseca, HB9SLV and J.P. Morel, HB9RKR

Trials and Modifications of a 23cm Amplifier

Here is a brief description of a small UHF amplifier for raising the output of a 23cm transverter from 30mW to around 500mW, the latter being the input drive level which a hybrid amplifier module such as the M57762 requires.

As we always do under the circumstances, we searched through our back numbers of VHF COMMUNICATIONS and immediately found what we were looking for in the article "Linear Amplifiers with Transistors for the

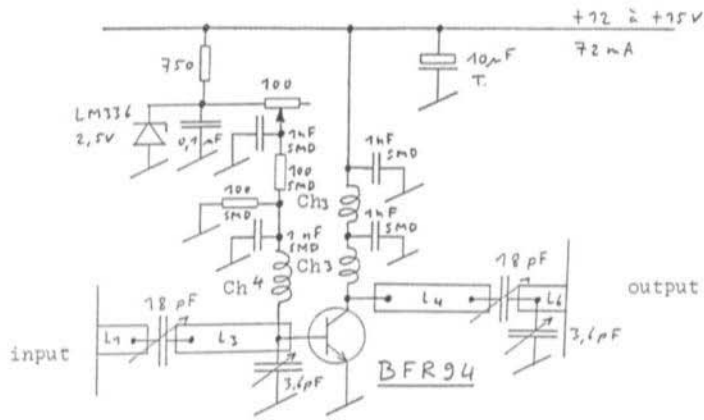

Fig.1 Circuit of the Wideband Amplifier for 200-1400MHz
L1, L6: 5 x 2.5 (50 ohms); L3, L4: 11.5 x 2.5 (50 ohms)
Ch3: 3turns 4mm long; Ch4: 4 turns 6mm long, both 3mm inner diameter, silver-plated wire of 0.7mm diameter.

23cm Band'' by DF8QK and DC0DA in (1). Several amplifiers were described in that article and we selected the very first one. With a modified bias arrangement (fig. 1) we constructed it.

We were fortunate that modern Hewlett-Packard test gear was at our disposal; what we discovered we are pleased to pass on. We think some experimenters will be interested in how the S parameters in this work out.

1. S-PARAMETERS, A BRIEF REFRESHER COURSE

S parameters can be defined as follows:

Si = S11 = Input reflection co-efficient

This is the relationship of the reflected wave to the forward wave at the input, under the conditions where Z1 = Z0 =50 ohms. We can also present this as an expression of the input VSWR.

Sr = S12 = Reflected transmission co-efficient

This is the relationship of the wave produced at the input to the forward wave at the output, with Zs = Z0 = 50 ohms.

Sf = S21 = Forward transmission co-efficient

This is the relationship between the wave produced at the output to the forward wave at the input, with Z1 = Z0 = 50 ohms. This describes, for this special application, the amplification of the circuit.

S0 = S22 = Output reflection co-efficient

This is the relationship of the reflected to the forward wave at the output, with Zs = Z0 = 50 ohms. In other words this S parameter describes the output standing wave ratio.

2. VALUES MEASURED ON THE AMPLIFIER IN FIG. 1

The circuit in fig. 1 gave the measurements shown in figs. 2, 3 and 4. Some comments are offered in case they are a bit too abstract for one or two readers.

The reflection co-efficients S11 and S22 are represented as polar diagrams, similar to the Smith chart.

