



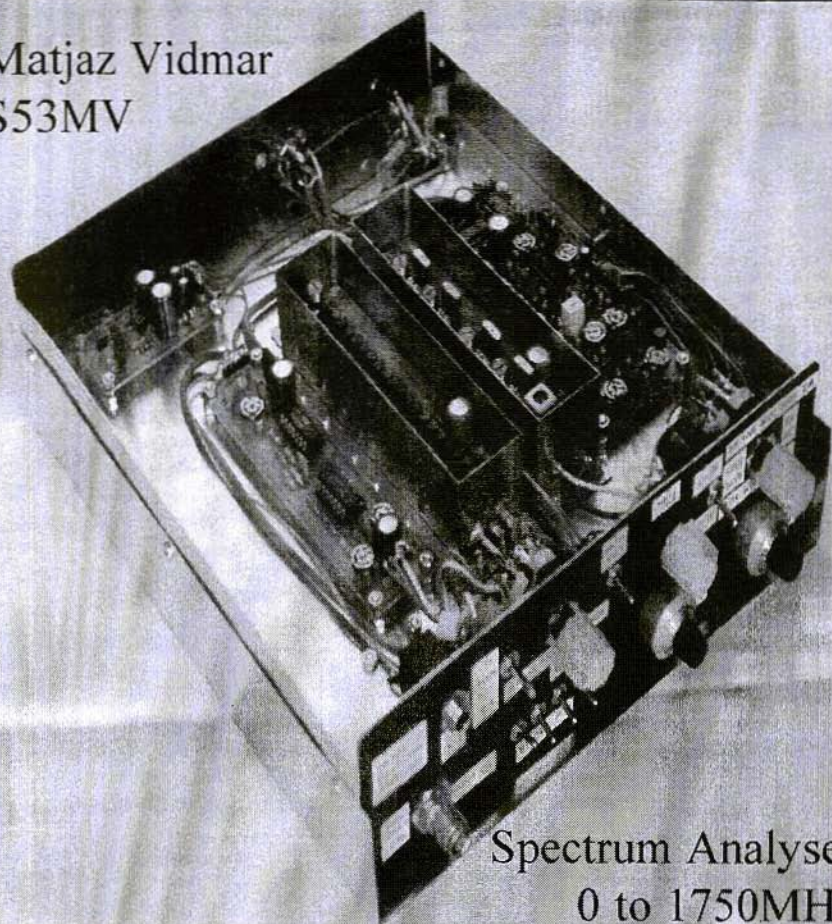
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Matjaz Vidmar  
S53MV



Spectrum Analyser  
0 to 1750MHz

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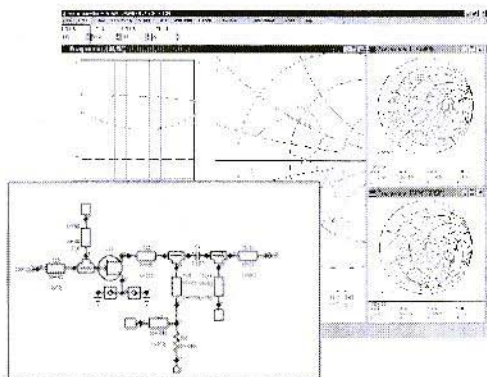
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Matjaz Vidmar, S53MV

# Spectrum Analyser

## 0 to 1750 MHz

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### 1. SPECTRUM ANALYZER DESIGN

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Wideband (panoramic) receivers and RF spectrum analysers are usually designed as multiple-conversion receivers with the first IF above the maximum input-signal frequency. Amateur designs are mainly limited by the performance of the VCO used for the first frequency conversion. The 2-4 GHz VCO presented in [1] and [2] allows the design of a simple spectrum analyser covering the frequency range 0 to 1750 MHz in a single span.

The resulting spectrum-analyser block diagram, shown on Fig.1, was already briefly described in [1] and [2], although only the microwave VCOs were described in detail. In this article the remaining building blocks of the above mentioned spectrum analyser will be described in detail, as well as the overall

assembly and tuning of the completed instrument.

As already mentioned in [1] and [2], the described spectrum analyser is a triple-conversion receiver with the corresponding IFs around 2.1 GHz, 70 MHz and 10 MHz. Since both the first LO and the second LO are VCOs, the first IF may be made variable. This may be useful to shift some spurious responses of the first mixer in some difficult measurements

The described spectrum analyser is designed to use a standard XY oscilloscope display. Additional control signals are provided to drive different displays as well as frame memories, storage normalisers, marker generators and/or trigger frequency counters. The outputs of both VCOs are also made available to drive a tracking generator or a frequency counter. Some of these additional circuits will be described in future articles.

