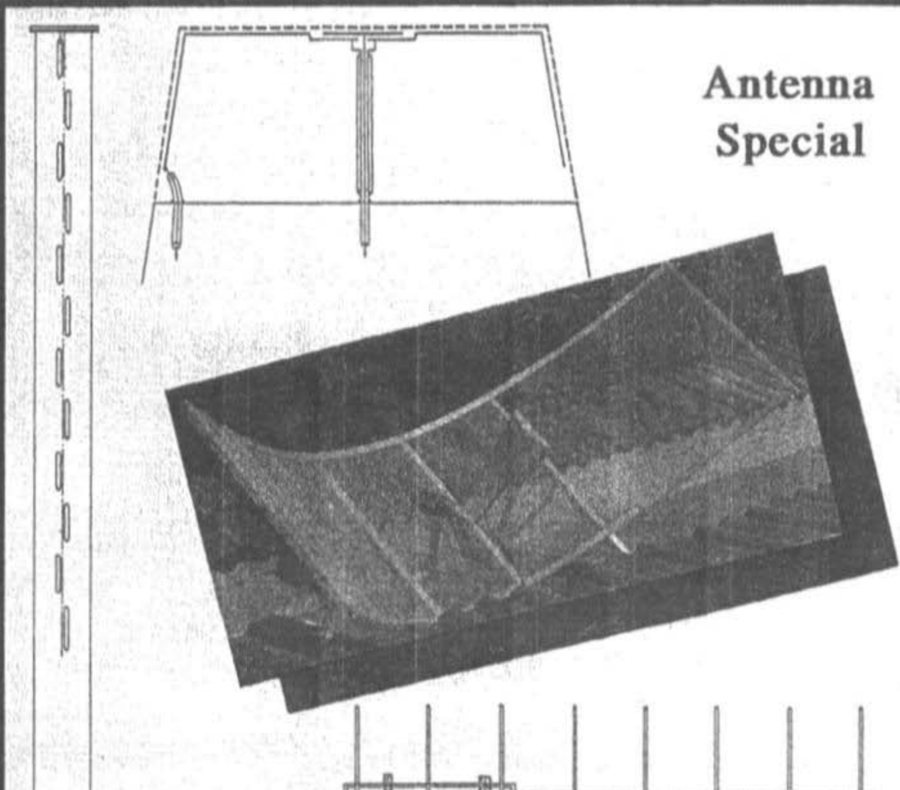
A Publication for the Radio Amateur Worldwide

Especially Covering VHF, UHF and Microwaves

VHF *COMMUNICATIONS*



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Volume No. 23 . Winter . Edition 4/1991

Published by: KM PUBLICATIONS, 5 Ware Orchard, Barby, Nr. Rugby, CV23 8UF, United Kingdom.

Publishers: KM PUBLICATIONS

Editors: Mike Wooding G6IQM, Kim Wooding

Advertising Manager: Mike Wooding G6IQM

VHF COMMUNICATIONS

The international edition of the German publication UKW-BERICHTE is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is published in Spring, Summer, Autumn and Winter under licence in the United Kingdom by KM PUBLICATIONS.

The 1991 subscription price is £12.00, or national equivalent per year. Individual copies are available at £3.50, or national equivalent each. Subscriptions, orders of individual copies, advertisements and contributions to the magazine should be addressed to the national representative, or - if not possible - directly to the publishers.

Back copies, kits, as well as the blue plastic binders are obtainable from your national representative or from KM Publications in the U.K.

Translated by Mr.A.Emmerson G8PTH,71 Falcutt Way, Northampton, NN2 8PH, U.K.

Printed in the United Kingdom by: Apex Printers, 1 Avon Industrial Estate, Butlers Leap, Rugby, CV21 3UY.

Please address your orders or enquiries to your representative.

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ISSN 0177-7505

Contents

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ATV with twin Sound Channels Part-1

The following article presents a concept for the production of an ATV signal with twin sound channels corresponding to the proper standards or norms. Building on the 70cm ATV transmitter by DJ4LB described in VHF COMMUNICATIONS (1), the author endeavoured to achieve the concept with standard modules and simple circuitry.

In this due regard was had on the one hand to the technical data laid down in the CCIR norms, and on the other to the dual sound technique developed by German broadcasters and introduced in 1981. (Note this is not the same as NICAM.)

Conformance to this enables not only considerable standardisation but also the use of mass-market (low-cost) television receivers and video recorders to the extent these have been adapted for stereo/dual sound operation.

1. REQUIREMENTS

The introduction of a second sound channel unfolds new opportunities for the development of amateur television, in particular the following applications

- two separate and simultaneous QSOs, independent of each other and possibly in two different languages, on a single ATV frequency;
- experimental radiation on of stereo sound transmissions, to establish matrixing characteristics;
- use of one channel for data transmission of all kinds, retaining the other channel for voice commentary.

Further applications will suggest themselves with use.

The transmission of two separate audio flows demands very high crosstalk suppression

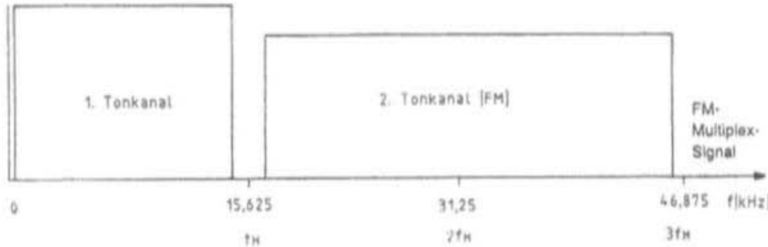


Fig.1: Multiplex System (Tonkanal = Audio Channel)

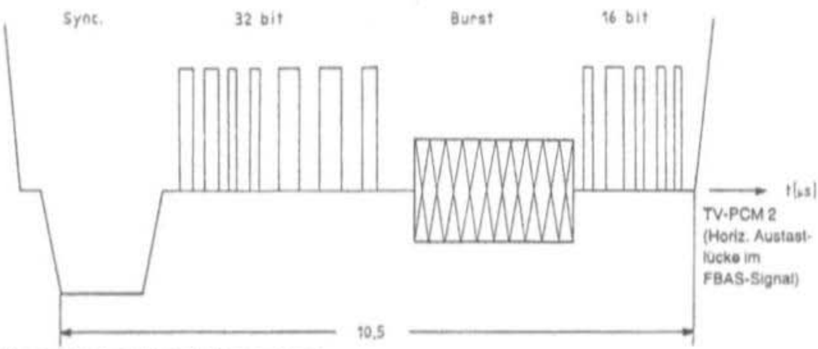


Fig.2: Pulse Modulation system (Horiz. Austasteluecke im FBAS-Signal = Horizontal Blanking period in the composite colour video signal).

between the two audio subcarriers: with regard to the sensitivity of overhearing, this mist be more than 60dB. If stereo is to be practised, on the other hand, the demands are reduced and an impression of stereo is already achieved with a channel separation of just 15dB.

Insignificant differences in phase and amplitude frequency characteristics must be considered together with small differences in deviation between the two channels at the same modulation frequency; that said, only so long as no matrixing for experimental stereo operation is intended.

To avoid audio disturbance, care must be taken over the following.

Non-linearities and/or overloading in the transmission elements can cause intermodulation between the two sound subcarriers or with the picture signal.

If the difference frequency of the two sound subcarriers is reflected in the picture carrier, moire patterns can be expected on the screen, which will vary in sympathy with the audio. Audible intermodulation in the sound channel can also occur between the colour subcarrier and a picture signal of 1.07MHz.

With sound demodulation using the intercarrier method audio interference can also occur due to phase errors in the vision carrier caused by over-driving. The sound subcarriers undergo phase modulation by the vision carrier; this is an undesirable possibility concomitant with all amplitude modulation.

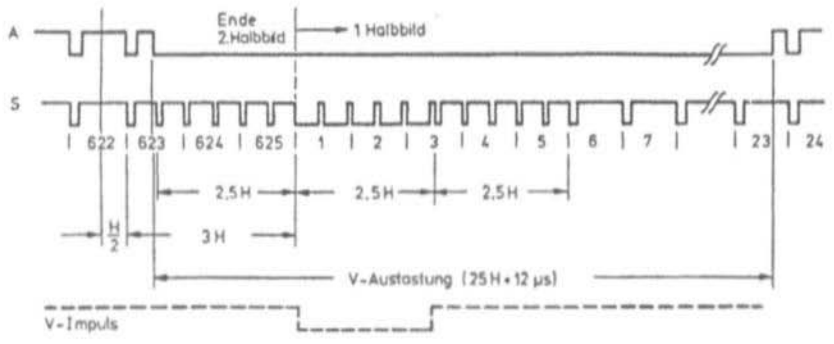


Fig.3: Transmission in the Vertical Blanking interval (Halbbild = Frame; V-Austastung = Vertical Blanking).

Unwanted audio interference also occurs from harmonic distortion components of vision signals whose frequency is the same as the sound subcarrier.

These possible interferences must be kept as small as possible by targeted measures both at the transmitting and at the receiving end.

At a later point in time it would be thinkable to define identification markers for different audio systems via a pilot tone. These considerations should be set aside for now and left until they become necessary as a result of everyday usefulness.

2. TRANSMISSION TECHNIQUES

With regard to the requirements discussed we can now describe some possibilities for putting them into practice.

2.1 Multiplexed FM on FM

This technique resembles stereo multiplex transmission of VHF radio. The second audio signal is modulated onto an auxiliary subcarrier with doubled line frequency and added to the existing audio subcarrier (Fig.1).

2.2 Pulse modulation (TV-PCM2)

Both audio channels are accommodated in the blanking region of the video signal (Fig.2). An advantage is the complete disappearance of the audio IF production in the TV transmitter.

2.3 Transmission during the vertical blanking interval

In this technique the second sound channel information is accommodated in an empty line during the vertical blanking interval (Fig.3). As the sound is transmitted only during one line per field (about every 50 microseconds), this information must be stored in the receiver in the meantime and replayed during the transmitted field.

2.4 Twin subcarrier transmission

For transmitting the second audio channel information an extra sound subcarrier is used, also frequency modulated and positioned about 250kHz above the first sound subcarrier. Power-wise, it lies about 7 to 10dB below the first sound subcarrier.

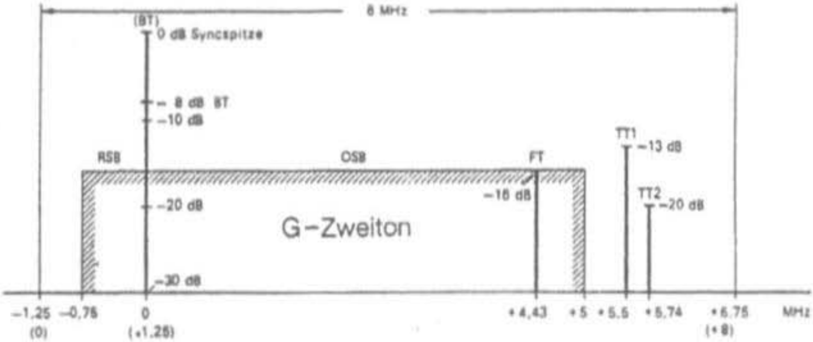


Fig.4: Situation of Vision Carrier, Colour Subcarrier, Audio Subcarrier 1, Audio Subcarrier 2, also vestigial and complete sidebands for Standard G. (BT = vision carrier; FT = colour subcarrier; TT = audio subcarrier; RSB = vestigial sideband; OSB = main sideband; Syncspitze = sync tip).

2.5 Selection of a system

Following detailed consideration of the possibilities indicated only the two subcarrier and multiplex techniques appear suitable for our application.

Pulse modulation is excluded on compatibility grounds, as it would cause interference to the synchronisation in line with the modulated audio. Also the bandwidth of the signal to be carried would cause difficulties.

Transmission in the vertical blanking interval would be extremely expensive to introduce,

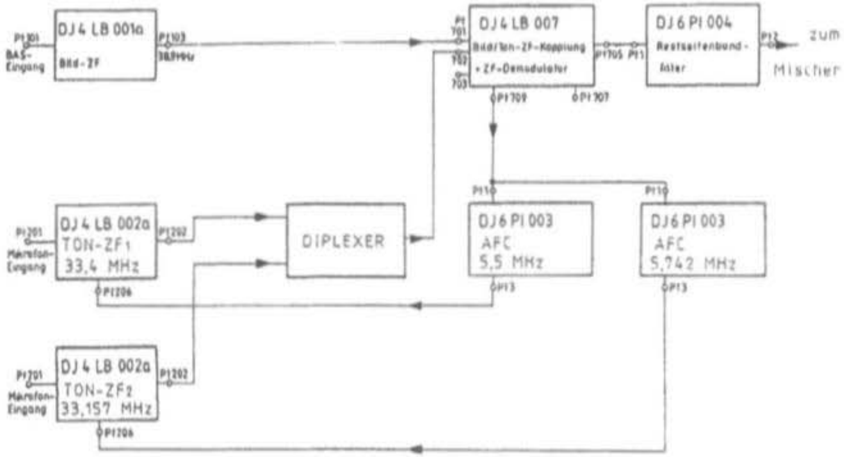


Fig.5: Expanded Block Diagram for ATV Transmitter (BAS = composite video; Eingang = input; zum Mischer = to Mixer; Bild = video; Ton = audio; ZF = IF; Kopplung = combining; Restseitenbandfilter = VSB filter).

although it would not cause compatibility problems.

Comparative investigations between the multiplex technique and the twin subcarrier system later exploited commercially pointed in almost all considerations towards the latter variant. Measurements both of crosstalk and harmonic distortion favoured it significantly.

For amateur television transmission with the twin subcarrier system higher insensitivity to interference is important, particularly in difficult reception circumstances such as reflections (ghosting).

3. PRINCIPLE OF THE TWIN-CARRIER TECHNIQUE

3.1 General

Fig.4 shows the amplitude spectrum of the ATV channel in the 70cm band.

In standards B and G using vestigial sideband practice, the first audio subcarrier lies 5.5MHz above the amplitude-modulated vision carrier. Its power level is around 13dB below the peak sync power of the video transmitter. Additional to this first audio subcarrier, which is present on all transmitters, is a second audio subcarrier, which is higher in frequency but with a power level 20dB below the video carrier.

If we select the difference between the two audio subcarriers to be in the region of 250kHz and an uneven multiple of the half line frequency, the possibility of interference due to unmodulated sound subcarrier is reduced (half-line offset). Thus possible difference frequencies are:

a) $31 \times [fH/2] = 242.1875$kHz

or

b) $33 \times [fH/2] = 257.8125$kHz.

The resulting separation of the second sound subcarrier from the vision carrier is determined as 5.7421875MHz or 5.7578125. In consideration of this interference, both frequencies are equally valid, but standard B/G prescribes frequency (a) as the difference frequency.

3.2 Carrier production in the transmitter

The block diagram (Fig.5) shows the alterations necessary for twin-sound subcarrier operation.

Up to now only the first audio IF oscillator was frequency-modulated with the modulation voltage and this IF taken to module DJ4LB007. Now both audio IF subcarriers are combined in a diplexer and presented together to the audio input of DJ4LB007. The two channels differ only in the oscillator frequency and are otherwise identical. This is true also for the automatic frequency control, which is provided twice and differs only in the alignment criteria.

New to the circuit is the diplexer. It is constructed very simply (just one transistor) and enables the adjustment of the sound subcarrier amplitudes.

3.3 Frequency control using phase locked loops

To maintain the exact difference in frequency between the vision carrier and the sound subcarriers I have used the tried and tested AFC circuit of DJ6PI003. This was in fact conceived for a single sound subcarrier and is not designed to handle the separation of the twin sound subcarriers.

An elegant solution is the linking of the two sound subcarriers to the crystal-controlled video IF. The block diagram in Fig.6 should be seen as a starting point, as the development of a PLL circuit would exceed the space available for this article. Perhaps there are amateurs who would feel motivated to examine this task.

3.4 Twin-audio reception

The transmitters which we use for ATV transmission produce a signal corresponding to the CCIR norm, which can be received on any home TV receiver (and converter, if used). If the home TV receiver is equipped for stereo sound, then no additional modifications are necessary for receiving twin-channel sound. The only thing to note is that the TV is in "dual sound" mode, to by-pass the stereo matrixing in the receiver. If this mode is not selectable manually (seldom the case) then internal modifications must be made at the appropriate part of the circuit. I do not wish to go further into this aspect.

In the production of multi-channel audio in conventional receiver practice, there are fundamental differences between intercarrier receivers and parallel-sound receivers.

TO BE CONTINUED

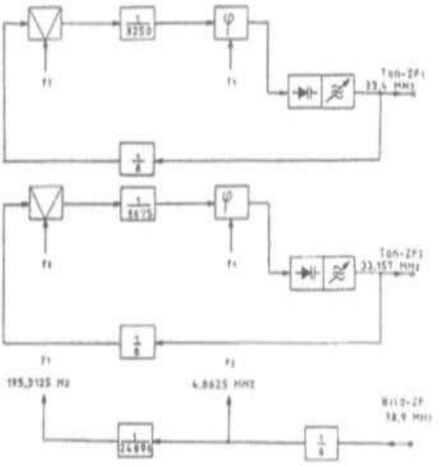


Fig.6: Suggestion for a PLL for frequency stabilisation of the audio IFs.

Oliver Nell, Klaus Solbach DK3BA, Jochen Dreier DG8SG

Omnidirectional Waveguide Slot Antenna for Horizontal Polarisation Part-1

Onidirectional antennas are used in telecommunications when it is desired to transmit (or receive) in several directions or when the heading of the other station is unknown. The latter occurs in amateur operation when calling CQ and particularly during contest operation.

The largest use of omnidirectional antennas is in mobile and portable operation, where the use of beam antennas would be impractical on account of their dimensions. In this case vertically polarised aerials are used, being the simplest to realise, from shortened rods through half-wave dipoles to stacked dipoles (mainly at fixed stations).

For operation in horizontal polarisation, especially DX and contest traffic, use is commonly made of beams which are stacked or bayed in the elevation and azimuth planes in order to achieve gain, for instance stacked Yagis and two-dimensional arrays.

In situations where weak signals can be received only on the main lobe of a narrow beam, an omnidirectional antenna with high gain would be highly desirable. In contests this would enabled the band to be searched with the omni antenna and once a station was found, then switch to the beam antenna to raise the signal out of the noise and make the QSO. In this kind of antenna diversity there appears to be a gain difference of about 5 to 10 dB between omni and rotatable beam antennas. So with the commonly used small-to-medium antenna system with a gain of 20dB this would represent a difference of the order of 10 to 15dB.

Another important application for omnidirectional antennas with high gain is horizontally polarised repeaters (such as amateur television) and beacons. On the microwave bands these use stacked radiators of the "Big Wheel" variety, which all the same present problems, both mechanical (size) and electrical (feeding). In addition they are not easy to fabricate or maintain.

On the other hand slot antennas (Ref.1) can be used. The dimensioning of these invites dimensioning errors, however, with the result that the chosen slot arrangement does not give the omnidirectional characteristic expected. Furthermore a "round" radiation pattern can only be achieved with flat waveguide. Since the omnidirectional slot-radiator antenna appears in principle to be very suitable for amateur purposes it was decided to research this type of aerial some more. The aim was to develop a mechanically and electrically foolproof realisation that was cheap and achievable with amateur resources. Significant research on dimensioning and the development of a sample antenna for the 23cm band was carried out as a diploma task in the antenna laboratory of Telefunken in Ulm, Germany. The following is an outline of the theoretical basics and the concept of the antenna arrangement.