The amplifier is switched on and is provided at its input with a signal whose

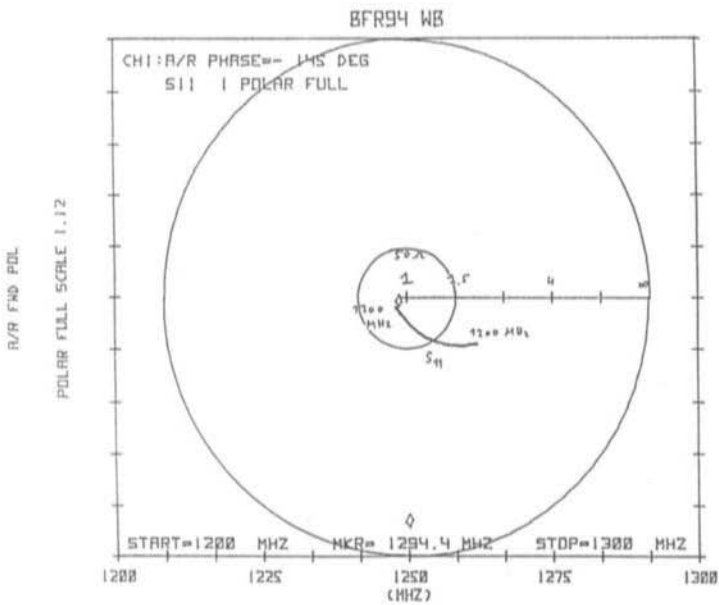

Fig.2 Input matching S11 of the Wideband Amplifier in Fig.1

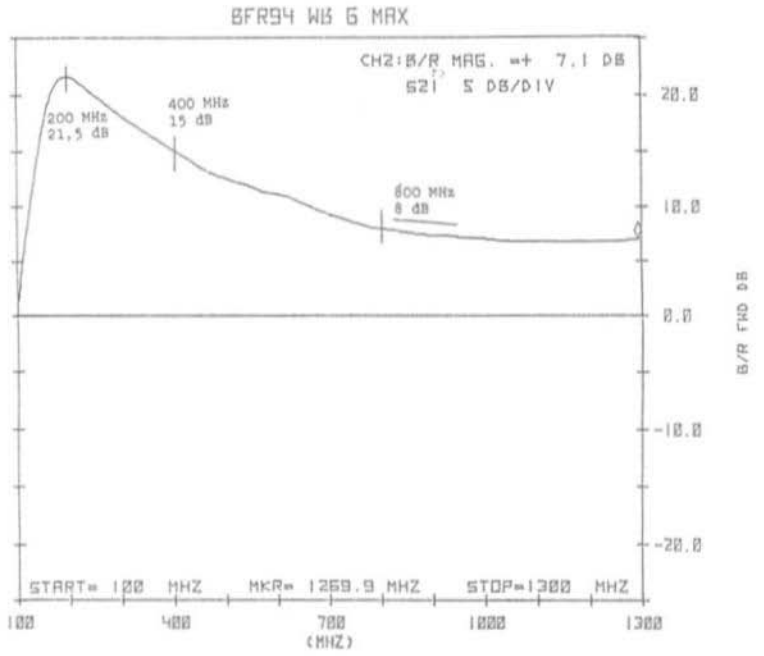

Fig.3 Amplification S21 of the Wideband Amplifier in Fig.1

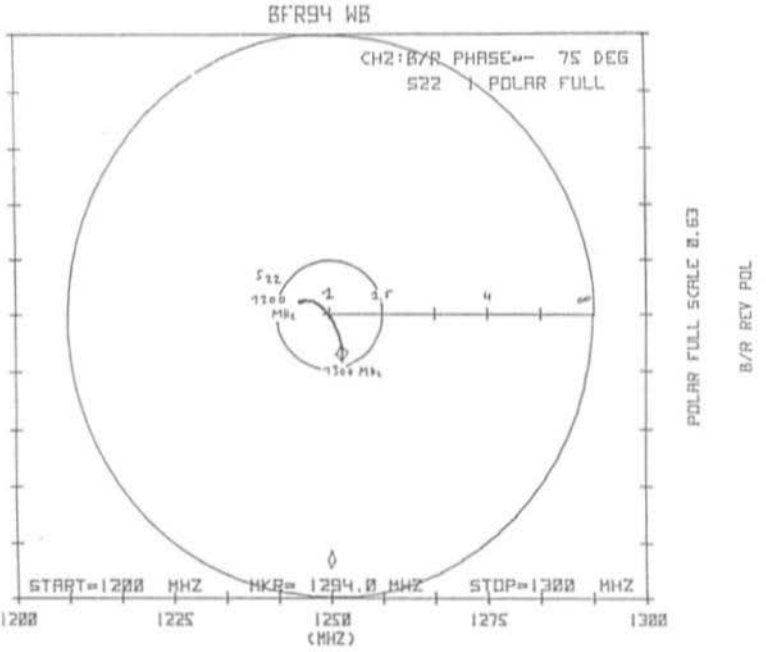

Fig.4 Output matching S22 of the Wideband Amplifier in Fig.1

frequency lies between the values noted on the X axis (in this case 1200-1300MHz/100-1300MHz). The curve drawn has two ends, on both of which the frequencies are noted. These are the beginning and end frequencies of the sweep.

You can also see little diamonds on the curves: these are markers for frequencies of particular interest. Now for some details on these illustrations.

- Every point in the polar diagrams above the horizontal line through the centre point describes an inductive reactance, and vice versa, every point below the line is a capacitive reactance.

- Every point inside a circle whose centre lies in the centre of the diagram and whose radius in the horizontal radius is scaled from zero to infinity represents a standing wave ratio equal to or smaller than the value of the circle's radius. A perfect non-reactive load with 50 ohms impedance would give us a curve whose points would all lie precisely in the centre of the concentric circles.

Our graphics have two circles: the smaller one contains all points with an SWR equal to or less than 1.5 and the larger one contains with all points with an SWR between 1 and infinity. Points outside the two circles naturally have no significance.

Now for the test results.

We can see immediately that the amplifier is very broadband. The amplification shown in fig. 3 is 7.1dB at the marker frequency of 1269.9MHz and 21.5dB at 200MHz. The amplification drops off at 6dB per doubling of frequency (= 6dB per octave), as one would expect.

The output matching is good - SWR equal to or less than 1.5 at frequencies between 1220MHz (and probably even lower) to 1300MHz. At these frequencies the input appears capacitive -the whole curve lies below the horizontal centre line. However, the values in most of the amateur band are still good enough.

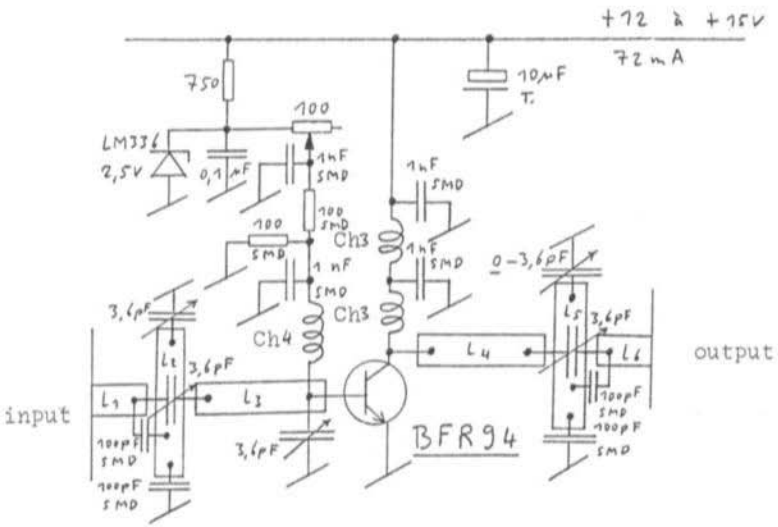

Fig.5 The Narrowband version for 1296Mhz
L1, L6: 5 x 2.5 (50 ohms); L2 - L5: 11.5 x 2.5 (50 ohms)
Ch3 and Ch4 as in Fig.1.

