

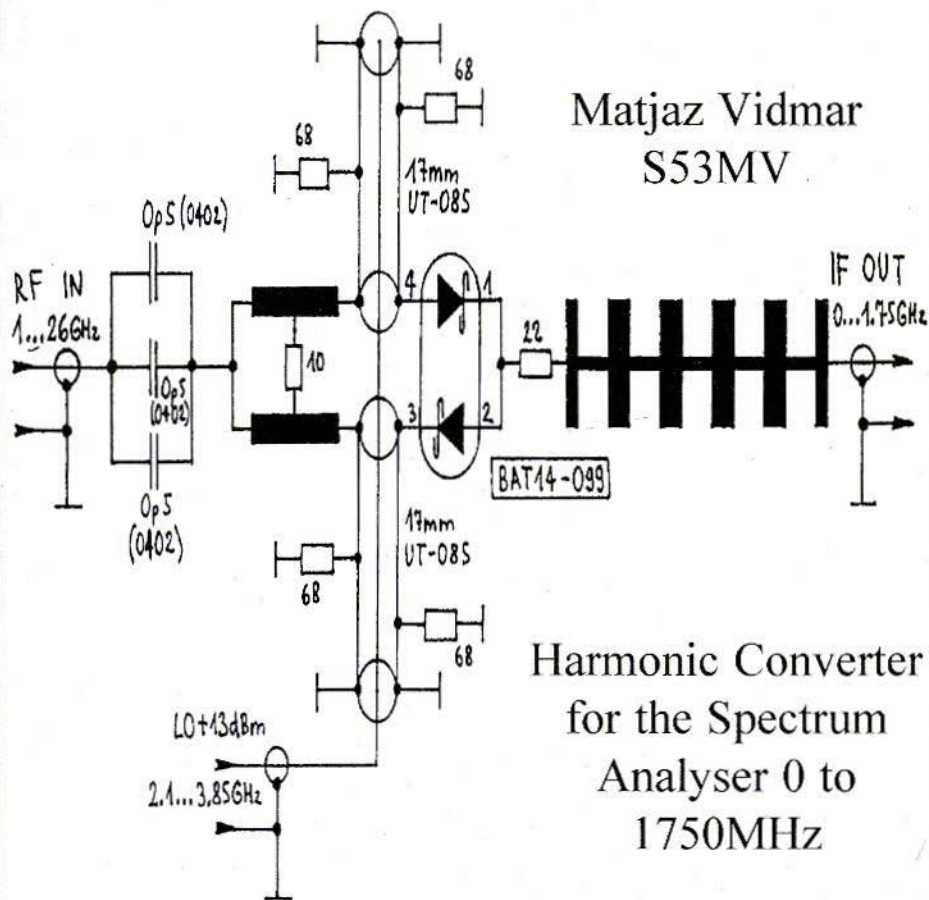


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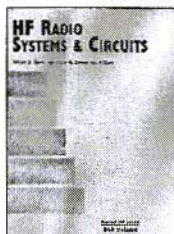
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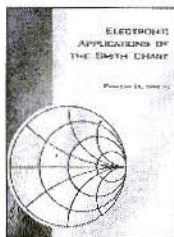
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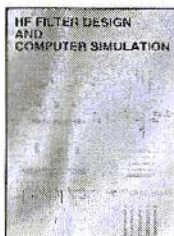
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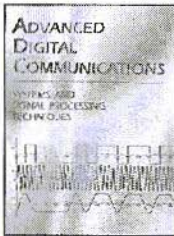
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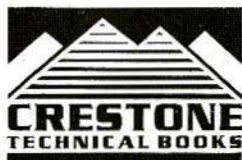
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There is also a final special offer 'Closing Sale' of back issues of VHF Communications on page-164.



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Ron Purkey, W9NUP

Another RF Combiner/Splitter

Combine or split any number of modules with minimal loss and great port isolation.

The use of combiners to make a higher power amp from several lower power modules is nothing new. What is new, or at least a different approach to an old design, is a combiner that will accept odd or even number of modules with minimum power loss and fault protection.

1.

History

Wanting a little more power, and having a free broadcast pull, I built an 8877 tubed kilowatt amplifier for 2 meters. Since my HF amp uses eight MRF-422 transistors and four quiet muffin fans, I had forgotten about the noise from a blower adequate to cool a tube amplifier. The shack is an 80 square foot building and the blower noise was irritating at best.

The HF amplifier uses 28 volt transistors and as the power supply was already made, the choice of suitable transistors was limited.

Not wishing to re-invent the wheel, I looked for a ready to go unit, unfortunately I could find nothing on the market using 28 volts at the kilowatt level.

Checking Motorola's web site revealed they are recommending the MRF-141G for higher power at 28 volts. Rated at 300 watts, these Gemini devices are not without fault; the chief one being a lot of power dissipated in a small footprint. Application note AR-313 describes the construction of a 10 MHz to 175 MHz broadband amplifier.

This unit is also sold by Communications Concepts [1] in kit form. Four to six of these modules combined would fill the bill. The question was how best to combine them without putting a high priced transistor at risk. Also, would it be possible to make this an "add a module as finances permitted" project without losing power in a dummy load by employing an odd number of modules.

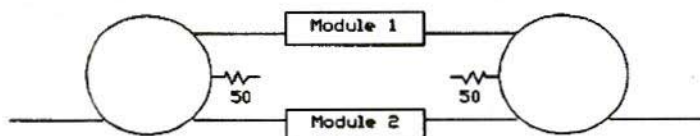


Fig.1:
Hybrid Ring

2.

A Study In Frustration

The most common form for safely combining power amps or modules at VHF and above is the 1-1/2 wave hybrid ring as shown in Fig 1. Theory and construction has been well documented previously [2]. Two rings are required, one to split the input power and the other to combine the outputs of the modules.

Claimed port-to-port isolation approaches 30dB with negligible loss of power. If one module is goes faulty, half of the remaining modules power is fed to the output and the other to the dummy load. The operating amp still sees 50Ω. So far so good.

Figure 2 illustrates the requirements for four amplifiers. Adding two more modules, however, has increased the number of rings from two to six. In addition, if three modules operating at full power

would fulfil the requirement, the fourth would be necessary as one half of the third modules output would be dumped to the load resistor.

Several "authorities" I consulted were clueless. I was desperate.

3.

The "N-Way" Combiner

Searching the web for ideas, I found Myat Engineering [3]. Among other things, they manufacture a high power multi-port combiner for FM broadcast named the E*Star (TM). Comprised entirely of 1/4 wavelength sections of transmission line and an appropriate number of loads, this combiner offers outstanding port-to-port isolation and the ability to use an odd or even number of modules without sacrificing performance.

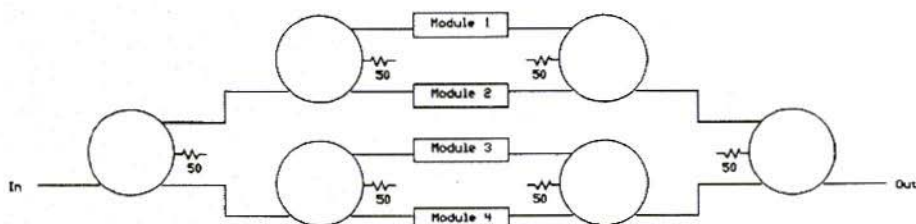


Fig.2: Four Amplifier Combiner

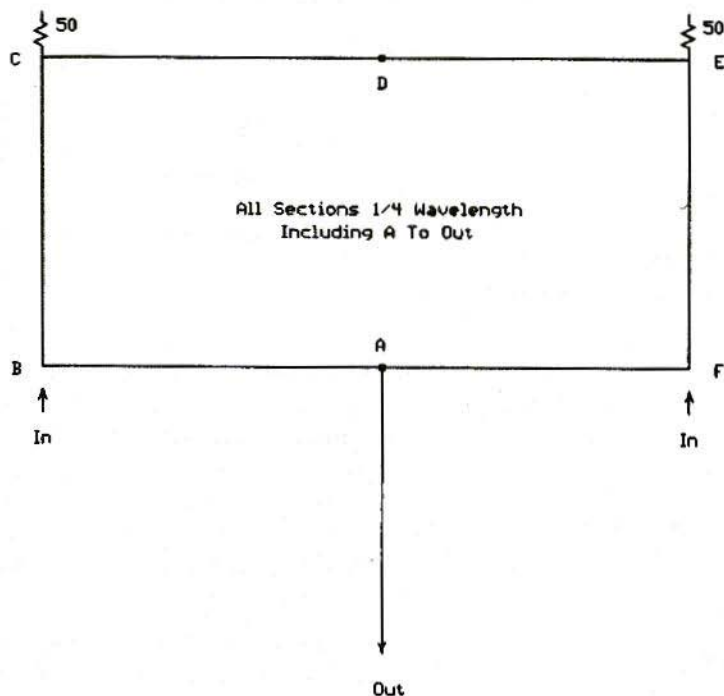


Fig.3: Two Module Combiner

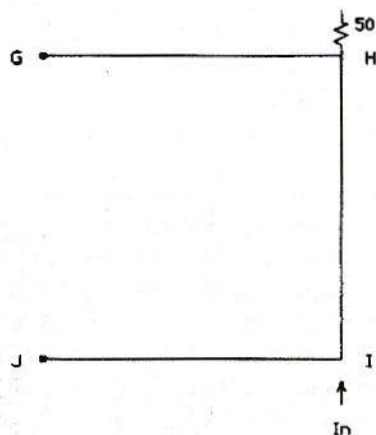


Fig.4: Adding an Additional Port

Figure 3 illustrates the concept based on two modules. Like the hybrid ring, port-to-port isolation is accomplished by RF arriving 180° out of phase from the opposite port. Here again, faulting one port allows continued operation at half power.

Unique to this design is the ability to add a third, fourth, fifth or sixth port without significant loss of power.

One additional port is shown in Fig.4. Connections G and J are tied to D and A (Fig.3) respectively to add the third port. Duplicating Fig.4 again would allow the fourth port to be added and so on. The quarter wave section from A to Out is used for matching to a 50Ω load.

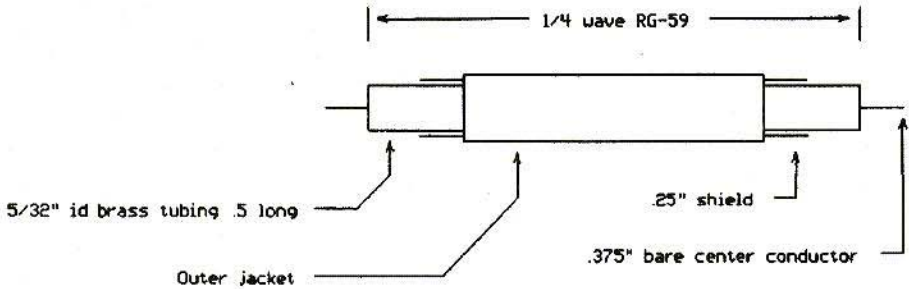


Fig.5: 1/4-wave Stub Construction Details

The impedance of this section is determined from the formula $(A \cdot 50)$ where $A = 50/\#$ of ports.

For example:

In a two port device, $A=25$

$$\sqrt{(25 \cdot 50)} = 35.36\Omega$$

A three port would be $A=16.67$

$$\sqrt{(16.67 \cdot 50)} = 28.87\Omega$$

Four ports = 25Ω

Five ports = 22.36Ω and

Six ports = 20.41Ω

the shields are connected close to the connectors. The output matching section is constructed in the same fashion. Two parallel connected 75Ω lines (37.5Ω) seem to work well with two ports. Similarly, a 50 and 75Ω line yield 30Ω for a three port and a pair of 50Ω cables equal 25Ω for a four port.

Closer impedances can be obtained by constructing the matching line from copper pipe at the expense of making the unit larger.

4. Construction

RG-59 coax is used throughout with the exception of the output matching line. Each 1/4 wave section is made as shown in Fig.5. I used solid dielectric cable and found 1/4 wave at 144.2 MHz equalled 13.5 inches. I chose to make my combiner rack mountable as shown in the photos.

Any layout should work well as long as

5. Bench Test

When construction is complete, terminate the output and all input ports except two inputs into 50Ω low power loads. Apply a signal generator to one input and a receiver to the other. Measure isolation between all ports in this fashion. I measured about 38dB with a Cushman CE-3 and an FM receiver set at critical squelch. A quick check with the engineers at Myatt confirmed this is a correct number for their units. Lacking lab equipment, I



cannot state mine is an absolute obtainable number, but it is reasonably close.

Next, terminate the inputs with dummy loads of appropriate power and apply power to the output port. The power should be evenly divided at the input ports with $< 2\%$ loss. Once you have obtained good port to port isolation and equal power division, the project is complete.

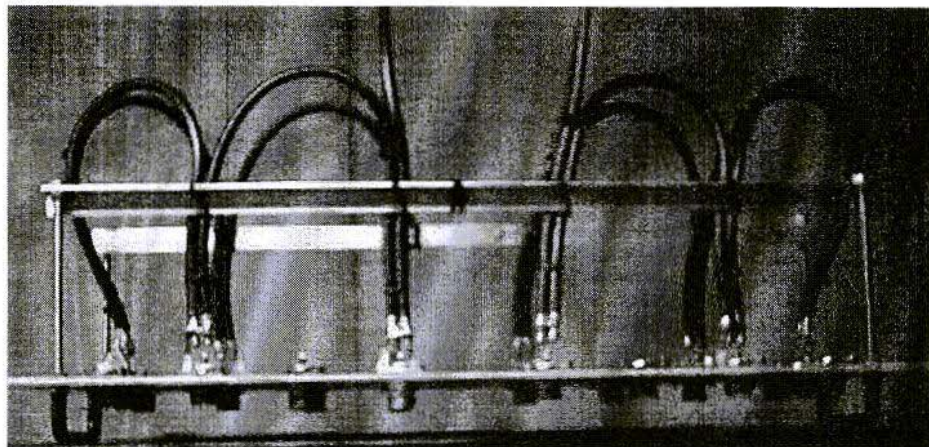
6. Conclusion

This approach to combining lends itself well to easy expansion. By simply adding three quarter wave sections and changing the output matching line impedance, additional modules can be added as finances permit. Remember, as with any combiner, the coax from the modules to the combiner inputs **MUST** be of equal length.

Also, if a module is pulled from service, the input port must be terminated in 50Ω . That's it. Enjoy being able to run an odd or even number of modules without wasting power or worrying about frying expensive transistors should a module fail. If you are unfamiliar with combiners, I strongly urge you to visit Myatt's web site. They have an excellent tutorial on the subject.

7. References

- [1] Communications Concepts, 508 Millstone Dr, Beavercreek, OH 45434-5480
- [2] HY-brid HI-power, Tom Pettis, QEX, January 1990
- [3] Myat Engineering, Norwood, NJ 07648-0425 (www.myat.com)



A Completed Combiner

Matjaz Vidmar, S53MV

Harmonic Converter for Spectrum Analyser 100 kHz to 1750 MHz

1. EXTENDING THE FRE- QUENCY COVERAGE OF SPECTRUM ANALYZERS

RF spectrum analysers are usually designed as scanning receivers with a high first intermediate frequency. If the first IF is set above the input frequency range, unwanted mixing products can be removed with a lowpass filter. Similar designs are therefore used also in other test equipment, like signal generators.

A high first IF implies that the circuits inside the test equipment should operate at relatively high frequencies. However, there are several technological constraints limiting the first IF to below about 5 GHz. The most important constraint is probably the accumulation of phase noise of the oscillators used in the different up-conversions and down-conversions.

Practical spectrum analysers therefore have the first IF in the range between 2 GHz and 3 GHz. For operation above

3 GHz the input lowpass filter is replaced by a tuneable (YIG) bandpass filter. Even without any input filtering, a spectrum analyser is still able to provide useful information by carefully considering all mixing products with both the fundamental oscillator frequency and its harmonics.

For operation at very high frequencies (in the millimetre-wave range), commercial spectrum analysers are usually supplied with external (waveguide) harmonic mixers, as shown in Fig.1. Of course the spectrum analyser should provide both a LO output and an IF input to connect an external mixer. Some harmonic mixers may even require a bias adjustment to optimise the conversion efficiency at a particular harmonic.

In a home-made spectrum analyser one cannot probably afford an expensive YIG-tuned preselector. Although there may be differences in the conversion efficiencies of different mixers, harmonic mixing almost always provides useful results.

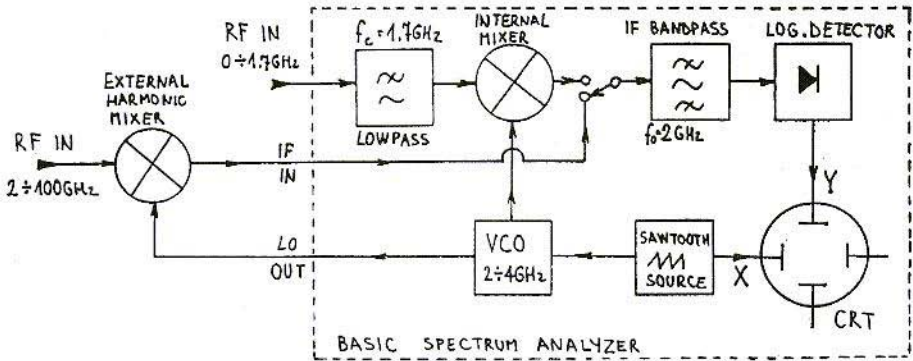


Fig.1: Extending the Spectrum Analyser coverage with a Harmonic Mixer

The only remaining problem is the identification of the different mixing products when a tuned preselector is not available.

A very efficient solution is an additional harmonic converter, including a harmonic mixer and an adjustable local oscillator, as shown in Fig.2. The adjustable conversion oscillator allows an additional degree of freedom to shift the harmonic-mixer responses. The latter allows identification

and convenient separation of the different mixer responses even without a tuned input preselector.

The spectrum Analyser same external (harmonic) converter can also be used to extend the frequency coverage of a tracking generator as shown in Fig.3. In fact the tracking generator itself is no longer needed, since the first LO output of the spectrum analyser is used as the tracking generator. The additional (harmonic) converter is then used to translate the 2...3.7 GHz band back to 0...1.7 GHz, where the basic spectrum analyser is receiving.

In the case of a typical spectrum analyser with a 2 GHz first IF this approach allows an additional coverage of 2...3.7 GHz besides the usual range of 0...1.7 GHz offered by conventional tracking generators. However, two constraints have to be met. First, the additional converter has to be tuned precisely to the spectrum analyser first IF. Second, sufficient filtering has to be provided so that the converter oscillator does not enter into the spectrum analyser IF. A separation buffer amplifier for the spectrum analyser LO signal is required for the same purpose.