At the same time approximation formulae are given for the maximum permissible number of slots and the achievable lobe width/gain. The dimensioning of the slots is described in sufficient detail for individual variations to be considered. Finally we describe the dimensioning, production and test results for 23cm and 13cm antennas.

1. CONCEPT OF THE WAVEGUIDE SLOT ANTENNA

The configuration shown in Fig.1 lends itself to an application needing an omnidirectional antenna with horizontal polarisation and a narrow vertical (elevation) radiation pattern with high gain. Here a waveguide is erected vertically with slots running also vertically.

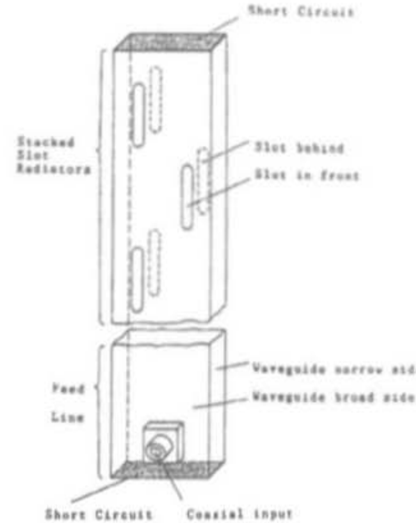


Fig.1: Waveguide Slot Antenna for omnidirectional radiation

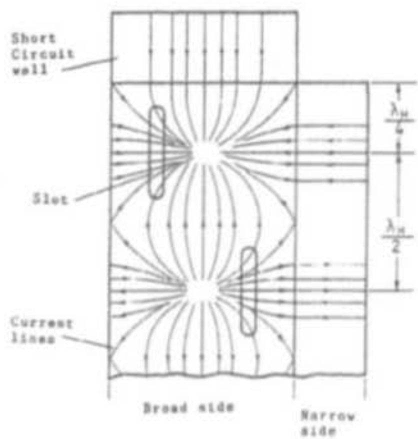


Fig.2: Sketch of the current flow lines in a waveguide slot antenna. View of broad side; the narrow side and short circuit are drawn as if they have been folded back.

The slots lie opposite one another on the front and rear broad sides, arranged either side of the centre line. The waveguide is short-circuited a quarter wavelength beyond the last slot. The inter-slot distances along the wave-guide are half the wavelength of the wave-guide. The antenna is fed via a coaxial-to-waveguide transition at the bottom end of the waveguide.

The mode of operation of the antenna goes under the name of "Resonant Array". In the absence of the slots a standing wave would occur in the whole waveguide (the end is short-circuited!). The current distribution resulting on the inner side of one of the broad sides from this is sketched in Fig.2.

At a distance of a quarter waveguide wave-length ($\lambda_H/4$) from the short circuit, the current follows a purely transverse direction and at the same time the current component dis-appears in the axial direction. The pattern of the current repeats itself exactly at distances of a half waveguide wavelength, however with the mathematical sign reversed. Equally, on the opposite side of the waveguide the same current distribution occurs, only in the reverse direction, that is with the mathemati-cal sign reversed.

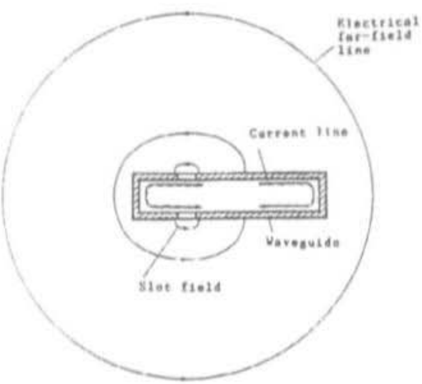


Fig.3: Excitation and radiation from opposed slots in the broad side of the waveguide.

Slots in the wall of the waveguide interrupt the flows; the currents flowing vertically excite electrical fields vertically over the broad side of the slots. This can be seen as the source of radiation into free space.

For vertically-arranged slots to produce hori-zontal radiation we need to find the locations where the currents in the wall of the wave-guide are flowing horizontally. In this, the distance from the centre-line of the waveguide determines the "strength" of the power that can be decoupled, since the horizontal flow components increase towards the sides (they are zero at the exact centre).

The interlacing of the slots about the centre line is necessary to equalise the sign-reversal of the current flows in the wall of the waveguide. In this way we can arrange a vertical stacking of slots which will produce excitation in equal phase and thus produce the narrow beamwidth and hence increased antenna gain desired. The more slots, the narrower the beam and hence the higher the gain.

As an approximation, the lobe width in the elevation [delta theta] is determined by the length of the radiating part of the antenna, i.e. the number of slots N and the stacking distance $\lambda_H/2$,

$$\Delta\theta \approx 50.7^\circ \cdot \frac{\lambda_0}{N \cdot \lambda_H/2} \qquad \text{(Gl. 1)}$$

in which λ_o is the free-space wavelength and N the number of slot-pairs.

Correspondingly, the antenna gain is given, by way of approximation, as

$$G \approx N \cdot \lambda_H/\lambda_0 \qquad \text{(Gl. 2)}$$

Unfortunately for practical constructions we cannot increase the number of slots ad lib. since the usable bandwidth of the antenna decreases with the number of slots. This effect is due to the production of a standing wave and resembles the behaviour of a resonator.

In an approximate fashion the number of slot-pairs is limited by the increasing mismatch and deformation of the radiation diagram at the frequency band edges as follows:

$$N_{max} \approx 100 \cdot \frac{0.5}{\Delta f/f_o} \qquad \text{(Gl. 3)}$$

wherein N_{max} is the highest number of slot-pairs usable in an antenna with a frequency bandwidth of delta f and a centre frequency of f_0.

The slots are dimensioned so that at the design frequency, the standing wave from the upper short circuit fades away at the lowest slot, which means the antenna is matched ahead of the first slot. The waveguide is matched at the transition to coaxial feeder and thus the length of this feeder region can, with due regard to material and fixing considerations, be of any length, allowing the waveguide to be used simultaneously as a mast and reducing the length of coaxial feeder.

The placing of the two opposing slots represents a special problem. Fig.3 demonstrates why in contrast to (1) the slots should lie directly opposite one another.

For the production of a closed, ring-shaped line of the electrical far-field, both slots must present opposing fields as they would excite at directly opposite locations on account of the natural current flow in the waveguide wall.

A further factor in ensuring the electrical far-field line has an equal field-strength is that the waveguide should be as flat as possible, that is the narrow side should be as small as possible. In (3) it is shown that below a height of the waveguide of about 0.15 wavelength (about half the standard height of waveguide) the circular diagram shows a ripple below 1dB (that is deviations from the ideal constant field strength along the far-field line).

For the realisation of antennas with amateur means this is no problem, however. If expensive precision waveguide with half-standard height is unavailable, cheap aluminium extrusions can be used, and despite their dimensional tolerances they can be used as flat waveguide in the lower microwave bands.

2. DIMENSIONING THE SLOTS

Dimensioning uses the equivalent circuit set out by Silver (4).

A single longitudinal slot in the broad side of a waveguide can be represented as a complex

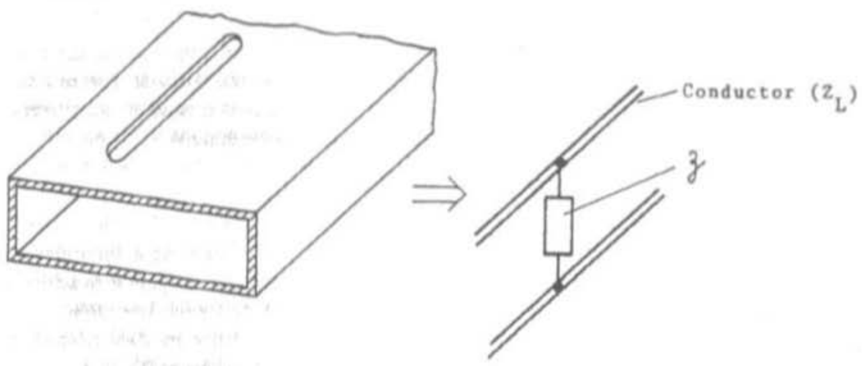


Fig.4: Equivalent circuit of a lengthwise slot in the broadside of the waveguide

impedance Z parallel to the conductor (Fig.4). From this equivalent circuit results the common description "shunt slot". With a slot length of approx. $\lambda_0/2$ the imaginary component becomes zero and the slot is in resonance. The real component of the slot impedance representing radiation of power into free space is calculated according to Silver as:

$$\frac{Z_L}{R} \approx 2.09 \frac{\lambda_H a}{\lambda_0 b} \sin^2\left(\frac{\pi x}{a}\right) \cos^2\left(\frac{\pi \lambda_0}{2 \lambda_H}\right) \quad \text{(Gl. 4)}$$

In the above λ_H is the waveguide wavelength, $_0$ the free-space wavelength, a x b the cross-section dimensions of the waveguide and x the centre-line of the slot. The following are also valid:

$$\lambda_H = \lambda_0 / \sqrt{1 - (\lambda_0/2a)^2} \quad \text{(Gl. 5)}$$

$$\lambda_0 \,[\text{mm}] = 300/f \,[\text{GHz}] \quad \text{(Gl. 6)}$$

Z_L is the reference resistance of the waveguide, and through the quotient formation of Z_L and R the normalised slot conductivity is formed. The arguments of the angle functions are drawn as a circular measure, so corresponds to 180 degrees. In the case presented here of the double slots in a flat waveguide (with height about a quarter of the width) the two slots work out almost as an ideal parallel circuit, in which the normalised slot conductivity is doubled. The best agreement of the results from a series of measurements is achieved with the following equation, modified correspondingly from equation 4.

$$\frac{Z_L}{R} \approx 3.5 \frac{\lambda_H a}{\lambda_0 b} \sin^2\left(\frac{\pi x}{a}\right) \cos^2\left(\frac{\pi \lambda_0}{2 \lambda_H}\right) \quad \text{(Gl. 7)}$$

To use these dimensioning measurements in antennas with several elements or slot-pairs it is important that the coupling of the slots along the waveguide axis is sufficiently small that the characteristics of the slots even in large groups (many slot-pairs) are worked out sufficient for our purposes according to equation 7.

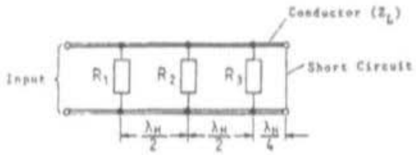


Fig.5: Antenna with 3 slots as an example for calculating input resistance.

Still missing for dimensioning the antenna is a determination of the centre line offset of the slots and their length.

To achieve the maximum possible gain all the slots should be driven at resonance at the design frequency of the antenna. This means the impedance of the slots could be replaced by the appropriate real component (resistance R). Moreover at this frequency the slots lie exactly half a waveguide wavelength apart and the final slot is exactly a quarter wavelength ahead of the short circuit (Fig.5). In this example the short circuit transforms itself into an open-load parallel to the last slot resistance, this last resistance transforms itself without alteration parallel to the next-previous resistance, and the parallel circuit of these both transforms itself without alteration parallel to the first resistance. That means that all slot resistances at the design frequency are connected in parallel, i.e. their conductivities add up.

To achieve matching at the entry to the group of slots it is necessary that the resulting resistance of the slots is equal to the reference resistance of the feedline.

$$Z_L = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}} \quad \text{(Gl. 8)}$$

For our application there is a simplification that the maximum antenna gain is to achieved with a given number N (equation 3) of slot-pairs. All slot-pairs are dimensioned the same with the slot resistance R

$$\frac{Z_L}{R} = \frac{1}{N} \qquad \text{(Gl. 9)}$$

With this result the slot centre line offset from equation 7 can be calculated.

Example:

The waveguide cross-section dimensions are a x b = 172mm x 42mm, wall thickness 4mm, frequency f_0 = 1.27GHz, N = 12 slot-pairs.

Equation 9 gives Z_L/R = 1/12 = 0.083
Equation 5 gives $\lambda_H/_0$ = 1.3765

With equation 7 the course of the normalised slot conductivities is calculated as a function of the slot centre line offset and as Fig.6 represents graphically. We are looking for the value of x which takes the value of 0.083 for the function, namely x = 8.6mm.

The slot length for resonance for the slot-pairs under discussion lies close to $\lambda_0/2$, however, there are dependencies on frequency, the wall thickness and height of the waveguide and above all, the centre line offset of the slots. Moreover, the shape of the ends of the slots plays a role: the rectangular slots with straight ends normally researched in the literature can scarcely be made in practice. The easiest way of making slots is from above with a milling cutter whose diameter corresponds to the width of the slot; the entry and exit points at the end of the slots will therefore be semi-circular.

The slot lengths finally resulting for resonance must be determined experimentally for the waveguide in use and the frequency region of interest in conjunction with the width and the form of the ends of the slots. Fig.7 shows the result for the waveguide in the first example. The slots are produced with a milling cutter of 10mm diameter, although good results approximating to these were also achieved with slots 50 per cent wider or narrower. The graphic also shows the slot length formed on the free-space wavelength L/λ_0 (from end to end) as a function of the slot centre line offset x. The central solid line curve is valid for the design frequency f_0 = 1.27GHz, while the pecked curves give the frequency-dependence results for 5 per cent higher or lower frequency.

With this representation the length necessary for slots dimensioned in the example can be found.

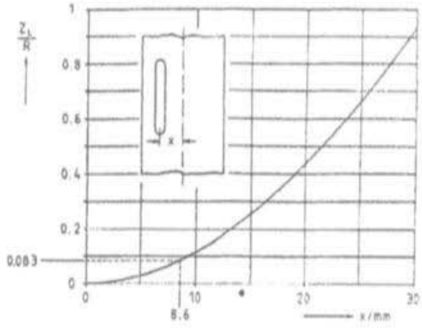


Fig.6: Representation of the normalised slot transverse conductor value as a function of the slot centre partition for the geometry in example 1.

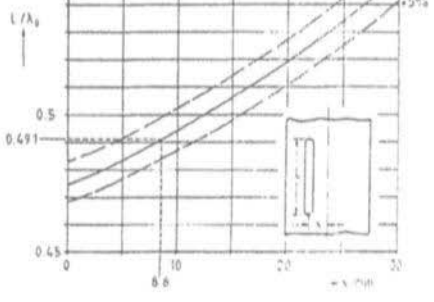


Fig.7: Slot length for achieving resonance as a function of slot centre partition for the geometry in example 1.

Heinz Kriegelstein, 8044 Unterschleibheim

Simple Doubling of the Data Storage Capacity of the DSP-Computer.

The DSP computer described by Matjaz Vidmar YT3MV (reference 1) is equipped with four data storage cards as standard, providing together 1MByte of static RAM. This is quite a lot so long as only part is used for the software currently employed and only figures and text are being processed, as for example with RTTY, packet radio or satellite telemetry. As soon as weather pictures are stored, such as lengthy METEOSAT image sequences or APT or even HRPT pictures (reference 2), the limits of the storage capacity are reached.

Fortunately YT3MV has already provided on the PCB decoding for 2MB, so that doubling the storage capacity is relatively simple. The changes are limited to fixing the memory board in piggyback fashion and expanding the decoding as necessary.

Since only two lines of the bus carry an additional loading, no special measures are necessary in this respect. The DSP computer here has been working several months without problems after fitting the doubled memory at a speed of 14.6MHz. So I'd now like to describe the expansion step by step.

1. EXTENDED CIRCUIT

Close study of the original circuit diagram for the YT3MV 006 board reveals that Store Enable [bar E] is controlled by decoders 02 and 03 of the HC138 chip. I have shown the numbers on the original diagram (Fig.1). Incidentally, following YT3MV's suggestion, I have used exclusively HC chips.

The expansion now consists of doubling the memory with piggyback chips and realising the additional enable lines needed with two extra HC138 chips. The only problem is the control of the bus-driver. It can be solved with a 74HC00 - or better still - a 74HC08, which

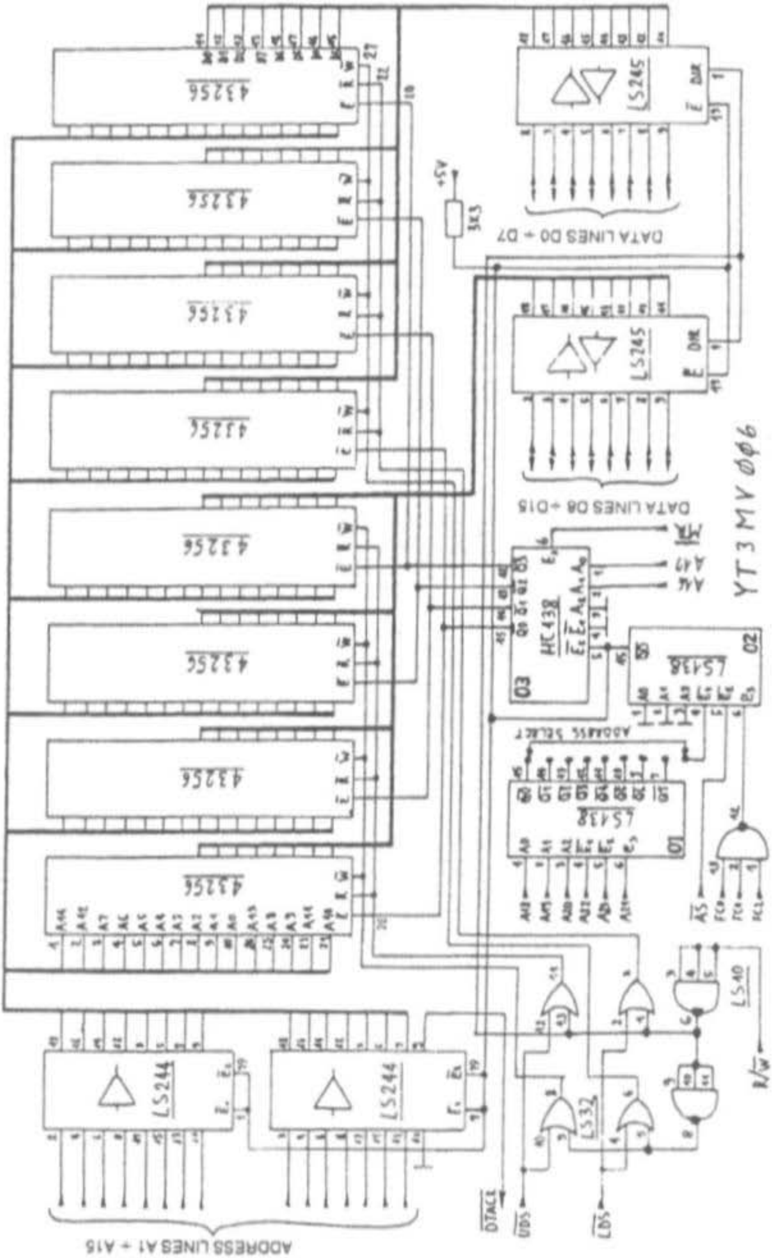


Fig.1: A Memory Card YT3MV 006 with 256k CMOS Memory

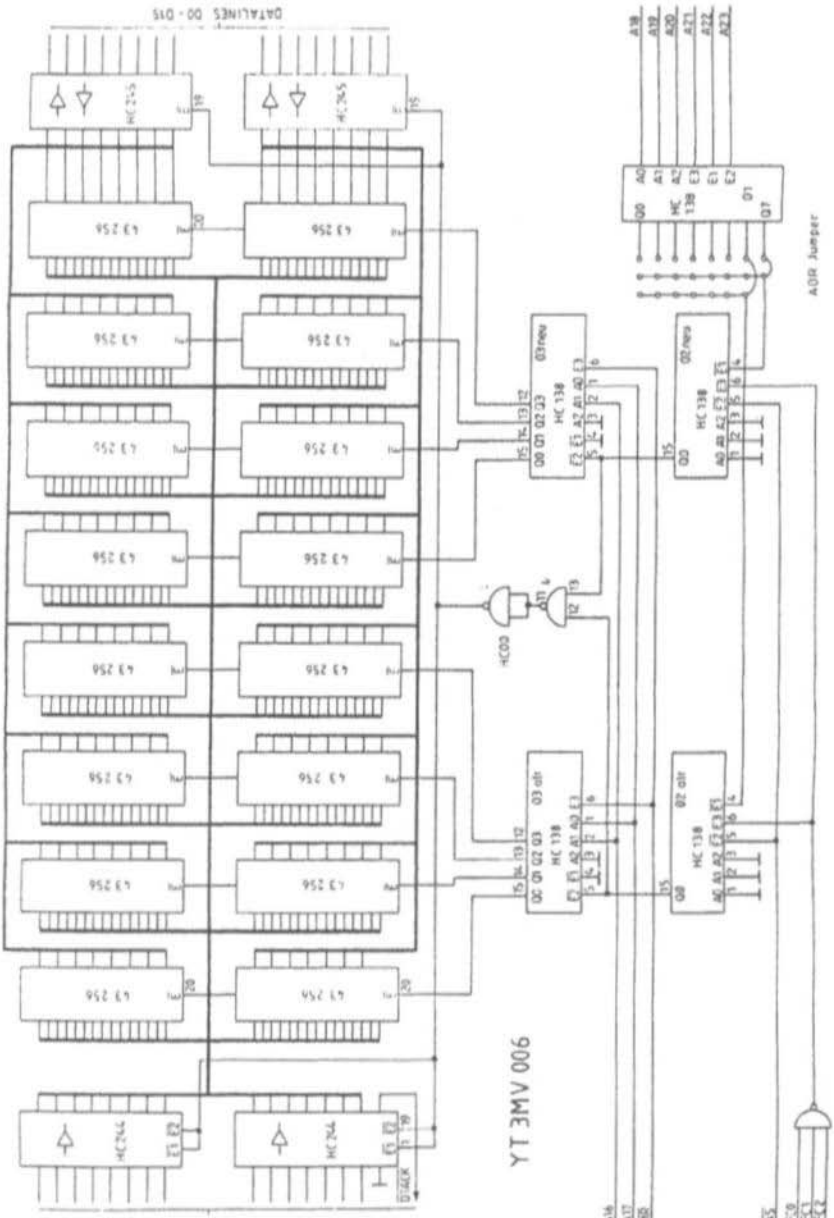


Fig.2: A Memory Card YT3MV 006 with storage doubled to 512k.

