



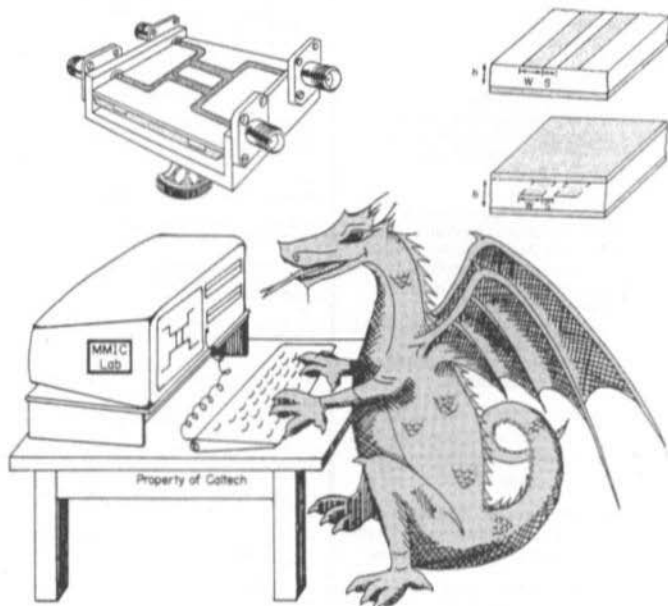
*A Publication for the  
Radio Amateur Worldwide*

*Especially Covering VHF,  
UHF and Microwaves*

# VHF COMMUNICATIONS

Volume No. 23 . Summer . 2/1991 . £3.50

## Microwave Design Magic



courtesy of 'PUFF'



# VHF COMMUNICATIONS

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

Volume No. 23 . Summer . Edition 2/1991

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Robert E. Lentz, DL3WR

## PUFF - A CAD Program for Microwave Stripline Circuits.

Back at the beginning of 1990 the authors Richard Compton (Cornell University), Scott W. Wedge (Hughes Aircraft Company) and David Rutledge (California University of technology) brought out version 1.5 of a remarkable versatile software package. Named after Puff, the magic dragon in the Peter, Paul and Mary song, the software was conceived especially for educational use. It was first used on microwave engineering courses at Caltech and at Los Angeles university; later it spread to many other universities and colleges and was also put to practical use by engineers and scientists. More than 6,000 copies were sent out worldwide of the first version alone.



PUFF represents an interactive circuit design environment for producing stripline and microstripline circuitry. Written in Turbo-PASCAL, it is fast, easy to learn and simple to use. Components are placed on the screen with cursor-key movements, followed by analyses in the time or frequency domain at the touch of a key. The results of the simulation can be seen as a plot, in a Smith diagram or in tabular form. Fig.1 shows the PUFF working screen with a simple example:

Top left is a square circuit board with four connections. At port 1 we have a an angled piece of 50-ohm stripline, next a quarter-wavelength piece with Z=87 ohms, and finally a 4mm-long 150-ohm resistor, connected to ground at the far end.

Below this we have (in this case) letters a to f representing components at our disposal (tline = transmission line; device = a FET on this occasion; clines = couples lines). The design impedance is indicated (50 ohms here) and the design frequency (5GHz this time).

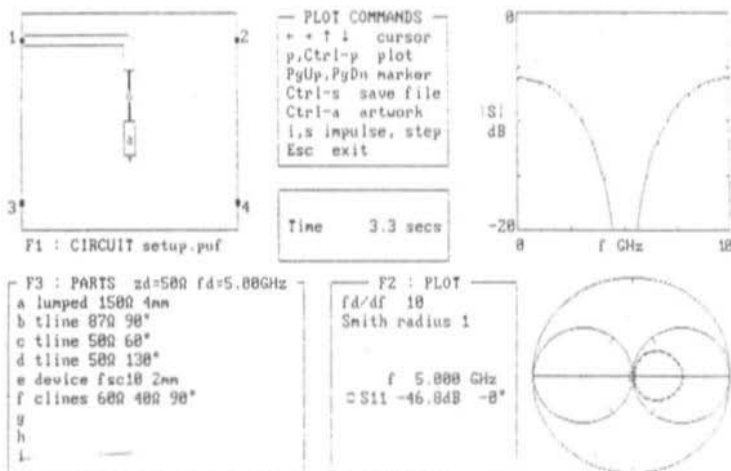


Fig.1 Screen shot of the Program in use

At top centre we have the commands at our immediate disposal. These vary according to whether we are working on a layout (f1), carrying out a simulation (f2) or arranging components (f3).

Top right we see the plotted results of the simulation -frequency response, impulse results or behaviour over time. Below is the Smith diagram for the S parameters selected.

Right in the middle is the time taken for the last calculation, below this the legend for scaling the Smith diagram, the design frequency and the measured value at the design frequency (which is not readable on the plot in the example).

After playing with the examples supplied and mastering the method of operation, you are ready to enter some values of your own. The printed circuit board can be altered according to material (thickness and  $\epsilon_r$ ), dimensions, number and location of ports; the resulting solution can be re-worked to suit components already to hand or a particular frequency required. PUFF calculates all dimensions and avoids crossings of tracks or drawing outside the edge of the board.

Striplines or microstriplines (they should not be confused!) can be characterised by impedance and electrical length.

Concentrated components are SMD-resistors, -capacitors or -inductors. They are entered with their value and mechanical length. PUFF understands four units for impedance or admittance: ohms, Siemens, Z and Y. Reactance values require a 'j' prefix or suffix, with the unit for an impedance following; for example 25johms and 0.5jZ specify a reactance of 25 ohms at the design frequency fd.

Devices are components with several ports, defined by S parameters. In the example we have an FSC10 FET from Fujitsu, with S parameters from 0 to 12GHz.

If you are happy with the results of your simulation, you can print out the enlarged layout, photograph and print it to the right scale and produce the PCB by photo-chemical methods.

At Caltech students have a two-hour practice session where they use PUFF to produce the following: solutions to a design exercise, producing film and PCBs, checking PCB track widths with a microscope, introducing



concentrated components and "devices", checking circuitry on a test station, and finally comparing simulation and actual results, which usually turn out very good.

The exercises are: a coupler, a low-pass filter, a band-pass filter, a low-noise FET amplifier and a FET oscillator. Fig.2 shows the comparison of a low-pass filter calculated by PUFF (solid and pecked lines) and the values measured (dots and crosses).

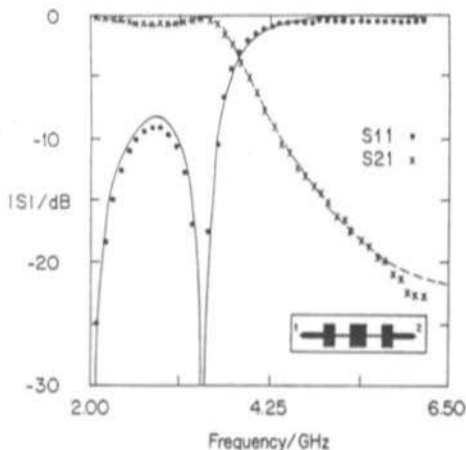


Fig.2 Comparison of calculated and actually measured values

PUFF does not contain an automatic optimisation procedure. The authors have, however, devoted a lot of effort into making it as fast as possible for users to check their circuits manually.

All measurements are given in mm, frequencies in GHz, and characteristics in the form of S parameters. <CTRL>-P is the command for printing a layout, enlarged to a (selectable) scale "p", on an IBM Proprinter. Naturally only the tlines and clines are printed, and bends are automatically chamfered (compensated).

Hardware requirements are IBM PC, XT, AT or compatible with EGA graphics. Slow computers should have a maths co-processor

fitted, while on faster models the emulation built into the software works fast enough.

The finer details of the program are too varied to mention all of them here: with the 5.25" 360k floppy you get a neatly printed handbook with 36 pages and many illustrations. Fig's.1 and 2 in this article came from this handbook. The handbook is remarkably clearly written and easy to understand. Studying it is just as much a pleasure as working with PUFF is. The manual also contains a short tutorial with formulas and diagrams on dimensioning striplines, together with comprehensive references and keyword index. It is clear that extremely experienced college tutors have worked on PUFF and its manual!

Even if PUFF is no substitute for advanced (and expensive) microwave design programs, it is no plaything either and clearly more than "just" educational software. In the USA the software and manual cost just \$10! Ordering information (mention size of disk required) from:

PUFF Distribution, Electrical Engineering M/S 116-81, California Institute of Technology, Pasadena, CA 91125, USA.

I would like to offer sincere thanks to Carl G. Lodstroem, SM6MOM/W6. Carl aroused my appetite for this software, obtained it for me and also supplied a disk full of "devices" - a whole host of transistors by HP and Motorola!

---

*VHF COMMUNICATIONS has placed a trial order of the PUFF software packages, prices yet to be announced. For further information on availability and prices please contact KM Publications direct at the address shown on the inside front cover of this magazine.*



Peter Vogl, DL1RQ

## A 6cm Transverter using Stripline Technology.

### Part-2 (conclusion)

As presented at the 1990 VHF-UHF conference in Munich

6.

#### FINAL STAGE DL1RQ 661

##### 6.1. Description

At the input of the final stage module (Fig.6.1) we find a further stripline filter, which suppresses all unwanted frequencies to below the 60dB level. My own preparatory tests indicate the filter's attenuation in through-pass mode to be approx. 2.5dB.

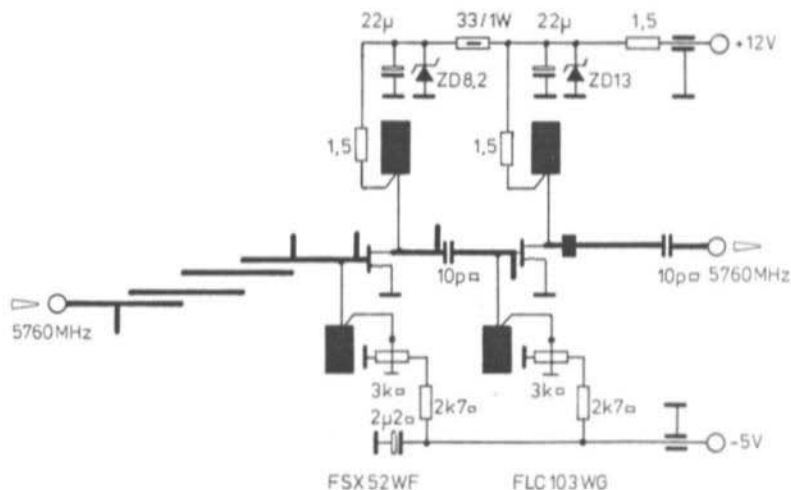


Fig.6.1 Circuit of the final amplifier DL1RQ 661

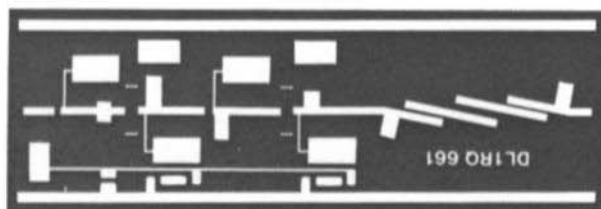


Fig.6.2 PCB layout of final amplifier DL1RQ 661

108 x 34mm; material: RT-Duroid D-5870; 0.439mm

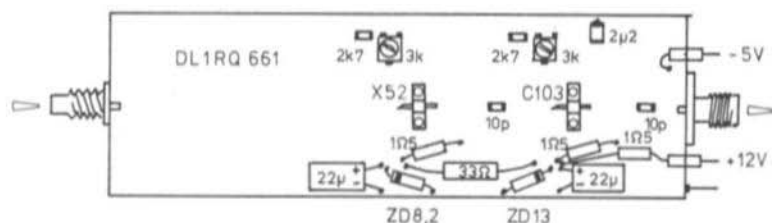


Fig.6.3 Component Layout DL1RQ 661

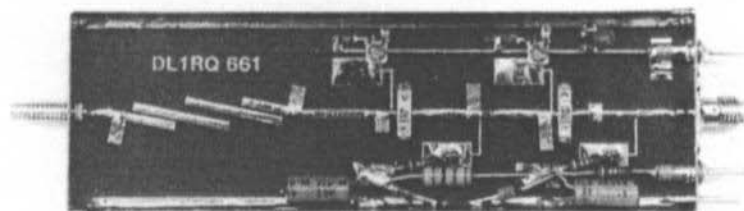


Fig.6.4 Completed module DL1RQ 661 in tinplate enclosure.

Lid measurements 37 x 111 x 13mm



Fig.6.5 End plate and Connector details DL1RQ 661



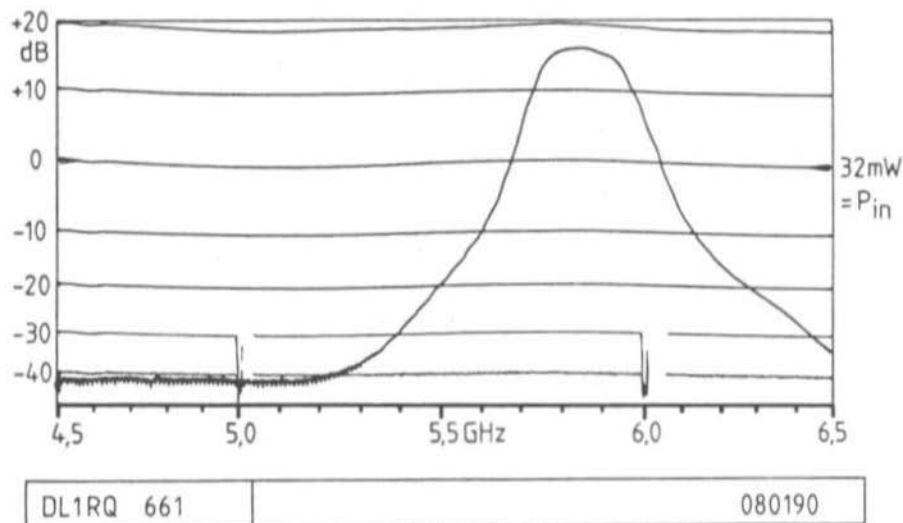


Fig.6.6 Frequency response of the final amplifier DL1RQ 661

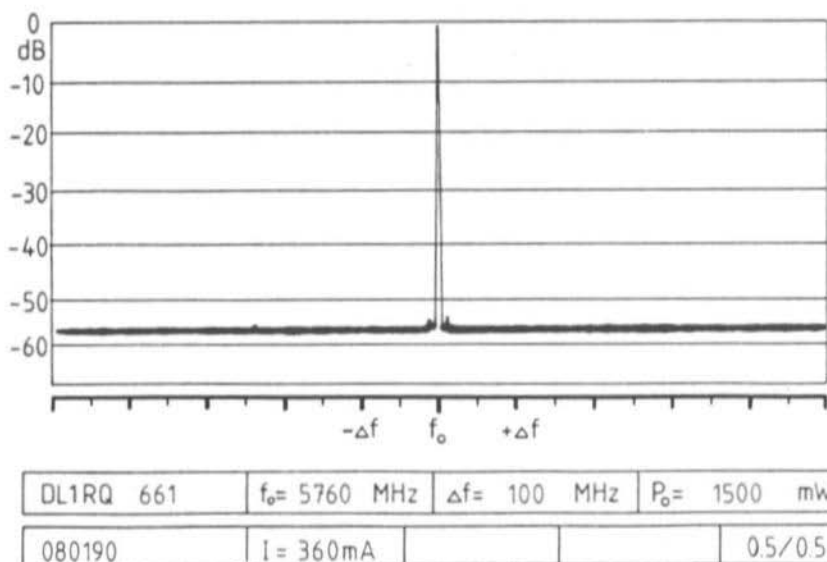


Fig.6.7 Output Spectrum of the final amplifier DL1RQ 661



Following the filter is a two-stage power amplifier. The medium power GaAsFET FSX52WF by Fujitsu is used for the driver stage. The final stage is Fujitsu's power GaAsFET FLC103WG.

The module achieves 17.5dB gain, filter attenuation included. Achievable output power is a maximum of 1.6 Watts.

## 6.2. Constructional information

When mounting the power transistors on the PCB (Fig's.6.2 and 6.3) two points should be noted in particular:

- as ideal as possible RF transfer from the groundplane of the PCB to the source flange of the transistors;
- as ideal as possible conduction path for the heat produced.

The author found the following construction convenient. The PCB is soldered with its groundplane onto a 0.5 or 1mm thick piece of copper. Good tinning in advance and plenty of pressure during the soldering operation are necessary to avoid air bubbles.

After soldering together a 2.5mm milling bit is

used to mill out two oval slots for the transistors in the Teflon material. The milling must cease at the depth when the copper of the heatsink is visible. After the PCB treated in this way has been soldered "water-tight" in the enclosure all components up to the transistors are put in place. The last-mentioned are finally fastened in the slots with appropriate screws. The transistors were soldered in by the author with a low-temperature iron (140 degrees) under closely controlled temperature conditions.

If the copper surface is vertical this should suffice in operation for cooling. If, however, the module is fixed flat onto a chassis, the gap between it and a matching copper plate must be filled with heatsink compound to make up the difference. In no circumstances should the heatsink plate touch against a cover.