A simple, tuneable harmonic converter for spectrum analysers will be presented in this article. Although this converter was originally designed for the spectrum analyser shown in [4] or [5], it will operate with any spectrum analyser having the first IF between 2GHz and 3GHz. The detailed block diagram of the harmonic converter is shown in Fig.4.

Besides a harmonic mixer the converter includes a tuneable oscillator covering approximately the same frequency range

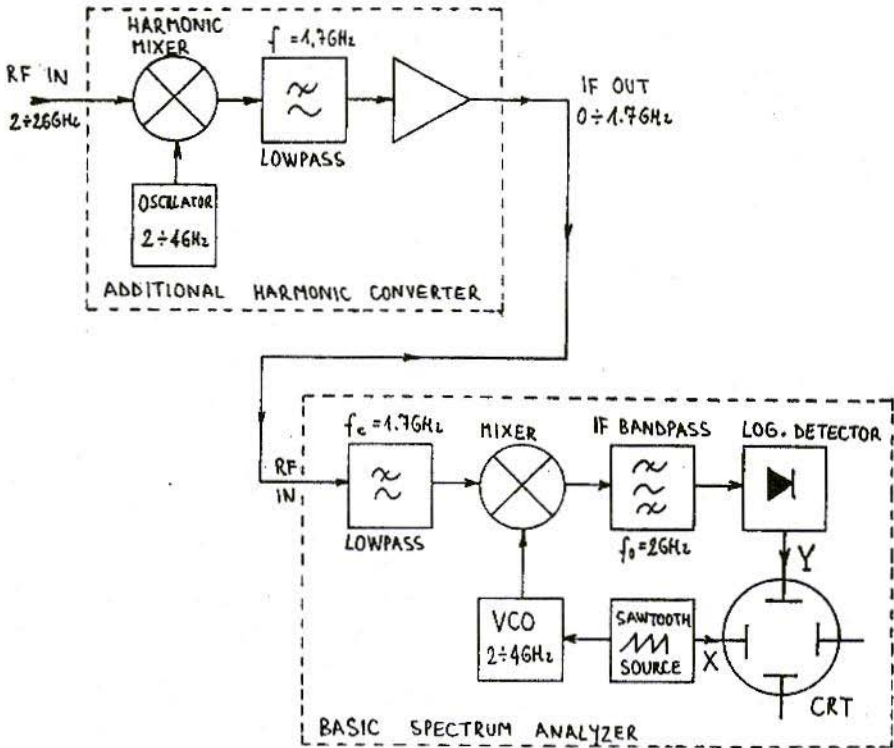


Fig.2: Extending the Analyser coverage with a Harmonic Converter.

as the VCO inside the spectrum analyser (2.1...3.85 GHz). The latter allows selecting the best mixing harmonic to obtain a spurious-free response and a good conversion efficiency at the same time. Of course the frequency coverage should include the spectrum Analyser first IF to allow the extension of the range of the tracking generator.

The (IF) output of the harmonic converter includes a wideband amplifier and several lowpass filters. The amplifier is required to compensate for the high conversion loss of the harmonic mixer as well as poor noise figure of the following spectrum analyser.

Lowpass filtering has to be performed in several stages to remove all unwanted mixing products as well as to prevent the harmonics of the conversion oscillator from entering the basic spectrum analyser.

Finally, the harmonic converter includes all required power supplies, since it is intended as a stand-alone unit. The VCO and output amplifier require a +8V DC supply obtained with a 7808 regulator. A DC/DC converter with an additional stabiliser is used for the VCO tuning voltage 0...30V.

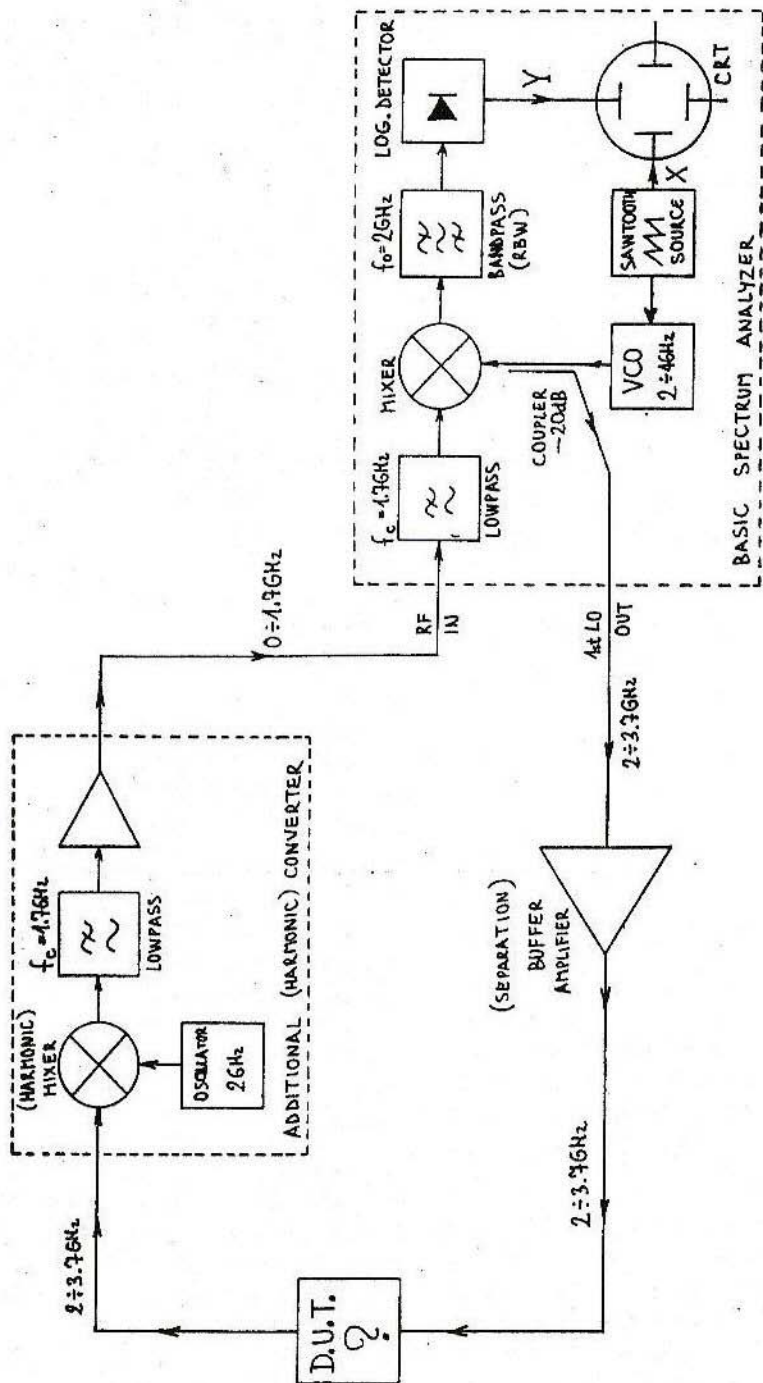


Fig.3: Extending the Tracking Generator coverage with a (Harmonic) Converter

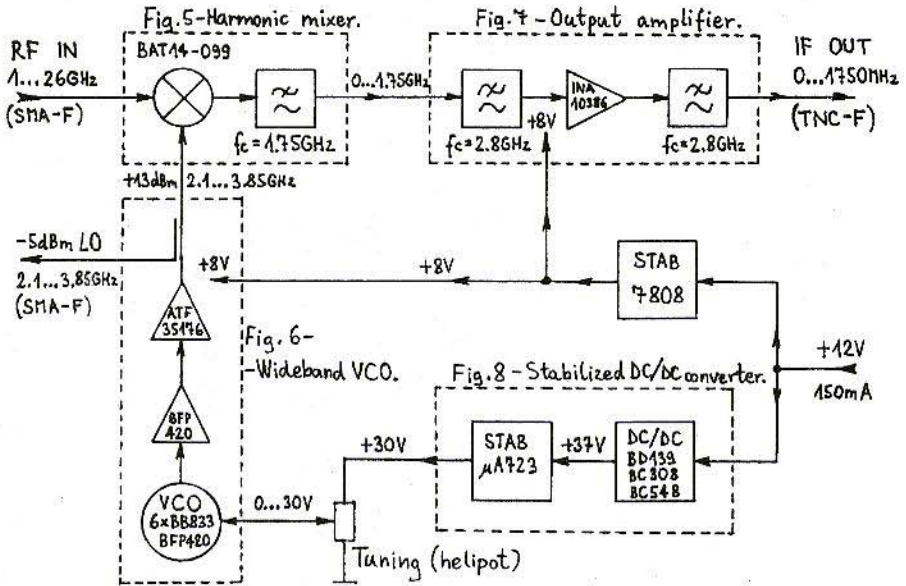


Fig. 4: Harmonic Converter detailed Block Diagram

2. HARMONIC MIXER

The circuit diagram of the harmonic mixer is shown in Fig. 5. In fact the harmonic mixer does not differ much from other mixers, like the first and second mixers in the spectrum analyser shown in [4] or [5]. The dual Schottky diode BAT14-099 produces useful harmonics up to at least 30 GHz while driven with a large (+13dBm or 20mW) LO signal in the frequency range 2.1...3.85 GHz.

Since a flat response across the whole microwave frequency range is desired, particular attention has to be made to suppress any parasitic resonances of the circuit. The input signal is therefore coupled using very small 0402 SMD

capacitors. Since the width of a single capacitor is 0.5mm, three such capacitors are installed in parallel to match the 1.5mm width of the 50Ω microstrip line.

Unwanted resonances of the balun made from UT-085 semi-rigid cable are suppressed by four 68Ω SMD resistors. The 10Ω and 22Ω resistors are also used to suppress circuit resonances. Finally, the whole shielded module requires a piece of microwave absorber (antistatic foam) under the cover to suppress cavity resonances.

The harmonic mixer does not include any attenuator on the input. Since it is difficult to build good attenuators for frequencies above 10 GHz with conventional SMD resistors, it is recommended that external SMA attenuators are used when required.

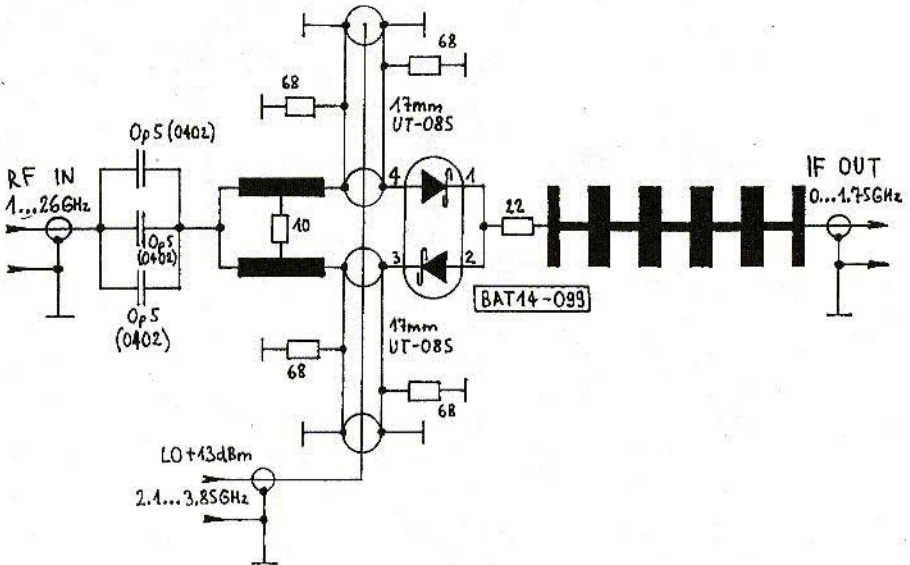


Fig.5: Harmonic Mixer

The SMA connectors themselves limit the frequency range of the harmonic mixer to about 26 GHz. Although there is no input filter, waveguide transitions were found to be excellent high-pass filters when required.

The main output of the VCO (+13dBm) drives the harmonic mixer directly. The auxiliary output (about -5dBm) is made available on the back panel. The latter may be used to check the conversion frequency with a counter or as a 2...3.85 GHz signal source for other purposes.

3. WIDEBAND VCO

The design of a suitable wideband VCO was already shown in [1], [2] or [3]. The circuit diagram in Fig.6 has just a small modification. A 10uF capacitor is added to the tuning voltage line for additional filtering, since inside the harmonic converter the VCO frequency is only set manually with a precision Helipot potentiometer.

For operation with the spectrum analyser described in [4] or [5] or other spectrum analysers with the first IF around 2 GHz, BB833 varactors should be used in the VCO to allow a frequency coverage from about 2 GHz to about 3.85 GHz.

For spectrum analysers with the first IF above 2.5 GHz, BB857 varactors are recommended (VCO range 2.4 GHz to 4.6 GHz).



4. OUTPUT AMPLIFIER

The circuit diagram of the output amplifier is shown in Fig.7. The amplifier is designed around an INA10386 MMIC that provides a flat gain of more than 20dB and a noise figure better than 4dB across the whole frequency range up to 1.75 GHz. The INA10386 provides just the correct amount of gain to compensate for the mixer conversion loss and Spectrum Analyser noise figure.

The output amplifier is necessarily installed in its own shielded enclosure, since unwanted signal filtering is very critical. Besides the lowpass filter inside the harmonic mixer module (cutoff about 1.75 GHz), two additional lowpass filters with cutoff frequencies of about 2.8 GHz are installed before and after the INA10386 amplifier. To suppress higher-order responses of the microstrip circuits as well as enclosure cavity resonances, microwave absorber foam is also needed under the cover of the amplifier module.

The INA10386 MMIC requires DC decoupling capacitors on both the input and output. Since the harmonic converter is usually used at relatively wide spectrum Analyser resolution bandwidths (100 kHz or more), the IF range can be limited to 100 kHz on the lower end. Even in the latter case finding suitable coupling capacitors is not simple. The best results were found with a parallel connection of a low-loss, 100pF NP0 0805 size capacitor that is soldered on the PCB first. Then a lossy 100nF (Z5U ceramic) and larger (1206) capacitor is soldered across the smaller 100pF capacitor.

5. POWER SUPPLY

Like the spectrum analyser described in [4] or [5] or the companion tracking generator described in [6], the harmonic converter requires an unstabilised +12Vdc supply. The latter is stabilised to +8V with a 7808 regulator to supply

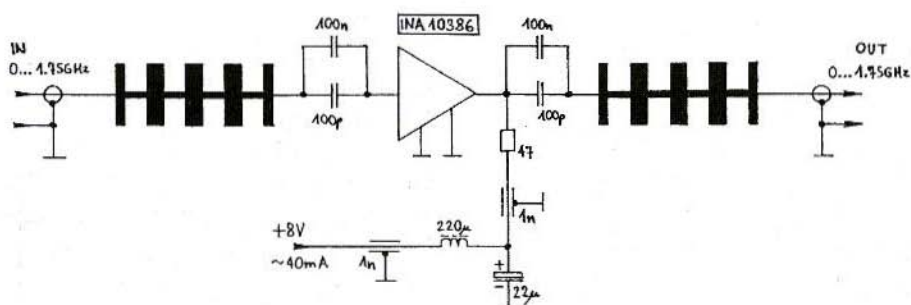


Fig.7: Output Amplifier

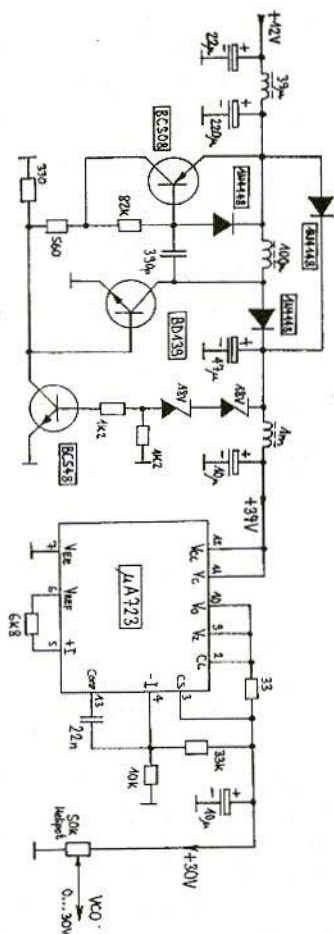


Fig.8:
Stabilised DC/DC Converter

the VCO and output amplifier. The 7808 regulator is wired exactly in the Spectrum Analyser way as in the spectrum analyser or tracking generator. The harmonic converter usually remains on all of the time to minimise frequency drifts, so a turn-on switch is usually not required.

The BB833 (or BB857) varactors inside the wideband VCO require much higher

tuning voltages up to +30V. Further, the tuning voltage has to be well stabilised and filtered, to avoid frequency drifts and excessive phase noise. Both unwanted effects are further multiplied by the order of the harmonic inside a harmonic converter!

The tuning voltage supply is shown in Fig.8. The circuit includes a flyback DC/DC converter with the transistors BC308 and BD139. The rough voltage regulation provided by the two 18V Zener diodes and BC548 feedback transistor is far from being sufficient for a tuning voltage. The resulting +37V is further stabilised down to +30V with a uA723 precision regulator. Due to the low current drain at +30V, no external transistors are required around the uA723. However, the uA723 still requires some feedback resistors to set the output voltage, overcurrent protection and a capacitor for frequency compensation.

The actual VCO tuning voltage is obtained through a precision 50kΩ Helipot potentiometer from the well-regulated +30V supply. The potentiometer is installed on the front panel and is equipped with a revolution-counting dial knob.

Since there is no tuning-voltage lineariser, the frequency scale is quite non-linear. The latter may even be an advantage in a harmonic converter, where fine frequency adjustments are usually required at the upper end of the frequency coverage while using high-order harmonic mixing at input frequencies above 10 GHz.



6. CONSTRUCTION TIPS

With the exception of the DC/DC converter, the harmonic converter only includes microstrip circuit boards. The harmonic mixer and output amplifier boards are shown in Fig.9 while the wideband VCO has already been described in [1], [2] or [3]. All microstrip boards are etched on 0.8mm thick, double-sided FR4 glass fibre-epoxy laminate. The DC/DC converter is built on a single-sided board as shown in Fig.10, etched on 1.6mm thick FR4 laminate.

All three microstrip boards are installed in shielding boxes made from 0.5mm thick brass sheet just like the modules of the spectrum analyser [4] or [5] or companion tracking generator [6]. All

three brass boxes require a microwave absorber (antistatic foam) under their covers to suppress cavity resonances.