makes the otherwise necessary inverters superfluous. The new circuit for the memory card is shown in Fig.2. For four memory cards, that is a total of 2MB, the following additional components are required:

32	uPD43256C-12L or equivalent*
8	74HC138N
4	74HC08N

* 120ns access time in static RAM is adequate for a processor speed of up to 12.5MHz.

2. CONSTRUCTION

Fig.3 will assist orientation of the work in hand. It is the original Fig.3.7 with added wire connections and piggyback ICs.

On the solder side of the PCB YT3MV 006 the connection must be broken between old IC3 (I03, HC138) and HC244/pin 19 directly beneath IC3.

Then a new HC138 is prepared. To stack it above IC2 pins 4 and 15 must be bent upwards and pins 7, 9, 10, 11, 12, 13 and 14 should be removed.

Now the new IC2 can be soldered on top of the old IC2 (pins 1, 2, 3, 5, 6, 8 and 16). If the IC was socketed, remove it first to avoid solder being drawn into the socket and making it unusable.

Now another HC138 is prepared. Bend up the following pins: 5, 12, 13, 14 and 15. Cut off pins 7, 9, 10 and 11. Using the remaining pins 1, 2, 3, 4, 6, 8 and 16 solder the new 138 to the old 138.

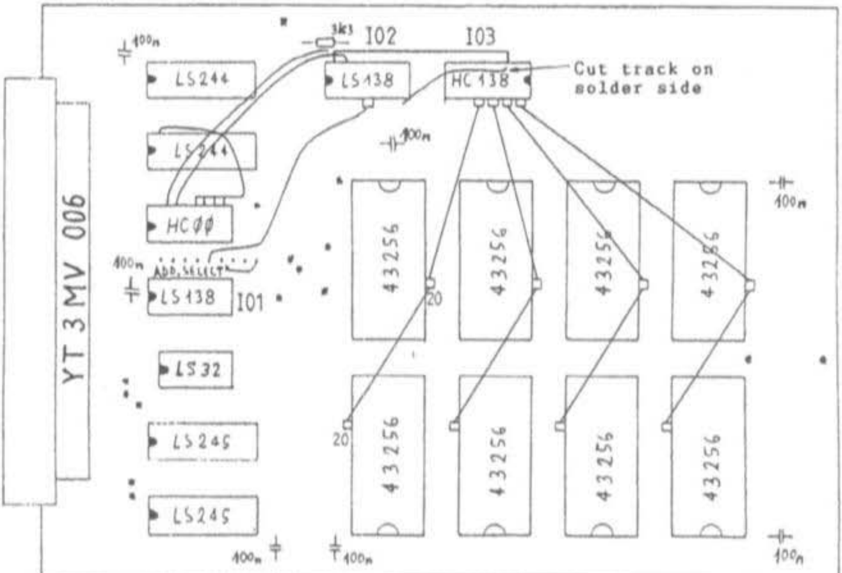


Fig.3: Connections on the expanded memory card.

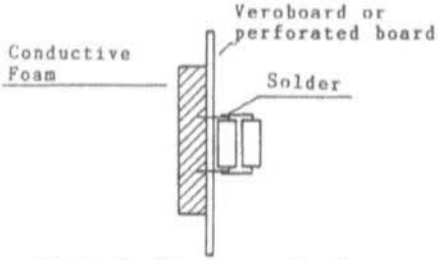


Fig.4: Auxiliary preparation for piggyback soldering to the memory IC.

On the memory ICs, only pin 20 needs to be lifted; all the others must be soldered to the pins already built-in. For working with these CMOS chips the familiar precautionary measures should be heeded.

If the memory ICs are in sockets, then they can be taken out and placed in a piece of perforated board or Veroboard, with a piece of conducting foam glued behind, as shown in Fig.4. In this way each pair can be turned round easily and soldered.

Of the new 74HC00 or HC08 all the pins up to 7 and 14 should be bent upwards. With just these two the chip is soldered to the existing 74HC10.

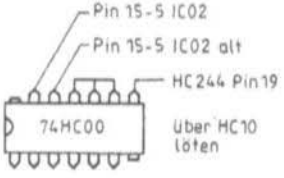


Fig.5: Connections to additional 74HC00

2.1 Wiring

From each of pins 12, 13, 14, 15 of the new IC3 take a wire to pin 20 on the new memory chips, similar to the track layout on the PCB (Fig.3).

Pin 15 of new IC2 is connected to pin 5 of new IC3 and further to pin 13 of the 74HC00 or 08.

To the existing track with pin 15 of old IC2 to pin 5 of old IC3 solder a wire and connect this to pin 12 of 74HC00 (or 08).

If a 74HC08 is being used, a wire can be taken direct from its pin 11 to pin 19 of 74HC244. On the other hand, if a 74HC00 is used, the next gate must be inverted (Fig.5).

This leaves just one more wire, from pin 4 of new IC2 to the Address Select (Fig.2).

When the DSP computer is started again following memory expansion, the operating system must be informed: the command is N 200000 200000.

3. LITERATURE

(1) Matjaz Vidmar, YT3MV: Digitale Signalverarbeitungs-Techniken fuer Funkamateure:

Part 1: VHF COMMUNICATIONS 2/88.
Part 2: VHF COMMUNICATIONS 1/89.
Part 3: VHF COMMUNICATIONS 2/89.
Part 4a: VHF COMMUNICATIONS 3/89.
Part 4b: VHF COMMUNICATIONS 4/89.

(2) Matjaz Vidmar, YT3MV: DSP Computer, Update 1. VHF COMMUNICATIONS 3/91.

Dipl.-Ing. Detlef Burchard, Box 14426, Nairobi, Kenya

A Cylinder-Parabolic antenna with compact Meteosat Converter

A weather picture on the TV screen always fascinates my visitors amazingly. It is not only useful and educational, but also in contrast to the local broadcast (''live'') television, not censored or subject to any political propaganda. Three years' reception of orbiting satellites led to the desire to try for METEOSAT as well.

The serious attempt to cross new frequency frontiers (for me) required suitable test gear: signal generator, noise source, counter and analyser. Here these are two or three times more expensive than in the first world, because of prohibitive import duty. Also there is no leasing company or person for thousands of kilometres around who could loan such apparatus for a while.

So, I was in the same position as many amateurs and had to improvise. The compact converter designed by Althaus (reference 1) made the project feasible.

First came a study of all the sources: back numbers of VHF COMMUNICATIONS, Funkschau, R&S News were all to hand. I won't recount everything here, only the questions that remained unanswered.

* What is polarisation?

* Which METEOSAT channel is the more important?

* Is the line start pulse compatible with the orbiting satellites?

* What sort of variation of level during the day and the year should one expect?

* How much noise does the sun contribute?

Most of these questions I can now answer and this may interest other readers.

It is quite impossible to fabricate a parabolic dish here, also the construction of such a

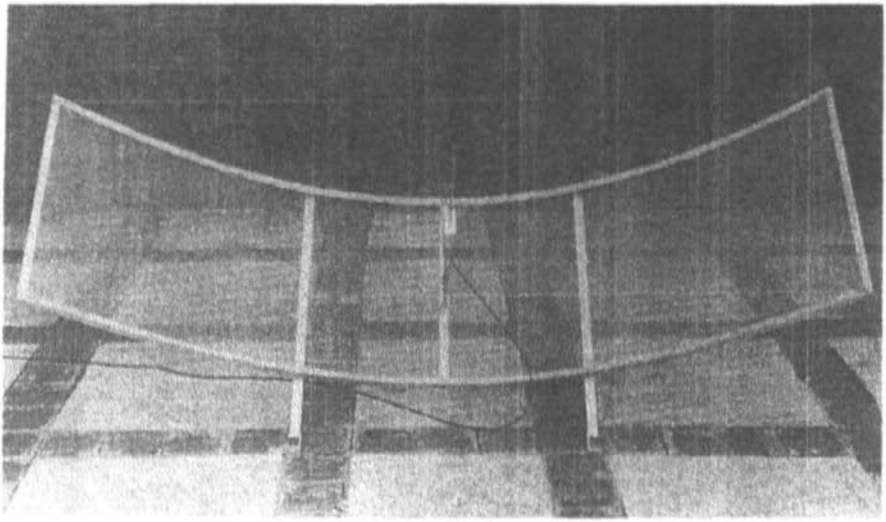


Fig.1: The almost complete Cylinder-Parabolic antenna during measurement

monster (which would be bound to attract the attention of the neighbours - and the authorities) appeared unnecessary. So I experimented first with long Yagis to Schaumburg's design (reference 2) one metre long. The results were so poor, however, that even lengthening them to 2 or 3 metres did not seem advisable. Any kind of corner reflector antenna, which is much easier to make, would be the same. These experiments did at least demonstrate polarisation; here in Kenya it is unquestionably vertical, with no component in the horizontal plane.

A receiver for 136 to 138MHz already existed and was going to be re-used. That it had scanning and AFC was not seen as a disadvantage. The METEOSAT signal of course has no Doppler effect, but the oscillator in the converter has a temperature drift of around 10 parts per million, which corresponds to 17kHz frequency shift and is not tolerable. This receiver also has correlation detectors for the line start, separate for METEOR and NOAA satellites. Neither of these is suited for METEOSAT. I will mention here at the outset what happened: the line start pulse of METEOSAT comprises six oscillations of 1/3 of the subcarrier frequency. Observation led me to presume that the oscillation is phased locked directly to the subcarrier, which is not the case with the orbiting satellites.

1. EXPERIENCE WITH THE CONVERTER

The kit for the converter gave no problems with construction or alignment. Later some inadequacies presented themselves, which were sorted out after a few weeks of testing.

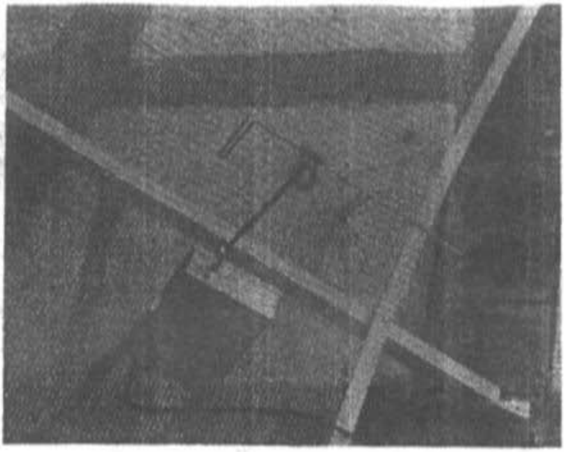


Fig.2: Details of the Colinear Dipole and the Balun

First to appear was that only channel 1 (1691MHz) sends weather pictures 24 hours a day. If this channel is to be worked upon in this receiver, the 97.09375MHz crystal must be changed. The data for this crystal are not given in reference 1. A TQ330514, that is with 10 ppm tolerance and temperature characteristics, turned out to be adequate for this application.

The current drawn by the converter was initially 85 to 105mA in the ambient temperatures of +10 to 60 degrees. This seemed somewhat high and could not be delivered by the existing receiver. Without noticeably altering the characteristics of the converter, the current consumption could be reduced to 55 to 58mA by selecting different operating points of transistors T3 and T4. This was achieved with new values of R3 (6k8), R4 (470R), R5 (220R), R6 (3k9), R7 (470R) and R8 (82R). The reduced attenuation of L3 must be produced afresh by a further 220R resistor from the L3 tap to ground. A test into reducing the current taken by T5 from 20 to 17mA (R9 = 56R) proved to be unfavourable. Amplification was reduced and noise level rose. The same applies for T6.

2. CHOICE OF ANTENNA FORM

A cylindrical parabola is a much more complete reflector than a corner reflector. At the same time it is easy to fabricate because it can be unwound. Virtually all the energy hitting the vertical surface is reflected onto the focal line, where a colinear dipole of appropriate dimensions can be located.

Behind a long yagi this kind of reflector can be made quite roughly with an area 0.5 x 0.5 meter of chicken wire mesh fixed in a wooden frame. A threefold colinear dipole has a radiating resistance of around 300 ohms, and with a shunt loop balun this can be transformed to a quarter of this figure, presenting a by no means bad match to the converter. With this definitely better reception was possible immediately, and it was possible to work out how much larger the eventual antenna would have to be. An improvement in receive performance of 6dB appeared adequate, necessitating an area of 1 square metre. A breadth of 2 metres and height of 0.5 metre were selected.

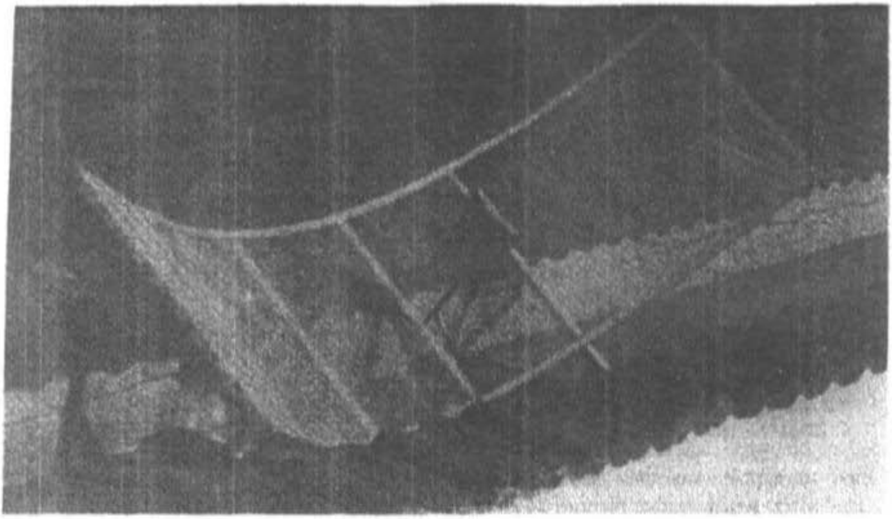


Fig.3: The Antenna on the house roof. For test operation the converter is concealed in a pocket of waxed cloth, joints and connectors are covered with silicone rubber. A piece of fishing net has been installed to prevent monkeys from making attempts at *'improvement'!*

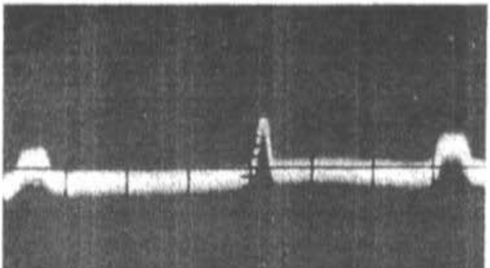


Fig.4:
Signals at IF level.
Far left: raw data.
Centre: WEFAX.
Rigbt: PDUS.
Far right: MDD
X: 1MHz / division
Y: 10dB / division
Analysis width 10kHz, video filter 1kHz, time 1 second.

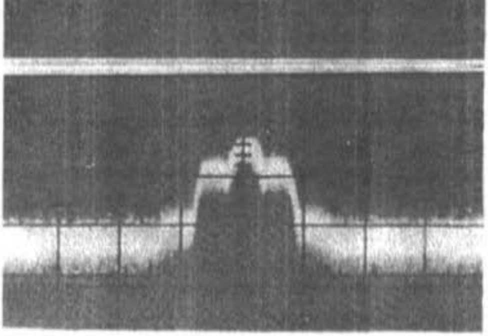


Fig.5:
The WEFAX signal of 1691MHz during the stop tone.
X: 20kHz / division.
Y: 10dB / division.
Analysis width 10kHz, video filter 1kHz, time 1 second.

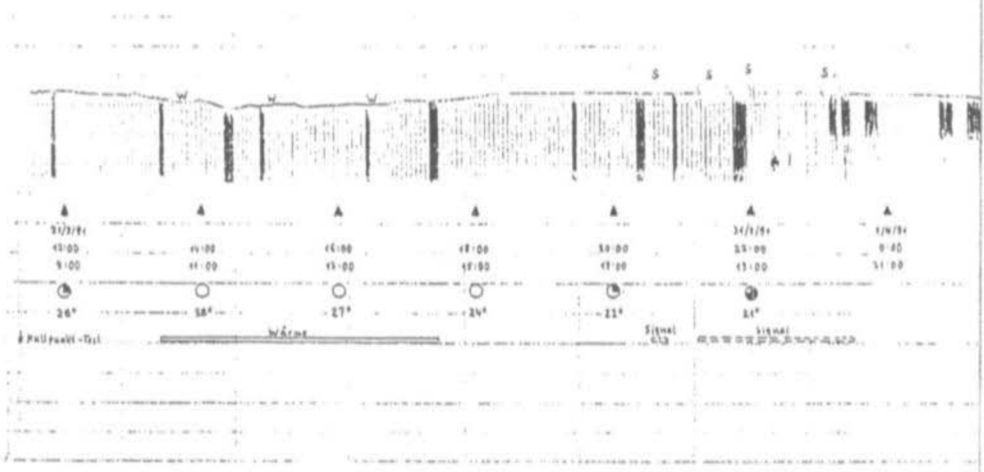


Fig.6: Record of reception over three days.