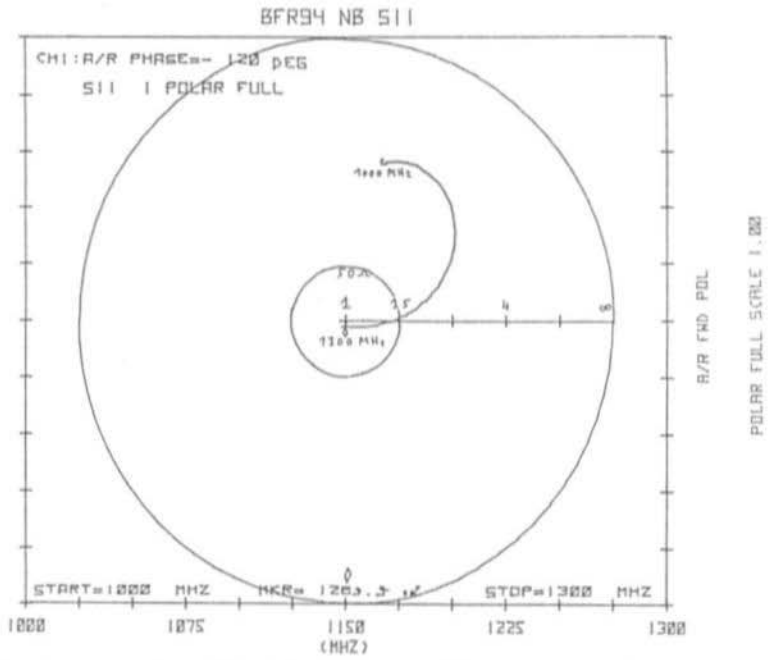

Fig.6 Input matching S11 of the Narrowband Amplier in Fig.5

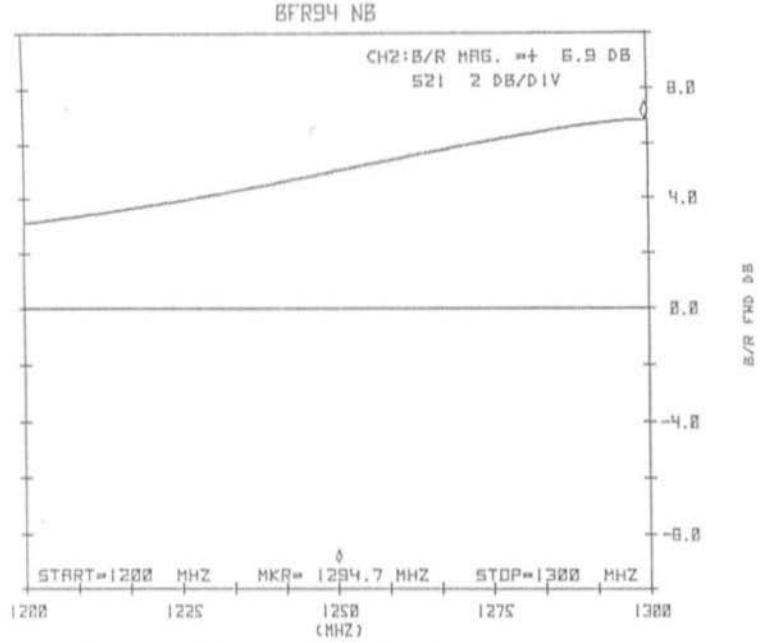

Fig.7 Amplification S21 of the Narrowband version

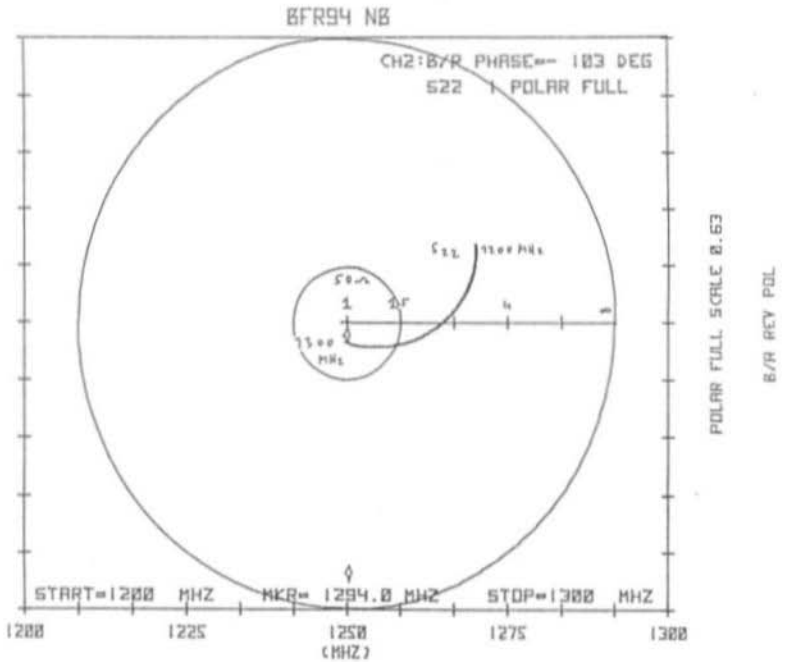

Fig.8 Output matching S22 of the Narrowband version

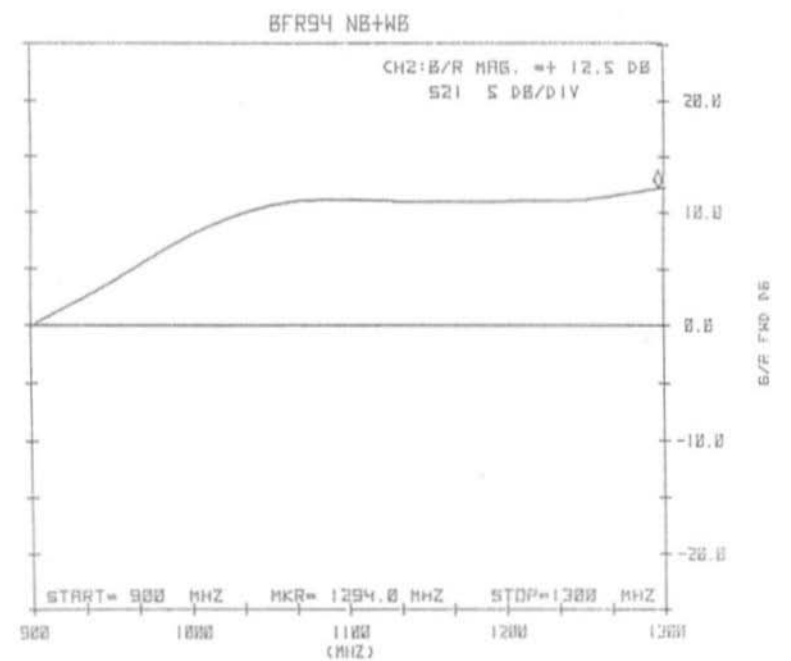

Fig.9 Amplification of both versions in cascade circuit

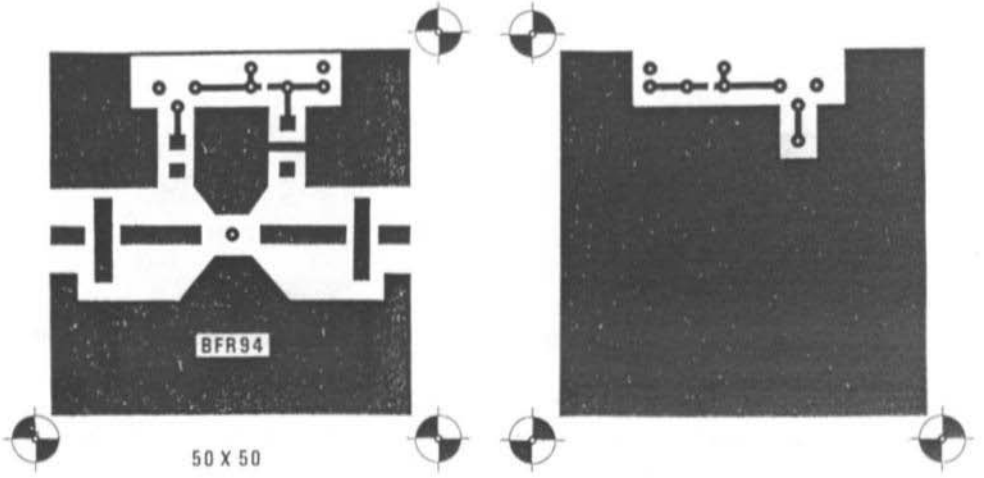

Fig.10 The layout for upper and lower sides can be used for both versions

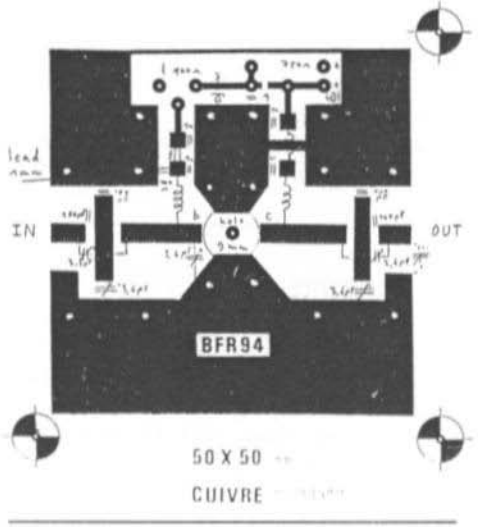

Fig.11 Arrangement of the components of the Narrowband version (upper side). For the Wideband version the layout is the same except that the capacitors on L2 and L5 are not required and are replaced by trimmers between the output and ground. Values shown on the circuit diagram.

3. A NARROWBAND VERSION