## 6.3 Alignment information and test results

The quiescent currents of the transistors should be set as follows:

Driver stage FSX52WF: approx. 70mA, corresponding to 105mV on the 1.5 ohm drain resistor.

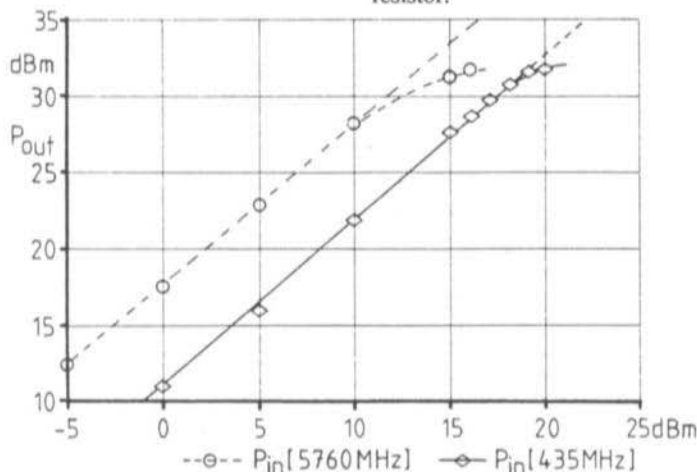


Fig.6.8 Linearity behaviour of the final amplifier module DL1RQ 661 (pecked line) and the complete transverter from 435MHz to 5760MHz (solid line)



Final stage FLC103WG: approx. 240mA, corresponding to 360mV on the 1.5 ohm drain resistor.

Fig's.6.6, 6.7 and 6.8 are measurement charts of test prototypes.

7.

## LITERATURE AND SOURCES OF SUPPLY

(1) Isel side profile 1 no. 190835 (from Isert-Electronics, W-6419 Eiterfeld 1, Germany).

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

(3) 10GHz Transverter using Stripline Technology. Peter Vogl DL1RQ. DUBUS 2/86, pp 115 ff.

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

(5) Redesigned 10GHz SSB Transverter by DL1RQ. Peter Vogl DL1RQ. DUBUS 4/86, pp 310ff.

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Bernd Neubig, DK1AG

# A Modern, Professional look at the Design of Stable, Crystal Oscillators Working at High Frequencies

## Part-2 (conclusion)

### 2.

#### CRYSTAL OSCILLATOR CIRCUITS FOR VHF AND UHF (CONT.)

##### 2.5 Huth-Kuehn/LC-Colpitts Special circuit

A circuit not to be recommended, which despite its problems shows no signs of dying out, is shown in figure 9. It was criticised already in 1979 on account of its unreliable operation. At this stage I can only recommend reading the article in VHF Communications (2) again.

##### 2.6 VHF/UHF crystal oscillators with MMICs

Monolithic microwave ICs (MMICs) are the simple equivalent of universal OpAmps for higher frequencies with just four connections (input, output, volts and ground): they are also very attractively priced now (Mini Circuits,

Avantek, Q-Bit et al.). For low phase noise the amplifiers should not be driven to self-limiting, since their  $1/f$  noise starts to increase a lot close to the (1dB) compression point (8), whereby the carrier-near noise (30dB per decade) of the oscillator becomes very bad. An example circuit for around 80MHz is indicated in Fig.10.

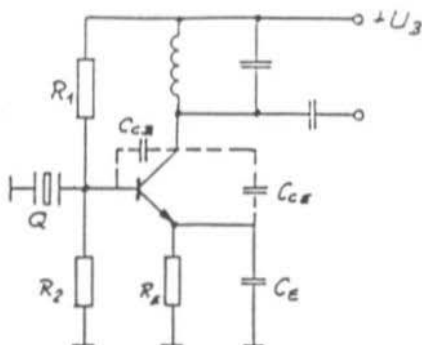


Fig.9 Overtone oscillator circuit (not recommended)

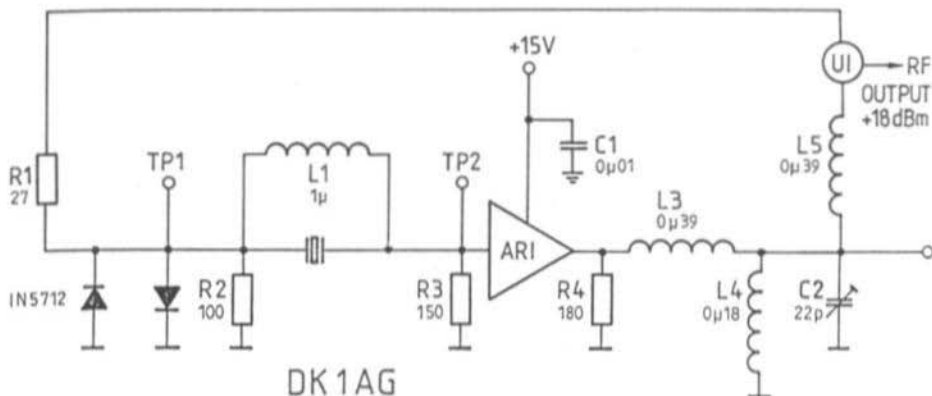


Fig.10 MMIC oscillator circuit according to (8)

The crystal operates with C0 compensation in series resonance as a longitudinal element in a resistor-PI element made up of R2 and R3. The (very high) output power is coupled through a zero-degree power splitter (UI). At the output of the amplifier (in this case Q-Bit QBH-150 or QBH-374) lies the resonant circuit L4, C2 for selecting the overtone. With L3 and L5 the not ideal phase shift of the MMIC can be corrected by a total of 360 degrees. Resistors R1 and R4 serve for correct termination of the power splitter and MMIC. Amplitude limitation is definably pre-set with the anti-parallel diodes.

The originator of this circuit reports on the excellent values of phase noise achieved (Fig.11), a satisfying finale and high point for his experiments!

### 3.

## SOME IMPORTANT VIEWPOINTS ON CONSTRUCTING STABLE AND LOW-NOISE CRYSTAL OSCILLATORS

### 3.1 Handling sub-resonance

Since the crystal's sub-resonances lie close to the main resonant frequency, false operation of the oscillator cannot be prevented by additional coil circuits. If you are contending with only moderately suppressed subsidiary oscillation - particularly with frequencies above 100MHz - then take note that:

- the amplification in reserve is as little as possible. This means that feedback is strong enough only to ensure satisfactory oscillation.
- the optimal point of operation of the resonant circuit is when the oscillator works at the nominal frequency with the crystal by-passed. The circuit should be detuned from this position only very slightly. If this is not

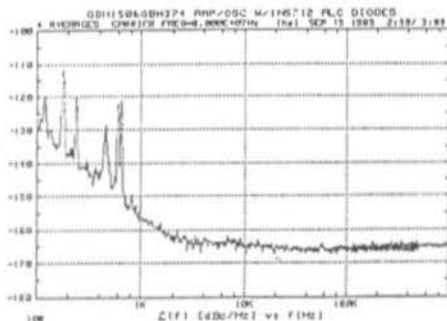


Fig.11 Phase noise of the circuit in Fig.10





probe out of a circular core and coupling coil, but then you would have the problem of calibration.

Measurement of the crystal loading can also be accomplished with an RF millivoltmeter or with an oscilloscope of adequate upper frequency limits. Dependent on the particular oscillator circuit there are a number of methods available (3):

#### a) Colpitts and Clapp-Guriel oscillator

The only precondition is that the resonant resistor R1 of the crystal is known. Using a probe of the lowest possible capacity, measure the RF voltage UBMeff between base and ground. Using an oscilloscope, the effective value is derived from the peak-to-peak value:

$$\begin{aligned} \text{UBMeff} &= \text{UBMpp}/(2 * \sqrt{2}) \\ &= \text{UBMpp}/2.82 \end{aligned} \quad [16]$$

From this the voltage UW is calculated thus:

UW =

$$\text{UBM} * \cos(\arctan((\text{Ca} + \text{Cb})/(\omega_0 * \text{R1} * \text{Ca} * \text{Cb}))) \quad [17]$$

from which the crystal loading PQ is derived:

$$\text{PQ} = \text{UW}^2/\text{R1} \quad [18]$$

Correspondingly the crystal current is:

$$\text{IQ} = \text{UW}/\text{R1} \quad [19]$$

In the Clapp-Guriel (overtone) circuit Cb should be replaced by Cbeff according to equation [8].

#### b) Clapp (grounded base) oscillator and LC Colpitts circuit

In this situation the two connections to the crystal are "hot" from an RF viewpoint, so the measurement of the crystal loading must follow a different course. The precise resonant frequency fs (fr), as well as the resonant resistance R1 (or Rr) must be known. The LC circuit is adjusted so that the circuit is oscillating exactly on the resonant frequency. Then the crystal should be replaced with an ohmic resistance of exactly the same value as

R1; if the C0 compensation coil Lp is provided, it must be removed. The oscillator must then oscillate in the region of the crystal frequency (plus or minus a few per cent). With a low-capacity probe the voltage UR is now measured across the substitute resistor Rers. Even better is to measure the voltage to ground of the two ends of the resistor and form the difference (best-wise using the built-in difference function of the oscilloscope). During the measurement the effect of the probe should be evened out by gentle adjustment.

The crystal loading is

$$\text{PQ} = \text{UR}^2/\text{Rers} \quad [20]$$

and the crystal current is once more

$$\text{IQ} = \text{UR}/\text{Rers} \quad [21]$$

### 3.4 Low-noise crystal oscillators

This theme has already been treated by me extensively in VHF COMMUNICATIONS (see (2) and (6)). Supplementing these still current designs I should like to add the following. The more the oscillator frequency is multiplied, or the more the final frequency is divided back in a PLL to a reference crystal, so the higher are the demands imposed on the phase noise close to the carrier of the crystal oscillator signal. Phase noise close to the carrier is determined in the first instance by:

- the operating quality of the crystal in the circuit. Reference (6) has already gone into this thoroughly.
- the 1/f phase noise of the active stage (transistor, MMIC) which without feedback increases by 10dB per decade and with feedback makes itself felt in the oscillator phase noise with 30dB per decade.
- the influence of noise carried on the power supply. Here you are cautioned about the modern three-leg CMOS voltage regulators, which contribute significantly to phase noise.

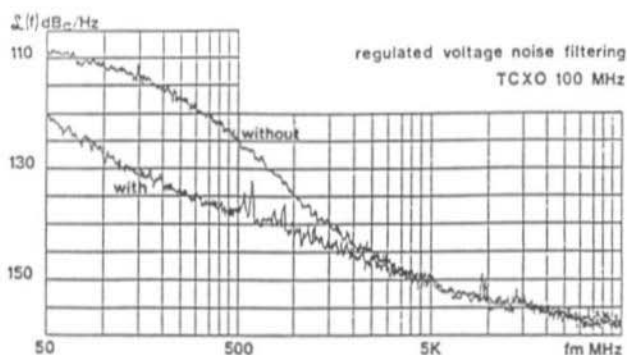


Fig.14 Influence of the voltage regulator noise on the phase noise of an oscillator

● the dependence of the oscillator frequency on the operating voltage, which can modulate external noise from the power supply and other sources. The voltage-sensitivity can be reduced by the use of a more robust crystal, for instance with a higher overtone. Capacitive diodes for drawing off or modulating should be dimensioned for minimal draw sensitivity.

Next to the careful selection of suitable components and a low-noise regulator (e.g. type REF-01 from PMI or equivalent types from Linear Technology), good use should be made of devices for blocking low frequencies before and behind the regulator together with additional filtering elements. The influence of filtering is indicated in Fig.14 (taken from (9)).

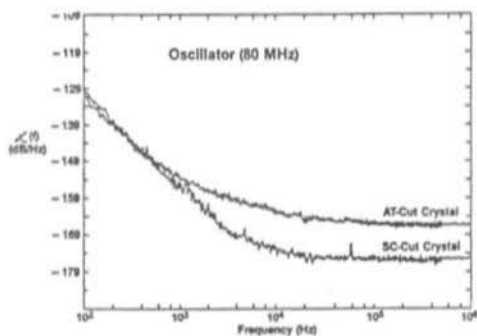


Fig.15 Difference in phase noise for crystals with AT and SC cuts

Crystal loading, which should be fairly high (several hundred microwatts, that is 2 to 5mA current for AT-cut), generally makes itself felt on the noise floor 10kHz and more from the carrier. Here the much praised but very expensive SC-cut offers definite advantages (Fig.15), which for most applications are offset by the much higher price and additional circuitry required.

### 3.5 Temperature stabilisation (crystal oven, OCXO)

For the highest stability it is necessary to provide a thermostat for the crystal or better, for all frequency-determining components. If possible the crystal should be operated exactly at its inversion point: for the AT-cut this is the minimum of the frequency-temperature curve, which should lie at least 5K and better 10K above the highest expected environmental temperature. A still higher inversion point temperature will raise current consumption unnecessarily through heating. Standard tolerance of the inversion point in crystal manufacture is plus or minus 5K.

A suitable circuit for thermal regulation is proposed by C. H. Schuebbe DJ6JJ in (7). To achieve a stable oven temperature which is in practice better than 0.1K it is also necessary to keep the temperature of





the OpAmp in the heat regulation circuit at as constant a temperature as possible. Otherwise the temperature drift of the offset voltage falsifies the regulation. Additionally you should place as close as possible together the temperature sensor, the heat source (power transistor) and the crystal. Ideally the crystal should not lie on the heat path between the heat source and sink. A particular danger occurs with small heatsinks when the thermal regulator can oscillate, noticed by "pumping" of the heating current.

When placing the crystal in circuit bear in mind that the connection wires of the crystal form a direct thermal path to the quartz disc. The electrical connection to the "cold" world outside should not be too short (opposite of RF requirements!) and should be have poor thermal conductivity (use thin wire).

With crystal ovens you should also observe

that when switching on after a power disruption that the oscillator frequency may not return to the previous value even after some time (the so-called retrace effect).

#### 4.

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8. Driscoll, M.M.: Low Noise Crystal Oscillators using 50-ohm modular amplifier sustaining stages. Proceedings 40th Annual Frequency Control Symposium, Philadelphia (1986), pp 329-335.

9. Girardet, E.: Limiting Phase Noise in TCXOs. Conference documentation Oscillating Crystals Symposium 89, Frankfurt (14.6.1989), pp 215-232. Publisher: ZVEI Fachverband Bauelemente der Elektronik.

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*Dr. Ing. Jochen Jirmann, DB1NV*

## Enhancements to the Spectrum Analyser

Three years have passed between first publication of the spectrum analyser design DB1NV 006-009 and now, and a few problem areas have shown up. At the same time further developments on the modules side has made possible some new variants on the circuit.

In particular the second overlay oscillator showed up as being relatively un-reproducible: the nominal frequency could often be obtained only after several changes of components since the tolerances of small capacitors are often quite significant. So we now present an interesting alternative using a ceramic resonator.

In addition the initial alignment of the first IF at 460MHz often seemed to cause problems on account of scattering in the helical filters, making it difficult to line up the centre frequency with the pass curve. A solution is to employ surface acoustic wave (SAW) filters from car telephones. The unevenness in the pass region is certainly not comparable with an optimally aligned helical filter but it is reproducible almost without need for alignment.

Also, the often-uttered wish for a digital framestore has not gone unheard, but because of the broad range of the description it is given in a separate article. The screen shots presented here were produced with this frame store and a graphics plotter using coloured inks.

---

### 1.

#### SECOND OSCILLATOR USING CERAMIC RESONATORS

---

When coaxial ceramic resonators became of interest to radio amateurs around 1988 there was initially only one type with a relative dielectric figure of 38 available, which set the lower frequency limit at around 900MHz (with sensible dimensions). In the meantime ceramics with a dielectric figure of 80 have become available, allowing the manufacture of 450MHz resonators. The primary application area is oscillators and antenna couplers in

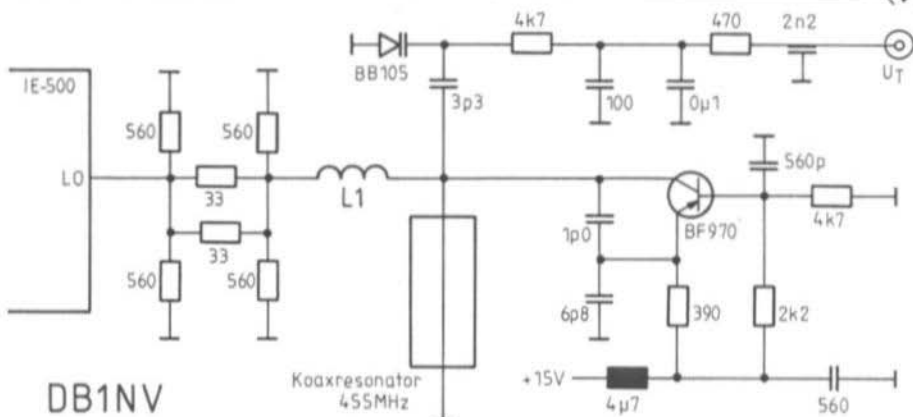


Fig.1 The second oscillator with Ceramic Resonator

L1 = 4.5 turns, 0.63mm varnished copper, 3mm inner diameter

mobile telephones (Netz C and NMT 450), yet their frequency characteristics also allow them to be used at 70cm and for the second oscillator of the frequency analyser.