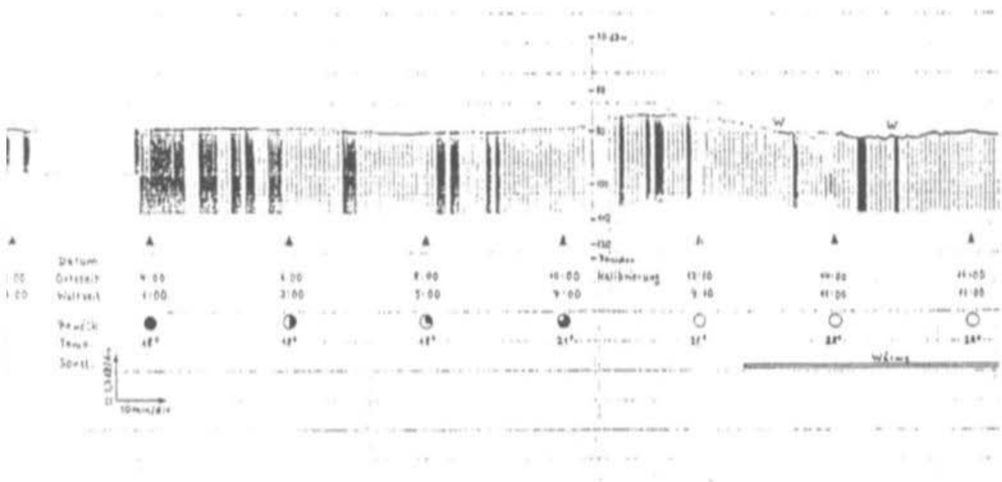
The finished antenna is shown in Fig.1. A framework of aluminium profile is covered with wire mesh. The centre strut has two supports for the converter. The input socket of the latter (changed here to SMA) carries the antenna made of thick copper wire via a short piece of rigid copper-sheathed cable. These details plus the balun can be seen better in Fig.2.

A choice may be had of several parabolas, of which only the aperture is the same. The selection here was of those whose focal line lay inside the area of the aperture. For bending the east-west profiles we draw the parabola on the workbench surface or on the floor of the workshop. The old maths textbook from school is of assistance here! Then we bend the profiles as closely as possible into these forms. A deviation of up to half a millimetre either way is permissible.

The remainder of the work is riveting or screwing together and offering up a suitable piece of wire netting, soldered at the cross points. Later on a covered the whole thing with a piece of fishing net as a troupe of monkeys (cercopithecus aethiops) were performing gymnastics on my house. Tensioning wires between the corners add stability.

METEOSAT here stands at 47.5 degrees elevation in the west. The antenna stands on the roof were already designed for this elevation before installation. This is adequate because the focusing of the antenna in the vertical plane is not very critical.

Two planks are laid on the roof in the east-west direction and the antenna is placed upon these (Fig.3). Only a slight correction of azimuth was needed to achieve optimum reception. After this the antenna was screwed onto the planks, which were loaded down with heavy stones.



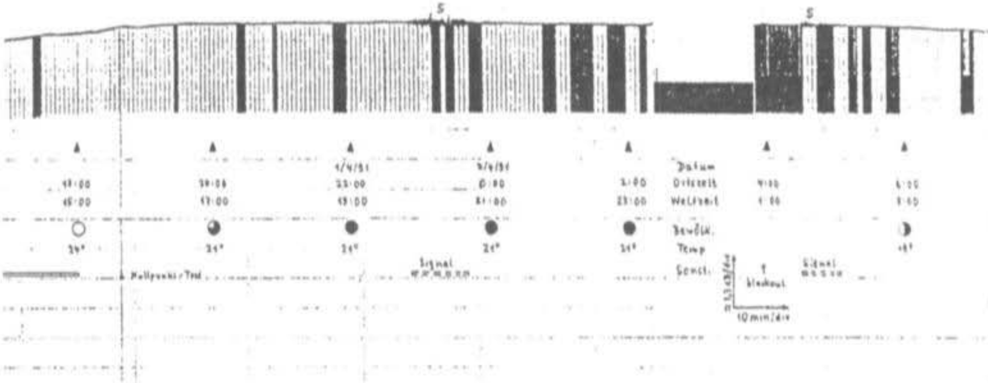
3. MEASUREMENTS

Naturally this antenna gives fault-free reception straightaway. An analyser on the converter output gives more precise information (Fig.4). The tall pinnacle in the centre is channel 1 on 1691MHz. At the analysis bandwidth selected, signal to noise ratio is around 10dB.

The broad peak to the right is channel 2 on 1694.5MHz. Most of the time it carries a modulation 0.5MHz wide, with has too little signal to noise for broadband demodulation. Also the carrier level cannot be determined at this analysis bandwidth. For the short periods that channel 2 carries WEFAX, it displays the same signal to noise as channel 1.

The small peak visible on the far right appears on high resolution to be two lines at 1695.73 and 1695.78MHz. Finally on the left is another signal like channel 2, but on 1687MHz. It is not transmitted all the time and never carries WEFAX modulation. Insiders will certainly know what these other channels are good for; I am unable to add any knowledge. (Editorial note: the relevant details are given in the picture captions).

The carrier of channel 1 reaches the analyser with -89dBm. Accepting that the converter has a typical amplification according to reference 1 of 26dB and the feeder cable (22 metres of RG-58) has a loss of 4dB, the antenna delivers -111dBm. However, since I cannot measure the converter gain accurately this is only a speculative, albeit good value. The operational bandwidth is 30kHz. It is easy to calculate then that the signal to noise ratio is 18dB. From Fig.4 it is additionally possible to discover that the amplification drops by 4dB at low frequencies. Tests to optimise the matching led nowhere. If the antenna is replaced with a 50 ohm load, the base noise goes up 8dB.



This shows the system noise temperature to be less than 50K (!).

A WEFAX signal with high resolution is shown in Fig.5. The unmodulated carrier lies 4dB above the central maximum level of this picture. The signal to noise ratio at 10kHz bandwidth is therefore 23dB.

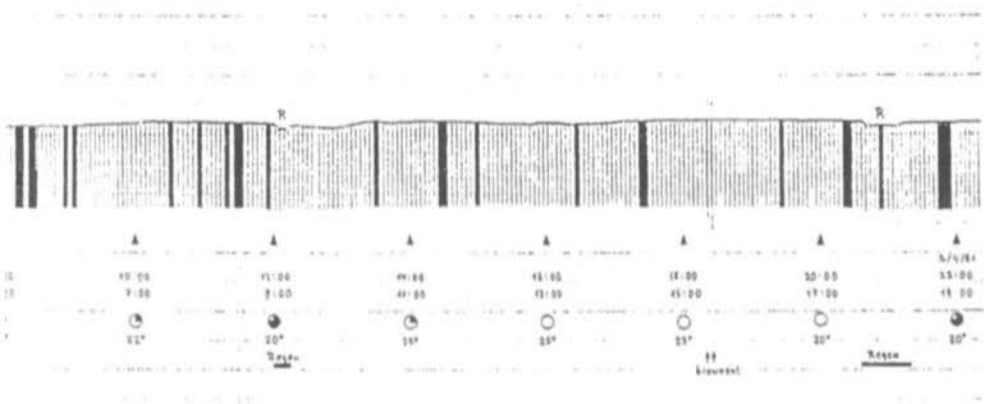
4. OPERATIONAL EXPERIENCE

It was possible to study the influence of the sun during the March equinox. On a total of 10 days light to heavy interference to the picture quality was detected between 14.00 and 17.00 local time. It peaked as total blackout of the picture on the 21st and 22nd March for an hour each time. The analyser showed then that base noise had risen by 14 to 15dB.

Since noise disturbance was frequently noticed on sunny afternoons and about 3 hours after sunset, a pen recorder was eventually connected which showed signal and noise levels over a period of three days (Fig.6). To enable this recording, I had to anticipate the working of the receiver to an extent.

The recorded voltage comes from the IF IC TDA1576 (pin 13) and is pretty linear to the dB. Compensation was made for its significant temperature dependence. The calibration was performed after the first day, so that each ordinate can be related to a particular input level on the receiver. This is identical to the converter output level less 4dB. The forward feed is consistently 10 minutes per section.

The receiver goes into scan search if the input level drops below -95dBm or if the deviation of the 2400Hz subcarrier drops below plus or minus 2.5kHz for more than 5 seconds. This happens regularly during the stop tone of the



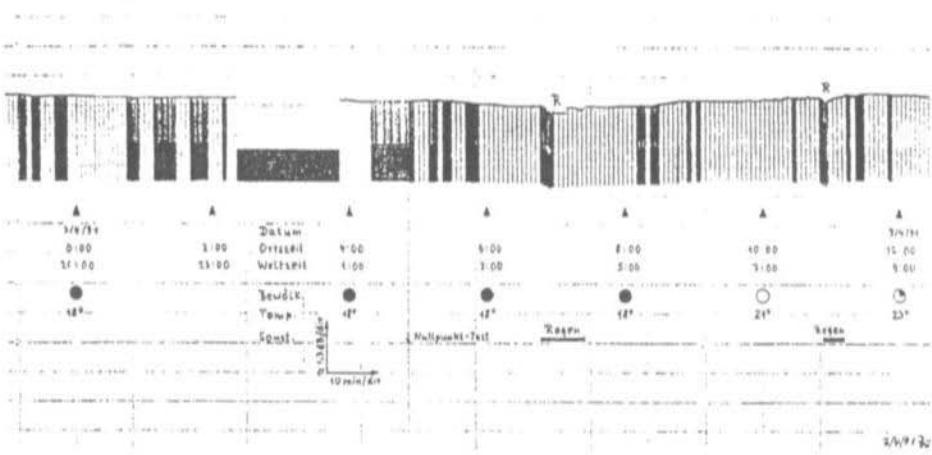
WEFAX signal. The lower limit for recording is thus the noise level, the upper is the signal level. The signal to noise ratio is simply read off from the level of the recording. One can further make out the individual pictures of the transmitted programme, also the times without picture modulation and without carrier.

The signal to noise ratio alters considerably with temperature, as can be seen by reading the notes of the local weather observations. Particularly when it is very warm (W) and when it is raining (R) the level alters. And then there are also spontaneous changes in signal level, marked S.

The W indications are explained by alterations in the gain of the converter, when the temperature varies significantly from the temperature at which the alignment was carried out. If the input transistor is warm, then the noise figure rises.

The R occasions are probably related to the fishing net, which , when wet, represents additional attenuation ahead of the antenna aperture. Whether I can remove it depends on the behaviour of the apes.

For the S occasions I have not yet found a reason. On the weather picture one notes in the middle of perfectly noise-free picture sections every 30 to 60 seconds one or two noisy lines. The usable signal can even lie up to 3dB above normal, then suddenly drop 10db below. Its course over time has a lot of similarity with multipath reception. The effect was noted on several evenings before recording began and it is also visible on the first day of recording. Unfortunately a recording over an extended time axis was missed. The next day the effect reappeared much reduced. The question arises whether something similar has been noted previously and what causes led to this.



Of course nothing has yet been reported about seasonal variations.

5. SUMMARY

A cylinder-parabolic antenna for METEOSAT reception is relatively simple to fabricate. Its receive sensitivity with a modem converter is significantly better than the combination named in reference (2). An antenna of the size given will normally provide signals with 4 to 5dB reserve over the FM threshold. All the same, in many circumstances this will not achieve undisturbed reception.

With this antenna small improvements are still possible. A reflector in front of the dipole would improve the illumination, and a lengthening of the colinear dipole to 5 elements would increase gain some more.

The converter should be protected from direct radiation from the sun to prevent overheating. It might be even better to remove the converter from the antenna and connect it by a piece of low-loss cable as short as possible.

6. LITERATURE

(1) M. Althaus, DF9DA: Compact Weather-Satellite FM Receiver. VHF COMMUNICATIONS 2/90.

(2) A. Schaumburg, DF7ZW:Receiving METEOSAT with Yagis. VHF COMMUNICATIONS 1/88.

Jim Toon, G0FNH

10GHz ATV The Easy Way Part-3

Following on from my previous articles in VHF Communications 1 and 2/91, "10GHz The Easy Way", I am now going to describe in detail how to manufacture the following:

Waveguide Flanges

The Dummy Load

The Cross Coupler

The Diode Detector

all of which are used in the main unit. The only tools I was *ALLOWED* to use by Editor Mike, were a file, a hacksaw and a drill and bits.

All of the materials used, excepting the waveguide and the detector diode are obtainable from your local hobby shop and DIY store.

1. WAVEGUIDE FLANGES

Brass waveguide flanges are very easy to make, all that is required is some 1/8 to 3/16" brass plate. With reference to Fig.1, carefully mark out the plate with a scribe and square. Drill the bolt holes for 3BA size bolts and then drill a series of small holes on the inside of the lines marking out the middle hole. Place the brass plate on a very firm surface, an anvil or steel plate for example, and with a sharp chisel, carefully cut through the holes and punch out the middle rectangle of brass. File the hole until all the angles are square, and then continue filing until the flange is a tight fit over the end of the waveguide. When satisfied with the fit solder in place **ON THE OUTSIDE ONLY**.

NOTE: The type of solder I use and recommend is **LEAD-FREE,** which can be obtained from your local DIY store or from a

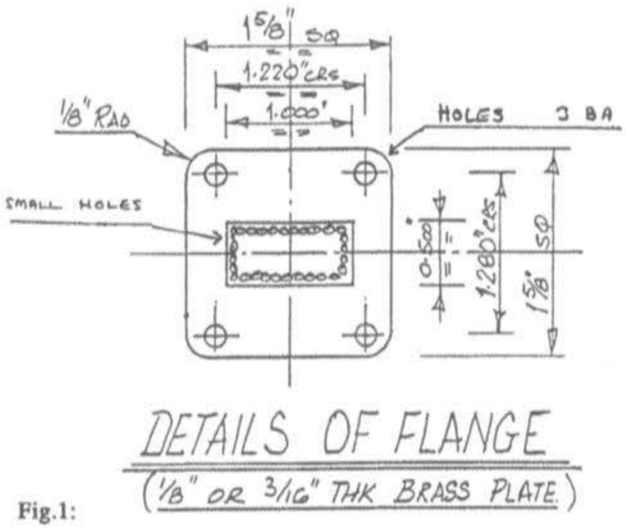


Fig.1:

friendly plumber. Remember, microwaves adore Lead, they will do anything to disappear into Lead and not come out again. Second only to open waveguide or LDM material (Lossy Dielectric Material - see next section), Lead is probably the most lossy substance you can use. Ensure that any solder containing lead is carefully scraped out from the inside surfaces of the waveguide.

2. DUMMY LOAD

Referring to Fig.2, you will require a 2'' section of waveguide with the ends nice and square, a flange and a small brass plate measuring 1'' x 1/2'' x 1/16''. Solder the flange on one end and seal the other end by soldering on the brass plate.

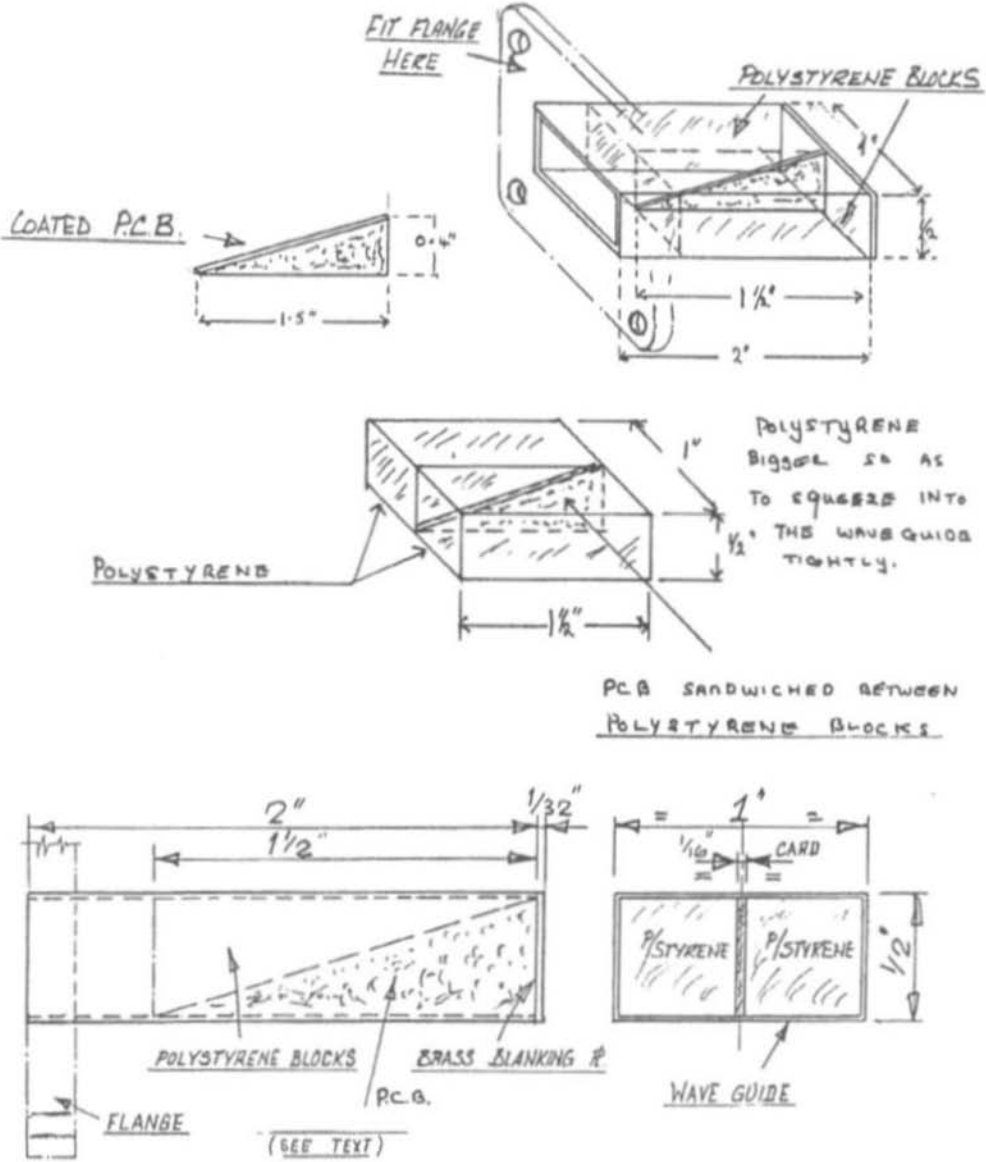
You will need to obtain some Graphite powder (approximately 70p per box) and ready-mixed wallpaper glue (**DO NOT** use the resin or solvent type, they will not work - only water-based paste) from your local DIY store. The wallpaper paste can be purchased in small handy-sized tubs.

Now to make your own Lossy Dielectric Material. Prepare a small piece of PCB (at least 2'' x 1'') and remove the copper surface. Mix together a small amount of the wallpaper paste and graphite powder to form a thick paste.

Cut the PCB to the size shown in Fig.2 and then apply a thick coat (1/16'') to both sides and all edges of the piece of PCB and make it as smooth as possible. Put in a warm place (the airing cupboard above the home-brew will do, but keep it clear of the washing!) for 24 hours until dry, then smooth it up with sand-paper to remove any irregularities (what a big word!). Cut some expanded polystyrene to the sizes shown in Fig.2.