We couldn't resist it! After we had already modified the pre-bias circuit we just had to try and modify the input and output circuitry. Our feeling was that the circuit was too wideband for us (had it been narrow-band, we would probably have tried to make it wideband!).

Anyway, fig. 5 is the result. We added tuned circuits on the input and output, namely L2 and L5, which are capacitive to ground at one end (an electrical ground connection must work just as well, but we didn't try this out) and tuned the other end with a 3.6pF trimmer.

The layout may look a little unconventional with the arrangement of the tuned circuit but the circuit does work very well, as figs. 6, 7 and 8 show.

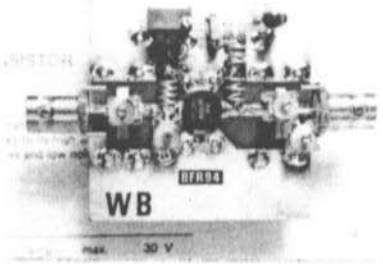

Fig.12 The Wideband Amplifier

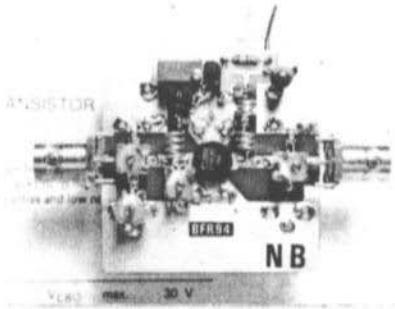

Fig.13 The Narrowband Amplifier

The amplification holds up practically unaltered at the upper end of the amateur band, whereas it drops rapidly at lower frequencies.