The harmonic converter module location is shown in Fig.11. The harmonic converter has the same depth (240mm) as the spectrum analyser [4] or [5]. The width is 100mm while the height is only 32mm, since all modules are located in a single plane. The bottom of the box is simply a piece of 1mm thick aluminium sheet, bent in the form of an "U". The cover is a similar "U" made from 0.6mm thick aluminium sheet.

The 7808 regulator is bolted to the back plate for heatsinking purposes. The two 470uF electrolytic capacitors, VK200 RF choke and 1N5401 diode are simply soldered between the +12V supply connector and the leads of the 7808 regulator. The overall current drain amounts to about 150mA.

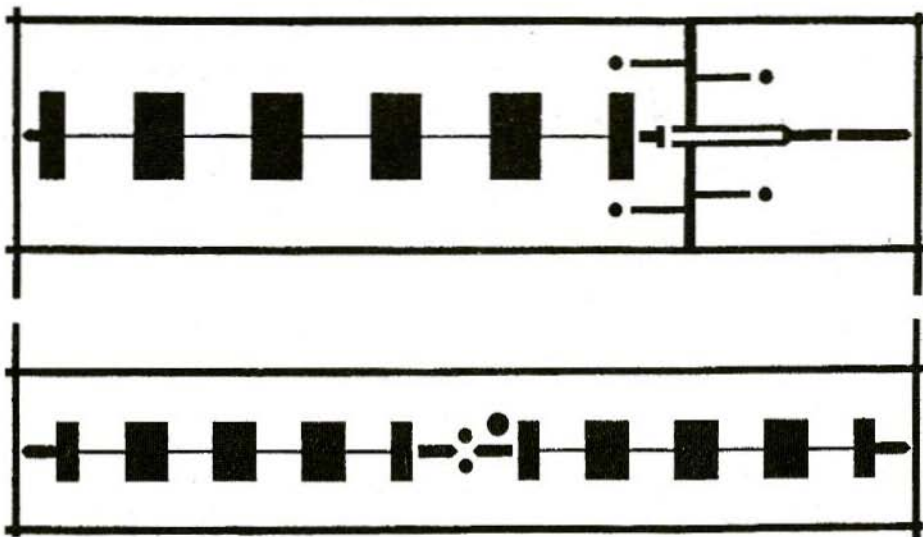


Fig.9: Microstrip Printed Circuit Boards
 Upper: Harmonic Mixer; 30 x 120mm
 Lower: Output Amplifier; 20 x 120mm

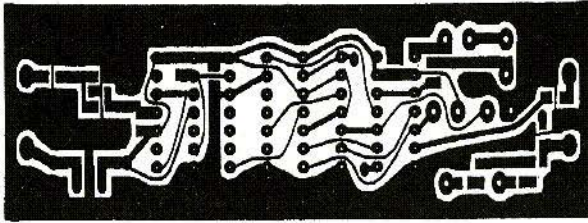


Fig.10: DC/DC Converter PCB; 30 x 30mm

7. ALIGNMENT AND CHECK- OUT

A harmonic converter is a much simpler piece of test equipment when compared to a spectrum analyser or companion tracking generator. The described harmonic converter has no tuning points. Of course, the correct operation of the power supply including the DC/DC converter should be checked first.

The most demanding module is certainly the wideband VCO. The latter should be checked for the frequency coverage as well as output-signal level. The latter should not drop below +10dBm on any frequency. Of course, if the spectrum analyser published in [4] or [5] is being built at the same time, it makes sense to install the better VCO inside the spec-

trum analyser and use the remaining VCO for the harmonic converter.

The spectrum analyser should be sensitive enough to display the noise of the harmonic-converter output amplifier even with an input attenuator setting of 10dB. This corresponds to a spectrum-analyser noise figure of about 20dB (without the input attenuator) that is a typical value for most spectrum analysers.

Finally the first LO output from the spectrum analyser is connected to the harmonic-converter. Additional attenuators may be required to bring the input signal level of the latter to about -10dBm. When the harmonic converter is tuned to the exact first IF of the spectrum analyser, the set-up can be used as a tracking generator.

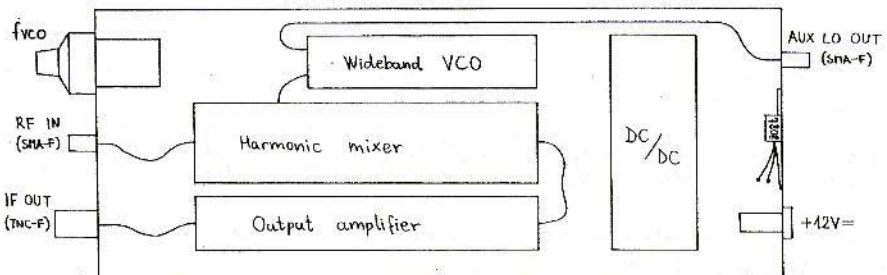


Fig.11: Harmonic Converter Module Location.



8.

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Dr. Ing. Jochen Jirmann, DB1NV

High-Frequency Measuring Instruments, Basic Circuits and Applications

Part-1

High-frequency technical measurements differ in many ways from conventional measurement procedures in DC or low-frequency engineering.

Anyone wishing to evaluate high-frequency measurements correctly needs some basic knowledge on the functioning of this kind of equipment, which the present article is intended to impart.

The article concentrates on the practical aspects. There is no attempt to go more deeply into theoretical matters such as, for example, wave propagation, Fourier transformation or S-parameters.

1. DIFFERENCES WITH CONVENTIONAL METROLOGY

The main source of errors in DC and low-frequency metrology is usually the resistive load on the test object. For example, the internal resistance of a standard digital multimeter in the voltage range is $10\text{M}\Omega$ and the apparatus loses 0.2 to 1V in the current range.

The main source of errors in the high-frequency and microwave ranges is provided by the input reactive impedances, which not only falsify the measurements but often also render the circuit incapable of functioning. Thus a normal 10:1 divider head on an oscillo-



scope has an input resistance of $10M\Omega$ parallel to $10pF$. At 100 MHz the direct impedance will be $-j160\Omega$ (capacitive) and at 1 GHz $-j16\Omega$!

Even an additional load of $10pF$ can change an amplifier frequency response or make an oscillator incapable of oscillating.

Line inductances are just as critical. An earth line 1 cm. long has an inductance of $10nH$, which at 100 MHz corresponds to a reactive impedance of $j6.3\Omega$, and at 1 GHz to $j63\Omega$.

Thus, using normal probe metrology is pointless at frequencies exceeding 100 MHz. There are two possible remedies:

1. Part of the measurement signal processing (impedance conversion or mixing) is transferred into an active probe. This is to force the input capacitance down to less than $0.5 pF$.
2. Measurements in the circuit are dispensed with, and the inputs and outputs of the circuit are laid out so that the coaxial cables can be connected with the correct impedance at 50Ω (general high-frequency technology) or 75Ω (telecommunications technology).

The measurement equipment used will then have suitable input and output resistances.

An important tip:

While normal multimeters or oscilloscopes are definitely overload-proof, so that the network voltage itself scarcely causes any damage at the input, both active probes and measurement equipment with 50Ω inputs are put at risk if the signal level is too high. DC voltages of 20V and high-frequency power levels of 100mW are normally permissible.

The precise values can be found next to a warning symbol on the front plate near the input socket. A few appliances, for example standard signal generators, have a reverse power protection system which protects at power levels of up to 50W. This is especially helpful in the testing of radio equipment, since an accidental pressure on the transmission key can otherwise destroy the measuring equipment.

In high-frequency and microwave technology, the accuracy of measurement is several orders of magnitude less than for low-frequency measurements, if we disregard frequency measurements. While digital multimeters in the medium price range can nowadays attain resolution levels of 5 to 6 places and precision levels of 0.005 %, an accuracy of measurement of 0.1dB or % is actually considered as a precision measurement for high-frequency circuits or amplifications.

The present article deals with the following measurement instruments and measuring procedures:

Voltage:	Diode voltmeters, sampling voltmeters
Current:	Current probes
Power:	Diode power gauges, thermo-electrical wattmeters, thermistor wattmeters
Frequency:	Microwave meters with frequency converters or transfer oscillators
Signal: sources	Standard signal generators, wobble transmitters

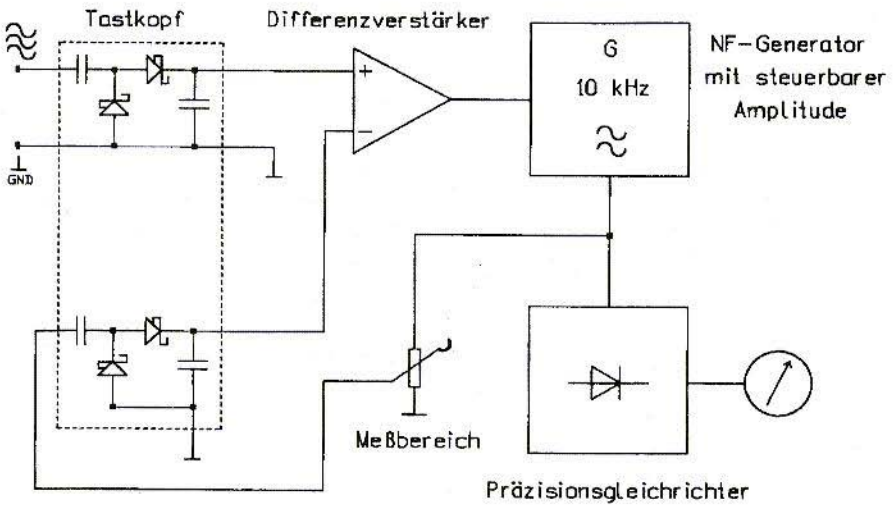
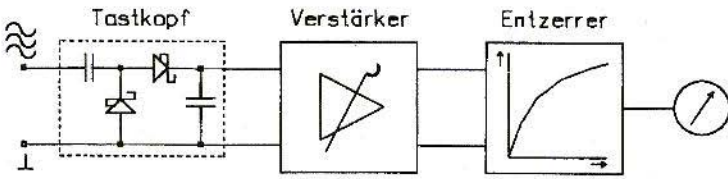


Fig.1: High-Frequency Voltage Measurement with Diodes (*a=upper; b=lower*)

Tastkopf = Probe, *Verstärker* = Amplifier, *Entzerrer* = Equaliser,
Differenzverstärker = Differential amplifier, *NF-Generator mit steuerbarer Amplitude* = Low-frequency generator with controllable amplitude, *Meßbereich* = Measurement range, *Präzisionsgleichrichter* = Precision rectifier

Spectrum : Selective voltmeters, audio analysers, high-frequency spectrum analysers

Noise: Noise sources, automatic noise factor gauges

Network: Bridge circuits, vector voltmeters, network analysers

Component: Automatic R-L-C test measurement bridge



2. HIGH-FREQUENCY VOLTAGE MEASURE- MENTS

At frequencies below about 10 MHz, high-frequency voltages can be measured with "broad-band voltmeters", which operate in a similar way to normal multimeters. The measurement amplifier and the subsequent precision rectifier are laid out only for much higher signal frequencies. To determine the effective value correctly, thermal conversion processes are often used, which operate by heating a resistance through which the amplified measurement signal is passed.

Special high-frequency voltmeters (typical frequency range up to 1 GHz) operate either using diode rectifiers or with sampling mixers, which are actually positioned within the instrument probe.

2.1 High-Frequency Voltmeter with Diode Rectification

A semi-conductor diode has an exponential characteristic curve which also makes it possible to rectify AC voltages which are considerably smaller than the "threshold voltage". Schottky or germanium diodes specially "bred" for this can rectify even AC voltages below 1mV. Of course, two effects should be taken into account: firstly, the diode is affected by temperature, and secondly, the series expansion of the exponential characteristic curve shows that the DC voltage generated at low high-frequency voltages (below app. 50mV) is propor-

tional to the high-frequency power, while at high levels of high-frequency voltage (exceeding 0.2V) it is proportional to the high-frequency voltage. This is why a characteristic curve equalisation built into the display section generates a display with a linear curve. To balance out the temperature, a calibrating generator with a known voltage is often present, and is connected to the measuring head before the measurement. Fig. 1a shows a typical circuit. The disadvantage is that the measuring heads can not be exchanged without calibration.

The circuit shown in Fig. 1b circumvents this problem in an elegant way. The measuring head contains two similar rectifiers. One is connected to the measurement voltage, the other to an internally generated auxiliary voltage, usually 10 kHz. A control circuit regulates the amplitude of the auxiliary voltage in such a way that the DC voltages of the two rectifier branches are identical. Then the auxiliary voltage is as high as the unknown high-frequency voltage. The auxiliary voltage can be measured using a normal precision rectifier and displayed. The trick is that the rectification ratio (the ratio of the high-frequency voltage to the DC voltage) of the rectifier is not factored into the measurement as long as both rectifiers have the same specifications. But temperature and ageing always influence both rectifiers in the measurement head, so that the influence on the accuracy is slight.

The best-known representatives of instruments of this type are the URV's from Rohde and Schwarz.

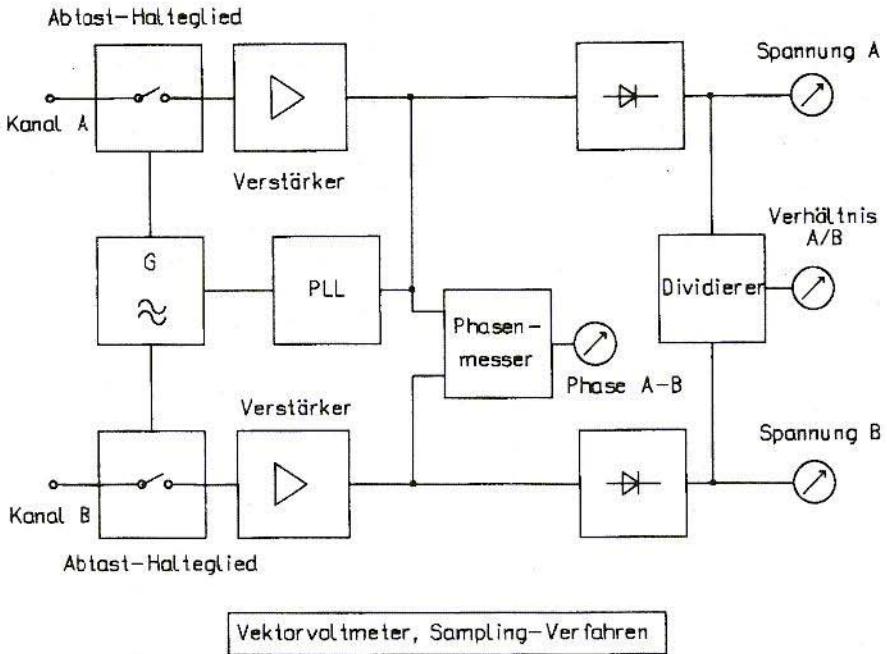


Fig.2a: Block Diagram of a Sampling Voltmeter

Abtast-Halteglied = Sampling gate, *Kanal* = Channel, *Spannung* = Voltage, *Verstärker* = Amplifier, *Verhältnis* = Ratio, *Dividierer* = Divider, *Phasemesser* = Phase meter, *Vektorvoltmeter, Sampling-Verfahren* = Vector volt meter, sampling procedure

2.2 Sampling Voltmeter

The sampling voltmeter uses a less well-known aspect of the scanning theorem. If a periodic, band-limited signal is scanned down, and if the sampling frequency is therefore less than double the highest signal frequency, the original signal can certainly no longer be reconstructed. However, reconstruction does provide a signal with a spectrum which certainly goes up to only half the sampling frequency, but the statistical properties of which, such as the curve

form, the effective value and the peak value, correspond to the original signal. Moreover, the phase position between two synchronously scanned input signals also remains the same.

Not only is this technique used in sampling oscilloscopes to represent procedures at extremely high frequencies, but it also reappears in high-frequency voltage meters. Dual-channel equipment (vector voltmeters, vector analysers) can measure, not only the voltages, but also the voltage ratios and the phase position.

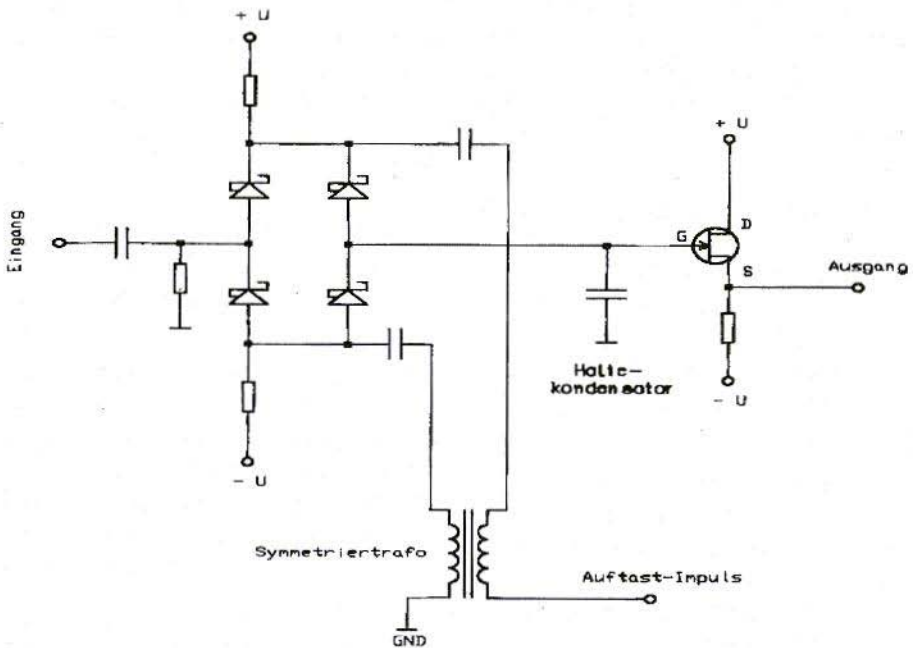


Fig.2b: Sampling Gate

Eingang = Input, *Ausgang* = Output, *Haltekondensator* = Holding capacitor, *Symmetriertrafo* = Balancing transformer, *Auftast-Impuls* = Trigger pulse, *Mit Diodenbrücke* = With diode bridge

We shall come back to the vector analyser again, in connection with the network measurements.

The core of a sampling voltmeter is a high-resistance, extremely rapid sampling gate, which is usually directly incorporated into the probe. The scanning procedure can also be interpreted as harmonic mixing, by means of which the signal frequencies (in the range around 100 kHz) are converted into an intermediate frequency position which is easy to process. The sampling gate usually takes the form of a diode bridge, with Schottky diodes as switches (the response time being tens of picosec-

onds). A high-resistance FET source tracker takes the voltage from the holding capacitor (in the order of 1pF). The extremely short scanning pulse (length: 10 to 100psec) is generated with a step recovery diode, which uses the idle/recovery time of a p-n junction. The scanning frequency can vary from a few kHz to around 100 MHz, depending on the equipment cost.