NOTE: Expanded polystyrene is transparent to microwaves. A useful tip is to fill the end of your Solfan (or whatever) heads with it - it will keep the warmth generated by the Gunn diode in and the cold, damp air out.

Sandwich the PCB coated PCB board between the two pieces of polystyrene and push the whole assembly into the piece of waveguide, making sure that it is a tight fit, and that it is the correct way round, as shown in Fig.2, and that's it.



DETAIL OF DUMMY LOAD

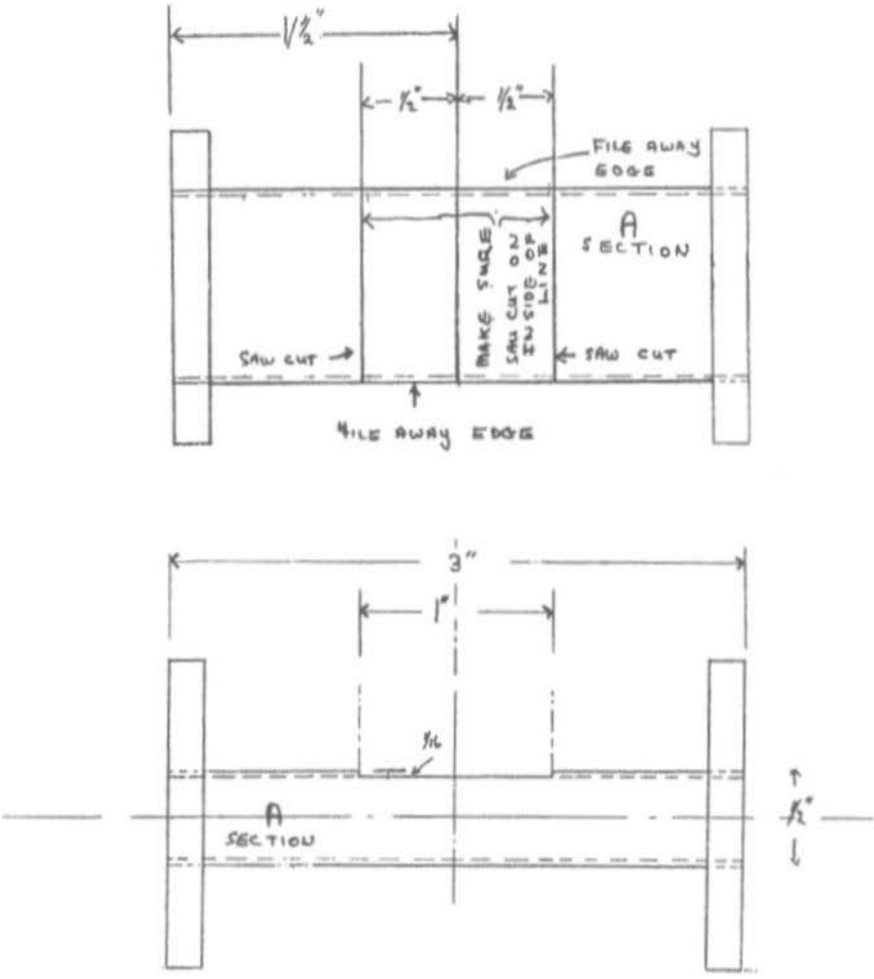
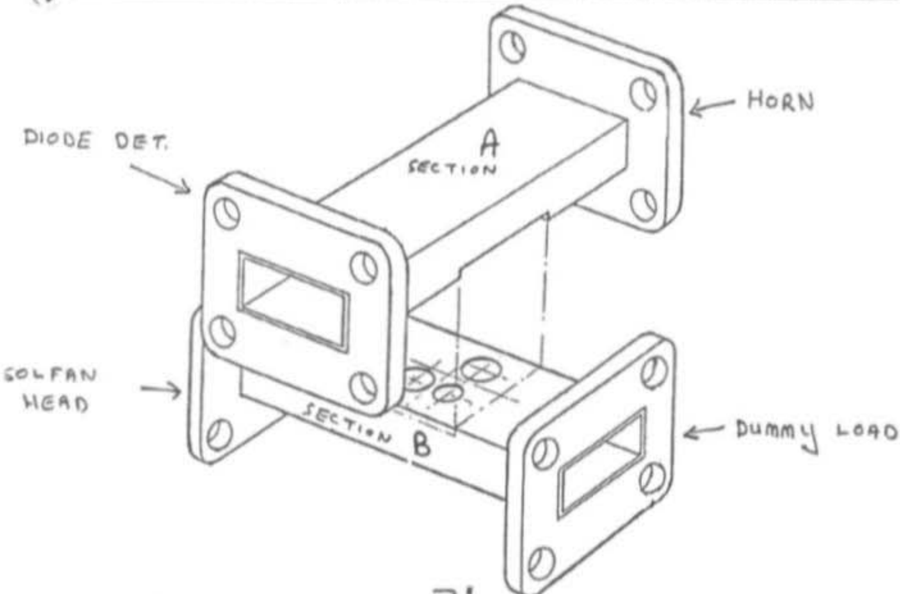

FILE AWAY EDGE
A SECTION
MAKE SURE
SAW CUT
SAW CUT
FILE AWAY EDGE
3"
1"
A SECTION

Fig.3a


HORN
DIODE DET.
A
SECTION
SOLFAN HEAD
SECTION B
DUMMY LOAD

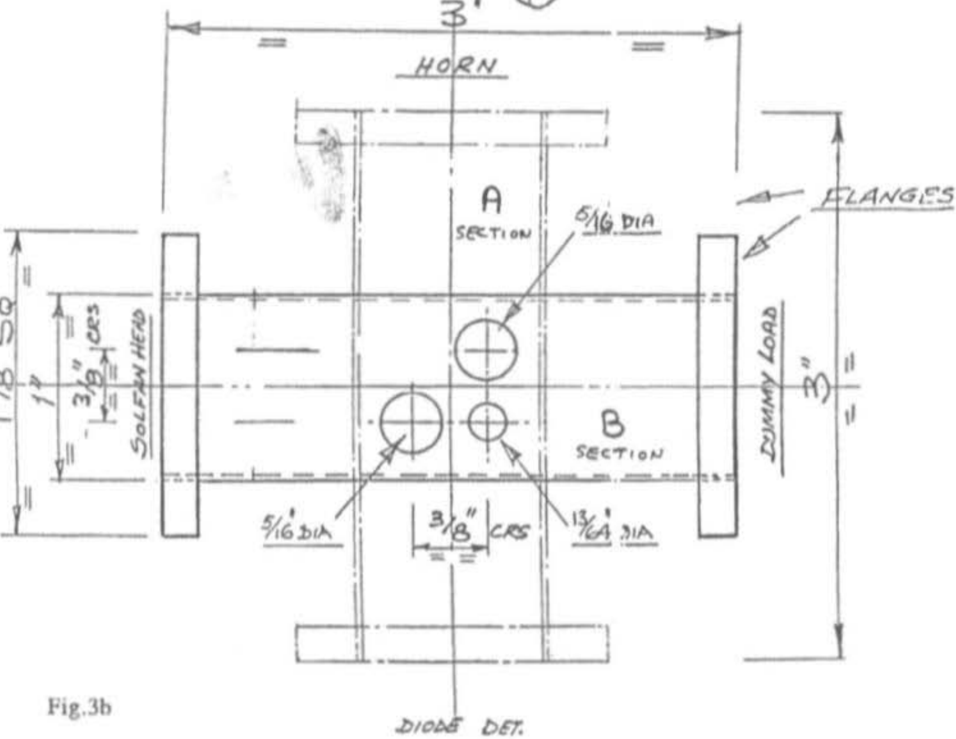

3'
HORN
A
SECTION
5/16 DIA
FLANGES
1 5/8" SQ
1"
3/8" CRS
SOLFAN HEAD
DUMMY LOAD
3"
B
SECTION
5/16 DIA
3/8" CRS
13/64 DIA
DIODE DET.

Fig.3b

3. THE CROSS COUPLER

The details of this unit are shown in Fig's.3a and b. For this unit you will need two pieces of waveguide, each 3'' long, and four flanges. Make sure that the end faces of the waveguide are nice and square. Study the drawings in Fig.3a and b carefully, do not be put off by the apparent complexity, you will find that it is really very simple. Take your time and mark out everything carefully, take care with the construction and the results will be excellent.

We will start with section-A. Mark off the middle point of the piece of waveguide (across the broad side) and then mark off 1/2'' on either side of the centre line. Mark down the sides of the waveguide 1/16'' and then scribe along. With a fine hacksaw carefully and slowly cut across the waveguide on the centre line side of the scribed lines until you just break through the waveguide.

Then, with a small fine file, file along the edge of of the waveguide (but not below the 1/16'' line) until a fine crack appears. Do this on both sides. With a small screwdriver at the corner, lift up the middle section, remove and discard. File the slot section carefully until the other piece of waveguide will fit across the slot with a nice tight fit and **FLUSH** on the inside.

Now to section-B. As with section-A, mark off the centre line on the broad side of the waveguide and then **VERY** carefully mark of the centres of the three holes at the positions shown in Fig.3b. Centre-punch the holes carefully and gently (we don't want to tune the waveguide!). Drill the holes with a 1/8'' pilot hole first and then the 13/64'' hole, and finally the two 5/16'' holes. Clean out any burrs inside and out.

Fit the two sections of waveguide together, again ensuring that the inside surfaces are flush. When satisfied solder the two pieces together on the outside. Solder on the four flanges and clean out all solder from the inside surfaces, and that's it.

4. DIODE DETECTOR

The details of this, the most complicated unit, are shown in Fig's.4a and b. To fabricate it you will need a type CV2154 coaxial diode, these are available at most rallies or from J.Birkett, Lincoln (0522 520767).

Also required are: 1 flange, a piece of waveguide 1.5'' long, a brass plate 1'' x 1/2'', a short length of 3/8'' outside diameter brass tube and a short length of 13/32'' outside diameter brass tube (the brass tube is readily available from model/hobby shops, as is the brass plate). The larger size of tube will just slide over the smaller size, and also the coaxial diode will just fit into the smaller tube.

A small length of 3/32'' brass rod (non-fluxed brazing rod is just right) and a 75 ohm BNC socket (MUST be 75 ohm) will also be required.

Starting with the BNC socket, cut around the small burr at the end as shown in Fig.4a with a small hacksaw, not too deep as the PTFE bush inside will be required later. Carefully push out the centre pin and the two PTFE spacers. Discard the case.

Take the piece of waveguide and square up the end faces. Scribe a line 3/8'' from one end across the broad side and also down the two narrow sides. Mark off the centres of the faces on the lines just scribed (i.e. 1/2'' for the broad side and 1/4'' for the narrow side). Carefully and gently centre punch the centre points and drill a 1/8'' pilot hole in each. Drill the holes in the narrow sides out to 3/16'' and the single hole on the broad face out to 3/8''.

Clean out any burrs and then solder the flange on the **OPPOSITE** end to the end with the holes. Clean out any solder from the inside of the waveguide.

Prepare a 5/8'' long piece of the 3/8'' OD brass tube and solder it into the 3/8'' hole in the broad side of the waveguide, ensuring that it is flush on the inside face. Prepare a 3/8'' long piece of 3/8'' OD brass tube and solder it centrally over one of the 3/16'' holes in one of the narrow sides.

Prepare the two **PTFE** bushes as shown in Fig.4b and fit one into both of the 3/16'' holes in the narrow sides of the waveguide (one of the bushes from the BNC socket should fit as is - if you have another spare BNC socket you will not need to manufacture the second bush from the spare block of PTFE, but just use the one that already fits from the second socket).

Cut a small round disc from a piece of singled-sided copper-clad board, and solder it into the piece of 3/8'' tube soldered to the **NARROW** side of the waveguide. Solder it flush with the top of the brass tube.

Now cut the 3/32'' brass rod to length (one inch and three eighths). Mark off 5/8'' from one end and centre punch. Drill a hole through the brass rod to the size of the centre pin from

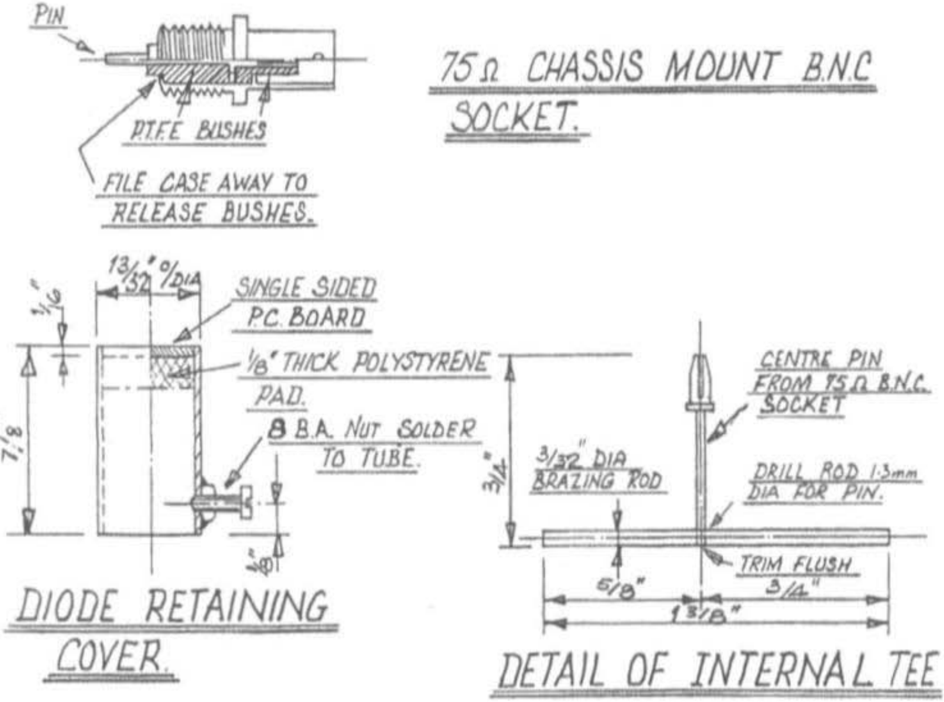
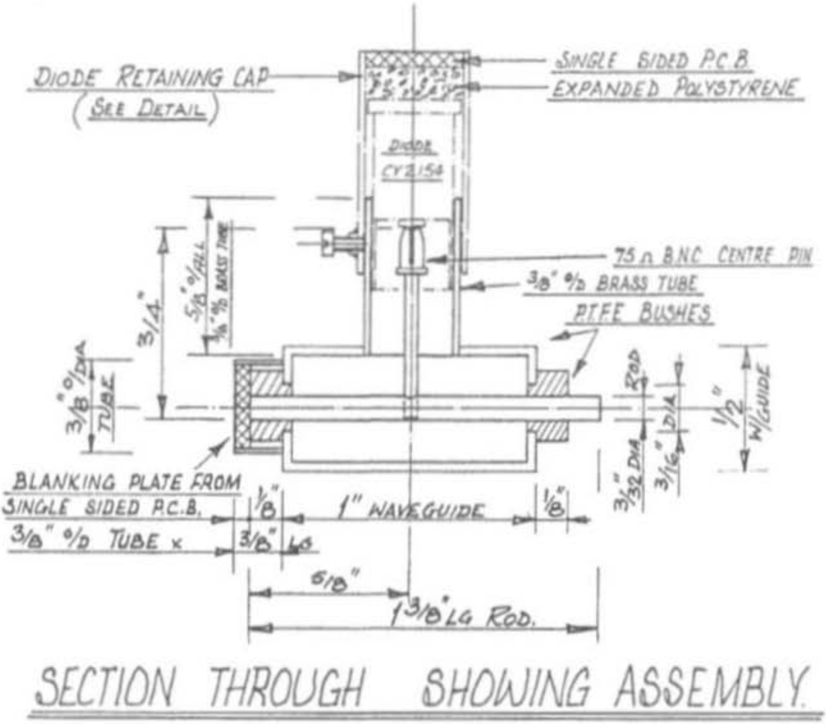


Fig.4a: The Diode Detector



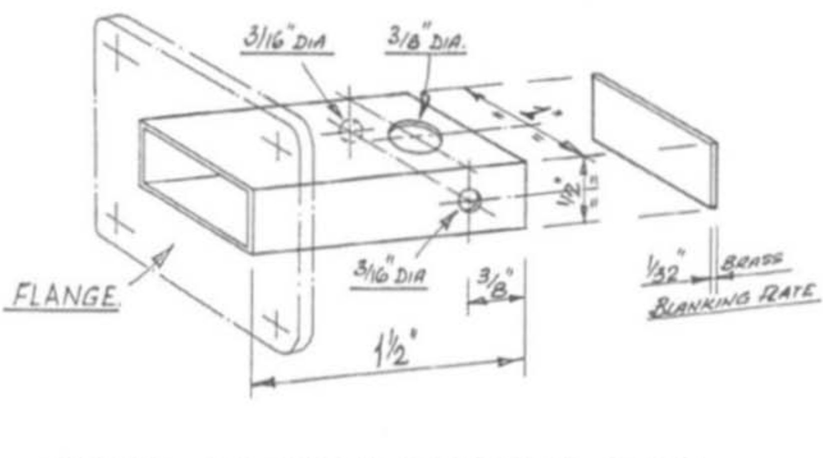


Fig.4b The Diode Detector

the BNC socket (a PCB size drill is the correct size). Fit the brass rod into the waveguide as shown in Fig.4b, pushing the rod through the holes in both PTFE blocks so that it is firmly in place.

Temporarily assemble the pin and the diode into the brass tube allowing the pin to pass through the hole in the brass rod (handle the diode with care -static charges may have a harmful effect - use a dud one if available). When the diode is fully home in the brass tube cut off the excess pin projecting through the brass rod.

Remove the diode and the pin, ensure that the hole in the brass rod is central with the brass tube, re-insert the pin and solder in place. Clean away as much solder from around the pin as possible. Now solder the end plate onto the waveguide being careful not to get any solder on the inside, as it will now be extremely difficult to remove - this is really a must for using lead-free solder as it will not matter.

All that is left to do now is is to make the top cover for the diode holder as per Fig.4a. Fabricate the cover from a 7/8'' long piece of the 13/32'' OD brass tube and solder an end cap on one end using a piece of copper-clad PCB or brass plate. Drill an 8BA clearance hole 1/8'' from the other end of the piece of tube. Solder an 8BA nut to the outside of the tube over the hole and fit an 8BA screw into the nut.

Fit the diode into the holder ensuring that it ''clicks'' into place on the pin. Place the retaining cover over the diode and tighten in place.

I suggest that you make a small box from brass plate to house the preamp (page-62 CQ-TV 152). Drill a hole through the bottom plate of the box to enable it to fit over the PTFE bush and the brass rod extending from the side of the detector unit. Solder or glue the box in position and connect the preamplifier input to the end of the brass rod. The IF output and DC power connections to the preamplifier should be made using coaxial cable.

NOTE: Don't forget the 10k resistor to be soldered from the diode connection at the input of the preamp to earth - strange video effects may occur otherwise!

I spray my units with matt black paint after first covering the flanges with tape to protect the inside surfaces. If you take care in building these units the end results will be very good.

I would like to thank Ted G4GLY for his help in the preparation of the drawings and the RSGB for the information gleaned from the VHF/UHF Manual. I hope that you have as much fun playing with microwaves as I do.

In the next instalment I shall describe how to manufacture a large horn antenna and a 10GHz Dipole antenna.

5. REFERENCES

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

2) Mike Wooding G6IQM, the British Amateur Television Club, The ATV Compendium.