In similar fashion input and output matching are very good at 1300MHz and soon deteriorate with lowering of frequency. Unfortunately 1300MHz was the top limit of our test gear, so we could not check what happened above the amateur band. All the same, we had what we wanted: a narrow band amplifier!

By the way it is also noteworthy that both versions of the amplifier turned out to be perfectly stable. As a final experiment we even connected the two amplifiers in series, the narrowband in front of the wideband: we found an amplification of 12.5dB at 1294MHz, dropping to 0dB at 900MHz, as fig. 9 shows. Additionally no tendency to self-oscillation was noted.

Since we are of the opinion that these amplifiers are still totally up-to-date, we present in fig. 10 our printed circuit board made of double-sided glassfibre-epoxy material, 50 x 50mm. Fig. 11 shows the component layout for the narrowband version. The wideband version has been built on the same PCB, only all capacitors connected with L2 and L5 are left out and there is a trimmer to ground from L6.

4. LITERATURE

(1) U. Beckmann DF8QK and J. Dahms DC0DA: Linear Amplifiers with Transistors for the 23cm Band. VHF COMMUNICATIONS 18 (1978) issue 4, pp 194-203.

Detlef Burchard, Box 14426, Nairobi, Kenya

The V-Antenna on the Car roof for mobile Direction-Finding

The following article is based on one in "Biotelemetry" (1) in which the use of the V-antenna was described for tracking wild animals equipped with a radio transmitter. Since antennas are fundamentally interesting and mobile direction-finding is especially interesting, I have adapted this detailed description by the author with further technical information. DL3WR

1. INTRODUCTION

Gazelles, rhinoceroses and similar animals are easy to fit with radio transmitters while one grasps one of their horns. A skilled team can fit them up in 15 minutes. The antenna glued to the horn ends up normally in a vertical position, so one can reckon on vertical polarisation. The directional effect of common VHF aerials is generally less pronounced in the vertical plane than in the horizontal direction - the HB9CV is an example of this; using it one gets very good direction-finding results in horizontal polarisation, but poor ones used vertically (fig. 1).

To increase the directional effect one can use an antenna with more elements, but the size rapidly becomes unmanageable. An alternative is to invert the polar diagram to get good directionality with small antenna dimensions; this technique was used here.

2. THE V-ANTENNA

An arrangement of three elements was developed, composed of two radiators and one reflector and optimised for vertical polarisation. Since the horizontal projection of the three elements forms a "V", the designation "V-Antenna" was chosen. Fig. 2 shows it on a small wooden mast.

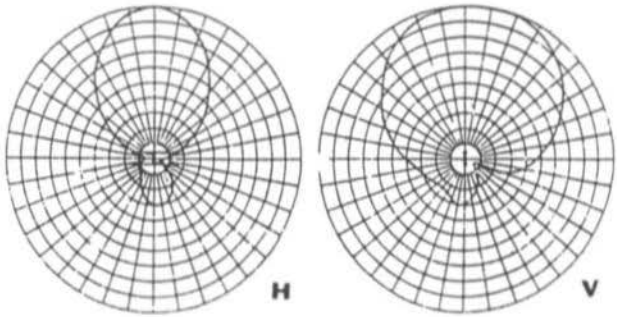

Fig.1 Directional diagrams of the HB9CV antenna
H = horizontal: V = vertical

If the RF energy received on the two radiators is added in a power combiner., we get the directional diagram shown in fig. 3a. If we put a suitable delay line in the two feeders the diagram shifts about 30 degrees to the left or right, as shown in fig. 3b. A comparison of the two voltages Vr and Vl gives a measure of the size and direction of declination between the two incoming wave fronts and of the antenna symmetry axis.

Switching between the two diagrams (directionality switching) and Vr/Vl comparison can be carried out electronically and the direction indicated on an instrument: this is like instrument navigation in a vehicle or aeroplane.

The radiators are connected to the gamma-matching elements with coaxial cable, so that the lower half of the V-antenna can be substituted by its mirror image without further ado. If desired, just the upper half can be mounted on a car roof, achieving the antenna sketched in fig. 4 without detriment to performance.

For an operating frequency of 174 plus or minus 1MHz the following dimensions and details apply:

Radiator length (i/c connector):	422mm
Reflector length (i/c connector):	470mm
Connector type:	PL259
Separation between radiators:	403mm
Separation between radiator and reflector:	403mm
Gain (including switching loss):	$4dB_d$
Matching:	SWR max. 2

The navigation arrangements comprise, apart from the V-antenna, three modules which are described next in detail: Directionality switch, Receiver Control and display circuitry.

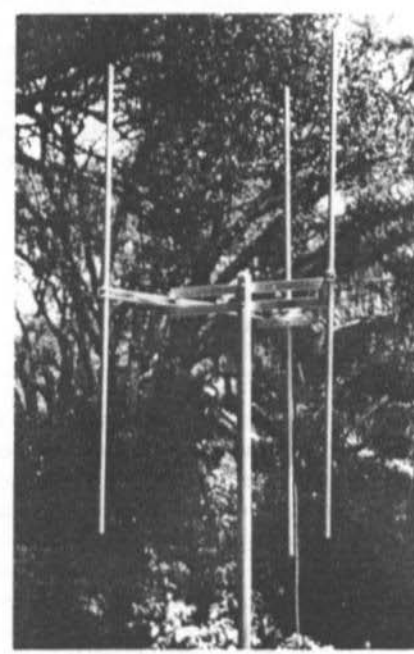

Fig.2 V-Antenna on wooden mast: cables are led down the reflector

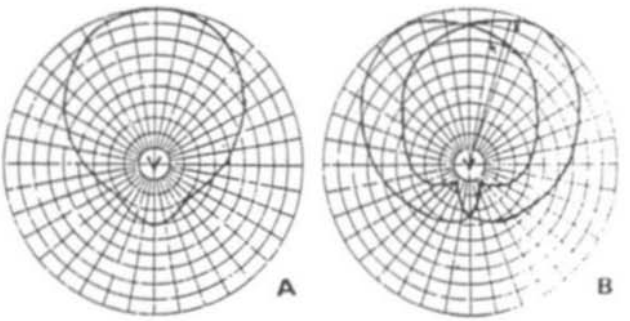

Fig.3 Directional diagrams of the V-antenna
A = zero degrees: B = plus or minus 30 degrees