Since the principles of constructing an oscillator with coaxial ceramic resonators has been described in issue 2/1989 (German edition), Fig.1 should suffice for illustrating the functions. As one sees, the collector circuit of the

oscillator transistor BF 970 is substituted by a resonator (Siemens B69500-T5723-F455, nominal frequency 455MHz). The feedback path leads via a capacitive divider 1pF/6.8pF to the emitter, so it's a case once more of a Colpitts oscillator.

Coupling from the mixer is done inductively (in order to suppress harmonics) via L1 (5 turns 0.7mm varnished copper, 3mm inside diameter). An attenuating element improves decoupling between mixer and oscillator.

For fine tuning a tuner diode BB105 is connected through 3.3pF. First tests indicated a centre frequency of a bit more than 450MHz and a tuning range of more than 5MHz. The oscillator power at the mixer input is ideal at 5mW. Unfortunately the characteristic curve is decidedly crooked, as Fig.2 shows: the tuning gradient alters over the tuning range by a factor of 3. Linearisation of the alignment characteristic curve is therefore necessary.

The clearest solution, providing the second oscillator with a frequency control loop, would demand too much effort. So two other techniques were tried out and found to produce the desired results without too much effort. One can either pre-distort the tuning voltage with a diode-resistor network or else

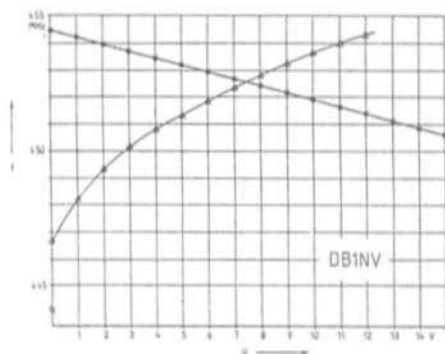


Fig.2 Alignment characteristic curves of the oscillator

Triangle = characteristic curve at oscillator input

Circle = characteristic curve at equaliser input (test pt. 21) of DB1NV 009

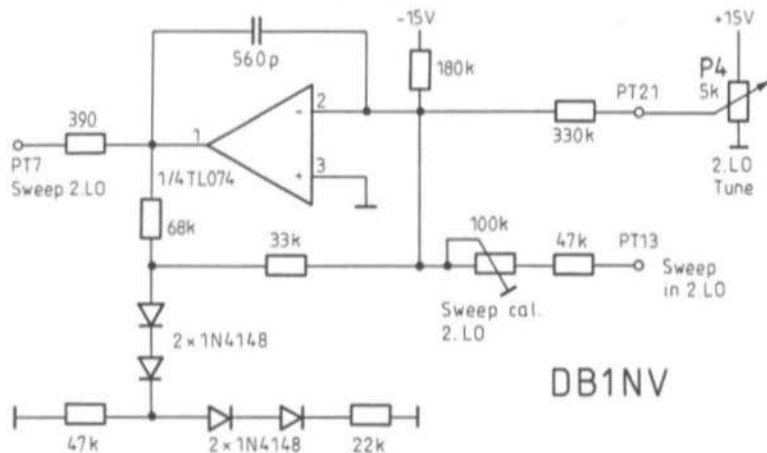


Fig.3 Equalising Network in module DB1NV 009

make the coupling of the tuning diode so frequency-dependent that it produces a linear frequency track.

The first variant calls for flow control (DB1NV 009). Looking at Fig.3, the circuitry of the operational amplifier I2 is changed at pins 1, 2 and 3 so that it produces amplitude-dependent negative feedback. With rising output voltage from the opamp the rising circuit reduces the negative feedback in two stages and thus increases the amplification for higher tuning voltages. The small number of components enables this circuit to be added easily to the PCB; it holds the gradient of characteristic curve constant within plus or minus 10 per cent, which is adequate for the present application.

A more elaborate lineariser can certainly reduce the error but the additional effort is not worth it. The result of linearising is seen as the second curve in Fig.2, which plots frequency against the tuning voltage at test point 21 of DB1NV 009 (input fine tuning).

The second method of linearising, frequency-dependent coupling for the capacitor-diode, is certainly more tricky but uses the minimum of components. One possibility is to add inductance to the capacitor-diode and make it a

resonant circuit, coupled to the oscillator.

Normally this kind of circuit is wishful thinking for oscillator constructors as it ends to develop feedback or to stop oscillating, for example because of the self-inductance of RF chokes. All the same, with careful construction it is possible to achieve an almost linear tuning curve over restricted frequency ranges. The attraction of this technique harks back to the valve days when strictly-linear tuning VFOs were called for (one need only think of the Collins receivers with mechanical digital displays). There was also often an additional demand for electronic tuning to a few kHz. e.g. for sideband tweaking or for achieving RIT independent of the main tuning.

Anyone interested can refer to (1). Reading almost forgotten circuit principles may also be worthwhile.

With a capacitor-diode the barrier layer capacity drops with the root of the tuning voltage, in other words the frequency gain becomes ever flatter with rising tuning voltage. For an oscillator with a linear characteristic curve we need diode coupling that becomes more pronounced with falling capacity and rising frequency: Fig.4 shows a possibility coupling a diode (shown symboli-

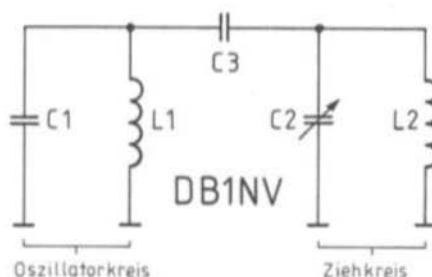


Fig. 4 Substitute circuit for alignment with a pull-in circuit

cally as variable capacitor C2) via a high-pass circuit L1/C1 (the ceramic resonator).

The functional method of this network can be illustrated with Smith diagrams but most readers will find a more descriptive explanation serves them better. With diode capacity C2 large and oscillator frequency low, the impedance of the parallel circuit C2/L2 is capacitive and small in comparison with the reactive impedance of C3. The voltage on the diode is small and the influence of the diode on the resonant circuit slight.

If C2 is now reduced, L2 begins to compensate a growing part of C2, which means the effective capacity of the parallel circuit L2/C2 drops more than proportionately. Then the voltage division is reduced with C3 and the

diode is ever more closely coupled to the resonant circuit L1/C1. In the extreme case the effective capacity in the parallel resonance is zero and the diode is completely in the resonant circuit.

In practice this situation normally leads to wild characteristic curves and is therefore to be avoided; the author has in his experiments set the resonant frequency of L2/C2 well below that of the main resonant circuit L1/C1.

The circuit of the oscillator is shown in Fig. 5; as one sees, the ceramic resonator forms the resonant circuit L1/C1 and L2 serves as feed for the diode bias voltage.

The result of the experiments is to be seen in Fig. 6: with correct alignment of L2 (second curve) a ruler-straight tuning curve is achieved over about 5MHz, which is fully adequate for the IF section.

In the valve era even more clever circuits were evolved: if this is your pleasure the circuit on page 34 of (1) would be worth looking at.

Using an inductively tuned VFO (Clapp circuit) the auxiliary circuit is led to the cathode line and serves simultaneously as feedback route. Current flow controls the detuning of the auxiliary circuit with a germanium diode and thus achieves constant

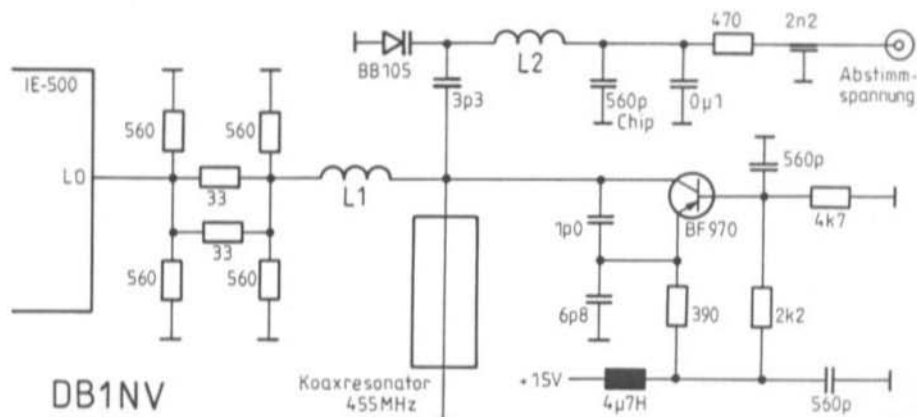
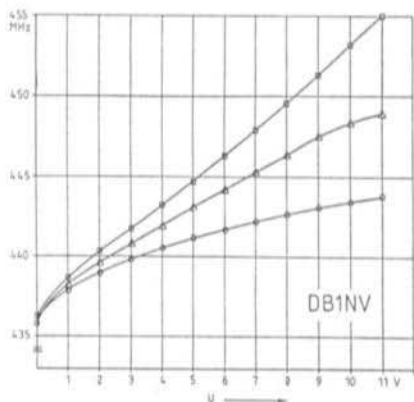


Fig. 5 The linearised oscillator



**Fig.6 Alignment characteristic curves for various values of L2**  
**Circle = L2 to infinity**  
**Triangle = L2 4.5 turns, 3mm diameter**  
**Square = L2 3.4 turns, diameter**

frequency movement over the tuning range of the VFO.

The primary purpose of altering the oscillator was to improve reproducibility. During tests it was found that both new variants improved the phase noise characteristics of the analyser.

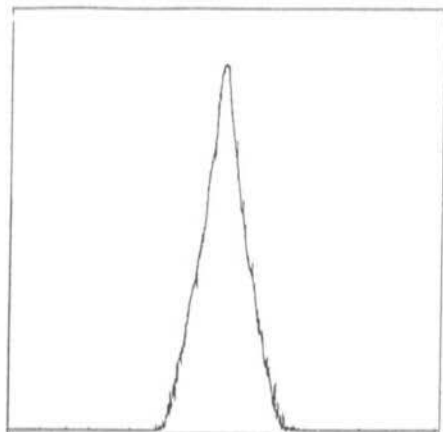
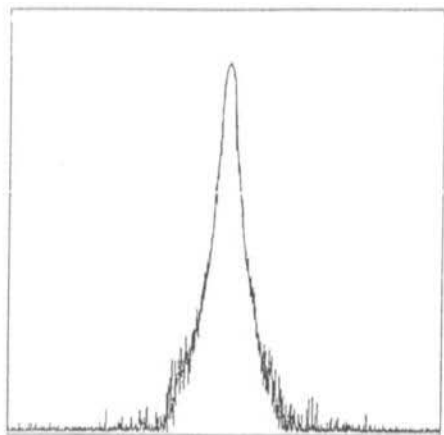
The improvement is not exactly large because the noise content is mainly determined by the (poor) first oscillator but it can become relevant if development of a low-noise and low-cost synthesiser for the first oscillator bears fruit.

The result is to be seen in Fig.7 which shows the 60MHz calibration oscillator with 10kHz/div and 70dB dynamic range. Here the spectrum analyser was driven with an external low-noise oscillator (HP 8640).

With the old oscillator circuit noise sidebands were still visible, but with the new circuit (variant 1) they are gone. The horizontal scale is incidentally not exactly equal, which should be considered during evaluation.

The illustrations were turned into hardcopy with an HP Thinkjet from the digital frame-store mentioned.

A method of arranging the second oscillator using the ceramic resonator on PCB DB1NV 006 is shown in Fig.8.



**Fig.7a (left) Noise sidebands of the original second oscillator**

**Fig.7b(right) Reduced noise sidebands with Ceramic Resonator**

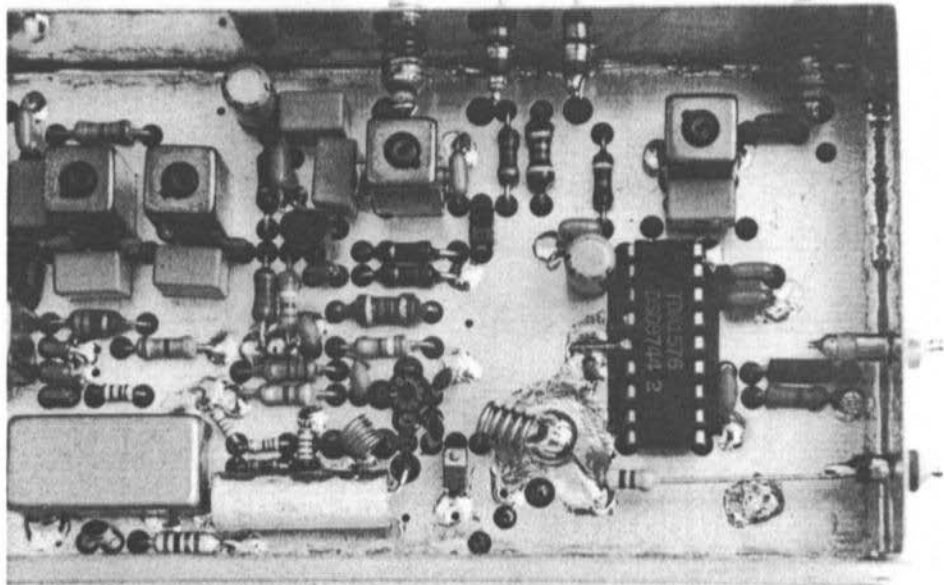


Fig.8 The new second oscillator on PCB DB11NV 006

2.

## FIRST IF WITH SAW FILTERS

In the first article on the subject of spectrum analysers already indicated that SAW filters would in principle have application as a means of improving selectivity in the first IF. At that time the only filters readily available were for satellite TV tuners. Their bandwidth of 27MHz at 479.5MHz and through-pass attenuation of around 20dB for selectivity of 40dB held no attraction for use in the analyser.

Meanwhile, however, new types of filter have appeared on the market which suit our requirements almost perfectly. Centre frequency lies at 455MHz with 4.5MHz bandwidth, the attenuation is less than 4dB and the wide-range selectivity is around 40dB. Input and output are fixed at 50 ohms. Their primary application is transmit filters in mobile tele-

phones, for instance in Netz C and NMT 450. Whereas the transmitter and receiver in car telephones were previously separate, each with their own synthesiser, space and current considerations have now brought about the return of the transceiver concept (as on Netz A phones such as the Tekade B95, still widely used in amateur circles. Between the transmit mixer and the final amplifier SAW filters are used for adequate suppression of harmonics.

Since adjusting helical filters in the IF has required a "magic screwdriver" and a lot of patience (centre frequency and filter curve both had to be spot on), the use of virtually fixed-tuned filters has seemed most desirable, especially for the less skilled constructor.

As basis of the experiments the Siemens filter type B4510 was selected, as it is obtainable and compares favourably with the price of helical filters used up to now. A glance at the data sheet (2) reveals the negative side: in the

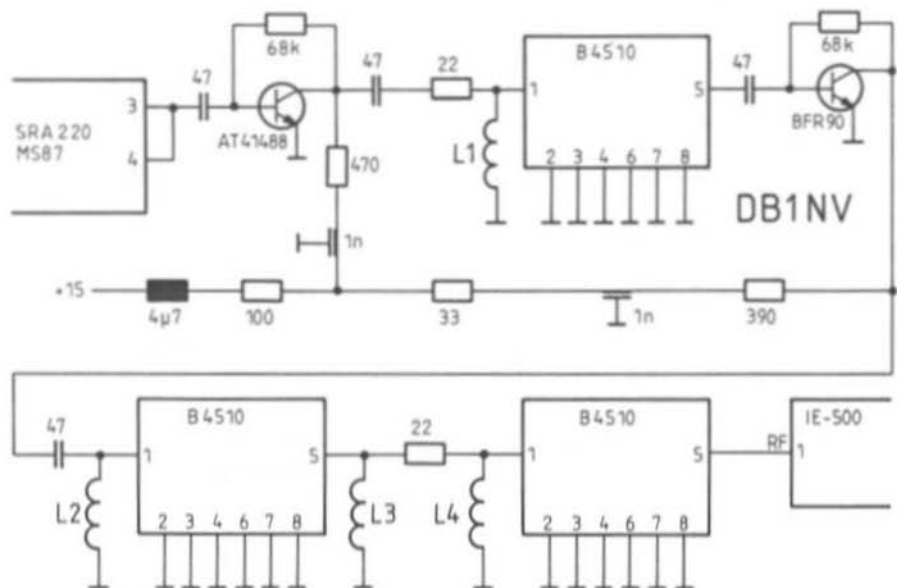


Fig.9 The partial circuit of the module DB1NV 006 with three cascaded SAW filters