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

Dr.Ing. Jochen Jirmann, DB1NV

A Digital Image-Store for the Spectrum Analyser Part-2

In issue 3/1991 the digital image-store of the spectrum analyser was presented for the first time, although at that time no PCB layout was available. The author spoke to a number of amateurs who had CAD facilities on their PCs, but these auto-routing programs were unable to create a layout for the components which did not rely on very fine track widths.

The author therefore designed a layout by hand which was not as elegant as that devised by computer but which fulfilled the following requirements:

* the PCB is, at 100mm x 150mm, only 30 per cent larger than the hand-wired version;

* fine track techniques are avoided, and only one track need be taken between each leg of the ICs;

* the PCB did not require plated-through holes, since all through-board connections could be made by hand;

* almost all signal racks run on the underside of the board, while the ground and operating voltages are laid out on the top side, allowing comparable results to be achieved with the first design, even if some signals were now not as clean as in the earlier construction.

1. CIRCUIT ALTERATIONS

In the disentangling process a 10-pin Molex-type contact strip was provided for assembling all the inputs of the operating elements.

The contacts to the printer were now connected so that direct connection to a 25-way Sub-D connector was possible.

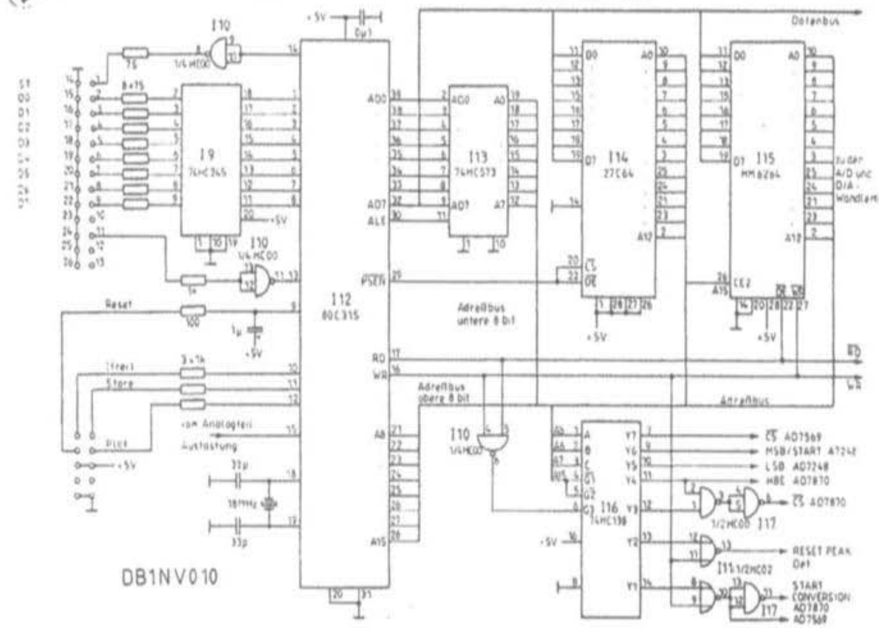


Fig.1: Digital Image-Store for the Spectrum Analyser.
Part 1: Processor, reworked for PCB DB1NV 010

Since the arrangement of some ICs has altered, the relevant parts of the circuit are shown once more in Fig's.1 and 2. The most important change is that in order to simplify the layout, two gates of a 74HC0 (I10 and I17) are now available on the board, although this could have been achieved with a single module of gates and wiring across the board.

2. COMPONENTS

To simplify the procurement of components, the latest parts list is given again; for uncommon components a supplier is given with which the author has had good experience in Germany. Readers in other countries may prefer to find a local distributor.

2.1 Integrated circuits:

I1, I2	TL074, TL084 (Texas Instruments)
I3	CD4053 (Harris/RCA), MC14053 (Motorola)
I4	AD7569JN (Analog Devices, from SASCO)
I5	AD7870JN (Analog Devices, from SASCO)
I6	AD7248JN (Analog Devices, from SASCO)
I7	79L05 (various suppliers)
I8	7805 (various suppliers)
I9	74HC245, 74HCT245 (various suppliers)
I10, I17	74HC00, 74HC132, 74HCT00 (various suppliers)
I11	74HC02, 74HCT02 (various suppliers)
I12	P80C31S (Matra-MHS, from Enatechnik) or failing this, 80C31 (see text of part 1)
I13	74HC573 (various suppliers)

I14 AM27C64-90DC or AM27C64-75DC (AMD, from Astronic)
I15 HM6264LP10 (Hitachi), MCM6064P10 (Motorola)
I16 74HC138 (various suppliers)

2.2 Individual semiconductors:

2 diodes 1N4001 or similar
1 diode 1N4148 or similar
1 zener diode ZPD4.3
1 zener diode ZPD5.1
1 N-channel switch-FET 2N4393 or BF247A

2.3 Passive components:

1 crystal 18MHz, HC49-U
2 chokes 68uH, 10mm lead spacing
4 precision trimmers 2Kohms, square upright shape, e.g. Bourns 3266W or Spectrol 64W
1 contact strip (Molex-type pins) 26-way
1 contact strip (Molex-type pins) 10-way

All capacitors are with 2.5mm or 5mm lead spacing. All resistors should be size 0204 if possible.

3. CONSTRUCTION OF PCB DB1NV 010

Refer to the layout plan in Fig.3 and first wire in the through contacts and the wire links. Any shortcomings in the through-board links will not be fun to correct later as many of them lie beneath ICs; maximum care is recommended. It took the author a whole weekend to de-bug the through contacts on a board (badly) made by someone else!

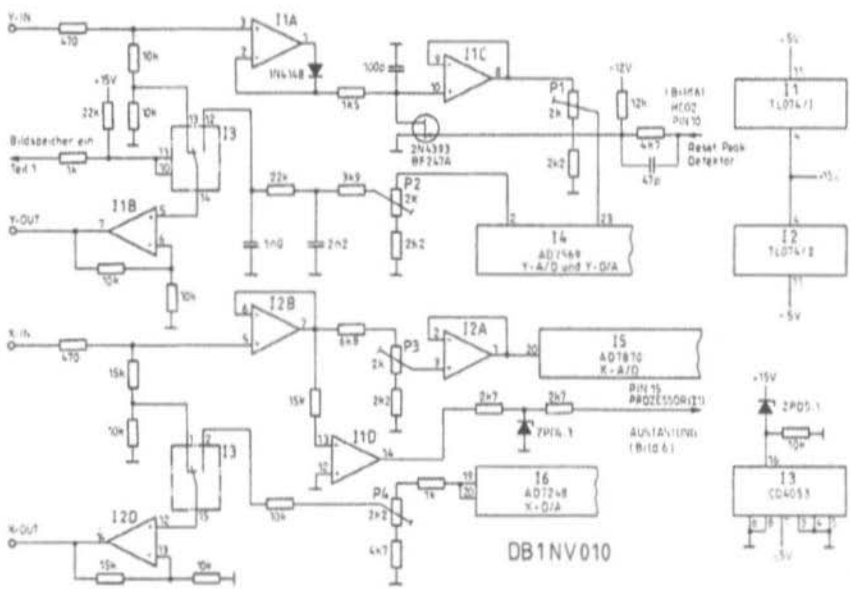


Fig.2: Digital Image-Store for the Spectrum Analyser. Part 3: Analogue circuits, reworked for PCB DB1NV 010

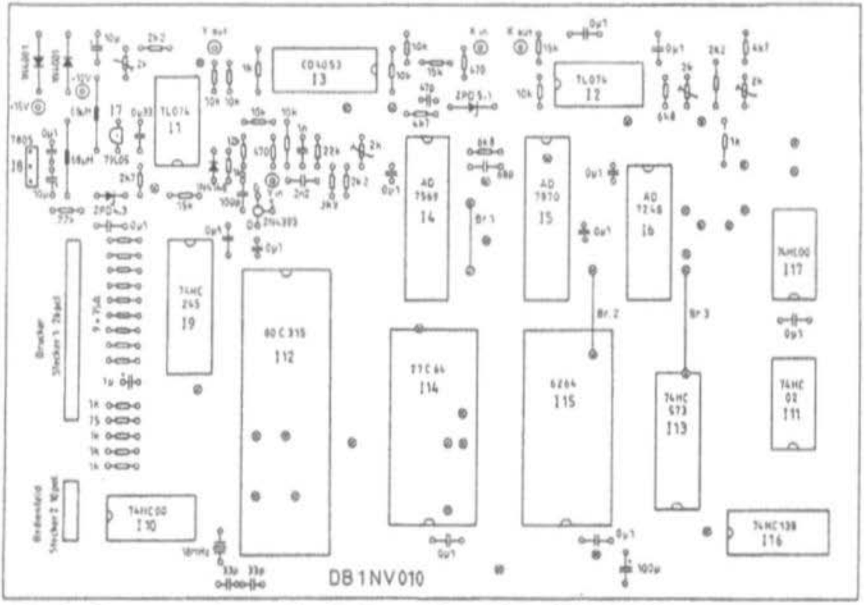


Fig.3: Component Layout for the Image-Store PCB DB1NV 010
O = throughcontact

After this process the other parts can be inserted, in more or less any order so longer as safety precautions are taken around the static-sensitive components. In this connection the author would like to give another warning about the dangers of 'socketitis', the excessive use of IC sockets. For a start, many of the components need to be soldered on both sides of the board, which the design of most sockets does not allow. Also cheap sockets are a source of poor contact and much aggravation. The author therefore recommends the use of sockets only for the EPROM (I14) and perhaps also the processor (I12) and RAM (I15), with the other ICs being soldered in direct.

For the capacitors a lead spacing of 2.5mm and 5mm is envisaged, and the resistors should if possible be size 0204. The larger 0207 pattern can be used if they are soldered in vertically. The holes for the trimmer pots have been designed to allow types with holes in-line in a row as well as those with the triangular arrangement.

A completed board, without the screening can yet fitted, can be seen in Fig.4. The sockets shown here for the processor I12 and the RAM I15 can be omitted.

After component fitting and visual checking for poor joints and solder splashes, the PCB can be commissioned as described in the previous part of this article. When making measurements bear in mind the altered arrangement of the gates and multiple op-amps.

Following functional checking, the unit can be soldered into a tinplate case for Eurocards, which will have room for fitting the printer

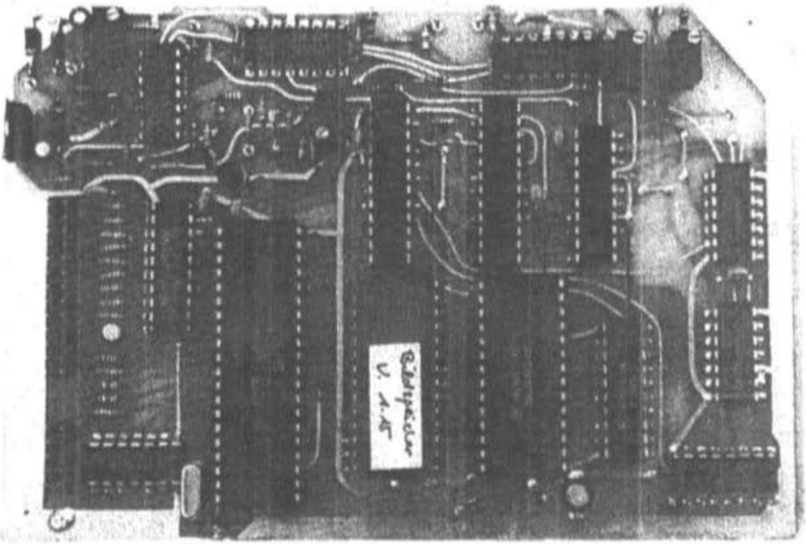


Fig.4: Sample construction of the Digital Image-Store DB1NV 010

connector if desired. The screening achieved is technically ideal, which is useful considering that the image-store will be mounted against the back of the unit, with the connector protruding.

Building the image-store into the analyser is now somewhat simpler since all the control lines are run through the 10-way contact strip and the 26-way connector now carries only the printer connection. If the printer connector cannot be built inside the screening can, the best way of making the printer connection is via a 25-pin Sub-D connector designed for flat ribbon cable and a matching piece of flat cable.

When making the connection to the 26-way contact strip, take care that the uppermost contact (pin 14 on the diagram) remains free, so that contact 1 of the Sub-D connector is wired to pin 1 of the image-store.

Lastly, an observation on the diagrams printed in Part 1:

In Fig.5 an error slipped into the address decoding. The "Start Conversion" signal for the AD7870 is, as drawn, taken from pin 1 of I11 (74HC02), but the corresponding signal for the AD7569 comes from pin 13 of I11.

Jürgen Langer, DJ5AT

UHF-Antenna with vertical Polarisation but no 'vertical' dimension.

The author revives an already known antenna form from obscurity, since it offers advantages for mobile operation in the UHF region.

1. PRE-HISTORY

I used to have a camping bus, the home-made extension roof of which was made of painted sheet metal - an ideal counterpoise. Now I am QRV on 70cm and would like to go out mobile as well. No problem, for me a 5/8 wavelength groundplane was the only choice. With a few breaks and bends this antenna achieved about 20km with 5 watts power in a partly wooded countryside.

The partner station radiated this power with a 6-element yagi (vertical polarisation). The theoretical, optical polarisation was about 14km away; topologically it lay much nearer. So this 20km seemed to be a practical benchmark.

If nothing else, the automobile manufacturer was a better metal-basher than I was, since the roof rusted faster than the rest of the bus. So I did what everyone else does and bought a plastic roof to stop the aggravation.

Now, however, the counterpoise lay some 60cm lower. And I would only pierce that thick, solid plastic skin under great pressure. Also it went across the grain with me to dive "through the bush" with the camping bus with the roof cluttered with wobbly, breakable and bendable bits and pieces.

The new antenna, if at all possible, should be under the roof but not take up any headroom and, because of the partner stations, be vertically polarised. The usable gain should be achieved only by vertical focusing with an equalised horizontal diagram.

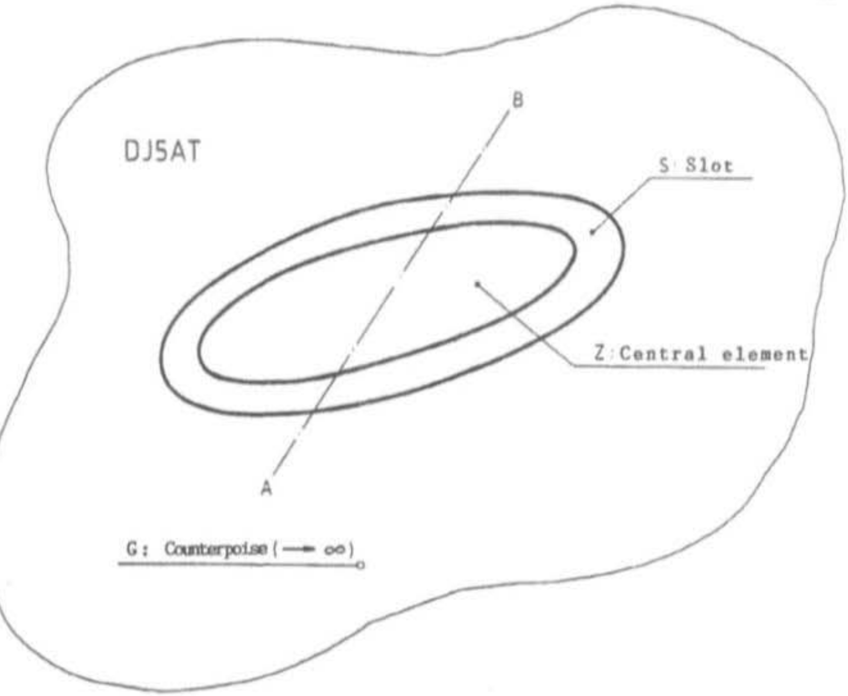


Fig.1a: Two-dimensional construction of a vertical radiator on infinitely large counterpoise

I attempted to use horizontally polarised, flat antennas such as the Maltese Cross (reference 1), hoping that reflections along the radio path would sort out the polarisation.

No luck, despite the best matching (s less than 1.5). The polarisation would have to match as well.

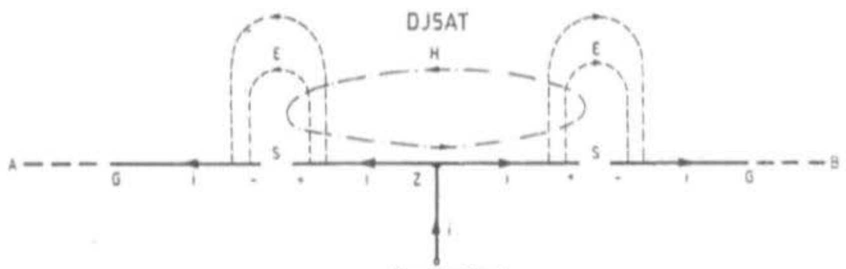


Fig.1b: Cross section A-B of drawing 1a. Field distribution above B, which must be considered to be symetrical below.

2. THE CIRCULAR, FLAT SLOT ANTENNA

During my researches I thought of planar antennas as well as slot radiators, as they are known in microwave technology. In reference (2) I struck lucky, even if I had nothing that resembled a counterpoise stretching to infinity. Figure 1 shows the principle (for vertical polarisation in this illustration). A horizontal, (for the time being) infinite and totally conductive surface (G) represents the "cold" counterpoise. Counterpoise to what? Well, to the whatyoumacallit, that is the central element or radiator, which only represents an antenna once it is surrounded.

The actual radiator is in fact a circular slot (S). The circular surface (Z), of which the inner dimension of the slot represents the outer boundary, has set measurements just like the slot itself. The energy that is transferred into space from the slot cross section is released there and must fall together in the correct phase with the diagonally opposite (across the diameter) radiation cross section. If the slot is activated in the same phase around its circumference, the diameter of the disc determines the sum of the phases and thus the radiation diagram which is now produced symmetrically above and below the counterpoise surface. The width of the slot sets the impedance of the slot, so long as it is significantly smaller than a quarter wavelength.

I had to make sure that too small a width (at the exact match) would also influence the radiation diagram. Precise knowledge was essential for optimisation here. Polarisation already makes itself felt in the course of the E-lines. It was even clearer to notice when the H-field was constructed. Current flowed radially from the centre towards the slot and from the other side of the slot back via the counterpoise (see Fig.1b). The sum of all the current paths on the inner (and outer) circumferences of the slot produce an H-field which runs parallel to the central plate (Z) and concentric to its centre. Exactly, then, the same result as a vertical radiator would achieve.

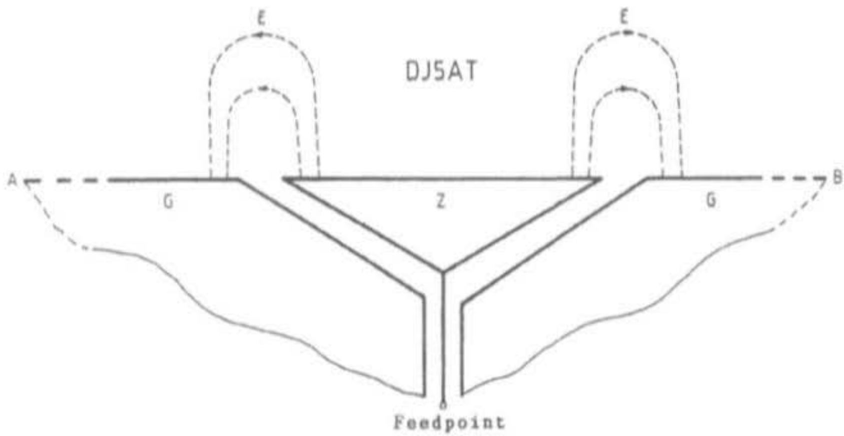


Fig.2: Principle of assymetrical radiation of the slot S and the consequence of the screening effect of the counterpoise G around Z.