3. DIRECTIONALITY SWITCH

Fig. 5 shows the principal elements of the V-antenna and directionality switch. The feeders to the two radiators R1 and R2 must be the same length. The switches are formed of PIN-diodes, the delay line DL is a short length of 50-ohm cable and the power combiner. can either be made of coaxial cable or purchased as a ready-made hybrid.

If logic level H is present on the two control lines c1 and c2, all diode switches are "on" and one gets the directional diagram of fig. 3a. If only c1 is at "H", PDS 2 and 3 are switched on and the diagram turns to direction R1 (fig. 3b). The opposite turn is made if PDS1 and 4 are switched on (c2 at "H").

3.1. Switching details of the directionality switch

The complete circuit is very simple and is shown in fig. 6. The V arrangement of the antenna comprises three quarter-wave whips, which are installed on an electrically conducting surface of at least 1m x 1m. A car roof is normally large enough for this. If for all this no conductive material is available a groundplane of this size can be made of wire netting and fixed above or beneath the roof - it must be well connected to the antenna connectors and this means soldering! The four diode switches are simply four switching diodes of the type BA244, which have lower through attenuation than the PIN-diodes used previously, plus six capacitors and eight resistors for separating the RF and DC paths.

The 30 degree delay line is a twelfth-wavelength long piece of 50 ohm coaxial cable and the power combiner. consists of two quarter-wavelength pieces of 75 ohm cable.

The associated 100R resistor absorbs power of unequal phase. The twin-conductor, screened control cable comes from the control and display unit.

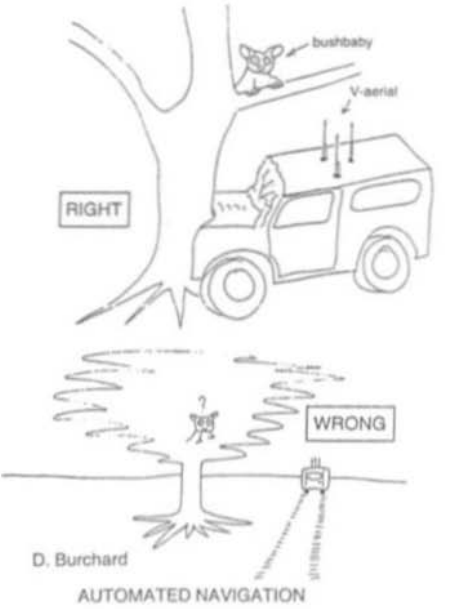

Fig.4 Halved V-Antenna in an African application

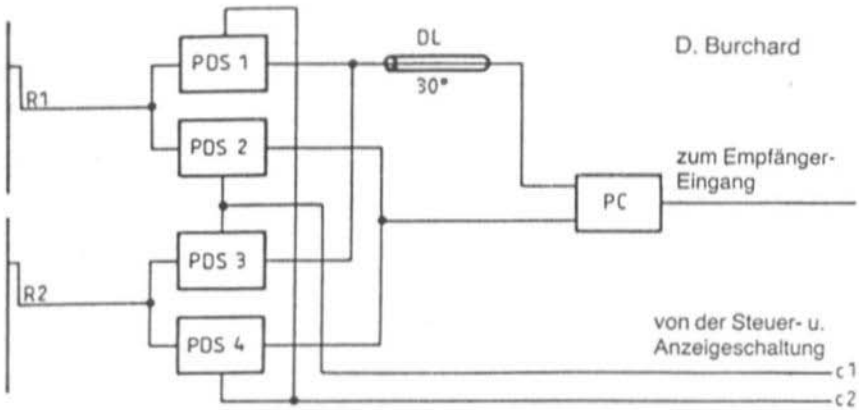

Fig.5 Principle of the Directionality Switch on the Antenna: R = radiator PDS = PIN-diode switch; DL = delay line; PC = power combiner

4. RECEIVER

Any kind of DF receiver with audio and trigger outputs can be used with the V-antenna if it has only slightly excess battery power. Automatic gain control is an advantage.

It is worth mentioning, to understand the necessary receiver characteristics, that the transmitters used in wildlife research operate in pulse mode in order to achieve an operating life of up to three years from a lithium battery. Typical values for pulse duration are 20ms, with a 1 second repetition rate.

Here we are using the model 287078 receiver made by B&R (W-7801 March-Buchheim, Germany) with the internal addition of the display circuitry (described in the next chapter) made on Vero-board and the display instrument glued to the side.

A more professional route would have been to construct a small additional housing for the electronics and mount this on the vehicle dashboard; this could be connected to the (unmodified) receiver by multi-core cable and connectors.

5. CONTROL AND DISPLAY CIRCUITRY

The block diagram in fig.7 should make things clear.

There is an analogue circuit section (above) and a digital one (below). Each pulse received activates the chain of timing elements (T1 - T3). Timing 1 is needed to start operation of the averaging filter, the other two are for switching the antenna diagram and for activating the sample and hold amplifier. The diagram is then shifted from the zero position to plus and minus 30 degrees, the analogue voltages proportional to Vr and Vl (fig. 3b) are stored in the two sample and hold amplifiers, and the difference indicated in the instrument.