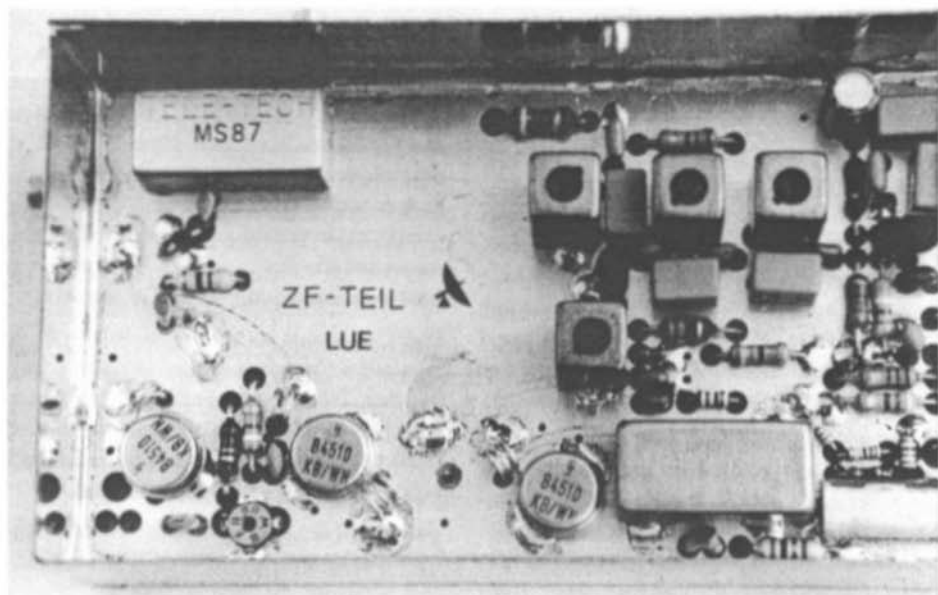


Fig.10 The UHF section of the module DB1NV 006 with the three SAW filters and the matching coils



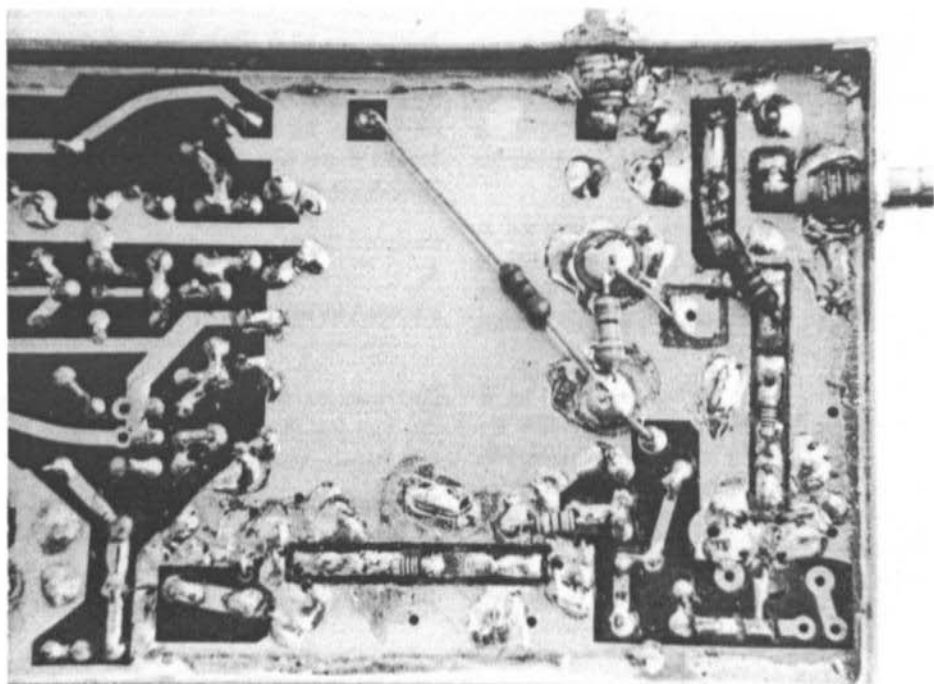


Fig.11 Conductor track side of DB11NV 006 with the hand-made 50ohm striplines

normal case the filter has a ripple of more than 5dB, which can be reduced to less than 2dB using auxiliary matching coils on the input and output. Unfortunately this matching reduces the wide and narrow range selectivity to around 10dB, the largest resonance peaks reach -30dB. The deeply serrated response of the filter curve is incidentally typical of SAW filters, as can be found out from (3).

For adequate selectivity of 80dB at the image frequency of 433.6MHz (second IF of 10.7MHz) three cascaded filters were introduced to be safe and sure, although a single filter demonstrates a selectivity of 45dB. It would be interesting to find out if secondary resonances vary much due to manufacturing tolerances. If there is no danger of the strong subsidiary resonance at 440MHz straying into the image area, then two filters would be adequate.

The prototype was built on PCB DB1NV 006 and had also the IF preamplifier from (German) issue 4/1989. As the partial circuit in Fig.9 shows, the IF preamp has been made with Avantek transistors AT41485 this time. Avantek recommends this transistor as an improved replacement for the BFR69; the smaller, lower inductance Micro-X outline with 2.5mm diameter and higher transit frequency of at least 7GHz allow use up to above 23cm. The price of around DM4 makes this type interesting for other UHF applications as well.

Following the preamplifier the first SAW filter is inductively matched with L1. The second amplifier uses a BFR90 as previously since an AT41485 would raise the amplification to an "unhealthy" 30dB. Two further SAW filters lie between the second amplifier and the second mixer (IE-500). Here the coils



L2, L3 and L4 serve as matching elements. All matching coils are made with 0.5mm wire on a 2.5mm former (shank of twist drill). L1 and L2 have 1.5 turns, L3 and L4 2.5 turns. These coils are soldered as close as possible above the PCB surface and placed to minimise feedback.

The completed UHF section of DB1NV 006 is visible in Fig.10; with a little skill the rebuild can be fitted on a standard PCB and the only extra work is to mill some 3mm wide 50-ohm lines (Fig.11).

The prototype construction achieved an IF bandwidth of 4MHz at 455MHz centre frequency, the ripple in the pass region was 3dB, albeit with the chaotic attenuation curve typical of SAW filters. Stray reception in the dynamic range of the analyser could not be detected.

All in all, it can be stated that the implementation of the SAW filters described produces a practically zero-alignment first IF amplifier, which ought to make construction easier particularly for less skilled builders. This is bought at the cost of less accurate measurement due to the greater ripple of the filters.

---

3.

### LITERATURE

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- (1) Friedhelm Hillebrand, Einseitenbandtechnik fuer den Funkamateuer. RPB Band 117/118, Franziz-Verlag 1968.
- (2) Vorlaeufiges Datenblatt Oberflaechenwellenfilter B4510, Siemens AG, 21.03.90.
- (3) Siemens-Datenbuch Oberflaechenwellenfilter fuer Fernseh-Anwendungen 1987/88.

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## Improved Air Cooling For 2C39 Power Amplifiers

Many people constructing power amplifiers for the amateur bands starting 144MHz and up still go for the 2C39 tube and its equivalents. While air cooling of the 2C39 types presents no problems on 2 metres and 70cm, the operating conditions become more critical at 23cm and the problem of ensuring adequate ventilation more difficult. Here, then, is a suggestion for a new method of cooling.

---

### 1.

#### THE PROBLEM

---

The valve data indicates an anode dissipation of 100 watts with maximum ventilation. This limiting value is exceeded by most users in the amateur field, sometimes considerably. Various authors have sought salvation: reference (1) describes oil cooling and (2) uses distilled water. In (3) DJ3EP demonstrates an improved system using aluminium heatsinks.

The disadvantage of the schemes mentioned, apart from the technical demands of (1) and (2), lies (in the author's opinion) in the difficulty of changing tubes. Using the cooling system of (3), the additional heatsink must be linked mechanically rigid to the PA unit, otherwise it can occur during installation after transportation that the tubes break in two.

---

### 2.

#### MY SOLUTION

---

The modification proposed here for the cooling surface of the 2C39 is (in the author's opinion) a good compromise between effort involved and the noise level of improved cooling.

Consideration of 2C39 and YD1270 tubes shows that they differ only in their heat sinks (Fig.1). The YD1270 offers an anode dissipation of 200 watts according to the data and its dissipating surface is correspondingly larger.



Fig.1 For comparison: left a 2C39BA, right a YD1270

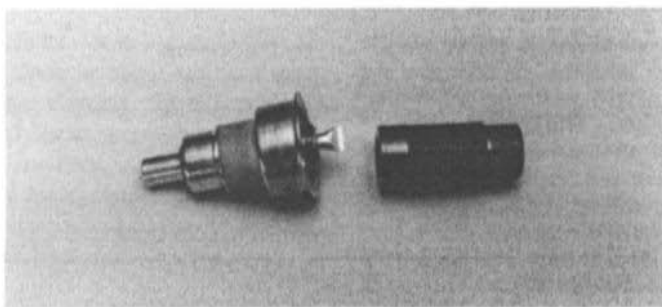


Fig.2 A 2C39BA with the coupler (author's photo)

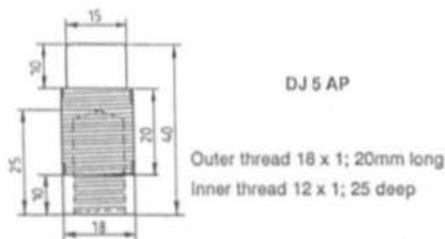
So we can remove the normal heatsink of a 2C39 and substitute the significantly larger one from a YD1270 - only it doesn't fit! Since the inner thread of a 2C39 heatsink has a smaller thread than the YD1270, a corresponding coupling piece or adapter must be made up.

This coupler, which for best heat conductivity should be made from copper, must be turned on a lathe (Fig.3). If no YD1270 parts are available but the necessary mechanical facilities are, a complete single-assembly heatsink

with 12 x 1 inner thread can be made from scratch from copper or aluminium. This can either match or exceed the size of that of the YD1270!

The heatsink of the YD1270 normally removes quite easily but the 2C39 may present more difficulty, according to its age and condition.

Before attacking and possibly ruining a good 2C39, it is worth practising on a defective or well-used one. Under no circumstances



DJ 5 AP

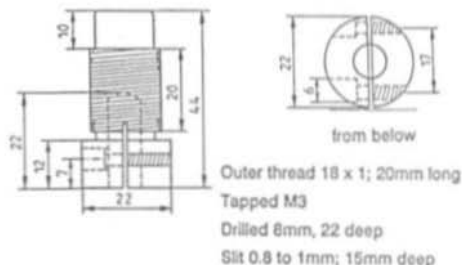
Outer thread 18 x 1; 20mm long  
Inner thread 12 x 1; 25 deep

**Fig.3a Coupler for improved cooling of 2C39 amplifiers with the fins from a YD1270 tube**

should the tube be placed in a vice and have its heatsink removed with greater or lesser force: instead use two pipe wrenches and strips of leather, one gripping the flange of the anode and the other heatsink. Then carefully ease the heatsink off.

The Telefunken version of the 2C39 has no screw thread but instead a stud of 8mm diameter onto which the heatsink is pressed with two set screws. These screws must be released first before the heatsink can be removed.

It is worth smearing the thread with white heatsink compound prior to screwing on the coupler.



from below

Outer thread 18 x 1; 20mm long  
Tapped M3  
Drilled 8mm, 22 deep  
Slit 0.8 to 1mm; 15mm deep

**Fig.3b Adapter for the Telefunken version of the 2C39.**

**Material: Copper, hard or half-hard, failing that Aluminium**

### 3.

### CLOSING CONSIDERATIONS

The described modification enables the heat dissipating surface of a 2C39 to be enlarged more than fourfold (Fig.4). For half-humane operation of the tube a relatively small blower will suffice; to reduce the noise during receive periods a resistor can be connected across it.

A reader inclined towards this scheme of things should nonetheless be clear of one thing: regardless of the heatsinking methods



**Fig.4 A 13cm PA with a modified 2C39BA**





Carl G. Lodstroem, SM6MOM/W6

## Measurement Arrangements for Complex Impedances

This article proposes and describes a scheme for measuring reflection coefficients of any load. They are indicated directly as a point on a Smith diagram. In principle one could take the values given by the measurement system and draw them directly on a Smith diagram using an X-Y plotter.

The measured voltages could also be taken to the X and Y inputs (equal sensitivity) of an oscilloscope and represent the locations of the reflection coefficients in real time. Additionally one could fix a transparent Smith diagram (or an enlarged section of one) over the 'scope's CRT.

I have said locations in the plural as it is possible to sweep the frequency - or vary the load - so that a line is drawn instead of just a point. This is an extremely capable tool for aligning circuitry. Fig.1 shows the overall scheme, which is composed in principle of an RF line with four detectors fixed at eighth-wavelength distances. Fig.2 shows a typical measurement result.

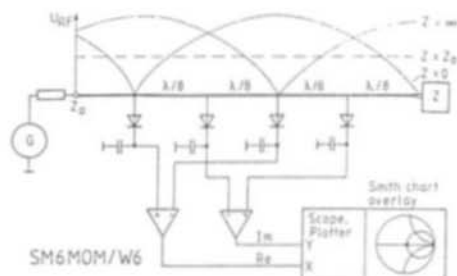


Fig.1 With two pairs of forwards-current diodes on an RF line, two amplifiers and an RF generator (approx. 1mW) complex reflection coefficients can be represented and measured. Because of the necessary 8th-wavelength distances, the frequency range of this arrangement is restricted: about one octave for antenna alignment: about +/- 5 to 10% around the design frequency for most measurements. Precision measurements can be carried out only at the design frequency.

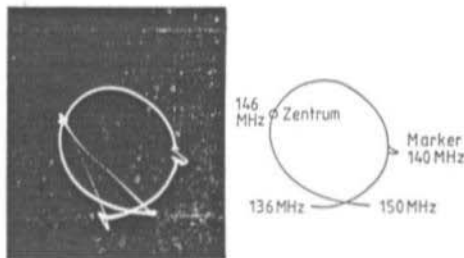


Fig.2 An HB9CV antenna, swept from 136 to 150MHz. Resonance lies at 146MHz (white spot at centre of screen and centre of Smith diagram). The marker on the outer right indicates 140MHz: the antenna is not good at this point.

1.

## PRE-HISTORY

Much has already been written about the Smith chart but the fundamental work is - not surprisingly - by Mr Philip H. Smith (1). The measurement instrument used here was calibrated by me in 1986 at a Design Contest and published in (2).

Later at the Foothill fleamarket in San Mateo, California I found a UHF admittance meter made by General Radio: with a receiver as detector it functions superbly. Doubtless there must be literature which I have not found yet for this device.

Finally there was the "Z-g Diagraph" from Rohde & Schwarz which showed some similarity and its description (3) can be read profitably in this connection.

In previous issues of VHF COMMUNICATIONS the function and construction of directional couplers (4, 5) and diode detectors (6) have been described (references are to the German edition); these publications give a lot of practical information.

For some decades now the Smith diagram has

been the means by which RF engineers illustrated complex resistances and achieved graphical solutions of matching problems, in other words an enhancement for the slide rule. These days the same problems (with many figures behind the decimal point) are more easily calculated with a computer and it can also represent the results on the screen in the form of a Smith diagram, which completes the circle.

Yet the computer is of little help with an "actual" problem! There on the bench sits a filter which needs aligning, there is an antenna that needs matching, how good is this attenuator or that load? If one could see their reflection coefficients on a Smith chart then one could tell how good or bad they were and what needed doing!

That is the application for this innovation.

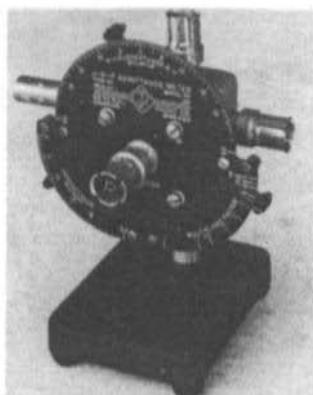
Ideas for this came from my excellent tutor Ingvar Svensson at the Engineering Academy in Goeteborg; he treated this theme (like all the others) as a figment of fantasy, as did Magnus Koch, an engineer with whom I was happy to collaborate several years ago. He is a measurement specialist and once said "If you can measure it with a bridge, do so."! I have merely combined the thoughts of these two men!

2.

## FUNCTIONAL METHOD

An RF generator  $G$  is connected to one end of an RF line and a load  $Z$  to the other, the length of the line being at least one wavelength (Fig.1). If the load is equal to the surge impedance of the line  $Z_0$  there is no reflection at the load; all power is absorbed in the load and converted to heat. At every position along the line one can measure the same RF voltage, assuming (as we do here) that losses in the measurement line are overlooked.





**Fig.3 The UHF Admittance meter by General Radio**

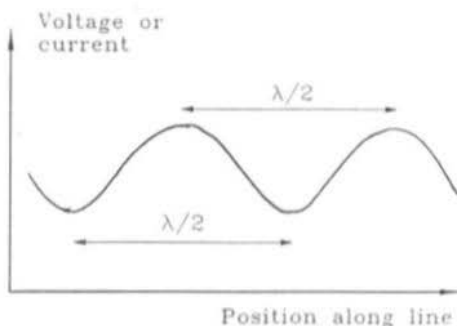
If the load is not equal to  $Z_0$  part of the power is reflected at the load. If the generator is also not properly matched then part of the reflected power is reflected again by the generator and the relationships become quite complicated. We shall assume that the generator is correctly matched.

The so-called mismatch can arise simply from a 50 ohm line being terminated in a 40 ohm resistor. Equally the termination might have too high a value or combine resistance with some inductance or capacity, either in series or in parallel. There are countless possibilities with mismatches.

Some forms of mismatch produce total reflection, certainly all those which have no resistive component: short circuit, open circuit, capacity, inductance and combinations of these.

If the load is not equal to the impedance of the line, the voltage will not be the same everywhere: instead it will vary periodically, giving us a "standing wave". Its form repeats itself every half wavelength along the line.