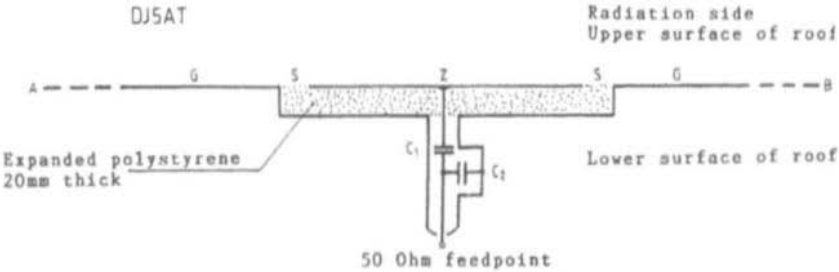


Fig.3: Cross-section A-B of the practical realisation with compensatory and transformation element C1, C2 with Z = 50 Ohms

2.1 First trial construction

My first trial construction, a rabbit cage wire surface in the vehicle with the dimensions 2m x 1.5m (these dimensions did not affect the frequency, they were chosen as large as could be fitted in), was provided with a slot gap of 3cm. Narrower gaps, e.g. 2cm, were much worse; larger ones were not tried. The formation worked extremely well, certainly in comparison with the 5/8 wavelength groundplane originally to hand.

This wire netting construction obviously could not remain loose in the car. I won't describe all my different trials and tribulations here in detail; nothing worked at first attempt, as some articles lead you to believe. Just this much: the antenna (at this stage fundamentally the surface of the counterpoise) was to be fastened to the plastic skin of the roof on the inside. The rabbit cage wire was easy to solder but turned out to be too inflexible to match all the contours cleanly. Fly-screen wire was very conformant, easy to glue as well, but would not take solder. My wish was self-adhesive copper foil but I could not find any. An alternative would be adequately-conductive paint - more of this later.

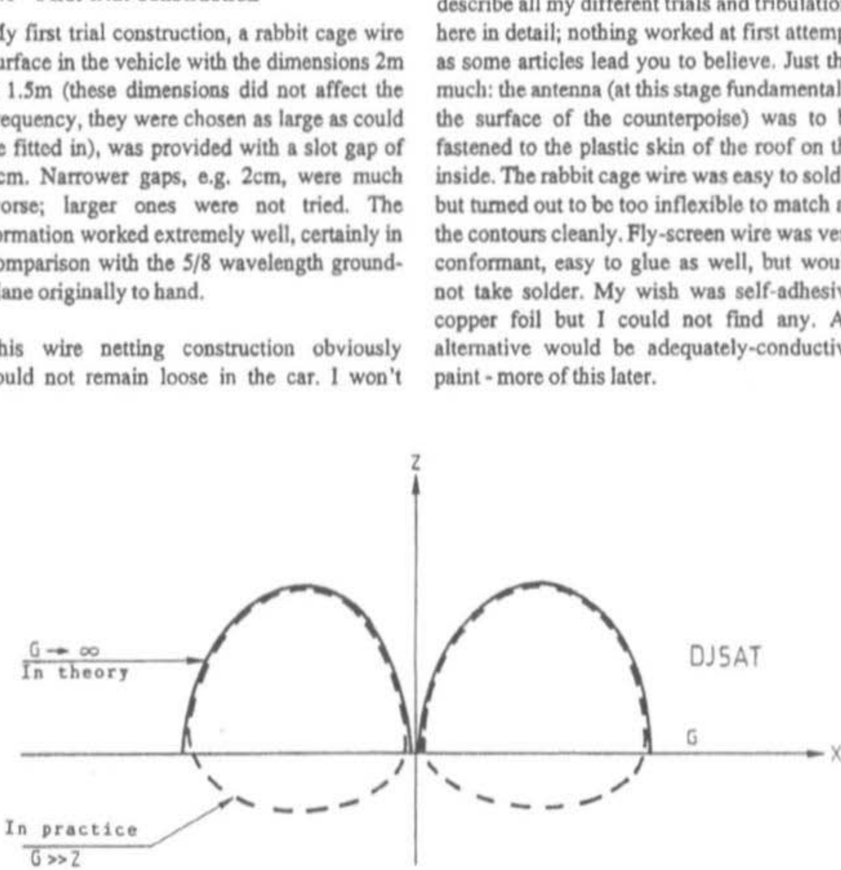


Fig.4:Radiation distribution with infinitely anf finitely (pecked line) counterpoises

Finally, I did find thin copper foil in a hobby shop. Gluing it on would have to come later. But up to now this tale has not been chronological. The questions of feeding and matching were yet to be solved.

2.2 Feeding

What is demanded is not any old excitation of the central, circular plate but the phase equality of the complete environment of the plate, as well as, of course, its opposite number, the counterpoise. The solution is already anticipated in reference (2), to broaden the inner conductor of the coaxial feedline conically until the diameter of the circle is reached (Fig.2). Clearly this cannot be achieved without some height.

My solution is show in Fig.3; the inner conductor is soldered to the centre of the round plate. What about the outer conductor? Because of the phase conditions, this too must be fed from the centre to the counterpoise, though isolated naturally from the central electrode. With this the symmetry of the upper and lower side is lost; radiation "downwards" is suppressed by the screening effect of the counterpoise below the slot and the central electrode. The theoretical vertical radiation diagram (with infinite counterpoise) is seen in Fig.4.

A finite counterpoise manifests itself in that the field meets itself again (pecked line in Fig.4). Here the original roof of the vehicle would be reflecting.

3. IMPLEMENTATION

The vehicle roof must be insulated against heat and cold. Expanded polystyrene is ideal for this and I use a thickness of 2cm. Since this material consists mainly of air, it should be good for 70cm too, and so I used it also for the electrical isolation of the slot and central plate. In this way the total height of the vertical radiator was determined: around 20cm.

The input impedance of the antenna is very low in its real component and noticeably inductive in its imaginary component. Transformation is achieved with low loss by simple capacitor matching. Compensation on the inner conductor uses a series capacitor (C1) to the 50-ohm coax cable. Further compensation is made with a parallel capacitor (C2) from the junction of the series capacitor and the centre of the cable to ground (Fig.3). In this way we have matching without problems in the context of s < 1.5 across the whole 70cm band. Both capacitors are high-quality tubular trimmers.

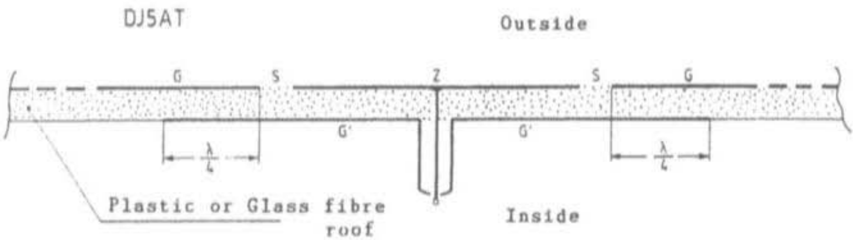


Fig.5: Construction without electrical connection of G, G'

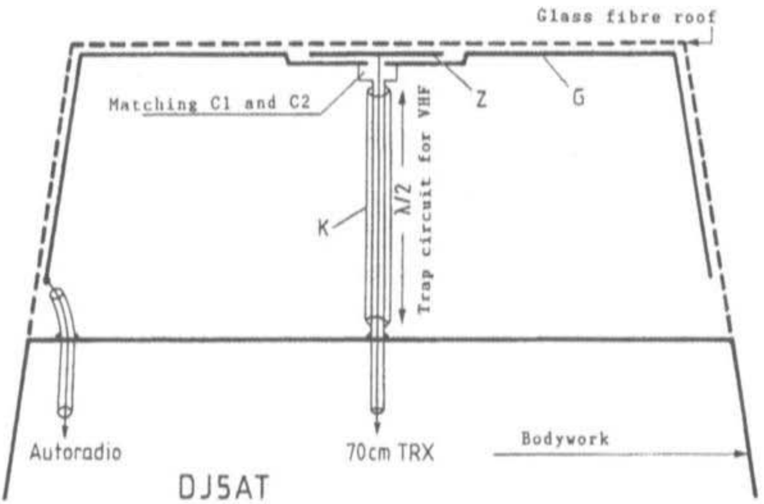


Fig.6: Cross-section across the short axis of the camping bus with circular slot antenna for the 70CM amateur band and car radio antenna (G) for the 3 metre band

3.1 Results

The practical results corresponded with the 5/8 wave antenna at the least. Subjectively I have the impression that radio contact is better, that is more secure. On the other hand, the performance of the 5/8 wave antenna is often influenced by the weather, and that is no longer so noticeable.

3.2 Further development

Now, in consequence of the inductive component of the input resistance, it is also conceivable that the chosen distance (height) of 2cm might be reduced. The space between the central electrode and ground leads the conduction-influenced wave and underlies the law of conduction. I am thinking there of a version as in Fig.5, which uses as electrical insulation only the already available plastic wall of the caravan, yacht, aeroplane or whatever; the shape of the antenna could be painted on this surface in conductive paint. The centre conductor connection must be led through the plastic. The necessary short circuit to the counterpoise is produced by an open quarter-wave overlap. The antenna weight can be neglected, also the additional geometric extensions.

Consequentially I have gone one stage further with my 70cm antenna. The initial problem with the amateur antenna, namely that it was on the outside and could be damaged, applied equally to the car radio antenna. So I have used the counterpoise surface as a car radio antenna. Clearly this means that the amateur coax cable must be trapped for VHF, since the amateur station is grounded to the vehicle chassis.

Simplest solution was to put a VHF trap circuit (K) in the transmit cable, as Fig.6 shows. This consists of a piece of screening braid that is a half wavelength at VHF and is open at each end, thus blocking VHF. The blocking resistance in this way is 120 ohms.

It could be higher but is not noticeable in practical terms because the large antenna surface is very effective, although I did nothing special to match it. A quarter-wave trap line connected to the outer of the cable (i.e. short-circuited) could possibly turn out to be of higher impedance but space problems in the vehicle militate against this modification.

4. ADVICE FOR DIMENSIONS

In reference (2) we find:

Diameter of the centre plate (Z):

$$d_Z = 2 * \lambda/pi$$

In the example selected:

$$d_Z = 45cm$$

Slot width (S):

$$S \approx 0.1 * d_Z/2 = 0.1 * \lambda/pi$$

yet $\ll \lambda$

In the example selected: S = 4cm

In the example selected: antenna height = 2cm

Effective antenna length lw:

$$lw = \pi^2 * d_Z^2/(4\lambda)$$

... for small diameters of the centre plate, with in consequence a vertical diagram free of side-lobes.

For the example with Zw = 50 ohms:
C1 = 1.7pF, C2 = 5.4pF.

5. LITERATURE

(1) Rothammel: Antennenbuch

(2) Bibliographisches Institut Mannheim, Wien, Zuerich Hochschultaschenbuecher-Verlag: Antennen, third part by Adolf Heilmann, 1970 (pp 134-139).

Dr.Ing. Ralph Oppelt, DL2NDO (ex DB2NP)

Universal 2:1 Economy Transformer for DC Part-1

The following description is of a voltage changing device which can be operated in both directions for low-loss doubling or halving of DC voltages. The name "economy transformer" is quite accurate since input and output share one pole in common and the voltage can be altered by cascading or tapping. Since over 95 per cent efficiency can be achieved the title economy is justified. The version described here is designed for a voltage range of 5 to 15 volts/15 to 30 volts (primary/secondary).

1. BASIC CONSIDERATIONS

For changing DC voltages one generally uses so-called upwards or downwards switching regulators (1). Since these use an inductance for voltage changing, there are inevitable losses in copper windings and in the ferrite material. An advantage, nonetheless, of this system is that in principle any voltage relationship desired can be achieved and the output voltage can be stabilised without further losses. That said, the direction of DC current cannot be reversed. Taking an example, a converter up from 5V to 12V cannot be used backwards to turn 12V into 5V.

If one wishes to avoid inductances to achieve greater efficiency, the only possibility at present is to exploit the charging of capacitors and switch round the voltages. This means, however, rather like line transformers in RF technology, only whole-figure transformation ratios can be realised, e.g. 1:N and N:1. By cascading a N:1 converter with a 1:M converter, any entirely rational transformation relationship N:M can be achieved.

Since switches can generally carry current in both directions, operation in the reverse direction (analogous to an AC transformer)

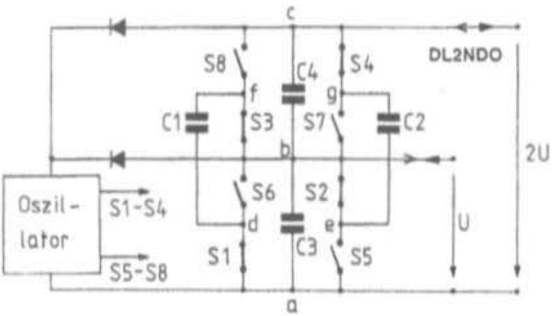


Fig.1:
Principle of DC voltage doubling/halving

must be possible. That means a 1:N converter must also work as an N:1 converter. Of course there is a condition that the control electronics for the switches in both directions must be provided with the corresponding voltages. The advantage of this bidirectional working is lost of course if fixed-direction switching elements (e.g. diodes) are employed. In the low voltage range, however, this is not a problem because of the poor efficiency.

2. PRINCIPLES OF SWITCHING

Fig.1 shows the principles of operation for a DC converter. The two output pulses of the oscillator, offset by half a period, control in turn the switch groups S1 to S4 as well as S5 to S8. By this means capacitors C1 and C2 are connected in push-pull fashion to the nodes a and b or respectively b and c. It is important that the switch-on periods of the two switch groups do not overlap, otherwise the voltages U and 2U would be short-circuited. This must in every case be guaranteed by a so-called anti-coincidence circuit.

If we now run the converter as a doubler (solid arrows) C1 is charged when switches S1 and S3 are closed with input voltage U, while the (already charged) capacitor C2 hands over its charge to C4 when switches S2 and S4 are closed, giving voltage 2U on the output. In the next half period C2 is charged, while C1 gives up its charge to C4. C3 functions there as a buffer during the charging process

Operating the converter as a halver (open arrows), the relationships are similar. Since S1 to S4 or respectively S5 to S8 are always closed at the same time, C1 and C2 are always in series with the input voltage 2U. At each individual capacitor the voltage U drops in this way.

To begin this exact voltage halving appears uncertain, since the capacitors in the series circuit must have a putative exactly equal loss resistance in parallel. In this connection one should think as an example of the series-connected capacitors in high-voltage power supplies, which have external shunt resistors for exact sharing of the voltage. In the present case, however, C1 and C2 are constantly being exchanged one for the other in their series circuit, i.e. C1 is one time "over" C2 and one time C2 is "over" C1. Consequently the voltages on the buffer capacitors C3 and C4 must be balanced out exactly.

In comparison with doubler operation, in this halving circuit the charge and discharge process is exactly reversed: the always "upper" capacitor C4 takes a charge, while the always "lower" capacitor C3 gives its

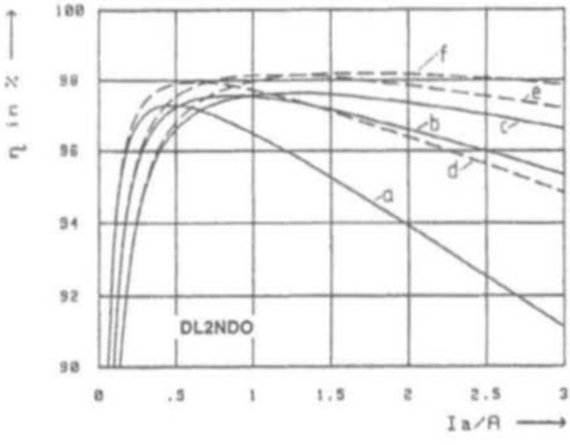


Fig.2: Efficiency of the Voltage Converter in doubler operation; efficiency is dependent on output current I_A. Various input voltages U are indicated. Solid line shows use of BUZ10 (R_{DSON} = 70 milliohms); pecked line shows use of BUZ11 (R_{DSON} = 40 milliohms).
a.d) U = 5V
b.e) U = 10V
c.f) U = 15V

charge to output terminals a and b. The fact with both modes of operation is that because of the push-pull operation, one of the two capacitors C1 and C2 is always charged while the other has just been discharged. This reduces the residual fluctuation on the output significantly.

As Fig.1 further shows, the supply for the oscillator and the control electronics are drawn not only from the rail with voltage U but also the rail with voltage 2U, which is symbolised by the two diodes provided as an OR gate. In this way the converter starts up without problem in either operational mode.

The converter in doubler mode can also be considered as a push-pull quadrupler using Greinacher's scheme (reference 2). Admittedly in (2) only the usual simplex version is described. S1, S2, S5 and S6 function so to speak as a rectifier, which produces between points d and e a rectangular AC waveform with voltage U. C1, C2 and C4 together with S3, S4, S7 and S8 represent a so-called Greinacher push-pull cascade, which adds the resulting DC voltage at C4 to the input voltage U. Consequently the four last-named switches can be replaced by a bridge rectifier with the AC connections made to points f and g. If only doubler operation is anticipated and efficiency aspects play a secondary role, this is very viable. Naturally, as in the classic Greinacher circuit, further cascading can be used to achieve tripling or quadrupling, etc.

3. EFFICIENCY ANALYSIS

To calculate the efficiency we can first set aside the losses in the capacitors and switches not in circuit as well as the fundamental loss component of the control electronics, confining our calculations to the most significant losses in the switches. In the efficiency curves shown later in Fig.2 the quiescent current losses (measured using test gear) are, however, considered.