The remaining signal processing is carried out in the intermediate frequency. Two precision rectifiers measure the voltage values, a phase meter determines the phase position of the voltages, a divider calculates the voltage ratio,

and analogue / digital converters digitalise the readings. The downstream computer can then display the measurement result, broken down by amount and phase or by real or imaginary sections.

Simple sampling voltmeters take broadband measurements of voltage (measurement band width, for example, 10 kHz to 1 GHz) and display the effective value. Since the scanning and the measurement signal are not synchronised, the procedure is known as incoherent sampling.

Two-channel sampling voltmeters normally use the signal in channel B to synchronise the measurement electronics. For this reason, only selective signal components are measured in channel A which are synchronous with the signal in channel B (coherent sampling). Thus interference some way away, such as noise or external voltages, is to some extent suppressed as far as the frequency is concerned (see Fig's.2a and 2b).

3. HIGH-FREQUENCY CURRENT MEASUREMENT

Direct current measurements can not be obtained in the high-frequency range by connecting up a meter, as with DC current or low-frequency current, since the case capacitance of the meter generates an external measurement error in the order of magnitude of 100pF.

High-frequency currents are therefore measured with current transformers, the best-known forms of which are the current probes from Tektronix. The body of the probes has a channel closed by a slide, into which the live wire is simply inserted. The current probes are primarily designed for use with oscilloscopes, but can equally well be used with high-frequency voltmeters or spectrum analysers.

We distinguish between two kinds of current probes. The simple version consists only of the current transformer, and can measure AC currents from a low-end frequency upwards. The better version also has a Hall generator in the magnetic circuit, which also captures DC fields.

These current probes also measure the DC fraction of the measurement current correctly. Depending on the format, high-frequency currents can be measured of up to and exceeding 50 MHz with special coaxial formats up to 1 GHz.

The current probe has a special pre-amplifier which can also be used to adjust the sensitivity (Fig.3).

Notes:

1. The pre-amplifier is usually calibrated for a load of 50Ω , so that a large measurement error arises when a high-resistance display unit is connected up.
A 50Ω adapted load at the meter input eliminates the problem.
2. The current conversion transformer core can be saturated by high levels of DC current so that measurement errors arise.



4. HIGH-FREQUENCY OUTPUT MEASUREMENT

In theory, high-frequency outputs could be measured using the current voltage method and the effective and reactive power levels could also be determined from the phase position. But since in the high-frequency range there is wave propagation on the circuits, the current and voltage amplitudes are dependent on location and thus do not provide any information. Thus absorption power meters are almost always used. These convert the high frequency fed in into heat through a standardised ohmic resistance (usually 50Ω or 75Ω).

We find three basic types in this group of instruments, the diode wattmeter, the thermocouple power meter and the thermistor power meter.

4.1 The Diode Wattmeter

As shown in Fig.4, the diode wattmeter consists of an ohmic resistance with a rectifier circuit in parallel. The display unit is calibrated directly in high-frequency power levels. This circuit is suitable for measuring power levels in the range above 1 Watt (transmitter outputs).

The curvature of the rectifier characteristic line is balanced out by giving the display instrument an appropriate scale. Of course, every measurement range then needs a scale of its own.

To measure lower power levels, the simple diode rectifier is replaced by one

of the high-frequency voltmeters described. This makes it possible to measure high-frequency power levels of less than 20nW (corresponding to 1mV in 50Ω). The upper limit of measurement can be pre-set only through the resistance rating and the electric strength of the rectifier. If necessary, the meter rectifier is connected to a tap of the load resistor. The disadvantage of this technology is that the voltage measurement error goes quadratically into the power level measured, and the accuracy of measurement is not particularly high.

It therefore makes more sense to measure the power level directly through the heating of the resistance.

4.2 The Thermocouple Power Meter

The measuring head of a thermocouple power meter consists of one or two thin-film resistors which are in direct heat contact with thin-film thermo-elements. The increase in temperature of the resistor measured by the thermo-elements, as against the ambient temperature, is proportional to the high-frequency power. Fig.5 shows the structure of a thermocouple measuring head.

Since the output voltages of the thin-film thermo-elements are extremely small DC voltages, the first stage of the display amplifier (usually a chopper amplifier) is inside the measuring head. The great advantage of the thermocouple principle of measurement is that the power meter can be calibrated using low-frequency power. The measurement range of thermocouple measuring heads is from 1W to 100mW . Higher power

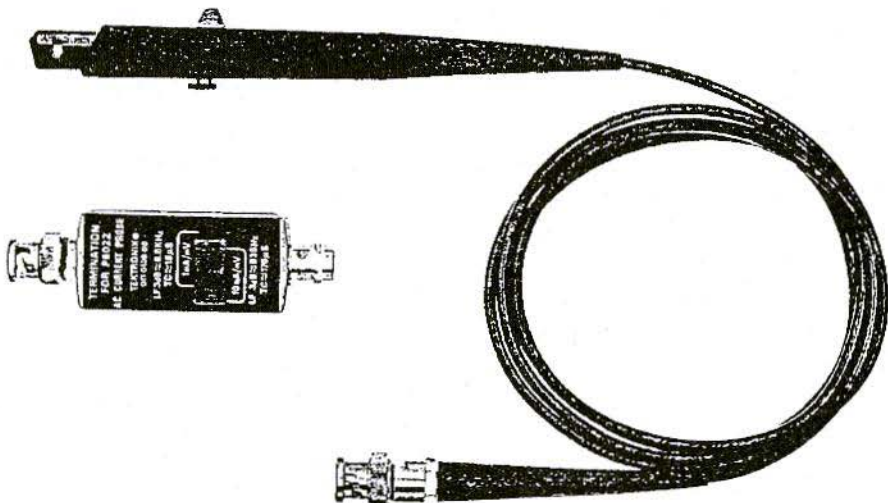
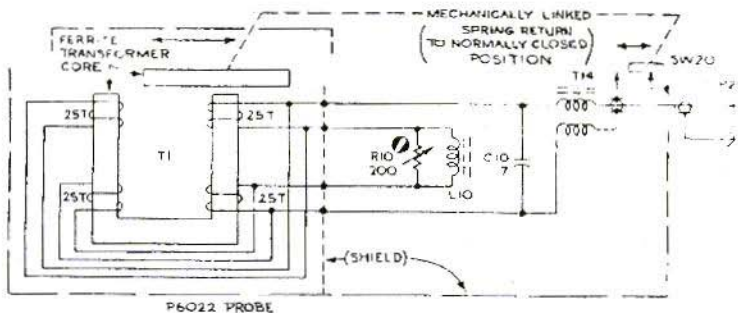


Fig.3: Structure of a Current Probe

levels are suitably reduced using attenuators or directional couplers.

Thermocouple power meters cover extremely wide frequency ranges in the coaxial versions; here are a few examples from Hewlett-Packard:

HP 8482A:
100 kHz...4.2 GHz, N-plug connection

HP 8485A:
50 MHz...33 GHz, APC 3.5mm plug

HP 8487A:
50 MHz...50 GHz, 2.4mm plug connection

In addition, there are also individual wave-guide measuring heads.

Since thermocouple power meters have a long-time drift, the display unit usually has a calibrating oscillator with a precisely known output power.

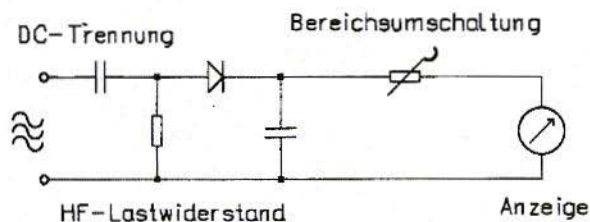


Fig.4: A Diode Wattmeter

DC-Trennung = DC separation, *Bereichumschaltung* = Range changing, *HF-Lastwiderstand* = High-frequency load resistor, *Anzeige* = Display

4.3 The Thermistor Power Meter

The thermistor power meter works on the principle of substitution. The heat output generated through the high frequency is compared with a heat output generated using DC current, and the DC current output is measured precisely. A temperature-dependent resistance, which nowadays is usually an NTC resistor in thin-film format, serves as a high-frequency load resistor and simultaneously as a temperature sensor hence the company name TFT (thin film thermistor) power meter. Fig.6 shows the basic structure.

The thermistor is operated in a self-balancing bridge circuit. The operational amplifier sets its output voltage in such a way that the thermistor is heated up to a resistance of 50Ω . The output voltage from the operational amplifier then acts as a measure of the power converted in the thermistor.

If the thermistor is now also heated with high-frequency current, then its resistance falls and the DC power is correspondingly reduced through the operational amplifier. The high-frequency power is replaced by a reduced DC power. The display instrument at the

operational amplifier output can thus be calibrated directly in high-frequency power.

But the thermistor can not differentiate between high-frequency power and an oscillating ambient temperature. For this reason, a second thermistor is incorporated in the measuring head, which is not high-frequency heated but is influenced only by the ambient temperature. A second self-balancing bridge thus captures oscillations in the ambient temperature and provides for temperature compensation.

The separation between the DC path and the high-frequency path through the choke and the capacitor represents a weak point in the circuit in Fig.6. Firstly, the choke is positioned in the DC circuit, and thus influences the precision of measurement through its resistance. Secondly, it must be free of resonance points and highly resistive over a wide frequency range. Practically structured measuring heads therefore use the circuit shown in Fig.7, which manages without a choke. With regard to DC, the two thermistors are connected up in series. With regard to high frequency, they are in parallel.

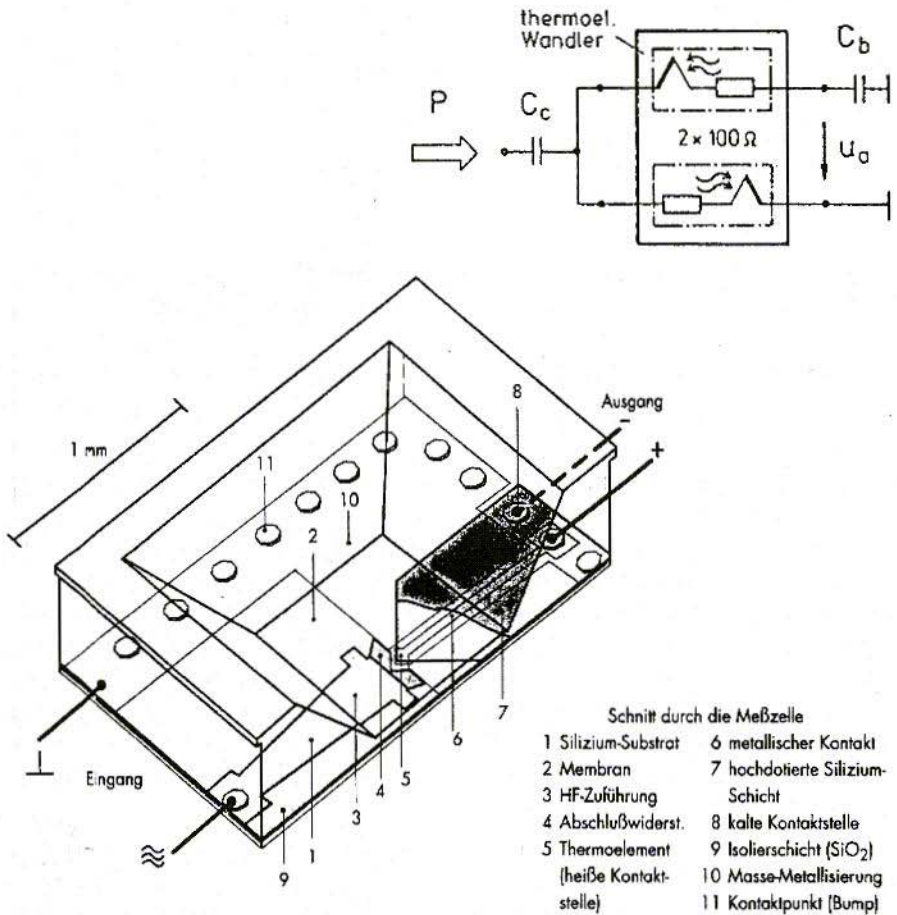


Fig.5: A Thermocouple Power Meter Head

Thermoel Wandler = Thermocouple converter, *Ausgang* = Output, *Eingang* = Input, *Schnitt durch die Meßzelle* = Section through the measuring cell, *Silizium-Substrat* = Silicon substrate, *Membran* = Membrane, *HF-Zuführung* = High-frequency feed, *Abschlußwiderst.* = Adapted load, *Thermoelement (heiße Kontaktstelle)* = Thermo-element (hot contact point), *Metallischer Kontakt* = Metallic contact, *Hochdotierte Silizium-Schicht* = Highly-doped silicon layer, *Kalte Kontaktstelle* = Cold contact point, *Isolierschicht* = Insulation layer, *Masse-Metallisierung* = Earth metallisation

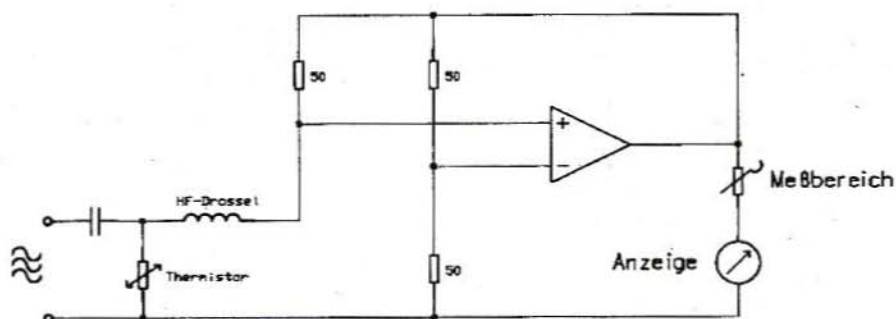


Fig.6: Principle of a Thermistor Power Meter

HF-Drossel = High-frequency choke, *Meßbereich* = Measurement range, *Anzeige* = Display

Thermistor power meters capture power levels of between 10W and 10mW. With series-connected instruments, the accuracy is approximately $\pm 0.2\%$, and is therefore better, by at least one power of ten, than that obtained using thermocouple power meters.

Coaxial thermistor power measuring heads such as the HP 8478B capture a frequency range of between 10 MHz and 18 GHz. There are also many special versions for dwarf waves with a wave guide input, which allow for power measurements of up to 300 GHz and more.

5. FREQUENCY MEASUREMENT

Only two procedures are still significant nowadays with regard to frequency measurement: the resonance procedure, using a calibrated oscillation circuit or resonator, and the frequency counter. Resonance procedures are confined almost exclusively to extremely high frequencies (dwarf waves, laser technology) where other procedures are too expensive.

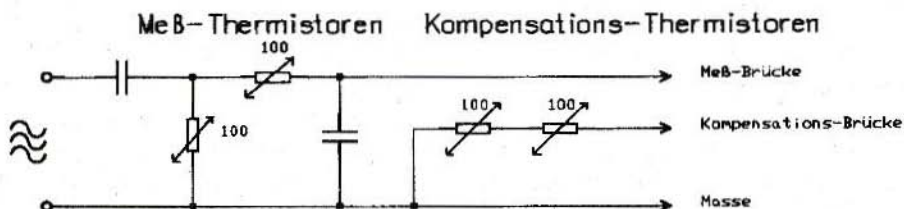


Fig.7: Practically Realised Circuit for a Thermistor Meter Head

Meß-Thermistoren = Measuring thermistors, *Kompensations-Thermistoren* = Compensation thermistors, *Meß-Brücke* = Measuring bridge, *Kompensations-Brücke* = Compensation bridge, *Masse* = Earth

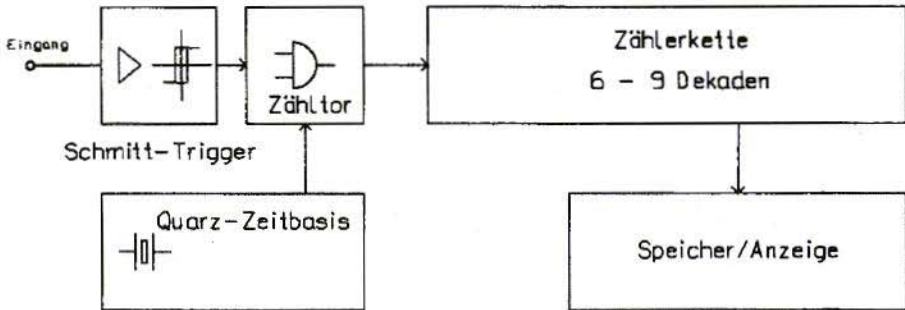


Fig.8: A Digital Frequency Meter

Eingang = Input, *Zähl tor* = Counting gate, *Zählerkette* = Counting chain, *Dekaden* = Decades, *Speicher / Anzeige* = Storage / display, *Quarz-Zeitbasis* = Quartz time base

The standard frequency meter used today is the digital frequency counter, as shown in Fig.8.

The input signal (voltage 100mV to 10V) is converted in a Schmitt trigger into a square-wave voltage, which is fed to a counting gate (AND gate). At the second input of the counting gate, a square-wave voltage is applied with a high time of, for example, 1 second, which is derived from a very stable quartz oscillator. The number of square-wave oscillations coming through the counting gate is detected using a binary coded decimal counter, and when the high time ends the reading is transferred into a storage and display record.

With this procedure it is clear that the high time must be extended to obtain an improvement in the accuracy of measurement, as this lengthens the measuring time. This is extremely annoying for calibration tasks.

For this reason, the meter manufacturers have thought up some significant improvements, which can also be used in combination:

Dual meters ensure that the instrument can still be used during the time required for display (and, if applicable, for converting the display). There are two counting gates and counting chains, which operate alternately. While one counting chain is counting the measurement signal, the contents of the second counting chain are processed and displayed.

The amount of data transferred is thus doubled, which is especially significant for short high times.

Reciprocal meters use a combined frequency and period measurement, as shown in Fig.9. The high time is generated using an adjustable timing pulse generator, which is not a high-precision component. In addition to the usual counting chain for the input signal, a second counting chain is present which is clocked with a very stable quartz clock (between 100 MHz and 1 GHz).