With a short circuit (or a load smaller than  $Z_0$ ) a voltage minimum and current maximum appears at the load end of the line. With an open circuit the opposite applies. Capacitive and inductive components lie in between.



**Fig.4 Voltage or current function along a line with a badly matched load**

When the reflection is pronounced - close to total reflection - the minima are sharp and drop nearly to zero volts. In significant cases of mismatching - as with a standing wave ratio (SWR) of 2 - the voltage variation along the line looks almost like a sine wave (Fig.4). With a perfect match there are, as mentioned, no periodic variations and at most a drop in potential if the measurement line has sufficient attenuation.

If one now measures the voltage on the RF line at four points each spaced an eighth-wavelength (45 degrees) apart, the four voltages should relate to one another in a sine/cosine fashion if there is a standing wave present. The four voltages are summed in pairs - without amplification to a resistor or at the output of a differential amplifier - and taken to the X and Y inputs of a plotter or oscilloscope.

If you now consider some extreme cases of terminations and work through the voltage relations resulting from them (Fig.1), you will quickly understand the arrangement and how you can use it.

We shall take a resonant circuit which at resonance gives rise to total reflection (Fig.5). The coil should be small and the (variable) capacitor screened, so that little radiation

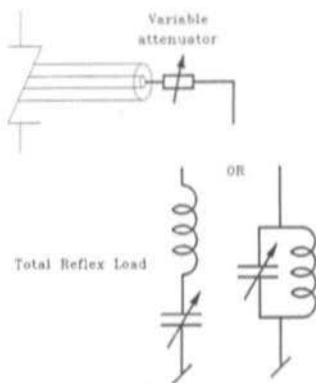


Fig.5 How calibration is carried out

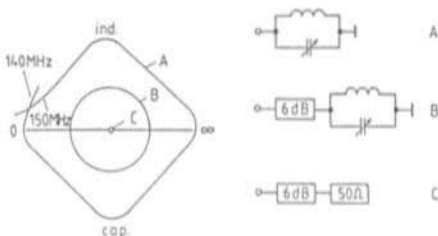


Fig.6 Three different terminations: A, B, C and the resulting indications - here for a measured frequency of 145MHz

If we tune the resonance (variable capacitor), the measurement point will wander around the periphery of the Smith diagram. The area around the short circuit (outer left) will remain clear with a parallel circuit as will the area around infinity (outer right) for a series circuit.

In reality we have not a circle but a square with rounded corners standing on one of these corners. This is because the voltages which determine the measured points do not have a sine wave form but pronounced minima. However, the better the match, the closer the shape is to a circle because the voltage response is nearer a sine wave (Fig.6).

If we now place a 3dB attenuator in front of the resonant circuit and tune it once more, the measured point now shows as a concentric, nearly round circle for 6dB return loss. With a 6dB attenuator we get a circle for 12dB return loss attenuation and with a 10dB one we have the circle for 20dB.

With an attenuator of, say, 10dB permanently on the output of the RF line the detector arrangement "sees" a return loss of at least 20dB, regardless of whatever is connected afterwards to be measured. This detracts from the appearance of sharp voltage minima and the diagram produced in this way will always be nice and round. If absolute sensitivity is required for specific measurements the attenuator should be taken out and you will have a 20dB enhanced measurement range. If the item being measured has a good match the diagram produced will be round in any case.

## 2.1 Operation

When we described the function of the detector arrangement previously we already mentioned the importance of calibration with the help of resonant circuits and attenuators. Practical use for the measurement of unknown impedances is not far removed from this.

First there must be a generator and measure-

occurs and as much power as possible is reflected. A (variable) attenuator ahead of the resonant circuit delivers calibration values.

A perfectly matched load (50 ohm instrumentation termination or attenuator of at least 30dB) gives the same voltage on each of the four diodes, everything is symmetrical and on the display we get a point in the centre of the Smith diagram.

With a resonant circuit at resonance used as load we get maximum asymmetry of the voltages of each pair and the point indicated moves out to the right (parallel circuit, infinity) or out to the left (series circuit, short circuit).



ment set-up for the frequency in question. The actual measurement involves comparing the display created by the item under examination with those given by the resonant circuit and attenuators.

The measurement arrangement is extremely sensitive so that even very long RF cables with measurable attenuation (which would otherwise have to be avoided) can be used without problem. Let us take for example a 30 metre length of RG-58 and measure it at 145MHz. The cable is 14.5 wavelengths long - that is 29 half-wavelengths. By varying the frequency by half of 1/29th of 145MHz (plus or minus 2.5MHz) we find out that the measurement points indicated describe a complete circle on the Smith diagram!

In this way it is possible to measure a remote object, for instance an antenna on the roof. This is the way it is done:

The antenna cable is first separated from the feed-point of the antenna and left open-circuit. Then we add lengths bit by bit at the accessible end of the cable until the display on the oscilloscope is out to the right, as for a parallel circuit in resonance. Instead of lengthening the cable one can also practise a small amount of frequency variation, as described in a previous section. Now we climb on the roof once more and connect up the antenna to be measured. Back downstairs we can see the complex reflection coefficients of the antenna and reflect on what is to be done (if anything at all is to be done). Or we can just sit and watch the 'scope display, noting when vultures come in and land on the antenna ...

In order to determine the amount of mismatching we use the arrangement in Fig.5. The value of the attenuator is varied until the amount (that is, the distance indicated from the centre of the display) is as great as that caused by the antenna. Then we double the value of attenuation engraved on the attenuator and that is the return loss of our antenna!

Given the opportunity we can also measure the attenuation of the feeder cable. If the roof end of the cable is open, total reflection occurs. Connecting the RF line we now have a signal travelling the length of the antenna cable twice and then being displayed (slightly) smaller as a total reflection. By adjusting the calibration according to Fig5 to give the same indication we learn the attenuation of the cable. This is directly equivalent to the attenuation read off, as in this case the wave also travels twice through the calibration set-up.

### 3.

## CONSTRUCTION

The measurement apparatus is made up fundamentally from a piece of RF line, four detectors and an indicating unit. My standard circuit is shown in Fig.7. According to frequency range and materials available, many widely-differing versions can be envisaged. Construction must follow good RF

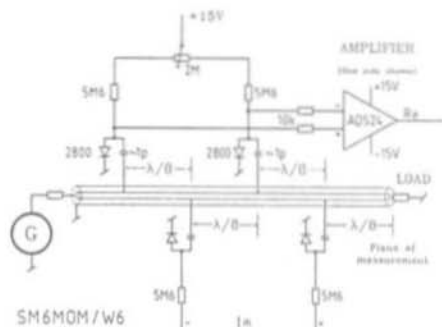


Fig.7 Circuit of the measurement bridge (second amplifier not drawn); in the arrangement shown the upper amplifier delivers the measured voltage of the actual component, the lower that of the imaginary component. Compare with Fig.1

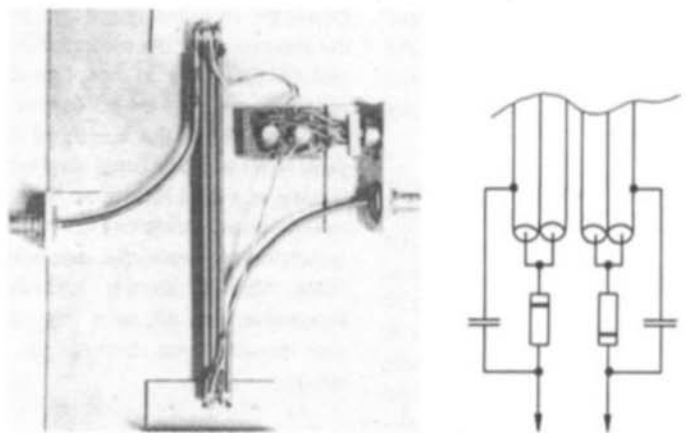


Fig.8 One of my first designs for the 2 metre band: diodes electrically connected and connected in anti-phase. On the right-hand side 8th-wave length pieces of semi-rigid cable next to one another

practice appropriate for the band in question of course and, in order to obtain repeatable measurements, it must be mechanically stable.

For each of the elements mentioned I will provide some information.

### 3.1 RF Line.

The RF line can be coaxial or take the form of stripline or micro-stripline, it can be parallel lines or waveguide. It only has to suit the use envisaged for it and be mechanically stable.

In most cases constructors will go for coaxial cable, and for the VHF, UHF and SHF bands a semi-rigid copper-sheathed cable represents a very good choice. The technique is not restricted to 50 ohms: for television we can take 75 ohm cable and for specialised measurement systems with an impedance of 25, 16.6 or 12.5 ohms for example we can use two, three or four lengths of 50 ohm cable in parallel. The detectors are then connected to only one of the cables.

For frequencies below 30MHz RG-58 or RG-8 is fine: for 7MHz an eighth-wavelength

is 3.5 metres long, which is still more or less manageable.

The pieces of cable can be laid beside one another as in Fig.8. For 145MHz as an example the eighth-wavelength pieces are 170 to 224mm long, according to the cable dielectric. Two of the detectors are each above and below: the small bumps caused by them are cancelled out because they affect the four quadrants of the Smith diagram to an equal extent. I always choose the blocking capacitors so that they are in series resonance at the frequency measured: for example at two metres 150pF with 15 to 20mm lead length.

A more elegant method is to coil (semi-rigid) cable, as shown in Fig.9. Each turn is an eighth-wavelength long, so that V:F7 all four diodes lie alongside one another. While the diodes cannot be made out in the picture, you can see the black blocking capacitors. The measured voltages are led to a DIN connector. In a separate case there are two multi-turn trim pots for balancing and summing; their wipers are connected to two op-amps of the type LM351.

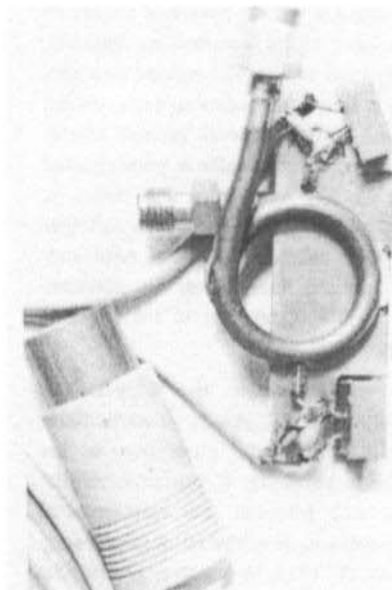
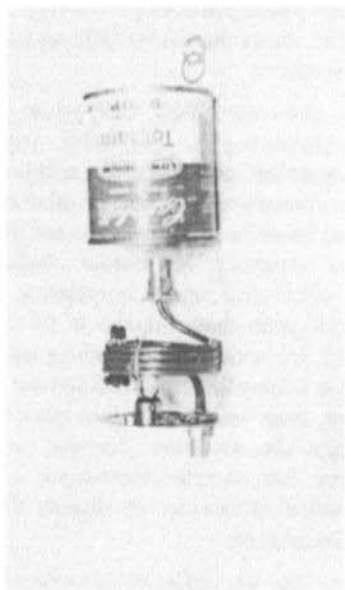


Fig.9(left) A subsequent version (for 145MHz) with coiled copper-sheathed cable, 8th-wavelength per turn, all four diodes on one side. A soup tin completes the arrangement mechanically

Fig.10(right) A version for antenna measurements in the 915MHz region; at higher powers used without amplifier. Each of the two turns is a quarter-wavelength

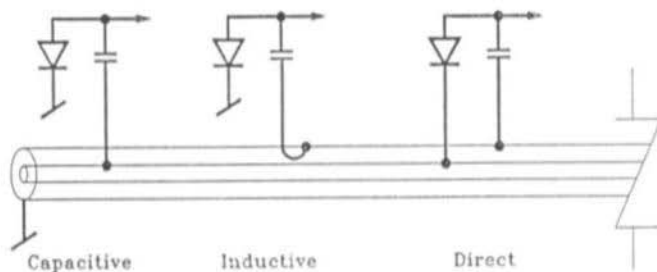


Fig.11 Three ways of coupling the detector diode to the RF line



Fig.10 shows a version for 915MHz. An eighth-wavelength here is only half a turn, so the diodes have to be mounted on opposite sides facing each other. The second turn and the two other diodes are out of sight below the screen panel (sheet brass or printed circuit board material), which makes a good ground plane. As an RF connector the SMA is particularly suitable; the measured voltages are once more measured in two multi-turn pots and summed, then taken via screened cable and a DIN connector to the separate amplifier box.

Naturally you can also use stripline or micro-stripline. In this form it is particularly easy to tailor the surge impedance to the application. In particular if you leave off the connectors and integrate the measurement set-up into one unit, you have some interesting possibilities. At lower frequencies the lengths involved could present difficulties.

### 3.2 Detectors

Either the electrical field can be coupled off the line with a capacitor or the magnetic field via a coupling loop; as Fig.11 indicates, at very low levels one can also make a direct electrical connection to the line. On the other hand, at very high RF power levels it is only necessary to place the diodes in the vicinity of the RF line.

In my first research set-ups I soldered the diodes direct to the centre conductor of the line, but this was not a good idea. For the rectified current there must always be a DC return path somewhere, and with the IC-202 I had problems with the direct current on the antenna input. It is better to connect the diodes to ground on one side and connect the other end loosely to the line via a capacitor, as shown in Fig.7.

Fig.7 also shows how to provide conventional Schottky diodes (e.g. HP2800) with bias current for achieving maximum sensitivity. At

the same time you can balance the diode pairs. In this way I have achieved good results up to 1300 MHz: above this the HP2800 appears to fall off rapidly.

For the short-wave and VHF bands germanium diodes (e.g. AA118, 1N34) - without bias - are no bad choice. For the middle and upper microwave regions, one the other hand, one must resort to the expensive and in all respects sensitive low-barrier Schottky diodes, which also require no biasing. The old silicon point-contact diodes of the 1N21 and 1N23 etc. series are also quite good, but they must be handled like MOS devices. For matching them beforehand, they should be first short-circuited, then clamped to the measuring line, then the short-circuit removed. Detailed information on detector diodes will be found in (6).

Initially I had the diodes in pairs connected head-to-tail, to read the differential voltage direct. At medium and high RF power levels it is then easy to connect them direct to the X and Y inputs of a plotter or oscilloscope or even to meters. But then the arrangement shown in Fig.7 turned out to be easy to copy and sensitive as well.

Fundamentally it is a good idea to pair the diodes. You can use a conventional digital multimeter: in the 200k-ohms range there is generally 10uA flowing; one reads the voltage drop in the direction of current flow and selects the diodes which are most similar.

### 3.3 Indication of measured values

For displaying the test results there is again a range of possibilities. One can use two test instruments having a centre zero position; one shows the real component, the other the imaginary component of the reflection coefficient.

One can also take an instrument with two movements and crossed-needles display, in



which case the cross-over point will indicate - somewhat coarsely of course - the location of the reflection coefficient on the Smith chart. One could bring into use a mirror galvanometer or servo motors with a mirror to cast a bright spot onto a large Smith chart on the wall (perhaps with HeNe laser for classroom use?). And finally you have the XY plotter or oscilloscope in XY mode already mentioned several times. Both can be very sensitive and offer adequate adjustment range for calibrating the display field.

### 3.4 Concluding remarks

For a broad range of applications and frequency bands it will be necessary to construct a series of measuring set-ups of various dimensions. It is not necessary, however, to equip each one with amplifiers and display components.

The user will do well to equip each detector unit with a four-core screened cable for connection to the amplifier module. For this I use normal audio plugs and sockets. All diodes can then be connected individually to one of the four pins; one resistor each of a few Kohms together with the capacity of the cable acts as a low-pass filter to stop RF interference from reaching the amplifier. The bias resistors and trimmer potentiometers are found in the amplifier unit. The full sensitivity will probably be seldom required - possibly for matching measurements at the inputs of receivers or when one wants to employ the motion detector scheme described.

For matching sensitivity to the measurement task you have at your disposal not only the output level of the generator but also the extent to which the diodes are coupled and the gain of the operational amplifiers. Expensive instrumentation amplifiers (e.g. AD524) can easily be pre-set to gain values of 1, 10, 100 or 1000.

[Text adapted by Robert E. Lentz, DL3WR.]

4.

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- (2) Carl G. Lodstroem: Complex Impedance Meter. RF Design magazine, December 1986, pp 50-53. Cacher Publications, Boulder, CO, USA.
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- (5) H. Tiefenthaler, OE5THL and B. Roessle, DJ1JZ: Mess-Reflektometer fuer VHF und UHF mit  $Z = 50$  ohms. UKW BERICHTE 13 (1973) Heft 2, pp 66-79.
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The Author's shack !



*Jim Toon, G0FNH*

## 10GHz ATV The Easy Way

### Part-2

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Continuing along from part-1 in the last issue of VHF Communications I will now describe the receive section of the system.

---

2.

#### THE RECEIVER

---

The circuit diagram of the receiver is shown in Fig.4, with the PCB layout in Fig.5 and the component overlay in Fig.6. Once the PCB has been loaded, checked for short circuits and mis-placed components, the Gunn diode bias supply is checked as explained earlier for the transmit modulator PCB. After confirming that the voltage swing available is correct reset the 10-turn potentiometer to give a bias of approximately 7 volts. The Solfan head should be course adjusted to 10.250GHz as explained earlier for the transmit head unit.