For the following calculation we are assuming doubler operation. Since the same switching losses occur in halving operation a separate analysis is not necessary. As Fig.1 shows, C1 or respectively C2 is constantly being charged or discharged via the switch-on resistance

R_{DSON} of the two switches. For the magnitude R worked out now the value $R = 2\ R_{DSON}$ should be inserted. If current I_A now flows at the output during a half period of duration T/2, the total charge during this period is

$$Q = I_a T/2 \qquad (1)$$

from the capacitor already undergoing charging (e.g. C1). In the build-up situation C1 has at the start of the charging a voltage always ΔU lower than the input voltage U. This voltage drop is entirely analogous to the voltage diminution during loading of a conventional AC transformer. The integration of the capacitor loading current

$$I_c(t) = \frac{\Delta U}{R} \exp(-t/RC) \qquad (2)$$

during switch-on duration T_E must then produce the charge set out in equation 1:

$$\int_0^{T_e} I_c(t)\,dt = \frac{\Delta U}{R} \cdot \int_0^{T_e} \exp\left(\frac{-t}{RC}\right) dt$$

$$= \Delta U C\left(1 - \exp\left(-\frac{T_e}{RC}\right)\right) \overset{!}{=} I_a T/2 \qquad (3)$$

From this we get ΔU as a function of T, T_E, R, C and I_A:

$$\Delta U = \frac{I_a T}{2C(1 - \exp(-T_e/RC))} \qquad (4)$$

Clearly the voltage shortfall is independent of the input voltage U which excludes a higher total efficiency η for higher U. Further, ΔU becomes minimal if the maximum possible value is selected for T_E, close to T/2. With variations of C a minimum limiting value for ΔU arises:

$$\lim_{C \to \infty} \Delta U = \frac{R I_a T}{2 T_e} \qquad (5)$$

C should therefore always be dimensioned so large that ΔU only turns out immaterially higher than this limiting value. In the present case (see dimensioning details in Fig.2) ΔU lies only 2.7 per cent (with BUZ10) or respectively 4.7 per cent (with BUZ11) above the theoretical minimum. Since the voltage difference ΔU appears both during charging and discharging, it follows for the output power P_A:

$$P_a = (2U - 2\,\Delta U)\,I_a = 2\,(U - \Delta U)\,I_a \qquad (6)$$

For calculating η we need to know the lost power P_V. This is worked out by the energy set free in the switch during the switch-on period T_E divided by the half period T/2. Note here that four of the eight switches are always carrying current during TE, since according to chapter 2 one capacitor is always just being charged and another has just been discharged:

$$P_v = \frac{2 \cdot R}{T/2} \int_0^{T_e} I_c^2(t)\,dt$$

$$= \frac{4\,\Delta U^2}{RT} \int_0^{T_e} \exp\left(-\frac{2t}{RC}\right) dt$$

$$= \frac{2}{T}\,\Delta U^2 C\left(1 - \exp\left(\frac{-2T_e}{RC}\right)\right) \qquad (7)$$

With the help of the last two equations the efficiency η is given directly

$$\eta = \frac{P_a}{P_a + P_v} \qquad (8)$$

as a function of U, I_A, T, T_B, R and C. It is indicated in Fig.2 for an output current I_A up to 3 amps with U as a parameter of the two transistor types BUZ10 and BUZ11. It is clear that with the 97 per cent switch-on period an excellent efficiency can be obtained even with the low-cost switch BUZ10. In contrast to the above calculation, allowance has been made for the fundamental power absorption influenced by the quiescent current (see table in chapter 6). This is why the curves in Fig.2 bend sharply as I_A approaches zero, whereas they would otherwise run together with η = 100 per cent.

The static internal resistance is related directly to equation 4, while the relationship $2\Delta U/I_A$ is formed. It amounts to 0.31 ohms (with BUZ10) or respectively 0.18 ohms (with BUZ11) for the doubler operation discussed here, assuming an ideal voltage source U at the input. Since the DC voltage converter is not a voltage stabilisation system but more of a kind of economy transformer, the internal resistance cannot be so extremely low. Further, the internal resistance of the source is transformed from the primary side to the secondary, as is familiar from AC voltage transformers. Determined by the buffer electrolytics between points a, b and c of Fig.1 (cp. also Fig.3), the dynamic internal resistance in the voice frequency range significantly lower all the same.

4. CIRCUIT DETAILS

The detailed circuit of the DC converter is given in Fig.3. For the control electronics CMOS ICs were chosen, since they distinguish themselves through low current consumption. The power supply for these is drawn via D1 and TR9 always from the side with voltage 2U (that is 10 - 30V). Only in the

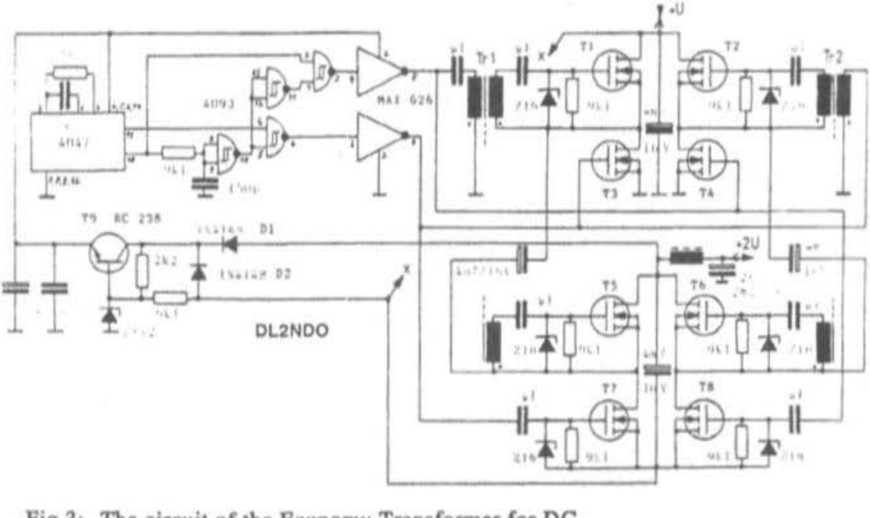


Fig.3: The circuit of the Economy Transformer for DC

case of voltage doubling is the supply voltage drawn first from the side with voltage U via the diode D2. Immediately oscillation has started D2 blocks, however, while TR9 conducts. A zener diode holds the voltage constant to around 8V regardless of whether the converter is being driven at the upper or lower end of its voltage range. Taking the supply for the control electronics from the 2U side ensures that even with the lowest input voltage U = 5V, the above mentioned voltage of around 8V is achieved, with the result that the power MOSFETs are driven with an adequately high gate-to-source voltage. With too low a control voltage, the switch-on resistance R[dson] of these MOSFETs and thus the switching losses would increase significantly.

The oscillator consists of a CMOS 4047 circuit connected as an astable multivibrator. Its output signal on pin 10 is shifted by means of an R-C element and a Schmitt-Trigger-Inverter by about 2 microseconds vis-a-vis the anti-phase output signal on pin 11. The remaining three Schmitt-Trigger elements are connected as NAND and OR gates, so that they produce two anti-phase rectangular pulses overlapping each other by about 2 microseconds. After passing through the inverting driver in the MAX626 module we achieve from this overlapping an interval in the keying, so that the two groups of MOSFET switches are never both switched on at the same time (anti-coincidence circuit). Since no counters but only gates are connected following the oscillator, no undefined switching sates can occur even in low supply voltage conditions. This is particularly important for operation with solar cells.

The driver module MAX626 connected ahead of the power MOSFETs is in a position to transfer a capacity of, for example, 1nF in just 20 nanoseconds. In the present case each driver controls four power MOSFETs, corresponding to a capacitive load of about 6nF. Rapid switching of these MOSFETs is absolutely essential, however, if small dynamic switching losses are to be maintained.

Only two switching transistors (TR3 and TR4) can be controlled directly; for TR7 and TR8 a 100nF capacitor suffices for changing potential. For the remaining four a small ferrite transformer must be provided on account of the floating source potential. A substitute using opto-couplers would certainly be possible but would take more quiescent current.

Connected to every switching transistor not directly controlled is a capacitor-zener diode combination (100nF/Z16), which shifts the square wave control signal into the positive, that means for example plus and minus 5V becomes 0 to 10V. At the same time the MOSFETs are protected against any over-voltage on their gates. This potential shift means that for secure switching of the MOSFETs no differing number of turns are needed on the ferrite transformers. Instead the three windings of the 1:1:1 transformer can be put on in trifilar fashion in one operation.

Recommended for the eight switching transistors TR1 to TR8 are low-voltage types (e.g. max U[ds] =50V), since these present the lowest switch-on resistance R[dson]. Naturally the price of such transistors is inversely proportional to R[dson] but already the type BUZ10 with R[dson] of 70 milliOhms will achieve very good efficiency (see chapter 3). Anyone requiring better can substitute the BUZ11 (R_{DSON} = 40 milliOhms) or the BUZ12 (R_{DSON} = 28 milliohms); additionally parallel circuits of several power MOSFETs are possible.

TO BE CONTINUED

Eugen Berberich, DL8ZX

Magnetically coupled Yagi antennas - overlooked by Amateurs?

The coupling to and from a yagi antenna and the cable is normally achieved using the electrical field - with the customary dipole (or folded dipole) - but this is not the only method. The magnetic field can also be used.

However, the author is sadly unaware of any amateur publication describing this kind of yagi antenna. Indeed it is a shame not to use this simple form (without a folded dipole) for do-it-yourself purposes.

The principle, used successfully by the Hirschmann company at VHF and UHF under the trade name Magneta, produces a very clean directional diagram (Fig.1), an excellent back-to-front ratio and narrow-bandedness. The restricted bandwidth (Fig.2) suppresses any spurious emissions effectively.

The Magneta principle gives high gain for relatively small dimensions. For instance a gain of 9dB is indicated for a 4-element antenna on European TV channel 6 - with a boom length of only 0.9 wavelength.

1. A TRIAL ANTENNA FOR 144MHz

Since no construction plans were available, the author recalculated the dimensions from the domestic television aerial (8-element version for channel E6) to the two metre amateur band. During sweeping the reflection attenuation (Fig.3) it became clear that resonance lay around 150MHz. The cause lay in the fact that the element diameter had also been scaled. Using the original diameter of 8mm, correct resonance was achieved at 145MHz. Figures 5 and 6 show all dimensions of this antenna.

The angled shape of the elements, a typical characteristic of this antenna form, was

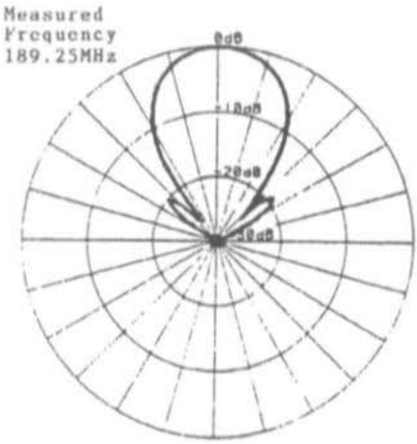


Fig.1: Typical Directional diagram of the Super-Magneta (Fesa 309AN K7)

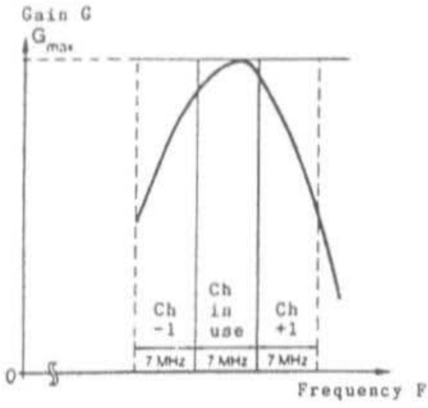


Fig.2: Typical selectivity curve of the Super-Magneta (Fesa 309AN K7)

achieved with a jig tool, in fact after construction using a template made of milled fibre (probably SRBP or Tufnol type -translator).

The rectangular boom (20 x 30mm) is drilled to receive the elements, which can then be fixed in place with stainless steel self-tapping screws.

The feed-loop is made of copper wire about 4mm in diameter. For symmetry a half-wavelength balun is made from thin Teflon coax cable. This is located inside a milled out block of fibre material at the input to the feed-loop. More exact details are probably unnecessary, since they depend on the available materials and tools.

2. GAIN RESULTS

Hirschmann indicate for their 8-element Magneta (1.6 wavelengths long) a gain of 12dB; for the 10-element antenna (2.14 wavelengths long) it is 13dB. With the 8-element antenna recalculated by the author a gain of 11.5dB was achieved. As reference standard for gain measurements a 4-element yagi antenna was used, for which the Wisi firm indicates a gain of 7dB. The horizontal aperture angle for the home-brew antenna turned out to be around 40 degrees.

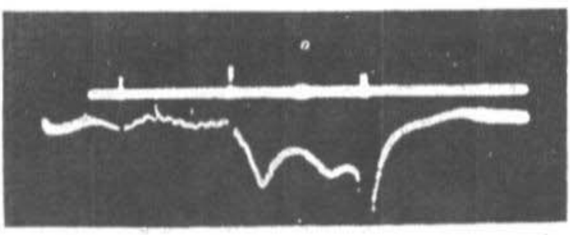


Fig.3: Attenuation of reflections in the first design with thick elements (markers every 10Mhz)

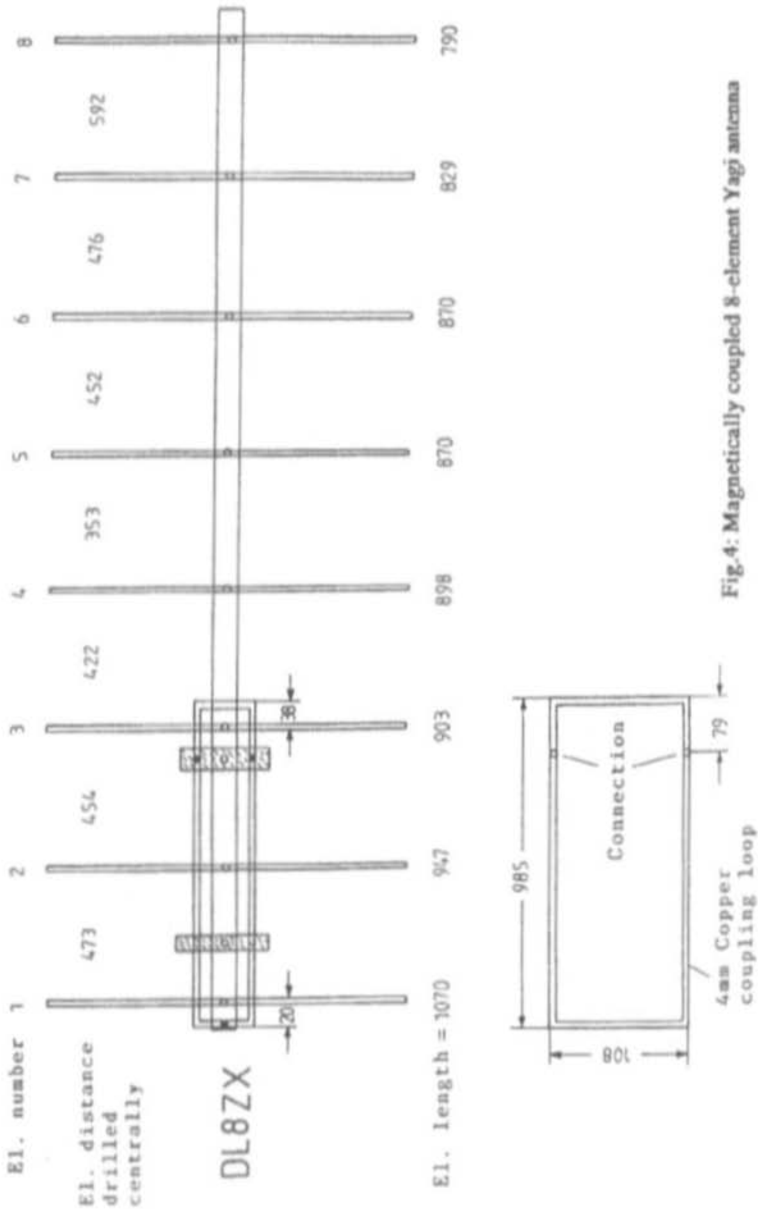


Fig.4: Magnetically coupled 8-element Yagi antenna

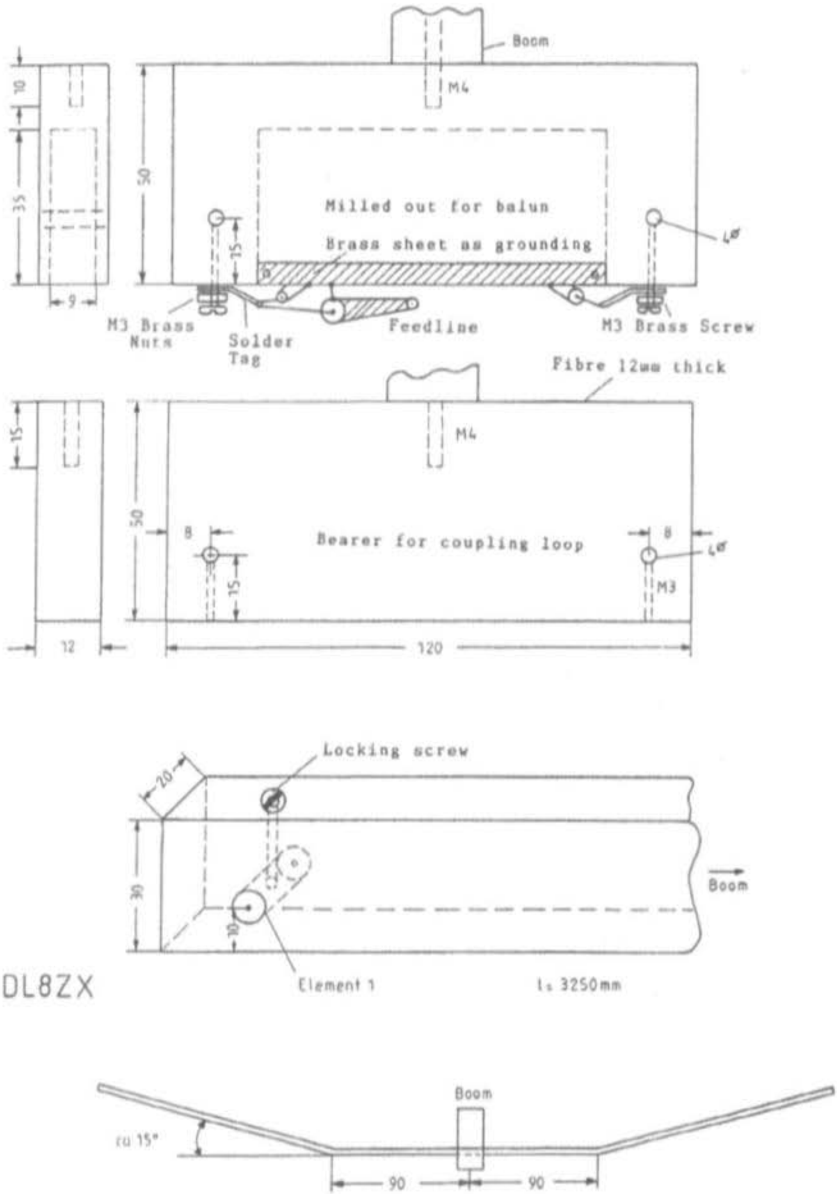


Fig.5: Details of the antenna in Fig.4

The next project will be a 5-element antenna, which stacked should achieve a gain of around 12dB. Tests will also be carried out with 70cm antennas on this principle, using UHF TV antennas as prototype.

3. CLOSING CONSIDERATIONS

For amateurs who are keen on experimentation a DIY antenna has been described which presents valuable properties and has some novelty. The author will be pleased to hear of constructors' experiences. The element diameter of 8mm should - as already mentioned - be retained without redimensioning. Instead of aluminium tube (the ends must be closed off) solid material would offer an advantage, because with the angling upward the sealing off becomes quite critical.

The feed loop is without special measures unearthed; that means it is only earthed via the coaxial cable. For lightning protection the centre of the loop on the reflector can be linked electrically to the boom, as shown in Fig.5.

4. LITERATURE

Antenna catalogue of the Hirschmann company.

MATERIAL PRICE LIST OF EQUIPMENT

described in VHF COMMUNICATIONS

DJ8ES	FM ATV Receiver for the 23cm Band		Art.No.	Ed. 1/1991
Converter				
PCB	DJ8ES 001	double-sided	6347	DM 22.00
Components	DJ8ES 001		6348	DM 179.00
Kit	DJ8ES 001	complete	6349	DM 194.00
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PCB	DJ8ES 002	double-sided	6350	DM 19.80
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Kit	DJ8ES 002	complete	6352	DM 110.00
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PCB	DJ8ES 003	double-sided	6353	DM 15.00
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PCB	DB1NV 010		6477	DM 44.00
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	DB1NV 010, incl. programmed EPROM		6478	DM 276.00
	DB1NV 010, EPROM only		10080	DM 44.00

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PCB	DD2EK 002	6150	DM 28.00
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