The S&H and DA are set up for unity amplification. The pulse duration must be at least 13ms for correct display and the impulse amplitude must be constant. The AGC necessary for this must come into operation within 3ms including the internal trigger delay time of the receiver.

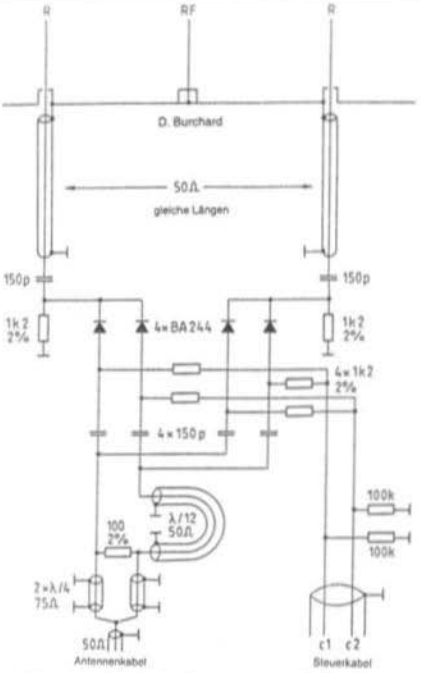

Fig.6 Circuit Diagram of the Directionality Switch

5.1 Details of the control and display circuitry

Fig. 8 shows a trial circuit with linear ICs and some CMOS logic. Only 5mA is drawn from the 12V operating voltage, most of which is taken by the switching diodes. Diverging from the block diagram, an internal trigger has been achieved, so that the DF receiver without its own trigger circuit can be employed.

The analogue circuit section - in the upper half of the diagram is equipped with low-power OpAmps all the way through. The first amplifier boosts the audio signal with amplification set coarsely by Rset and fine adjustment with P1.

The next two amplifiers form together a precision full-wave rectifier and the last is an active filter for averaging.

The output voltage of the filter is compared by the amplifier marked "trigger" with a DC voltage from P2. If the fixed level preset by P2 is exceeded, the LED lights up and the chain of timing elements is activated.

The output voltage of the filter also passes through electronic switches to two storage capacitors which are followed by OpAmps with high input impedances; these control the instrumentation. This circuit fulfils the function of the S&H/differential amplifier of fig. 7.

The last operational amplifier produces an auxiliary voltage of approx. 4.2V with low internal impedance.

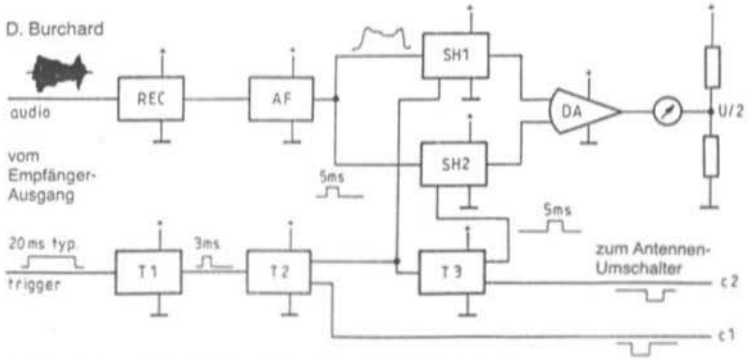

Fig.7 Principle of the Control and Evaluation circuit:
REC = receiver; AF = averaging filter; SH = scan/hold amplifier; T = timing circuit; DA = differential amplifier