The PCB should again be mounted inside a small die-cast box mounted directly onto the Gunn diode head

As stated earlier I prefer to use a head unit

without an in-line mixer diode. The Gunn diode head (the local oscillator) is mounted onto one port of a cross-coupler, with the opposite port terminated with a small waveguide load. The dish or horn aerial is fitted to one of the ports of the cross-coupled arm and a diode detector fitted to the opposite port.

Cross-couplers are not difficult to manufacture and several designs are featured in the RSGB VHF/UHF Manual 4th edition. Alternatively they can often be obtained at rallies for around £2 to £4 each. Similarly, diode-detectors are easily obtained at rallies. Once again do not forget to add the 470-ohm resistor and the 0.01uF capacitor across the Gunn diode, and on the diode-detector add a 10K resistor across the diode (see Fig.11).

#### 2.1 IF Preampfier

Now we come to the pre-amplifier, the circuit of which is shown in Fig.7, the PCB layout in Fig.8 and the component overlay in Fig.9. This unit must be constructed using double-







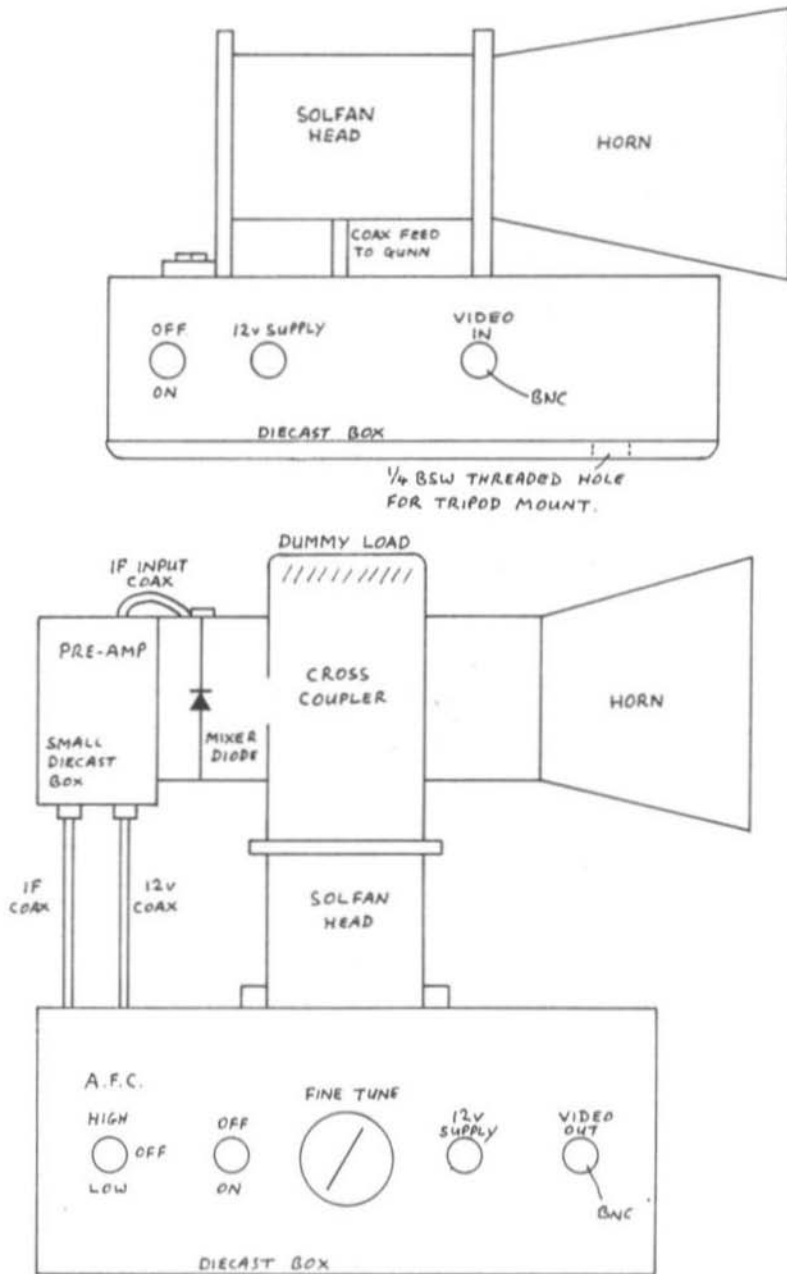


Fig.12 Mounting methods for the various assemblies

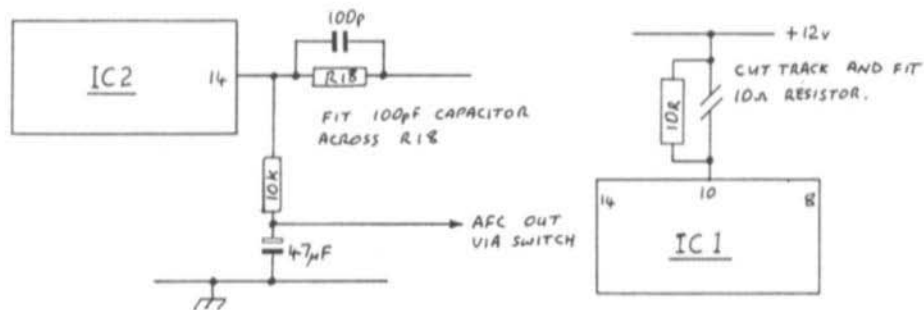


Fig.10 Modifications to the BATC FM-Demodulator

sided PCB material and fitted inside a small die-cast box. The unit should then be mounted as close as possible to the mixer diode in the diode-detector. The IF output from the pre-amp is taken via a coaxial lead to the BATC FM ATV demodulator PCB. The demodulator PCB is built according to the instructions supplied with it from BATC Member's Services, and then modified as shown in Fig.10.

Setting up the receiver is accomplished as follows: after ensuring that the local oscillator head is running at around 10.250GHz switch off the AFC and tune in the best possible picture from the transmitter by adjusting the brass tuning screw on top of the Solfan head.

NOTE: Ensure that the transmitter is some distance away from the receiver otherwise the receiver will be swamped and incorrect tuning

may occur. Also, it may be necessary to readjust the video deviation on the transmitter to achieve good results. Once the best possible picture has been tuned in using the brass tuning screws fine tuning is achieved by adjusting the 10-turn potentiometer on the receiver Gunn diode bias supply. Now switch the AFC on and select High or Low as appropriate and that's it! With a little practice you soon get the hang of tuning signals in.

Audio can be added to the system, but this has been covered many times in other 3CM projects, and there are many ideas around showing how this can be accomplished (see CQ-TV 136/137/141 or the BATC ATV Compendium for further information). Fig.12 shows how I mounted the transmit and receive units into their respective die-cast boxes.

That completes this simple, but effective, 10GHz ATV station. Very soon Solfan Heads, cross-couplers, waveguide, plumbing etc will probably be old hat - but I enjoy it! See you on the band.

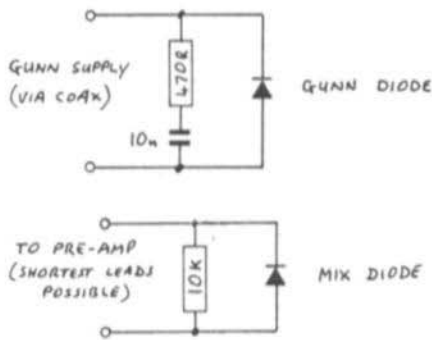


Fig.11 Connections to the Gunn diodes

3.

## REFERENCES

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



Werner Brueckner, DL6MDA

## RF Sweeping with the PC

There's no such thing as a free lunch and this applies with equal validity to usable test results at radio frequencies. All the same, there's no need to spend the price of a luxury limousine on an analyser if all you are doing is sweeping a VHF pre-amp. The following description discusses this topic impartially and comes up with a substitute for a spectrum analyser with an 80dB dynamic range.

1.

### CONCEPT

For checking the amplitude frequency response of passive or active circuits properly a spectrum analyser is really indispensable. A description of one, by DB1NV, has appeared in VHF COMMUNICATIONS. The demands are considerable, though, and certainly confined to a small number of amateurs.

The concept proposed here goes down a different route (Fig.1). The core is a straight-

forward-to-make PC plug-in card containing a D-to-A and A-to-D converter. The card is therefore very simple, as it is mainly digital and needs virtually no alignment.

All the same, it does need an external oscillator. You can either use a ready-made module, perhaps one you already have, or a circuit published previously in VHF COMMUNICATIONS or else the universal oscillator using an MC1648 to be described now.

I would like to point out that the idea for this article came from the article "Diode detectors in 50-ohm broadband technique" by Carsten Vieland (1).

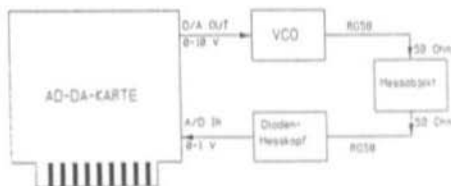


Fig.1 Overview of sweep set-up



Why do we need a computer as well now? Mainly because most people already have them. With a computer you can exploit several of its important properties:

- Keyboard: this makes potentiometers, variable capacitors, etc. superfluous;
- Screen: this makes readouts, 'scope displays, etc. unnecessary;
- Printer: substitute for screen shots

### 1.1 Which computer to use?

The description that follows is based around an IBM-compatible XT or AT fitted with a CGA colour graphics card (640 x 200 point resolution). All EGA and VGA cards can emulate this CGA mode. People using Hercules cards can use a CGA-simulator.

The big advantage of using this graphic mode is the built-in function in DOS to produce a screen dump by pressing <SHIFT> and <PRINT SCREEN> together (assuming a dot-matrix printer is connected). The only other requirement is to load the DOS GRAPHICS.COM program before using the sweep program (unfortunately not possible with Hercules graphics). All the curves reproduced here were printed out in this manner.

### 1.2 Which programming language to use?

The program presented here was written in TurboPASCAL 4. Even died-in-the-wool hardware fans will overcome their objection to software when they see how elegantly they can depict, say, the frequency response of their two metre preamp.

The program can easily be re-written in BASIC.

2.

## THE SWEEP OSCILLATOR

When measuring the frequency-amplitude response of a filter or amplifier you first need a variable frequency generator with an impedance of 50 ohms. It is possible to set the frequency manually and by diode rectification, note the voltage value. This can also be done more elegantly and much quicker with the help of a PC and the plug-in card described. Most frequency generators possess an input which enables us to control the output frequency, in other words the voltage-controlled oscillator input (VCO input). An example is the Function Generator 3310B by Hewlett Packard, variable from 0 to 5MHz.

Most amateurs have higher aspirations (in frequency). Up to 200MHz the circuit using the Motorola IC MC1648 is very suitable (Fig.2). It is extremely easy to hook up and very affordable. Data for this IC is collated in Table 1 (overleaf).

By simply altering the inductance the oscillator can be matched to the frequency band required. The RF circuitry is best made in an RF-tight enclosure, silver plated, and provided with a BNC connector for the RF output. Unfortunately the effective output of the oscillator is just 100mV or so into 50 ohms, but it does have a very linear amplitude response. This is the most significant criterion

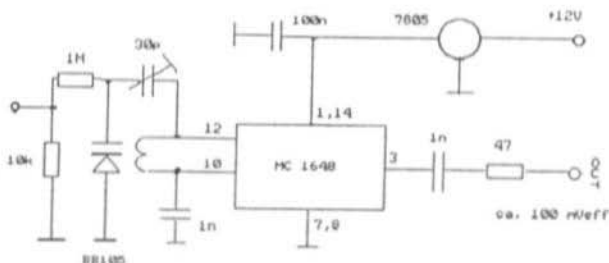


Fig.2 The VCO with the MC1648



with sweep generators for measurement purposes.

If you intend handling higher frequencies in broadband fashion the oscillators in (2) are recommended. DB1NV describes oscillators up to 1.5GHz there.

**TABLE 1: Technical data for MC1648**

Case:	Dual-in-line, 14 pins
Input capacity:	typically 6pF
Dissipation:	150mW at +5V
Max. output freq.:	225MHz
Coil inductance required:	
for 1-10MHz	100uH
for 10-60MHz	2.3uH
for 60-100MHz	0.15uH
for >100MHz	<0.15uH.

## 2.1 Component list for the VHF VCO

1	Motorola MC1648
1	voltage regulator 7805
1	Varicap BB105
1	trimmer capacitor 30pF
2	1nF
1	100nF
1	10kohm
1	1Mohm
1	47R
1	BNC connector
1	coil, 2 turns on 10mm former

## 2.2 Calibration

### 2.2.1 Calibrating the frequency axis

A problem arises from the use of a varicap diode as the frequency determining element of the VCO. The non-linear characteristic curve of this component must either be taken care of in software or else the frequency response of the oscillator must be measured with a

frequency counter in advance and made to correspond with the screen-display value. The author chose the second method.

So we produced the frequency-voltage characteristic curve manually (see Table 2). With the VCO used here 20 measurement points in 1MHz steps were adequate.

If we are using only 600 points of the ones available along the abscissa, the X value of the Line command of the computer can be calculated simply with the line (round (Volt) \* 60). Of course this will have to be repeated 21 times to reach the 20 divisions.

The block repeat commands available in Turbo-PASCAL are very handy for this.

**TABLE 2: Frequency-voltage characteristic curve of the VCO using MC1648 and varicap BB105**

Frequency MHz	Tuning voltage Volts
130	0.06
131	0.30
132	0.50
133	0.77
134	1.05
135	1.39
136	1.71
137	2.08
138	2.48
139	2.91
140	3.43
141	3.93
142	4.52
143	5.14
144	5.79
145	6.47
146	7.14
147	7.81
148	8.48
149	9.14
150	9.83

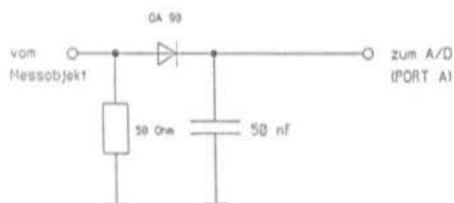


Fig.3 The diode measurement head

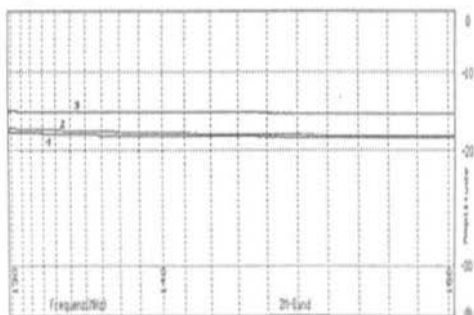


Fig.4 Measured voltage over frequency for: 1) AAZ15; 2) OA90; 3) OA5

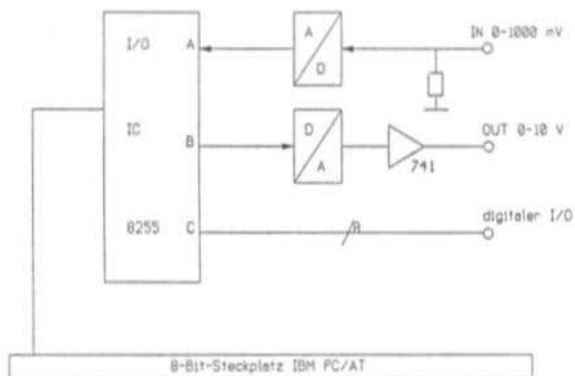


Fig.5 The A-D/D-A card

Bit 0	Bit 0 --- Direction from Port C (lower); 1=in, 0=out
Bit 1	Bit 1 --- Direction from Port B; 1=in, 0=out
Bit 2	Bit 2 --- Select mode of operation 0 or 1
Bit 3	Bit 3 --- Direction from Port C (higher); 1=in, 0=out
Bit 4	Bit 4 --- Direction from Port A; 1=in, 0=out
Bit 5	Bit 5 --- Select mode of operation:
Bit 6	Bit 6 --- 00=mode 1, 01=mode 2, 1x=mode 3
Bit 7	Bit 7 --- Operating mode identifier: 1=active

Fig.7 Controlling the port module 8255







### 2.2.2 Calibrating the amplitude axis

It is recommended you establish the characteristic curve of the measurement diode OA90 (AA112) manually. To do this we need a switchable attenuator to 30dB following an RF generator of at least 1 volt effective output. If any aberrations are established, you can easily change these in the procedure sequence Line (X1, Y1, X2, Y2) in the program listing.

### 2.3 Selection of measurement diodes

Since the author is only concerned with frequencies below 200MHz, a simple measuring head (Fig.3) with germanium diodes is used.

If you are interested in higher frequencies you should definitely look at DJ4GC's article (1). According to this the limit of reliability of a normal germanium point-diode is reached at around -50dBm. It can be used up to measurements of 500MHz or so. A sensitivity check can be made by connecting the oscillator output direct without sample object to the measurement head input. A comparison of several diodes of the same type gave astonishingly small difference in amplitude characteristic, mainly less than 1dB. On the other hand, with different types variations were noticed throughout. Thus with the OA5 there was around 3dB more voltage than with the diodes AAZ15 and OA90 (Fig.4). The popular SHF diode 1N21 gave values similar to the OA90.

---

3.