When a measurement begins, the counting gate is opened synchronously with the rising flank of the input signal, and

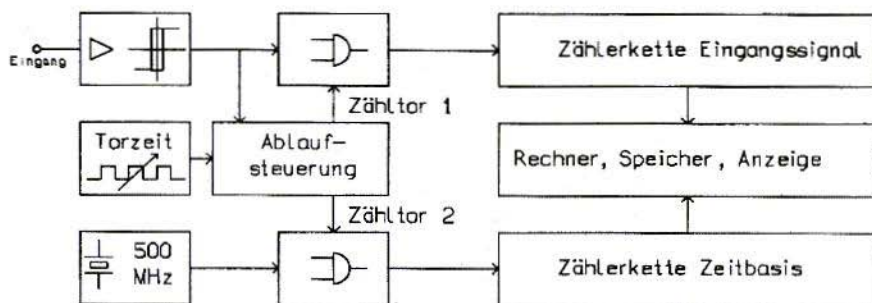


Fig.9: Frequency Measurement using a Reciprocity Meter

Eingang = Input, *Zähltor* = Counting gate, *Torzeit* = High time, *Ablaufsteuerung* = Sequence control, *Zählerkette Eingangssignal* = Counting chain input signal, *Rechner, Speicher, Anzeige* = Computer, memory, display, *Zählerkette Zeitbasis* = Counting chain time base

the timing pulse generator for the high time is started. At the same time, the counting chains for the input signal and the quartz clock are enabled. When the high time ends, both counting chains are stopped by the next rising flank of the input signal. Meter 1 contains the number of periods of the input signal, meter 2 the actual high time, with a resolution of from 1 to 10 nsec.. Dividing the input signal meter reading by the quartz clock meter reading supplies the frequency. In this way, the resolution improves by at least 2 to 3 decimal places for the same measurement time. With a high time of 1 second, the frequency can then be measured at 1 to 10 mHz and a measurement time of 1 to 10 msec. is sufficient for a resolution of 1 Hz. Top-quality meters like the Hewlett Packard HP5335A give a resolution of 9 digits per second, irrespective of the frequency. Reciprocal meters are certainly more sensitive than normal meters when signals are noisy or quivering,

since the noise is no longer averaged out over many signal periods.

The upper frequency limit of the meter is determined by the logic modules used. With modern modules of the 74AC or 74AS type series, counting frequencies of app. 150 MHz can be obtained. ECL flip-flops obtain counting frequencies of 1 to 5 GHz. Meters with clock frequencies of over 10 GHz have certainly been developed in the laboratory, but they are still a long way from being marketable products.

For frequency measurements in the microwave range (with series-connected instruments going up to 110 GHz), automatic frequency converters are used, which convert the measurement signal into a directly countable frequency plane (usually below 100 MHz). The two basic processes used involve the automatic transfer oscillator and the heterodyne converter. These two processes could not have been implemented before the appearance of integrated circuits

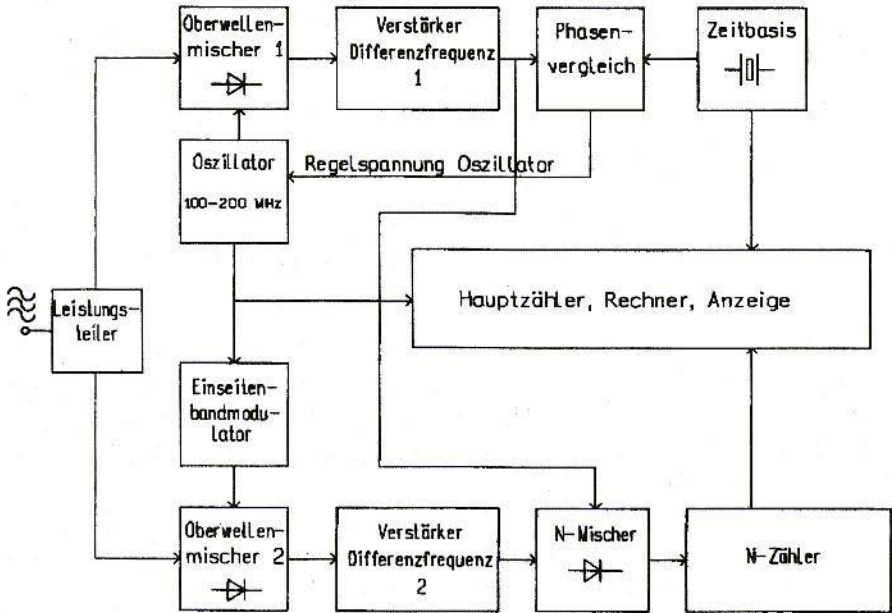


Fig.10: A Microwave Meter with Transfer Oscillator

Leistungsleiter = Power divider, *Oberwellenmischer* = Harmonic mixer, *Verstärker* = Amplifier, *Differenzfrequenz* = Differential frequency, *Phasenvergleich* = Phase comparison, *Zeitbasis* = Time basis, *Oszillator* = Oscillator, *Regelspannung Oszillator* = Oscillator phase control, *Hauptzähler, Rechner, Anzeige* = Main meter, computer, display, *Einseitenbandmodulator* = Single-side band modulator, *N-Mischer* = N mixer, *N-Zähler* = N meter

from about 1970 onwards. The automatic transfer oscillator was developed by Systron-Donner, and also forms part of the microwave meters from Hewlett Packard and Marconi. The heterodyne converter was developed by EIP and has also been used by Eldorado Electronics.

5.1 The Automatic Transfer Oscillator

The automatic transfer oscillator is based on having an easily countable oscillator (in the frequency range up to

100 MHz) synchronised with a subharmonic of the measurement signal. The core unit is a harmonic generator with a downstream mixer and an intermediate-frequency amplifier as in Fig.10. Let us look at the top half of the circuit first.

In the harmonic generator, the input signal is mixed with a harmonic, N,fo, from the internal oscillator. At the beginning of a measurement procedure, the internal oscillator is slowly tuned in from the bottom upwards. The intention

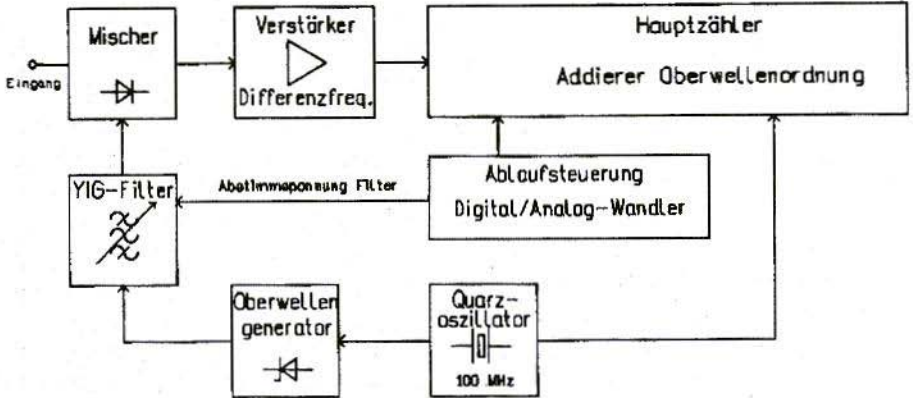


Fig.11: A Microwave Meter with Heterodyne Converter

Eingang = Input, *Verstärker* = Amplifier, *Differenzfreq.* = Differential frequency, *Hauptzähler* = Main meter, *Addierer Oberwellenordnung* = Harmonic order adder, *Abstimmspannung* = Tuning voltage, *Ablaufsteuerung* = Sequence control, *Digital/Analog-Wandler* = Digital / analogue converter, *Oberwellen-generator* = Harmonic generator, *Quarzoszillator* = Quartz oscillator

is to obtain a differential frequency, $f_z = (f_c - N \cdot f_o)$, at the mixer output, the rated value of which is, for example, 30 MHz.

A synchronisation circuit then controls the oscillator in such a way that the differential frequency, f_z , is adjusted to the (quartz-stable) rated value. This makes it possible to calculate the input frequency, f_c , in that the reading from the oscillator, f_o , is metered using a normal frequency meter and, in a second step, the harmonic, N , is determined which is used for mixing. N can be determined by de-tuning the oscillator by a small amount, f . The differential frequency will then alter by $N \cdot f$.

It is plain that this method functions only if the measurement signal is not frequency-modulated. But this represents a big limitation in practise. For this reason, a second harmonic mixer branch

is used to determine N , as in Fig.10. The output frequency of the oscillator is displaced in a single-side band modulator by an amount f (by 1 MHz), and is mixed with the input frequency in the second harmonic mixer. The differential frequency thus engendered, $f'z$, differs from f_z by an amount, $N \cdot f$. If we now mix f_z with $f'z$, the result is a value of $N \cdot f$.

An auxiliary meter uses this to determine the value of N . A frequency modulation of the input signal influences both differential frequencies to the same extent, and thus has no significance.

Microwave meters with transfer oscillators are certainly very sensitive, since the differential frequencies, f_z and $f'z$, can be amplified in the narrow band (less noise). Values of -35dBm up to 12 GHz and -20dBm at 26 GHz are typical. But since the reading is taken

from the internal oscillator, the measurement for a pre-set resolution, N , takes a little longer.

5.2 Microwave Meters with Heterodyne Converters

The microwave meter with a heterodyne converter also operates with a harmonic mixer, but in addition a variable-frequency filter is wired up between the harmonic generator and the mixer, as in Fig.11.

The mixing oscillator is a quartz oscillator with $f = 100$ or 200 MHz, and the differential, f_z , is read off directly. By tuning the filter, we can feed all harmonics of f_Q to the mixer. For the measurement, the filter is tuned from the

bottom upwards until the differential frequency, $f_e - N.f_Q$, is within the operating range of the meter.

The factor $N.f_Q$ can be read off from the frequency scale of the filter. If the meter reading is added to this, the input frequency, f_e , can be obtained directly. Measurement is thus simpler, by some way, than with the transfer oscillator and, since the differential frequency is read off directly, the full resolution is obtained immediately. The disadvantage is that the amplifier for the differential frequency, f_z , has to operate in the broad band, and thus generates more noise.

(To be continued)

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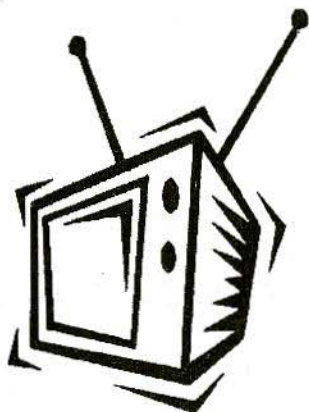
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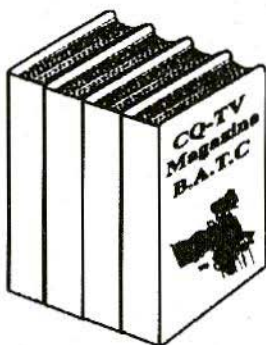
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Gerald Düllberg, DL8DAQ

Microprocessor-Controlled FM Receiver for Weather Satellites

Part-2 (conclusion)

Editor's Note: Owing to an error on my part, Figure 3 of this article, which appeared on page-120 of the last issue (1999-Q2), was reproduced mirrored. A correctly oriented version of this figure has been reproduced here on the next page ... Mike

3. DESCRIPTION OF SOFTWARE

The processor itself controls the station-finding function. The PLL sequentially obtains the I²C signals from channels 1 to 6 of the processor. When an APT tone is detected, the station-finding function is halted, so that the image signal received can be plotted.

In addition, the AFC is automatically activated when an APT tone is detected. This is especially necessary when signals are received from satellites in polar orbit, in order to balance out the Doppler effect. The AFC also makes its presence felt noticeably in a positive way if the LO in the converter should deviate slightly from its rated frequency.

Moreover, when an APT signal is present pin-1 of the processor goes to logic "1". This activates the emitter follower of the low-frequency amplifier.

If no APT signal is available, the receiver remains at the frequency for about another 8 seconds (fall-delay time) and then the station-finding function is re-activated.

The channels are shown on a simple 7-segment display, since the resources needed for an LCD display, in terms of

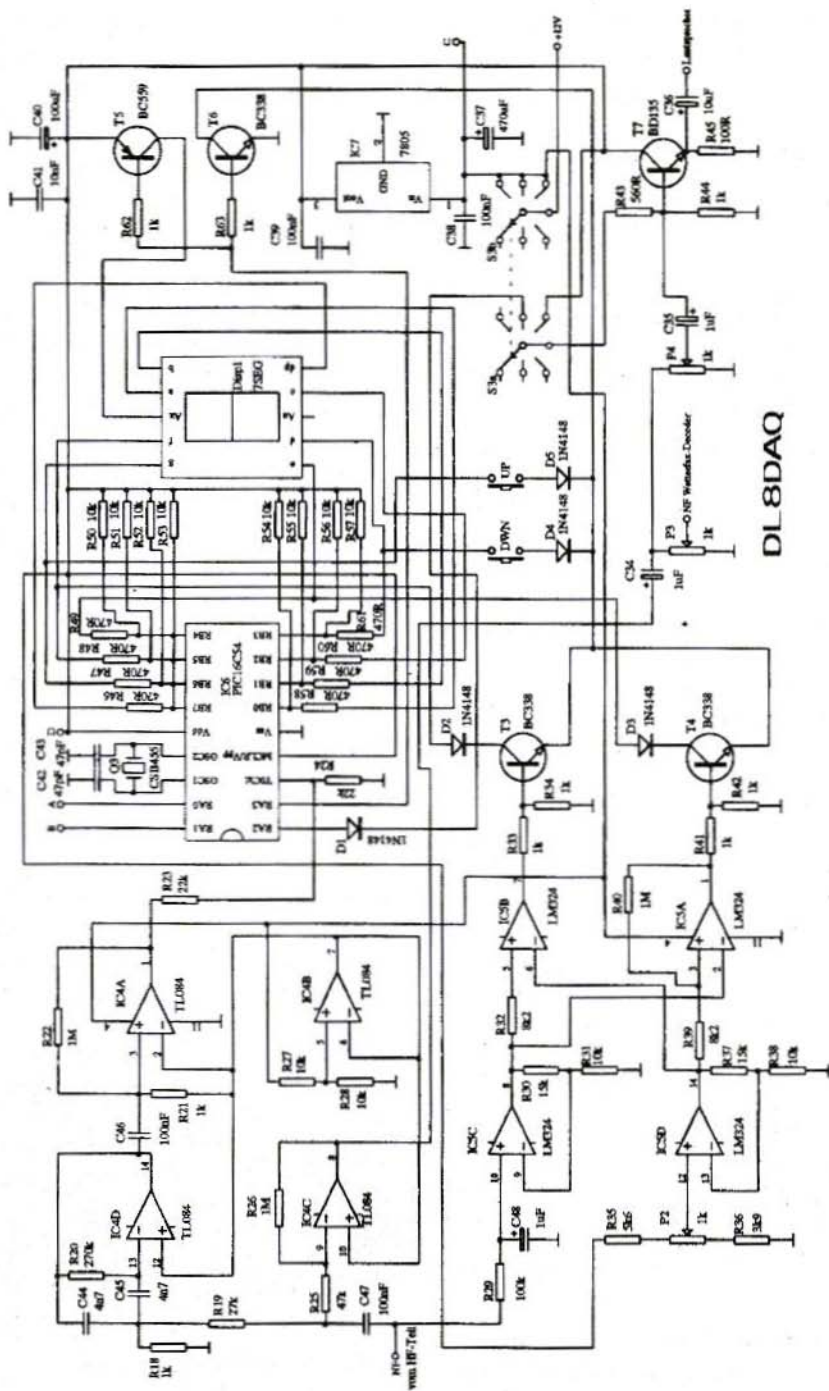


Fig.3: Control Electronics of the Weather Satellite Receiver

cost and programming, do not appear to be justified for this application.

When the operational voltage is switched on, the receiver automatically goes into the station-finding mode. The channels can naturally also be selected manually through the "UP" and "DWN" keys.

If one of these keys is pressed while the station-finding function is active, the receiver immediately switches to step-by-step operation.

Moreover, the two fine-setting frequencies can now be selected for Meteosat reception.

There seems little point in scanning these frequencies, since Meteosat transmits on an almost continuous basis, and with a converter connected the station-finding function would continually be dwelling on one of these two frequencies.

To re-activate the station-finding function, one of the two keys must be pressed for a period of about a second.

4. STRUCTURE

The entire receiver is mounted on a through-hole plated European standard size board, copper-coated on both sides (Fig's.6 and 7). The top side is a continuous earth surface, apart from two tracks.

The operational controls and the channel display are mounted on a second board

measuring 30mm x 80mm, which is mounted vertically to the main board (Fig.5).

With regard to assembly, it should be noted that the antenna relay is not soldered in until much later, since otherwise there could be problems with putting the antenna sockets in.

Special care is required when mounting the two SMD components, T1 and IC1, the position of which on the foil side of the board can be clearly recognised in Fig.8. The positioning of the other components is shown in the components drawing (Fig.9). An IC base is provided for the processor. The other IC's are soldered in directly.

The high-frequency transformer (TR1) consists of a 5mm ferrite bead, into which two turns of double-wound twisted enamelled copper wire are inserted (max. diameter 0.3mm) (Fig.10).

The chokes for the remote power supply are soldered in, if required.

With short antenna feeds, the polar satellites can be received without pre-amplifiers. In that case, make sure the remote power choke is not fitted, since many antennas form a DC short circuit.