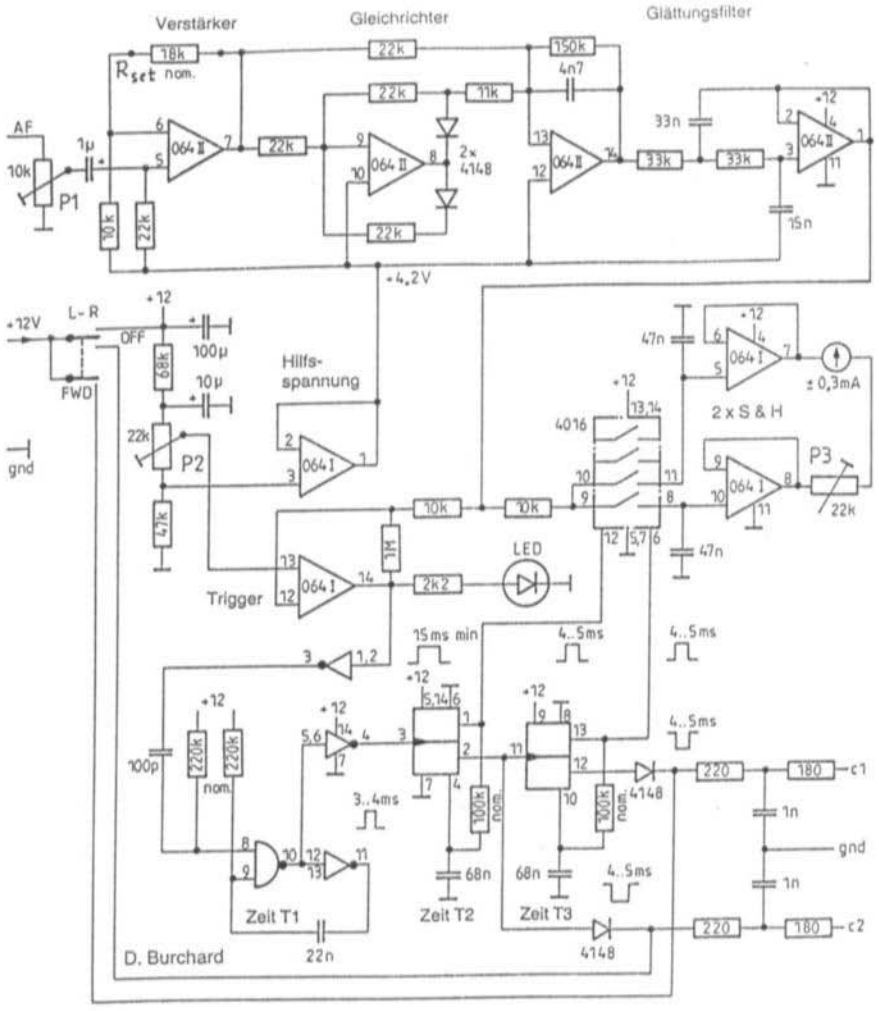

Fig.8 Circuit of the Control and Display electronics

Timing circuit T1 consists of two NAND gates of a 4011, the remaining two gates being employed as buffers before and after the circuit for regenerating the pulse edges. Timing circuits T2 and T3 are made up of D-type flip-flops, of which two are in one 4013. The Q outputs of both flip-flops control the electronic switch in the S&H circuit. The two bar-Q outputs control, via c1 and c2, the switching diodes in the aerial change-over switches. The R-C-R network in both control lines suppress high frequency interference from the digital circuit section.

When the double toggle switch is in the FWD (forward) position, battery voltage is applied to the two control lines and all diodes are switched for maximum sensitivity in the forward direction. With the toggle switch in the L-R (left-right) position, the control and display circuitry operates as follows.

The audio voltage at point AF is amplified, rectified and averaged - as already described. When the smoothed out voltage is sufficiently high, it triggers the timing circuit T1 via the comparator: T1 produces a pulse 3.5ms long and the LED lights up. The averaging filter needs this time to build up to the correct average value. After this timer T2 switched the antenna diagram into the -30 degree direction and switches the 4016 (or 4066) electronic switch shown top right to through-pass. After an operating time of 4.5ms the switched through voltage value is stored in the 47nF capacitor and read out on the side of the instrument.

Then, under the control of T3, the same thing happens with the +30 degrees direction and the lower S&H circuit, which now gives the stored voltage on the other side of the instrument.

The instrument gives the difference between both voltages, which gives a measure of direction of the oncoming wave front: with each pulse received a new direction value is displayed. The display sensitivity can be adjusted with P3.

6. OPERATIONAL EXPERIENCE

Open-air tests were carried out in the open bush land of the Kitangela Game Conservation Area. A tracking transmitter of the type 294079 (B&R) was used, with a radiated power of about 1mW (antenna height 0.5m).

The V-antenna was mounted on the back of a Land Rover, in such a way that it could be turned to pick up its diagram or orientated straight ahead for instrument navigation (fig. 9).

When we were at a distance of 10km from the transmitter the signal was far too weak (receive performance below -140dBm) to trigger the instrument's navigation, yet once heard we could track the direction coarsely (with the toggle switch in the FWD position).

At 5 or 6km distance the instrumentation tracking began to function but the instrument needle wavered about 5 degrees one way and the other because of the overloading receiver noise. Satisfactory results were achieved at 4km distance (receive performance above -120dBm), when the fluttering of the needle had reduced to less than 1 degree. With decreasing distance the display improved further, in which we determined with further tests that the instrument display was almost linear up to a tracking angle of more than plus or minus 20 degrees.

The system described takes about 10ms per tracking operation. Compared with the human method by ear, which takes at least 10 seconds per sounding, it is so fast that movement of the animals is almost inconsequential.

7. LITERATURE

(1) Burchard D.: The V-Aerial, A Navigation Aid for Tracking Down of Radio-Equipped Wildlife. Biotelemetry IX, pp 315-318, 1987.

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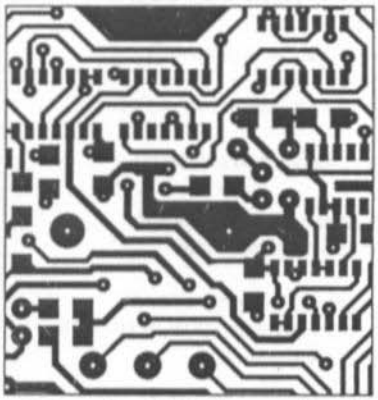

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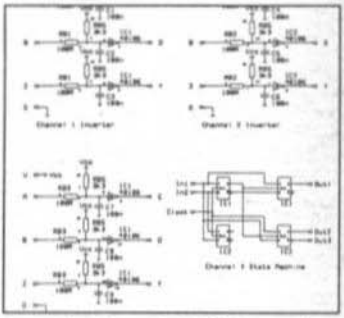

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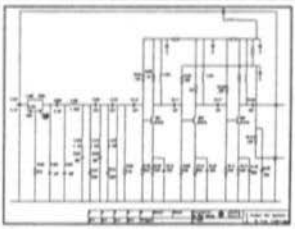

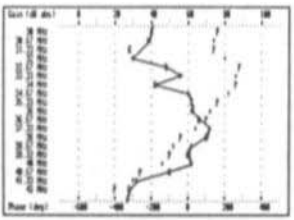

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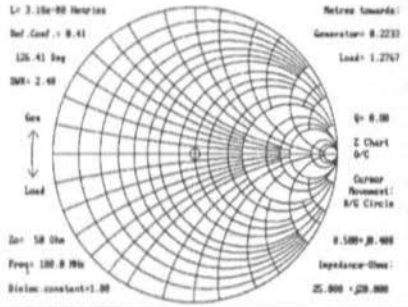

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