### THE A-D/D-A PLUG-IN CARD

---

Fig.5 shows the block diagram of the A-D/D-A card, and Fig.6 the complete circuit.

The kernel of the half-length card for the IBM PC/AT is the affordable and easily sourced

A-D converter CA3162E. It offers a display range of -99 to +999 millivolts, with overflow indication, and has a high temperature-stability for its internal voltage reference.

The measurement output for the three digits is produced in multiplex fashion with 96 measurements per second. Since a three-digit measurement output covers a range of  $2^{\cdot}10 = 1024$ , we can justifiably call it a 10 bit converter. It has a dynamic range of  $20 \cdot \log(1024) = 60.2\text{dB}$ .

On the other hand, the reverse direction, the D-A conversion using the ZN426E, is set out only for 8 bits wide. Consequently it is unequally faster. Conversion speeds of 2 microseconds, corresponding to 500kHz, are theoretically possible although the computer is scarcely capable of producing these. In any case, the D-A converter does not need to be faster than the A-D converter, at least not in our application.

Via Port B the frequency can be determined by the computer. By way of the object to be measured, which must be terminated according to its impedance, the measured voltage is rectified and passed to the A-D converter at Port A. Fig.7 demonstrates how the operating modes of the port module 8255 are controlled.

Graphical representation and the results on the printer are now purely for the software to determine.

I must make clear here that the VCO is driven by an analogue signal which has been quantised into 8 bits, in other words in 256 steps. This means that over very large sweep ranges, e.g. 100MHz in this example, the measured frequency is not entirely continuous and there will be gaps (at  $100\text{MHz}/256 = 0.39\text{MHz}$  steps).

For some applications this will be too great. A "slurring" of the steps and thus a continuous signal can be obtained with an R-C element on the D-A output. Fig's.8 and 9 show finally the component layout of the card and a prototype construction.





```
(*****)  
(* Wobbelprogramm 130-150 MHz mit *)  
(* AD-DR-Karte in Turbo-Pascal 4/5 *)  
(* für UKW-UCD mit MC 1648 *)  
(* unter Verwendung der CGA-Grafik *)  
(* (c) W. Brückner DL6MDR *)  
(*****)  
  
program wobbler ;  
  
uses crt,graph;  
  
var treiber,modus,wert: integer;  
    x,y: longint;  
  
procedure grafik: (-----)  
  
begin  
    treiber:= cga;  
    modus := cgahi; (640*350 Punkte )  
    initgraph(treiber,modus,'');  
end;  
  
procedure raster: (-----)  
  
begin  
    rectangle(0,0,600,199);  
    setlinestyle (1,0,1);  
  
    line (round(0.06*60),0,round(0.06*60),199);  
    line (round(0.3*60),0,round(0.3*60),199);  
    line (round(0.5*60),0,round(0.5*60),199);  
    line (round(0.77*60),0,round(0.77*60),199);  
    line (round(1.05*60),0,round(1.05*60),199);  
    line (round(1.39*60),0,round(1.39*60),199);  
    line (round(1.71*60),0,round(1.71*60),199);  
    line (round(2.08*60),0,round(2.08*60),199);  
    line (round(2.48*60),0,round(2.48*60),199);  
    line (round(2.91*60),0,round(2.91*60),199);  
    line (round(3.43*60),0,round(3.43*60),199);  
    line (round(3.93*60),0,round(3.93*60),199);  
    line (round(4.52*60),0,round(4.52*60),199);  
    line (round(5.14*60),0,round(5.14*60),199);  
    line (round(5.79*60),0,round(5.79*60),199);  
    line (round(6.47*60),0,round(6.47*60),199);  
    line (round(7.14*60),0,round(7.14*60),199);  
    line (round(7.81*60),0,round(7.81*60),199);  
    line (round(8.48*60),0,round(8.48*60),199);  
    line (round(9.14*60),0,round(9.14*60),199);  
    line (round(9.83*60),0,round(9.83*60),199);  
  
    line (0, 40,600, 40);  
    line (0, 90,600, 90);  
    line (0,166,600,166);  
  
    line (350,198,428,198); (2M-Linie)  
    settextstyle(2,0,4);
```

Fig.9 Program listing (continued overleaf)



```

    outtextxy(610,35,'-10');
    outtextxy(610,85,'-20');
    outtextxy(610,160,'-30');
    outtextxy(610,1,'0');
    outtextxy(610,190,'db');
    outtextxy(100,185,'Frequenz(MHz)');
    outtextxy(360,185,'2M-Band');

    settextstyle(2,1,4);
    outtextxy(610,100,'Amplitude');
    outtextxy(1,167,'130');
    outtextxy(200,167,'140');
    outtextxy(580,167,'150');

ends;

procedure einlesen; (-----)

var
    ein,ausgabe : longint;
    a,b,c : longint;
    volt, messwert : real;
    adresse : integer;

begin
    adresse:= 784; {768,772,776,780,784,788,792 oder 796}
    port[adresse+3]:=144; {Kontrollregister}
    setlinestyle(0,0,1);

    volt:= 0; x:=0; messwert:= 0;
    moveto(0,199);
    repeat
        repeat {Hunderter}
            ein:= port[adresse];
            until (ein and $70 = $40);
            ein:= port[adresse];
            a:= (ein and $0f)*100;

        repeat {Einer}
            ein:= port[adresse];
            until (ein and $70 = $10);
            ein:= port[adresse];
            c:= (ein and $0f);

        repeat {Zehner}
            ein:= port[adresse];
            until (ein and $70 = $20);
            ein:= port[adresse];
            b:= (ein and $0f)*10;

        messwert:= (a+b*c);
        if messwert >999 then messwert:= 0;
        if messwert < 0 then messwert:= 0;
        messwert:= 66.6666*(ln(0.000001+messwert)/ln(10));

        y:= round(messwert);
        lineto (x,200-y);

    volt:= x*255/600;

```

Fig.9 Program listing continued (continued overleaf)



```
ausgabe:= round(volt);
portu[adresse+1]:= ausgabe;   ( auf Port B)
portu[adresse+2]:= ausgabe;   ( auf Port C)

inc(x);
until x = 600;
end;

begin (----- Hauptprogramm -----)

grafik;
setviewport(0,5,639,349,true);
rectangle(0,0,600,200);
raster;

repeat
einlesen;
write (#?);                    (kurzer Ton)
until readkey = #27;           (warten auf ESC-Taste)

closegraph;

end.
```

Fig.9 Program listing, continued

With the control word used 144d (d stands for decimal), Port A is switched to input (A-D converter), Port B to output (D-A converter) and Port C again as an output. Port C is not used again here. Now the measured value is read in. To aid understanding it should be mentioned that the A-D converter CA3162 has its own built-in timing generator, which cannot be influenced by the PC.

This timing generator now sets out the three digits of the test result in order of hundreds, units and tens, BCD-coded on four output lines. The values must be fetched by the computer one after the other and then combined. Since the speed of this is determined by the timing generator of the CA3162, a fast AT is of no greater value than the slowest 4.77MHz PC. The maximum conversion speed of 96 measurements a second is thus a kind of unbreakable sound barrier. To tell the PC which of the three digits is being presented the three SELECT lines of the CA3162 are passed to the Port A line. By means of a logical concurrence, the measured value and digit count can be separated again.

The complete result is combined in the variable "Messwert" (measured value) and is then ready to be represented graphically as

long as the limiting values 0 and 999mV are not exceeded either way.

The powerful capabilities of the CA3162 to measure small negative voltages as well are also prevented from being taken into account.

Following this, the D-A converter is addressed. Its output value is raised in such a way that its highest value of 255 is achieved when the graph has traversed 640 pixels on the abscissa. The 741 op-amp following the D-A converter is aligned so that it then delivers 10 volts output. This same value is what the oscillator connected needs as maximum sweep voltage.

If the abscissa has not yet drawn 640 pixels, then the measurement process is repeated, otherwise the measurements are halted and paused until a key (any key) is pressed.

The whole measurement process takes 640 measurements/96 measurements per second = 6.6 seconds, a thoroughly acceptable result. Sweeping should not be carried out too fast, otherwise the integrating capacitor following the OA90 diode may not be fully charged (particularly at low test frequencies). Sweeping too slowly should also be avoided for fear of getting ripple in the test results.



5.

### ADDRESSING THE A-D/D-A CARD, WITH PROTOTYPE ADDRESSES FOR THE PC

To make possible the addressing of additional card, the PC is provided with 32 positions in the address range 768 to 998 (decimal). The 8255 i.c. on the A-D/D-A card itself requires four address positions, following consecutively. Here the lowest address is used to address Port A, the next higher for Port B, then the next for Port C and the fourth address is for receiving the control word. This should be output first from the PC to the A-D/D-A card, so as to prepare the three ports as inputs or outputs, as described.

An 8-pole DIP switch is provided on the card, enabling any choice of start address to be selected in a block of fours. In the example program the lowest address is 768 decimal. Table-3 below makes the connection clear.

Any block can be chosen, best of all the first one. It may be that other cards are already occupying this address, and if so one should study the handbook to select another. Finally set the corresponding DIP switch to ON, the rest OFF.

The computer can now have access to the four addresses of this block, that is to the three

input/output ports and to the control word register. With us this control word register is always set to the value 144 (decimal), allocating Port A to input and Ports B and C as outputs (Fig.7).

6.

### ALIGNMENT

The A-D/D-A card is programmed to address 768. After inserting the card we load Turbo-PASCAL 4.0 and type in the program in Fig.9. At the output of the card (pin 13 of the Sub-D connector) a sawtooth should appear, with a maximum amplitude of 10V. This is adjusted with potentiometer P1 (Fig.6).

For setting up the analogue/digital part of the card we apply alternately voltages of 1V and 0V to the input (pin 5 of the Sub-D connector). With 0V on the input potentiometer P2 (zero adjust) is used to align the resulting measurement line with the -35dB mark on the diagram. With 1V on the input potentiometer P3 (gain) is used to set the resulting measurement line to match the 0dB mark on the diagram. Each measurement run is started anew by pressing any key. You can break out of the program by hitting <ESCAPE>.

**TABLE 3: Programmable block addresses (decimal)**

1st block:	768 (769, 770, 771)	corresponding to Port A, B, C and the Control word
2nd block:	772 (773, 774, 775)	corresponding to Port A, B, C and the Control word
3rd block:	776 (777, 778, 779)	corresponding to Port A, B, C and the Control word
4th block:	780 (781, 782, 783)	corresponding to Port A, B, C and the Control word
5th block:	784 (785, 786, 787)	corresponding to Port A, B, C and the Control word
6th block:	788 (789, 790, 791)	corresponding to Port A, B, C and the Control word
7th block:	792 (793, 794, 795)	corresponding to Port A, B, C and the Control word
8th block:	796 (797, 798, 799)	corresponding to Port A, B, C and the Control word

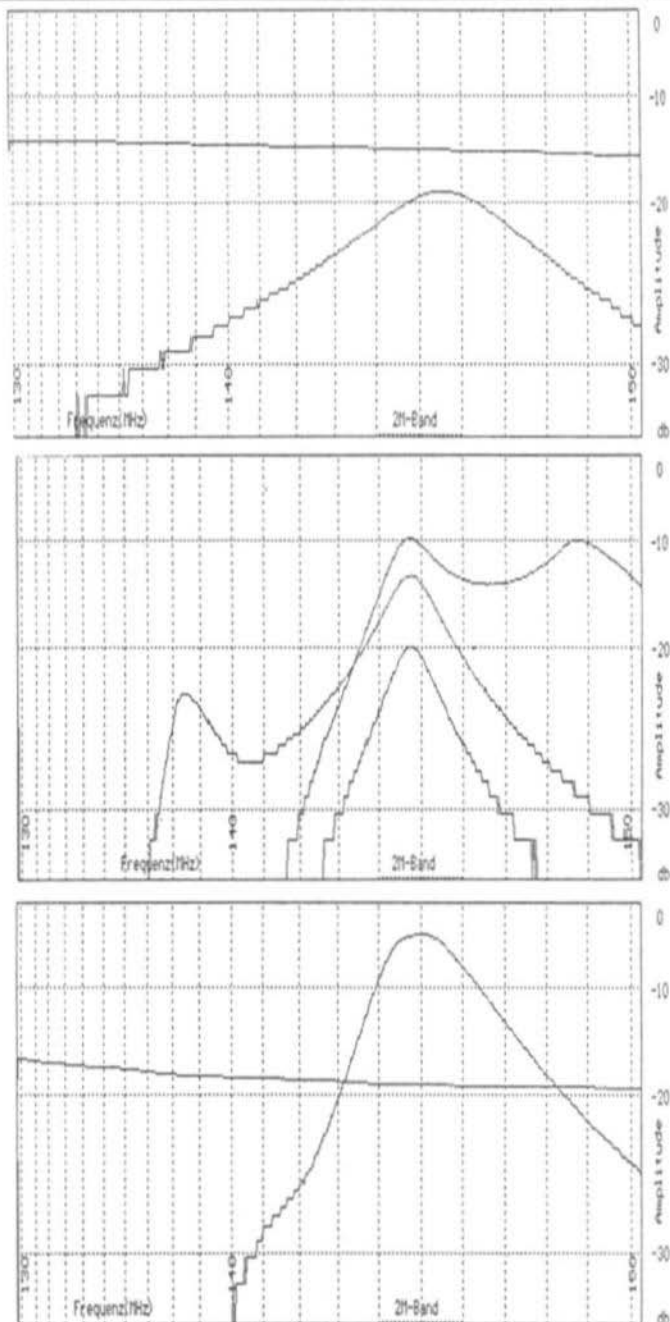


Fig.10 (top) Measurement curve of a simple LC circuit

Fig.11 (middle) Alignment of a 2 metre preamplifier

Fig.12 (bottom) Successful alignment of the preamplifier





7.

## EXAMPLES OF MEASUREMENTS

Fig.10 shows the curve obtained with a simple LC circuit. It is easy to see the approx. 3 to 4dB through attenuation and the relative flat fall-off of the flanks.

Fig.11 shows the lining up process of a two metre preamplifier with four stages. The ability to draw as many curves as desired over each other makes alignment simple.

Fig.12 shows the aligned preamplifier. Do not forget to produce a reference line by removing the test object and substituting a BNC coupler. Then it is possible to read off the amplification direct in dB, here about 15dB.

Special thanks are due to Mr F. Breuninger, DB9MZ, who contributed to this article in countless discussions.

8.

## LITERATURE

(1) Vieland, C., DJ4GC: Diodendetektoren in 50-ohm Breitbandtechnik. UKW-BERICHTE 27 (1987) Heft 4, pp 223-237.

(2) Jirmann, J., DB1NV: Breitbandig abstimmbare spannungs-gesteuerte Oszillatoren. UKW BERICHTE 26 (1986) Heft 3, pp 150-157.

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*John Barker*

## A Home Built Satellite Dish Steering System

Rather than fork out £600 plus for a multi-satellite system, John Barker built his own motorised polar mount from a few bits and pieces.

---

1.

### INTRODUCTION

---

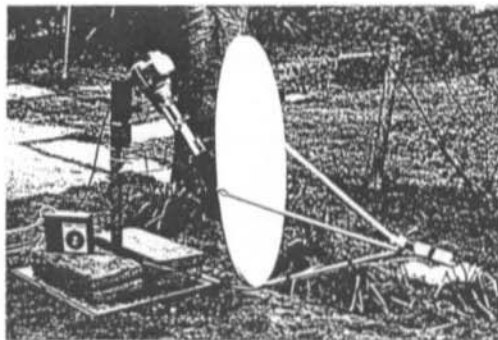
Although multi-satellite systems cost £600 and more, they are little more than Astra systems with a motorised dish and controlling positioner. With this in mind the temptation was strong for me to convert my existing Astra system to track the Clarke belt of satellites myself.

It took some work, but for an outlay of only £50, I now have a steerable dish, controlled from the armchair.

I wanted the motor controllable from indoors, and rather than attempt to build a control system myself, I made use of an aerial rotator. These are readily

available devices designed to turn TV or radio aeriels. They cost about £40-£50 complete with the controller box, from the likes of Tandy or MAPLIN ELECTRONICS.

The rotator motor unit is meant to drive a vertical mounting pole. This meant adapting the system to drive a satellite polar mount - not as simple as it first appeared. The secret, I soon found out, is to mount the rotator upside down and reverse the bracket on the back of the dish.



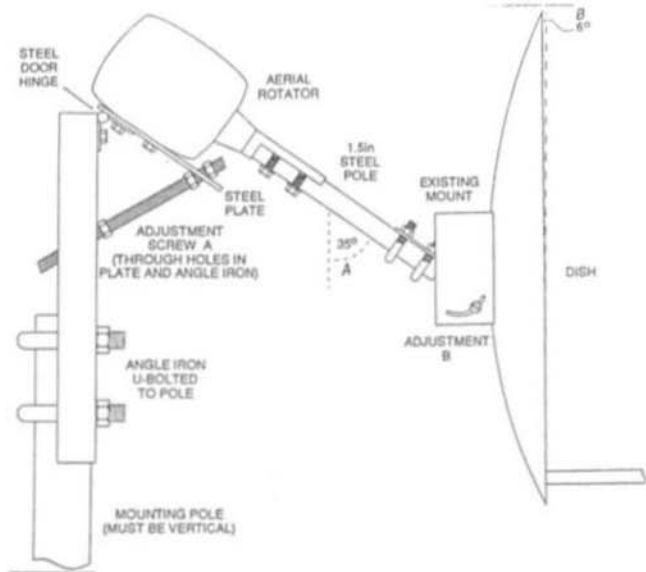


Fig.1 Construction details

Virtually any dish can be converted if it clamps onto a pole, although it may need altering so the elevation will go lower. In most cases, this will involve drilling a new pivot hole. If the back bracket can be reversed then so much the better.