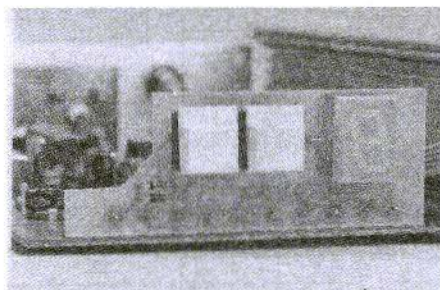


Fig.5: Operational Controls on Display Board

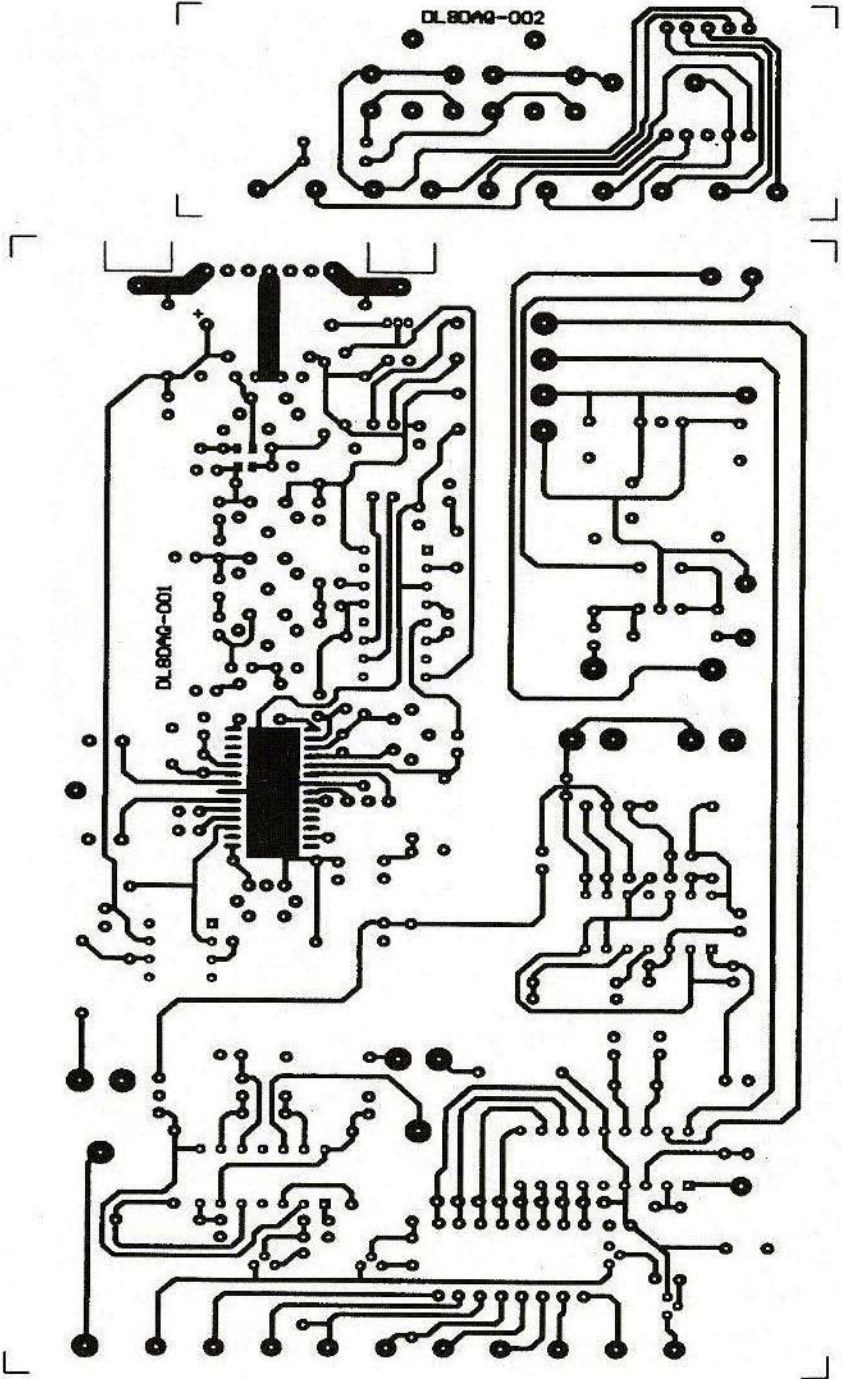


Fig.6: Component Side of the WX Receiver & Display PCB

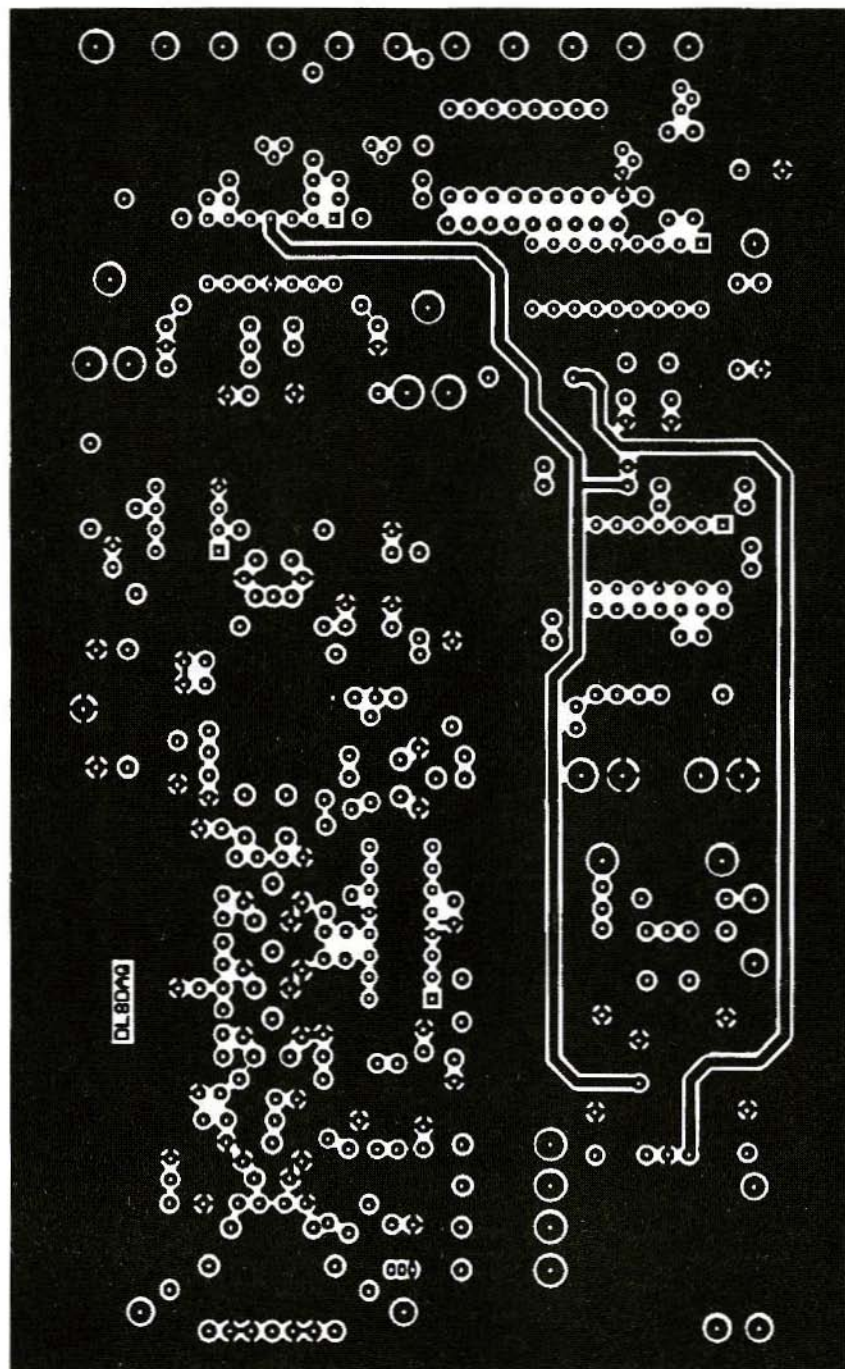


Fig.7: Earth Side of the WX Receiver & Display PCB with Two Tracks

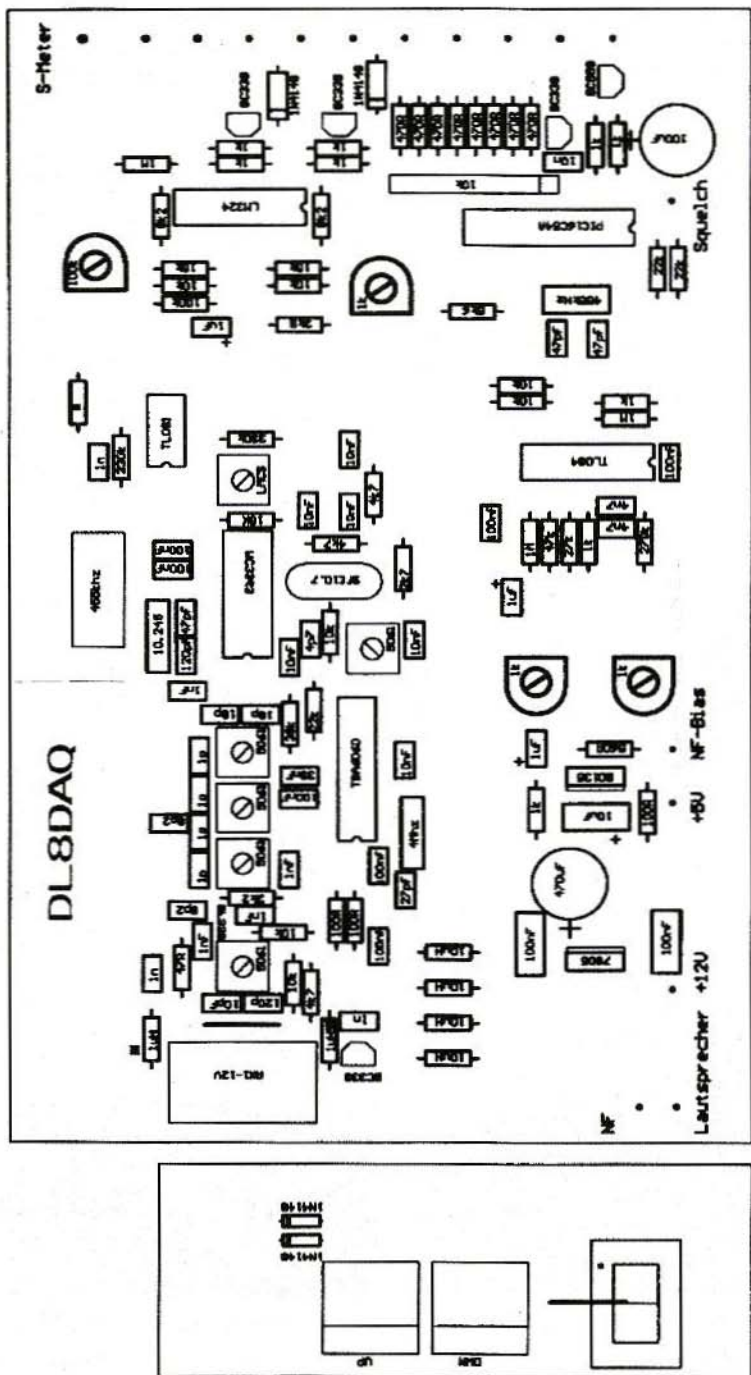


Fig.9: Component Overlay

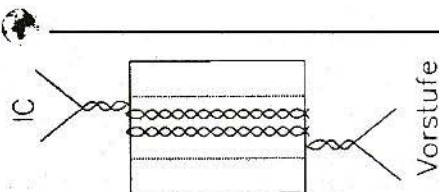


Fig.10: The High-Frequency Transformer (TR1)

When all the components have been mounted on both boards, right down to the antenna relay, the operational control board is soldered vertically to the main board, with small pieces of bent wire (Fig.11).

The prototype receiver is housed in an aluminium profile housing, such as is known from the TNC's. The BNC input sockets are fitted into the rear wall of the housing. They are also screwed down internally with a piece of tinfoil or with soldering lugs in order to obtain a good earth connection. The tinfoil is soldered to the board when the housing has been installed (Fig.12).

Only now can the antenna relay be soldered in!

Finally, the sockets and the circuit closer are wired up in accordance with

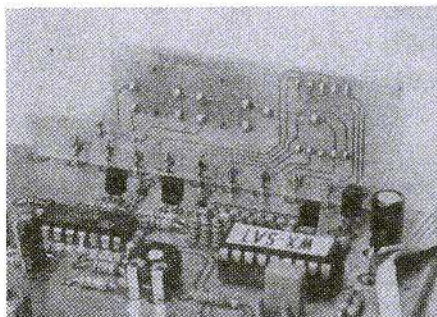


Fig.11: Detail Photo, Operational Board

the connection drawing (Fig.13). The circuit closer is a four-stage change-over switch with two levels. This switch has several functions. Firstly, it switches the operating voltage on. Secondly, it operates the external monitoring loudspeaker, which can have three settings (mute, squelch, always on).

Before commissioning, give the assembly another thorough visual check.

Fig.14 shows the general view and gives us a look into the open prototype.

5. BALANCING

Balancing is relatively simple, but special care should be taken in balancing the AFC!

First the PLL is put into operation. Tune to channel 3 for this (137.5 MHz). The oscillator coil (L5) is now set in such a way that a voltage of app. 2.5 Volts can be measured at pin-13 of the PLL-IC TSA 6060. If channel 8 is now selected

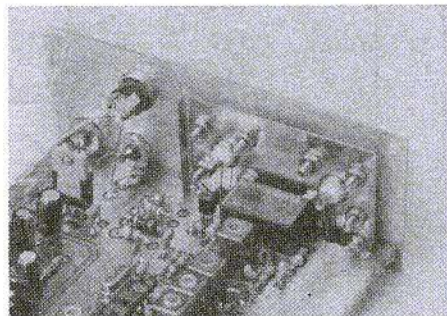


Fig.12: Detail Photo, Assembly of BNC Sockets with additional Earth Plate

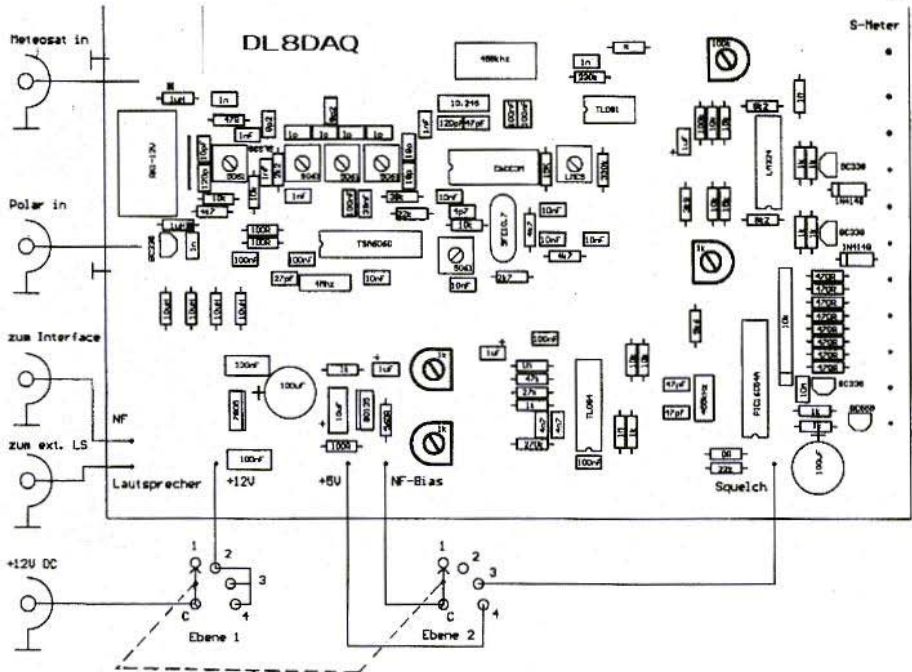


Fig.13: Wiring Diagram of Entire Assembly

Zur Interface = To interface, *Zum ext. LS* = To external loudspeaker, *Lautsprecher* = Loudspeaker, *Ebene* = Level

(141.0 MHz), the voltage is approximately 3.5 Volts.

If the channels are now switched through one after another, the tuning voltage should change correspondingly.

The pre-stage can be balanced with the greatest precision if a wobble meter is used (Polyskop).

The high-frequency signal is stored in the "Polar socket" (careful: use the DC block!) and the de-modulator is connected at the point between C14 and C15 through a 1pF capacitor.

However, in practise it has proved to be the case that balancing can also be done simply at the maximum signal, with the

measurement taking place at the S-meter output. A suitable signal generator is needed for this, which supplies a signal which is not too powerful, but is constant and unmodulated (approximately - 60dBm).

Balancing with the signal from a weather satellite can not really be recommended, because its signal strength oscillates too much as it flies over the receiver.

Channel 8 (141 MHz) is not as sensitive as the other channels. But this has no effect on reception through a Meteosat converter because the latter determines the aggregate noise factor.

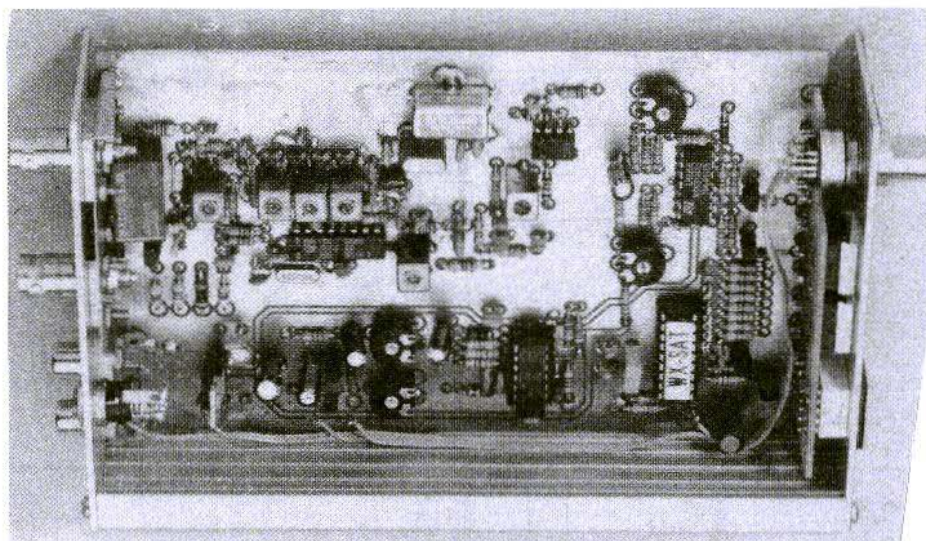


Fig.14: A Look at the Open Prototype of the 137 MHz Receiver

Balancing the AFC calls for some sensitivity, but it can be done with a simple multimeter. First the de-modulator circuit, L6, is hunted through without an input signal. It will be noted here that the noise becomes weaker in both directions. Now we search for the maximum noise point.

Then the voltage is measured at pin-8 of IC 5. The same value is now set at pin-14 with P2.

The volume of the monitoring loud-speaker can now be set with P4.

The signal level at the low-frequency output to the weather fax converter is set to approximately 350 mVeff, and can be adjusted with P3 if necessary, depending on the weather fax converter used.

This concludes the balancing.

6. PARTS LIST

IC's:

1 x	MC3363 SMD
1 x	TL081, DIL
1 x	TL084, DIL
1 x	TSA6060, DIL
1 x	LM324, DIL
1 x	7805, 5V voltage stabilising
1 x	PIC16C54A(WX-SAT V1.0) programmable, DIL

Transistors:

1 x	BF998, SMD
4 x	BC338 or similar
1 x	BC559 or similar
1 x	BD135

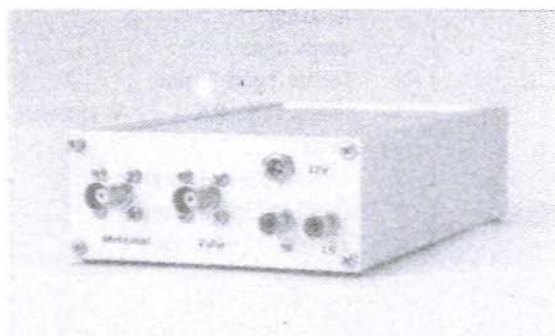


Fig.15:
Rear View with
Connection Sockets

Diodes:

- 5 x 1N4148
- 1 x 13.5mm 7-segment display, common anode
- 1 x 4 MHz quartz, HC-18/U, series
- 1 x 10.245 MHz quartz, HC-18/U, series

- 2 x 1 H chokes, RM 10mm
- 4 x 10 H chokes, RM 10mm

Trimming potentiometers: 5mm x 10mm, horizontal

- 3 x 1 k Ω
- 1 x 100 k Ω

Resistors: ¼ W, carbon layer

- 1 x 47 Ω
- 3 x 100 Ω
- 8 x 470 Ω
- 1 x 560 Ω
- 9 x 1 k Ω
- 1 x 2.2 k Ω
- 1 x 2.7 k Ω
- 1 x 3.9 k Ω
- 3 x 4.7 k Ω
- 1 x 5.6 k Ω
- 2 x 8.2 k Ω
- 15 x 10 k Ω
- 3 x 15 k Ω
- 3 x 22 k Ω
- 1 x 27 k Ω
- 1 x 39 k Ω
- 1 x 47 k Ω
- 1 x 100 k Ω
- 1 x 270 k Ω
- 2 x 330 k Ω
- 3 x 1 M Ω

Ceramic capacitors: grid 2.5 mm.

- 4 x 1 pF
- 1 x 4.7 pF
- 2 x 8.2 pF
- 1 x 10 pF
- 2 x 18 pF
- 1 x 27 pF
- 3 x 47 pF
- 2 x 120 pF
- 4 x 1 nF
- 2 x 4.7 nF
- 9 x 10 nF
- 1 x 39 nF
- 9 x 100 nF

Electrolytic capacitors: 16 V

- 3 x 1 μ F, RM 2.5
- 1 x 10 μ F, RM 5
- 1 x 100 μ F, RM 5
- 1 x 470 F, RM 5

Ceramic oscillator:

1 x CSB 455

Inductive resistors:

5 x Neosid 5061

1 x Toko LMCS 4102

Filters:

1 x SFE 10.7

1 x Murata CFL 455 AG2

Other:

2 x Digitast-type input keys

2 x Cynch panel jacks

1 x DC socket

1 x miniature rotary switch, 2 x 6
with stopper, e.g. Knitter

2 x BNC flanged bush

1 x aluminium profile or tinplate

housing

1 x earth plate

1 x ferrite bead 5 mm.

1 x Matsushita RK 1-12 V antenna
relay

1 x DL8DAQ-001 board

1 x DL8DAQ-002 board

1 x 18-pin IC base for PIC

7.

LITERATUREMotorola: Linear interface IC's
Device Data

Data sheet: TSA 6060

Microchip: Microcontroller (PIC)

Westerwele, H.: Weather Satellite
Reception Practice

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Ian Waters, G3KKD

Two Filters and a Diplexer for 23cm

In celebration of 50 years of the BATC, this article has been reproduced from the latest edition of CQ-TV. For further information about the BATC, see their advertisement on page-165 of this issue ... Editor

This article describes two interdigital bandpass filters for 23cm that have been made and tested to information contained in the ARRL publication UHF Microwave Experimenters Manual. The article also describes a diplexer that uses one type of filter to enable transmission and reception via a single antenna.

1. INTERDIGITAL FILTER DESIGN

Filter A is a fairly simple design of modest performance with three active elements. It has a -3dB bandwidth of 8.2%, or 105 MHz, centred on 1282.5 MHz and thus covers the whole 23cm amateur band. It has a VSWR of 1.5:1 and, when constructed from aluminium, an insertion loss of approximately 1dB.

Filter B is a somewhat more complex design, with five active elements and a much better performance. It has been designed to have a -3dB bandwidth of 1.1%, or 14.5 MHz, centred on 1316 MHz and thus covers the RMT-2 amateur television repeater output channel, although it can be tuned to cover any other 14.5 MHz segment of the band, for instance a repeater input channel. This filter also has a VSWR of approximately 1.5:1 and, when constructed from aluminium, a slightly higher insertion loss of 1.5dB.

The measured amplitude frequency responses of these filters are shown in Fig.1, while the measured response of Filter B is shown with an expanded frequency scale in Fig.2.

It should be mentioned that the ARRL information covers the design of 23cm filters from three to eight elements and percentage bandwidths from 1 to 10%. In total four filters have been constructed by the author using the ARRL

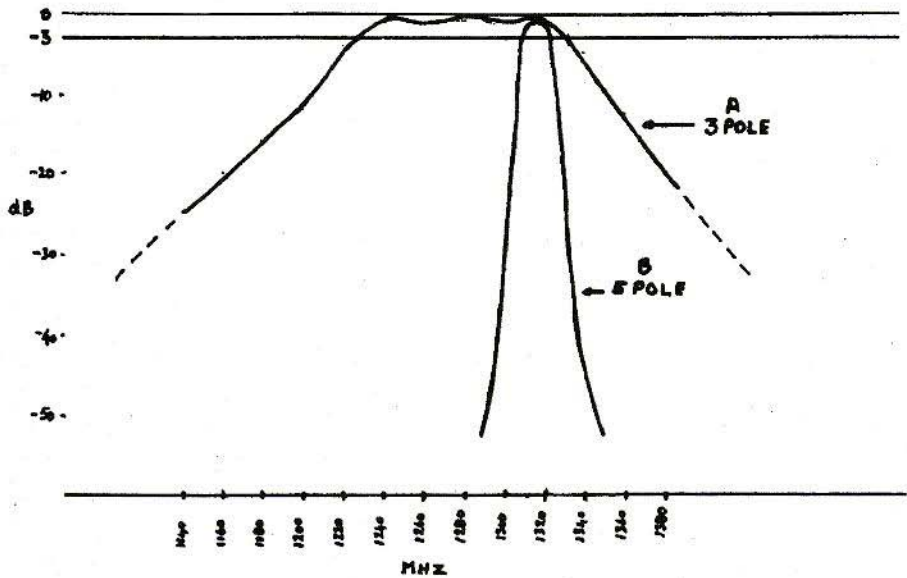


Fig.1: Amplitude Frequency Responses of Filters A and B

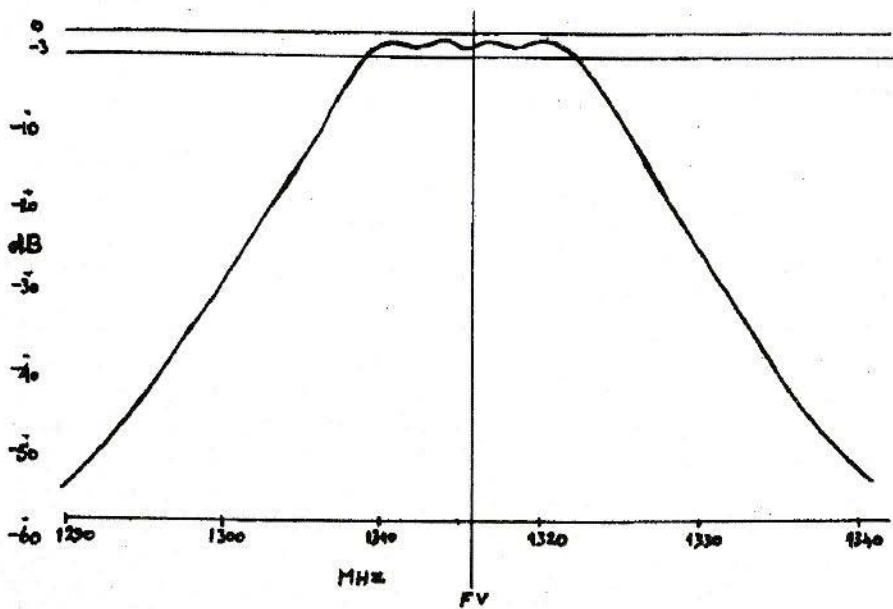


Fig.2: Response of Filter B with an Expanded Frequency Scale

data and they have all performed as expected and have tuned up without difficulty

2. CONSTRUCTION

The filters can be constructed from aluminium or silver-plated brass. Plated brass would give better performance, although it is difficult to say how much better, but would be much more expensive.

The mechanical details of Filter A are given in Fig.3 and those of Filter B in Fig.4. They are both quite simple to construct, provided that a small lathe and the usual hand tools are available. However, the work for Filter B is quite time consuming. If no lathe is available then the services of a friendly model engineer might be obtained.

The top and bottom plates are constructed from 1 inch and $\frac{1}{4}$ inch aluminium strip. The side plates are cut from 16 SWG or preferably 14 SWG aluminium sheet. The top and bottom plates should be drilled and tapped M3 or 4 BA and the sides attached with screws at 1 inch intervals. Care should be taken to remove any burrs and ensure good contact.

The rods are turned down from the next largest size of stock aluminium rod. They should be accurate to within ± 0.002 inches. They are centre drilled and tapped for attachment to the top and bottom plates using any suitable screws. The tapped ends are countersunk

slightly to ensure that electrical contact is at the circumference. The end rods, or input and output transformers, are drilled to accept the centre contacts of the input and output plugs. These are clamped by small screws in tapped radial holes at the ends of the rods.

The tuning screws may be of any size from $\frac{1}{4}$ to $\frac{3}{4}$ inch and can use any fine thread for which taps and dies are available. I have used 32 threads-per-inch which allows for a smooth and fine tuning adjustment. However, care must be taken as such fine threads can pick up stray material and seize quite easily. Plenty of cutting paste or oil is required. Lock nuts are also required, which are tightened once tuning is completed.

The connectors may be N, TNC or BNC to suit requirements and the method of fitting will depend on the connectors used.

3. TUNING UP

The author used a sweep oscillator with frequency markers to tune the filters, which permitted the passband to be set exactly as desired and the shape to be optimised with minimum ripple. One problem is to first tune the filter to allow enough sweep signals to pass through it to enable alignment to start. With the 3-element filter it is not too difficult to randomly turn the tuning screws until something is seen. The larger number of permutations possible

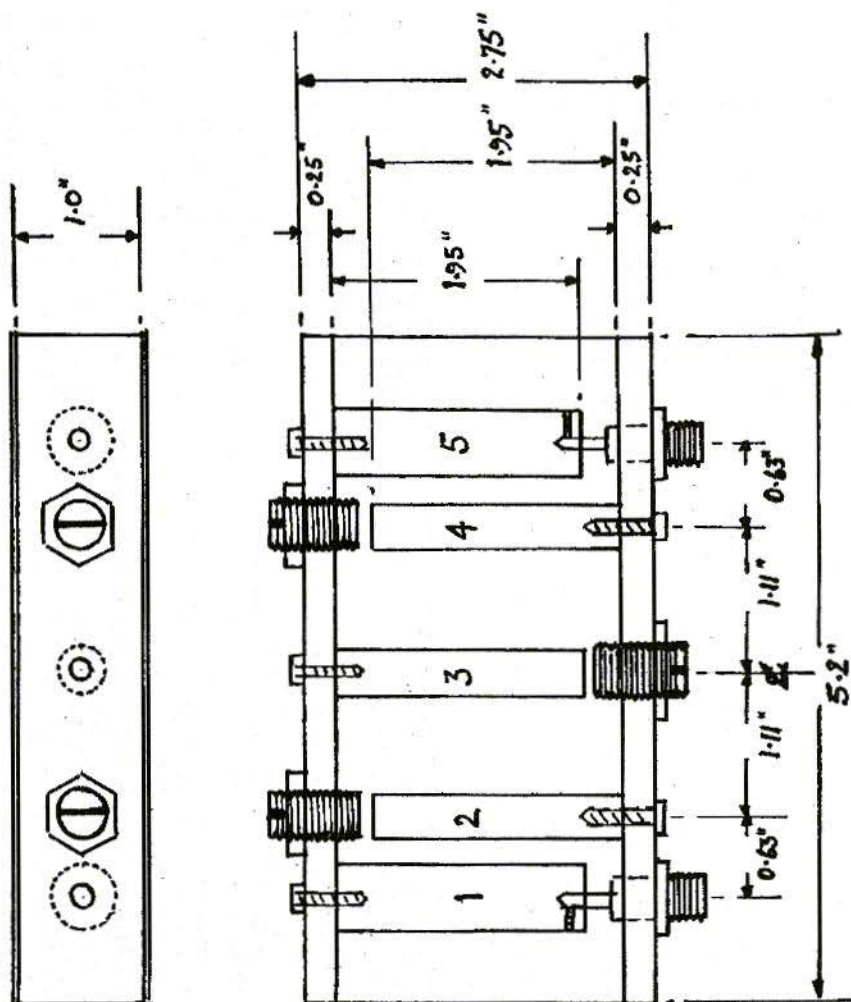


Fig.3: Mechanical Details of Filter A

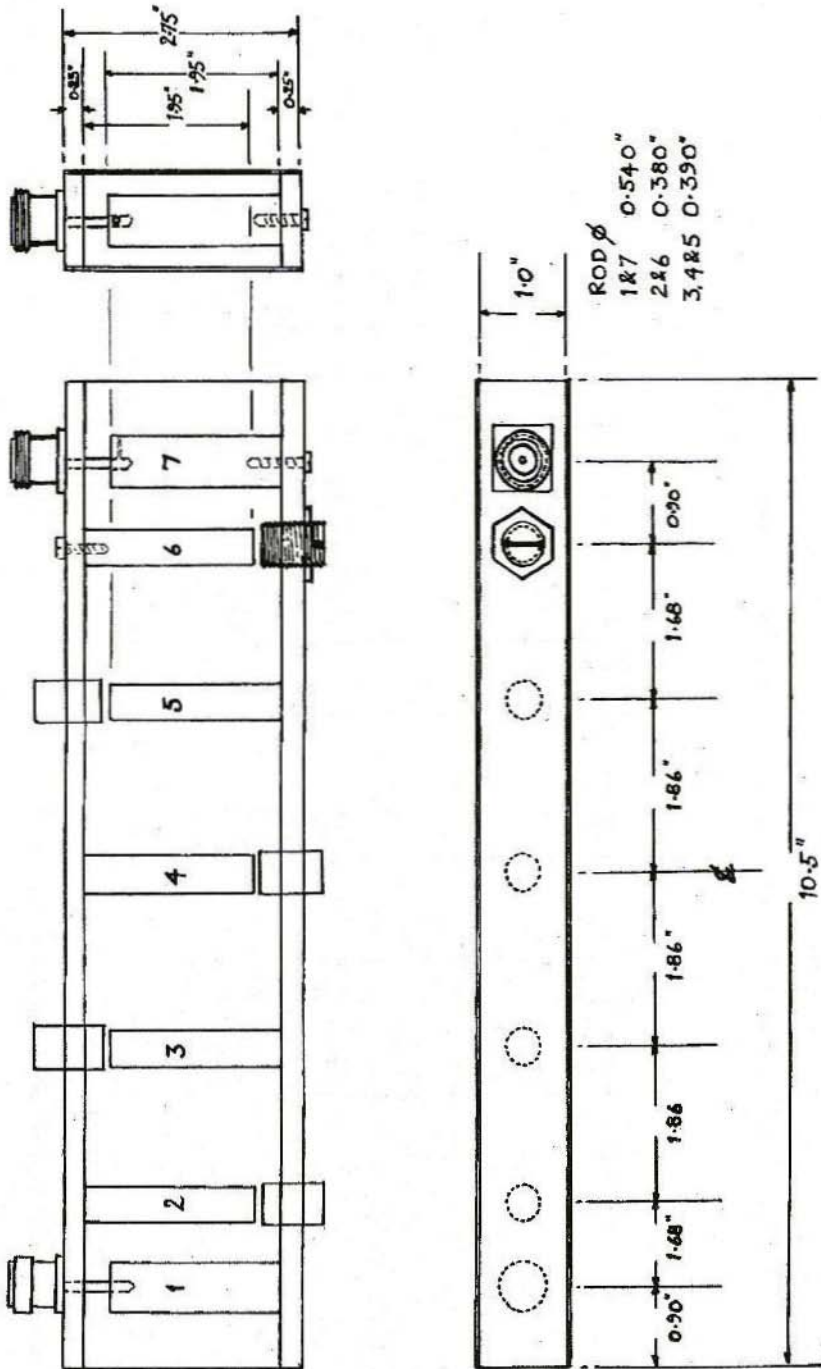


Fig.4: Mechanical Details of Filter B



with the 5-element filter make this much more difficult. One method to get a start is to inject a CW signal at the passband centre frequency and look for a signal at the output using a spectrum analyser. Once a signal is seen, even at -70dB, the tuning can be quickly optimised. Another method could be to use a communications receiver.

4. APPLICATIONS

Filter A was designed for use as an image noise filter placed between the RF stages and the mixer in a receiver, or after the up-conversion mixer in an IF modulated transmitter, to improve unwanted mixing products.

Filter B has been applied in a number of ways. It can be used at the output of a transmitter, perhaps a repeater transmitter, to remove unwanted modulation sidebands. If the insertion loss of 1.5dB is too much to lose, then assuming that the PA is a class AB linear brick amplifier, the filter may be placed between the drive stage and the PA.

A class AB amplifier, if not overdriven, will not reinsert the removed sidebands and the loss of drive power can usually be made up quite easily.

This filter can also be used at the input of a receiver to protect against adjacent channel interference.

Another application of Filter B could be in the aerial feeder of a look-through antenna to reject the sidebands of the

stations own transmission, that can degrade the look-through picture.

5. DIPLEXER

Two filters type B, one tuned to the required transmit frequency and the other to the receive frequency, connected together by a coaxial T-junction as shown in Fig.5, can be used to make a very effective diplexer, thus enabling transmission and reception via a single antenna.

The T-splitter is constructed from a N-type coaxial connector and two lengths of semi-rigid coaxial line fitted into male N-plugs. The combined lengths of these lines and connectors is such that the effective length from the centre point of the T-piece to the start of the filter input transformers is one wavelength at the centre frequency, taking the velocity factor of the coaxial line into consideration.

The theory is, each filter has a reasonable match and so presents a low impedance at its passband frequencies, but a high impedance at the other channel, some 65 MHz away in a 23cm Amateur Television system. The one wavelength T arms transfer these impedances to the junction point. An incoming signal thus sees a low impedance path to the receiver, while the other path appears as a high impedance. Conversely, the transmitter sees a low impedance path to the antenna, but the route to the receiver appears as high impedance.