Amstrad owners can obtain a converting bracket to enable their dish to clamp to a pole.

A 60cm dish will not give good results - you will find other satellites, but they will be "sparklie".

I chose an 80cm offset dish, which is relatively cheap, easily available, and had the elevation adjusted by a screw thread, which suited my purpose for experimenting.

The LNB is a 1.7dB unit with a ferrite polariser. This dish comes with a selection of clamps to take virtually any LNB, including the popular Marconi. The fact that the rear bracket could be turned upside down was just plain luck, and in hindsight makes it the perfect choice.

The results were surprisingly good, achieving virtually "sparklie-free" pictures on *RAI Uno*, *CNN*, *Super Channel* and *Discovery*, which was the one channel I really thought I was missing out on.

Most other channels, although suffering from "sparklies", are quite watchable. All the Astra channels are of course perfect.

The accuracy of the rotator is also surprisingly quite good.

## 2.

### CONSTRUCTION

Construction speaks for itself, but where I used welding a nut and bolt would have done just as well. The diagram in Fig.1 shows how mark two would look, which is more simple and better than the experimental model.

The whole thing needs to clamp onto a vertical pole, which can be a pole attached to base, a pole clamped to a good wall bracket, or just a pole sunk into the lawn. In fact, the particular dish I used comes with a wall bracket which would do the job perfectly.

Setting up requires a little patience and some understanding of what the object of a polar mount is. If you intend to wall-mount it, I suggest you have a practice on the ground first.

First, set your satellite receiver up with a



portable TV in the garden next to the dish and refer to channel information - frequencies and so on. A signal meter will not be required.

By turning the rotator controller, set the dish exactly central to the rotator (measure each side from the edge of the dish to the hinge). Now adjust angle **A** by adjustment of nuts **A**.

Adjust angle **B** using the elevation adjustment on the dish. Swing the whole assembly about the mounting pole to find a channel on 1 degree West. If it cannot be found, adjust angle **B**. When you have found 1 degree West tighten the clamps to the mounting pole. Do not loosen or move this again.

Now tune your receiver for a channel on each satellite along its arc. Turn the rotator controller to find if the arc is correct. When a new satellite is found manually, push the dish up or down to see if the signal improves, and therefore whether your arc fits the arc of the satellite. Fine tune the adjustments accordingly.

---

3.

## CONCLUSIONS

---

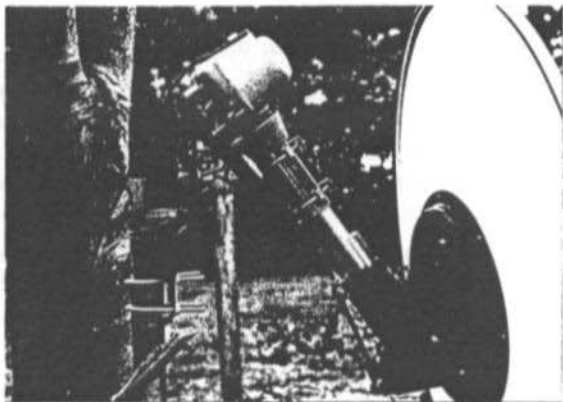
One very useful piece of information which I wish I had had the first time is a "marker" for each satellite. That is, one channel on each satellite which is a strong signal, in PAL, not scrambled, and in the normal frequency band, with on-screen identity and broadcast 24 hours, or showing a test card when off. Table 1 gives six channels I have found suitable for this purpose which can be used to set up this (or any other) polar mount.

The end result is, of course,

not in the same league as a proper multi-satellite system when it comes to convenience. You have to tune the receiver and then adjust the dish position with a dial, rather than the single push button most commercial systems offer.

However, this is a small price to pay compared to the savings made, and the rotator system, at the very least, gives you the opportunity to try out a multi-satellite system for very little cost before plumping for the real thing.

Pos'n	Channel	Freq.	Polaris'n
23.5 E	3SAT	11.598	Vertical
19.0 E	SCREENSPORT	11.214	Horizontal
13.0 E	SUPER CH'NL	11.674	Vertical
10.0 E	RAI DUE	11.640	Horizontal
1.0 W	TV NORGE	11.016	Horizontal
27.5 W	DISCOVERY	11.175	Horizontal





## READERS FORUM

---

### RADIO ASTRONOMY

---

**Hans Schlueter, MD, DJ7GK**

The idea of detecting a transit of the moon with my satellite TV set-up arose from a conversation with DK8CI, the author of "First Results of the Garching Radio Astronomy Observatory" (VHF COMMUNICATIONS 4/1990).

In the IF of my satellite TV receiver I checked for a stage which definitely had no limiting and fed out the noise voltage. E four-digit digital voltmeter indicated the voltage after rectification and filtering with  $C=1\mu\text{F}$ .

My parabolic dish has a diameter of 1.2 metres and the LNC according to the data contains a HEMT with  $F=0.84\text{dB}$ . On a starlight clear night with full moon I directed the antenna so that the shadow of the LNC fell

on the centre of the dish (with offset feeds this idea obviously won't work - editor).

The indicated voltage climbed reproducibly from 140.0 to 140.8mV. A calculation by DK8CI proves that the rise in noise level comes very close to the value calculated theoretically.

Moral: even with very simple set-ups extremely interesting tests can be carried out. It is possible that amateur EME signals can also be detected in this fashion.

---

---

### FM DEMODULATION FOR METEOSAT

---

**Hans Schlueter, MD, DJ7GK.**

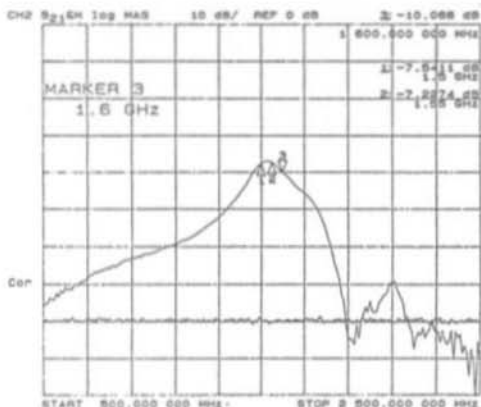
In VHF COMMUNICATIONS 2/1990 DF9DA published details of a "Compact FM Receiver for Weather Satellites". The coincidence demodulator used in this has pros and





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A recently introduced product in our range, the RPA-7006 1.6GHz narrowband masthead Antenna Pre-amplifier is proving ideal for reception of Inmarsat satellite signals. The full specification of the unit is given below. A power supply module (P-9028) is also available for providing the necessary DC supply for the pre-amplifier via the feeder cable.



## Gain versus Frequency

FREQUENCY RANGE	1.4 TO 1.6GHz
GAIN	30dB
3dB BANDWIDTH	200MHz
NOISE FIGURE	2dB
INPUT V.S.W.R.	<1.20:1 nom frq
OUTPUT V.S.W.R.	<1.50:1
THIRD ORDER INTERCEPT	+14dBm
SUPPLY (VIA COAX)	12 to 14 VDC
CURRENT	80mA
SIZE *	76 x 29 X 38mm
WEIGHT	200gms
INPUT CONNECTOR	N plug
OUTPUT CONNECTOR	N socket
FIXING SYSTEM - BASEPLATE 4 HOLES	

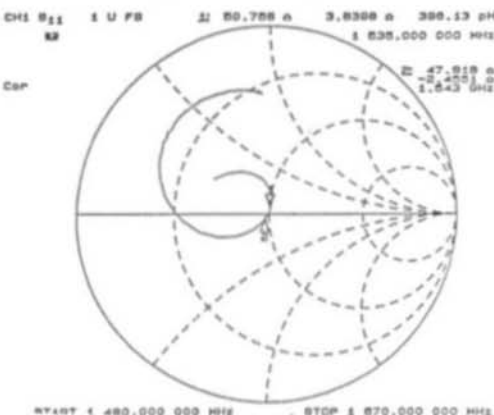
\* N connectors extend length from 76mm to 120mm. The fixing baseplate is within this length (97mm). THE AMPLIFIER IS NOT WATERPROOF IN THIS PACKAGING.

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The amplifier has a gain of 30dB and is unconditionally stable with any input or output load conditions.



## Input Match Smith Chart

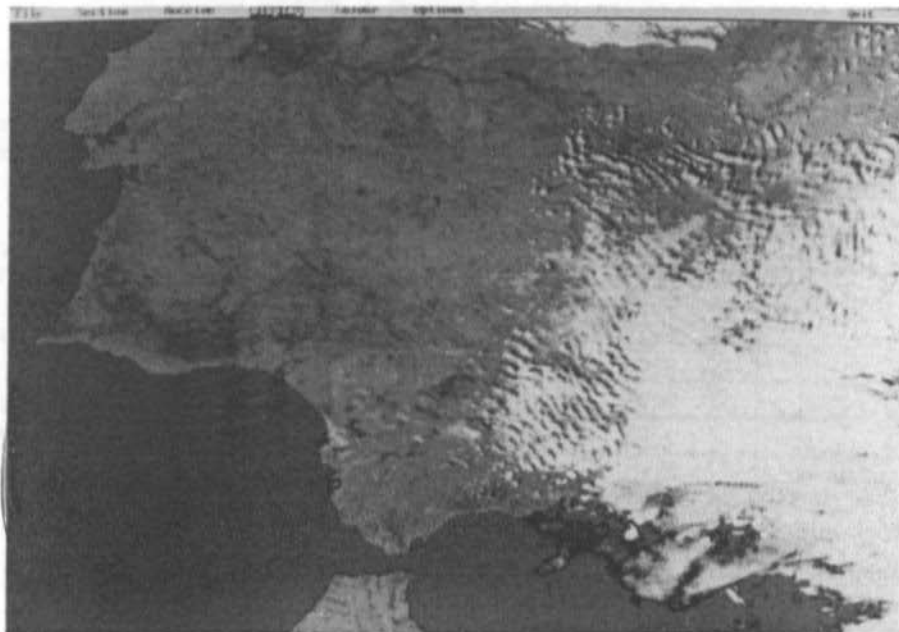
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# TIMESTEP



## HRPT System

Noise-free digital HRPT transmissions from NOAA, with a ground resolution of just 1.1km, allow images to be received in incredible clarity. Rivers, lakes, mountains, cities and even small towns can be seen on good days. Fishermen will appreciate the increased resolution of sea surface temperatures.

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Zoom to greater than pixel level is available from both a mouse-driven zoom box or using a roaming zoom that allows real time dynamic panning.

Sections of the image may be saved and converted to GIF images for easy exchange.

Latitude and longitude gridding combined with a mouse pointer readout of temperature will be available late in 1991.

Tracking the satellite is easy and fun! Manual tracking is very simple as the pass is about 15 minutes long. A tracking system is under development and expected by the end of 1991. A 4-foot dish and good pre-amplifier are recommended. The Timestep Receiver is self-contained in an external case and features multi-channel operation and a moving-coil S meter for precise signal strength measurement and tracking. The data card is a Timestep design made under licence from John DuBois and Ed Murashie.

Complete systems are available, call or write for a colour brochure.

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## VGASAT IV & MegaNOAA APT Systems

### 1024 x 768 x 256 Resolution and 3D

The Timestep Satellite System can receive images from Meteosat, GOES, GMS, NOAA, Meteor, Okean and Feng Yun. Using an IBM PC-compatible computer enables the display of up to 1024 pixels, 768 lines and 256 simultaneous colours or grey shades depending on the graphic card fitted. We actively support nearly all known VGA and SVGA cards. Extensive image processing includes realistic 3D projection.

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### Equipment

#### Meteosat/Goets

- 1.0M dish antenna (UK only)  Yagi antenna
- Pre-amplifier  20M microwave cable
- Meteosat/GOES receiver
- VGASAT IV capture card
- Capture card/receiver cable
- Dish feed (coffee tin type)

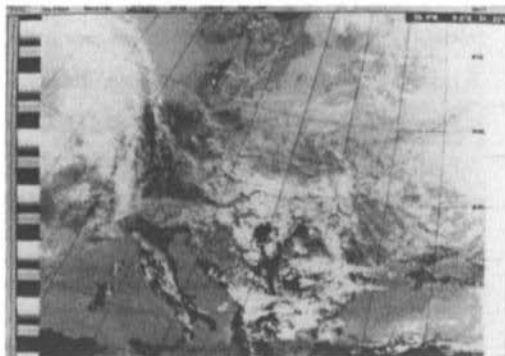
#### Polar/NOAA

- Crossed dipole antenna
- Quadrifilar Helix antenna (late 1991)  Pre-amplifier
- 2 channel NOAA receiver  PROscan receiver
- Capture card/receiver cable

*Call or write for further information.*

USA Education Dealer: Fisher Scientific, Educational Materials Division, 4901 W. LeMoyné Street, Chicago, IL 60651.  
Tel: 1-800-621-4769

USA Amateur Dealer: Spectrum International, P.O. Box 1084, Concord, Massachusetts 01742. Tel: 508 263 2145



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**MATERIAL PRICE LIST OF EQUIPMENT**

described in VHF COMMUNICATIONS

DJ8ES	FM ATV Receiver for the 23cm Band	Art.No.	Ed. 1/1991
<b>Converter</b>			
PCB	DJ8ES 001 double-sided	6347	DM 22.00
Components	DJ8ES 001	6348	DM 179.00
Kit	DJ8ES 001 complete	6349	DM 194.00
<b>Digital Frequency Indicator</b>			
PCB	DJ8ES 002 double-sided	6350	DM 19.80
Components	DJ8ES 002	6351	DM 94.00
Kit	DJ8ES 002 complete	6352	DM 110.00
<b>IF Amplifier</b>			
PCB	DJ8ES 003 double-sided	6353	DM 15.00
Components	DJ8ES 003	6354	DM 105.00
Kit	DJ8ES 003 complete	6355	DM 115.00
<b>Demodulator</b>			
PCB	DJ8ES 004 double-sided	6356	DM 19.00
Components	DJ8ES 004	6357	DM 136.00
Kit	DJ8ES 004 complete	6362	DM 150.00
<b>DL6MDA AD/DA Card</b>			
PCB	DL6MDA 001 though-cont. single	6366	DM 38.00
Kit	DL6MDA 001 complete	6367	DM 110.00
Floppy disk with .EXE programme on 3.5" floppy		6368	DM 20.00
<b>PUFF - A CAD-software for Microwave Stripline Circuits</b>			
PUFF incl. Handbook	1 floppy (360kB)	3407	DM 28.00
S-Parameter MOTOROLA	2 floppies (360kB)	3408	DM 20.00
S-Parameter SIEMENS	1 floppy (1.2MB)	3409	DM 15.00
S-Parameter PHILIPS	2 3.5" floppy	3411	DM 30.00

**Post and packing minimum charges DM 6.50**

The above items are all supplied by, and obtained from, UKW-Berichte in Germany.

To obtain supplies please contact your country representative for details of local prices and availability. Alternatively, you may order direct from UKW-Berichte or via KM Publications, whose addresses may be found on the inside front cover of this magazine.



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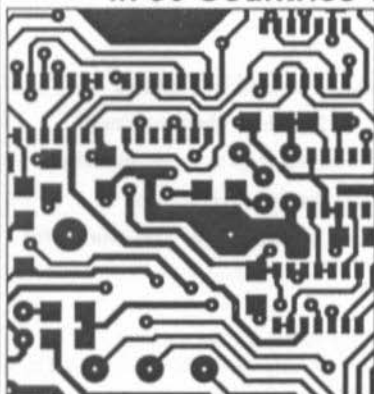


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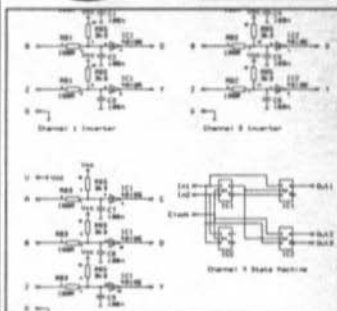
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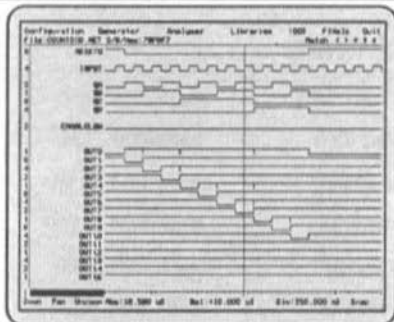
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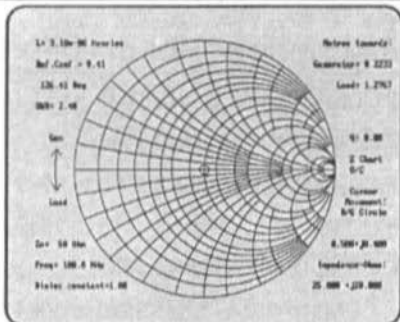
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