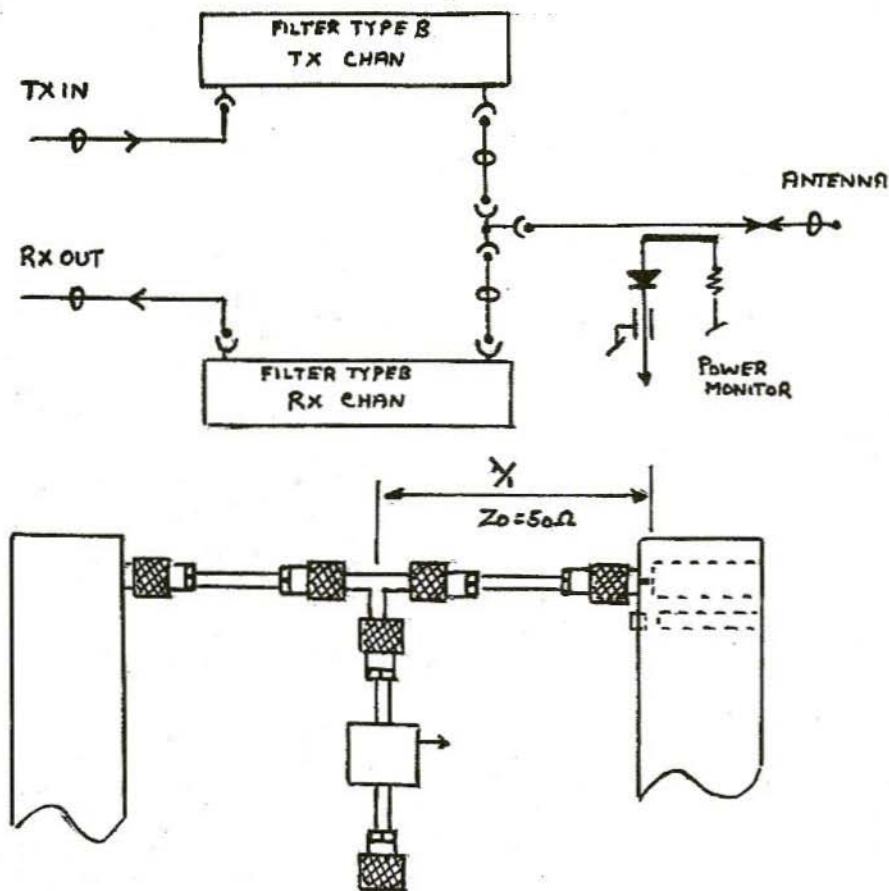


Fig.5: Construction Details for the Diplexer

The transmit filter suppresses out-of-band modulation products, particularly any that may be in the receiver pass-band. The receiver filter also suppresses the transmitted signal and any other signals beyond the wanted receiver passband. Although its insertion loss degrades the receiver sensitivity by 1.5dB, the improvement in received picture quality in the presence of inter-

fering signals makes this well worth while.

The transmit/receive cross loss is beyond the limit of practical measurement. When receiving a full-strength (P1) picture it is possible to switch the transmitter on and off without seeing any change in the received picture quality.



Graham Baker, ZL1TOF

UHF TV Masthead Pre-Amplifier

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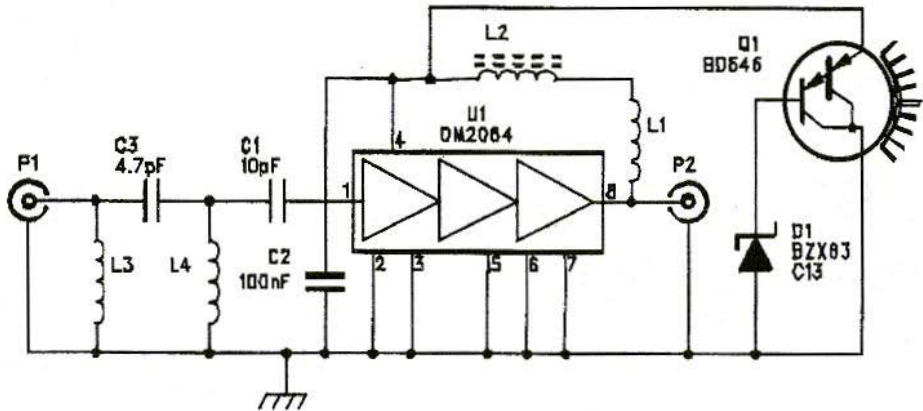
As narrow band aerials and amplifiers, normally used for amateur television, are not suitable for broadcast reception I erected a commercial UHF TV aerial and diplexer.

The next step was to get a decent pre-amplifier close to the aerial to deliver an adequate signal to run a number of TV outlets.

The Spectrum low-noise pre-amplifier has too little gain to deliver the necessary signal level and adding another stage such as the Wellington VHF premium performance pre-amplifier just increases the intermodulation distortion.

Commercial amplifiers have used the OM350 and OM361 devices since the early 1980s. These devices have noise figures of app. 7dB. Newer types of device are available and cost no more than the old types. The OM2064 (RS Components: 284-589) has a gain of app. 28dB and a noise figure of the order of 4.4dB. Under test it gave 107bBuV (-2dBm) for -60dB intermodulation in a three-tone test.

The device can be laid on a sheet of copper or brass with the ground connections bent into an L shape and soldered to the sheet. The other connections may then be made self-supporting or supported on small pieces of PCB glued to the ground sheet.



UHF Masthead Pre-Amplifier

I produced a double-sided PCB for the OM2064 with 75 strip-lines and BNC connectors at each end.

All the capacitors are 3.2 x 1.6mm surface-mount types. The amplifier is powered via the coaxial cable and mounted in a die-cast box (Farnell: 301-528 or DSE: H-221) as are the commercial models.

During testing I noticed ghosting on the VHF channels. Because the amplifier is broadband (40 to 860 MHz) it was amplifying the distorted VHF signals picked up by the UHF aerial and fed them through the diplexer into the receivers. A high-pass filter at the front end of the amplifier cured this problem. The shunt regulator was added for protection against a too high supply being applied.

Coil details:

L1 and L2 are constructed together, 3 turns of 0.35mm wire on a ferrite bead (DSE: R-5400) and 3 turns close-wound on a 3mm mandrel.

L3 and L4 are 1.75 turns of 0.5mm wire on a 3mm mandrel, spaced 1.5mm above the PCB surface.

The coils L3 and L4 must be mounted such that there is no mutual coupling.

A second amplifier was also constructed using the same design but utilising F-connectors. Using a spectrum analyser the high-pass filter was aligned to roll-off at as high a frequency as possible with minimum attenuation at 443 MHz. This amplifier is in use for amateur television reception and exhibits good performance.



Bob Platts, G8OZP

Dielectric Antenna for 3cm

In celebration of 50 years of the BATC, this article has been reproduced from the latest edition of CQ-TV. For further information about the BATC, see their advertisement on page-165 of this issue ... Editor

Dielectric antennas provide a simple means of achieving reasonable directional gain in a compact unit. They operate on the principle of refraction, but I shall not bore you with all the theory, its a little heavy and to be honest my middle-aged brain is unable to get around it as well as I could in my student days.

The design provides a gain of app. 20dB, with a match of better than 1.2:1 over the whole 3cm band.

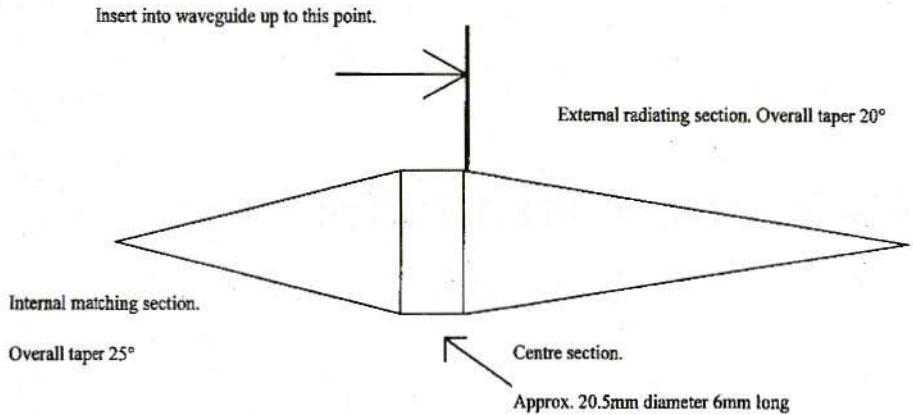
The waveguide is 22mm copper pipe (the standard plumbing variety) and has a transition to WG16. The polarisation is the same as the feeding waveguide.

The material should be Nylon 66, PTFE may be used, it is more difficult to

machine but provides improved performance. Nylon 66 is available from good engineering suppliers or RS Components and possibly Farnell.

Machine on a lathe very carefully with a sharp tool. Nylon 66 is naturally slippy. This means that it will not grip very well in the chuck. Also, as it is flexible it can grab, digging the cutting tool into the job, ripping it out of the chuck and throwing it across the workshop. I know from experience. A chuck rotation speed of 600 rpm is recommended, also clear the swarf away regularly whilst machining.

The parallel section should be a tight fit into the 22mm pipe, which will keep moisture out. As there can be a variance in the dimensions of copper pipe it is



Dielectric Antenna with app. 20dB Gain

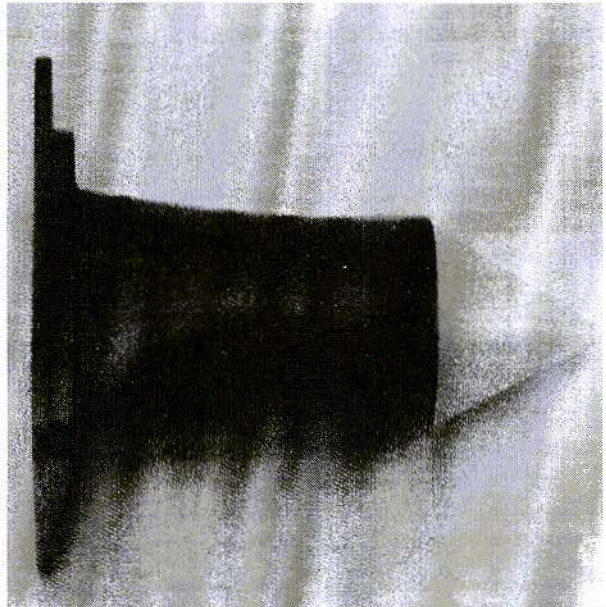
best to measure the internal diameter accurately before machining. There is no point in trying to glue the unit into the pipe as glues do not stick to Nylon 66.

To convert the 22mm pipe to the standard WG16 waveguide requires a transition. These are relatively simple to construct.

Approximately 120mm of 22mm copper tube should be annealed for app. half its length, by heating it to dull red and allowing it to cool naturally. Cut a 75 to 100mm long piece of hard wood to a rectangular section of 23mm by 10.75mm. Then, 50mm from one end of this piece of wood shave it down to a section of 6mm square.

The wooden former is the swaged (aka hammered!) into the annealed end of the pipe. As this is done gently hammer the out-

side to form the correct size and shape to fit a WG16 flange. Clean the pipe well before soldering into the flange. The other end of the pipe may need trimming back if damaged during swageing.





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ASSEMBLED GUNNMOD2 3cm ATV TRANSMITTER PCB & KIT

Gunnmod2 PCB forms the heart of a compact, fully featured FM. Gunn diode PAL, SECAM. or NTSC Amateur Television Transmitter with 5.5, 6.0 or 6.5 MHz intercarrier sound. The PCB is suitable for most types of 3cm and other frequency Gunn oscillators up to 25mW (up to 300mW Gunn oscillators may be used with the addition of one extra transistor). The single 3 by 2 inch quality PCB incorporates a stable, adjustable Gunn diode modulator with video pre-emphasis and adjustable deviation, together with a stable intercarrier sound generator with selectable pre-emphasis, audio limiting, input level control and output filtering. The PCB requires a single 10 - 15V DC unregulated supply and consumes only 20mA, excluding the Gunn diode. This makes it ideal for portable or mast head use.

PCB with full instructions - you supply components, Gunn Osc, case, connectors, etc..... £ 7.50

Kit with full instructions - you supply Gunn Osc, case, connectors, etc (Note 1)..... £ 25.00

Kit with Gunn oscillator - you supply case, connectors, etc. (Notes 1 & 2) £ 35.00

Note 1: Please state 5.5 MHz, 6.0 MHz or 6.5 MHz audio sub carrier frequency. (6 MHz standard) *Note 2:* At present Gunn oscillators are good tested surplus units. Gunn oscillators are WG16 square flange mounting, 8 - 12mW output and pre-tuned to 10.340 GHz, other frequencies at request. 8 - 12mW oscillators, when equipped with a suitable antenna and a low noise RX are capable of providing line-of-site transmissions well in excess of 100Km. With minor modifications this TX may be used to transmit data.

3cm LNB's

Low Noise Block receive converters have been the main contributive factor that has opened up the 3cm band to ATV and other modes. With the low receive noise figures now available, transmissions are no longer restricted to clear line-of-sight paths. Test transmissions have conclusively shown that over the horizon transmissions by various propagation modes are now possible. None line-of-sight transmission by means of scatter from rain clouds has also been achieved. Equipped with suitable antenna systems, etc., operation well in excess of 100km is readily achievable. Integral feed horn types will fit directly on to standard offset satellite dishes with a 38 - 40mm mount.

Brand new with integral feed horn 0.7dB noise figure £ 50.00

Brand new with integral feed horn 1.0. - 1.2dB noise figure..... £ 45.00

Brand new with integral 20dB antenna 0.8db noise fig. £ 60.00

Brand new 22mm circular input 0.7db noise figure £ 55.00

Brand new 22mm circular input 1.0db - 1.2db noise figure £ 50.00

All LNB's. are fully tested with the local oscillator at 9.0GHz $\pm 0.01\%$ (this is often better than original). Other local oscillator frequencies available by request. Horizontal or vertical polarisation is selected by applying either approximately 13 or 1BV DC.



ASSEMBLED DOVE ATV TUNEABLE IF / RECEIVER KIT

The Dove ATV Tuneable IF/RX forms the ideal heart of an ATV receiving system for either 24cm (see note) or the higher bands when used in conjunction with a suitable Low-Noise-Block Converter. It is designed as a compact easy to construct unit that requires no adjustment, setting up, or alignment. The kit contains all board mounted components and full instructions. The only other requirements are a case, power and output connectors, four potentiometer controls (all 10k lin) and a switch (SPST). The unit has been designed on a flexible modular concept. Plug-in modules are being developed which include: a fully tuneable sound demodulator, plus single and multichannel synthesisers. Unprocessed DC-coupled baseband output is provided for expansion into data reception or multiple subcarrier demodulation. An AGC output is provided for signal strength, etc. A divided by 256 local oscillator signal is also available.

Size:	105mm wide x 97mm deep x 50mm high
Power:	11 - 14v DC @ 400mA
Frequency Range:	925 - 1800 MHz
Sensitivity:	-57dBm (300µV)
Audio Subcarrier:	6 MHz
Video Bandwidth:	25 Hz to 3-5.5 MHz Variable
Video Outputs:	1V P-P Composite Unprocessed Baseband
Audio Output:	1.5 W into 4Ω minimum
Controls:	Main Tuning, Volume, Video Gain, Video Bandwidth, LNB Power ON/OFF
LNB / Preamp Power:	12V DC Switched
Connections:	RF Input F-type, all others 0.1" pitch (supplied)
Mounting:	Fixing pilot hole in each corner

NOTE: When used as a 24 cm Receiver unless the signal is a strong local one a preamp will be required.

DOVE RECEIVER KIT £ 57.00

24cm HEMT GaAsFET ATV PREAMP

Satellite receivers are basically tuneable IF units which require, for satisfactory operation, input signal levels of many millivolts. In a satellite installation most of the RF gain is provided by the dish mounted LNB. This means that if used on the 24cm band, unless they are used only for local working, they will require a high-gain pre-amp at the front end to compensate for the gain normally provided by the LNB. This 24cm low-noise, high-gain pre-amp is designed specifically for ATV use, but may be used on any modes if required. Micro striplines plus Helical band pass filter make alignment very simple and greatly reduces the possibility of interference from out of band signals.

1dB noise figure HEMT GaAsFET front end, 40db gain.

12 - 20V DC supply via coax. (from satellite RX etc.) or separate feed if required. Sturdy tin plate enclosure. High quality BNC connectors. Very simple to align.

Note: This kit contains some surface mount components. TX/RX switching is not included

Complete Kit..... £ 77.50
Fully assembled, aligned & tested..... £ 90.00

Shipping charges at cost

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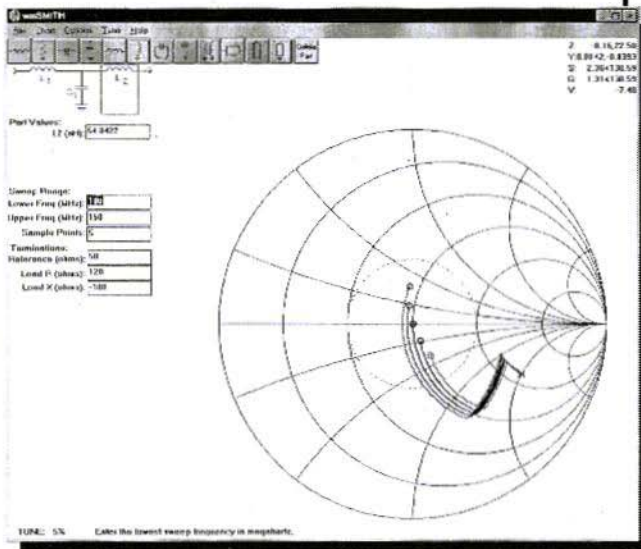
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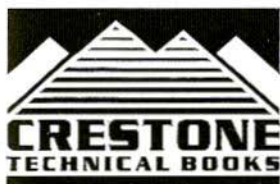


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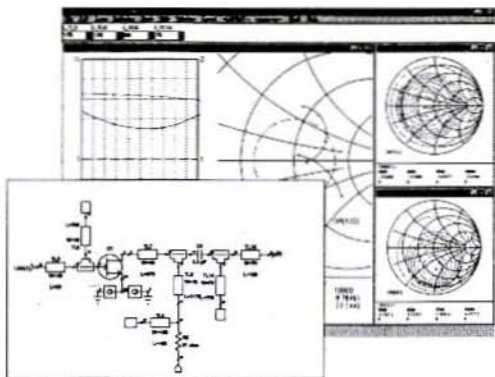
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