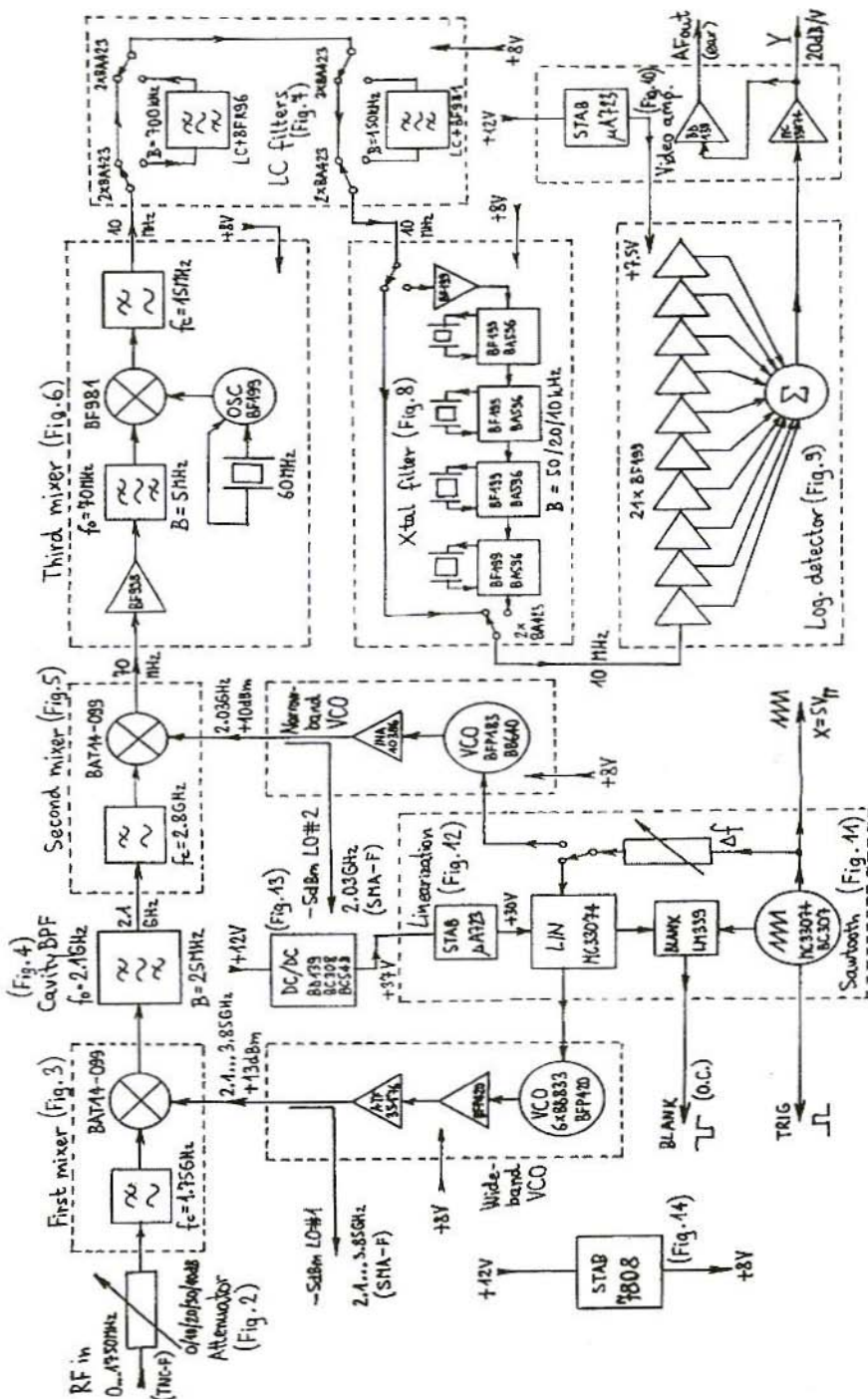


Fig.1: Spectrum Analyser Block Diagram

## 2. INPUT ATTENUATOR

The input attenuator is a simple yet important part of a spectrum analyser. The main function of the input attenuator is reducing the signal level to avoid overdriving the first mixer. The input attenuator should therefore only include very linear components like resistors and mechanical switches.

The basic attenuator circuit is a simple "PI" or "T" resistor network. The resistor values are selected both for the desired attenuation value and input/output impedance matching. Since a "PI" or "T" network contains three independent resistors, all three quantities can be adjusted independently within certain limits: attenuation, input impedance and output impedance.

Generally speaking, attenuators can be built for any frequency provided that the resistors (and switches) are sufficiently small with respect to the wavelength and/or are designed as parts of transmission lines with carefully controlled characteristic impedances. Professional attenuators, both fixed and adjustable, are available for frequencies up to 18 GHz and beyond, depending mainly on the coaxial connectors.

On the other hand, it is much more difficult to build good microwave attenuators from standard electronic components. Resistors with wire leads are only useful up to about 500 MHz. SMD resistors are much better due to their smaller size and can be used up to at least 5 GHz. Finding suitable mechanical switches for microwave frequencies is even more difficult.

The circuit diagram of the attenuator, shown on Fig.2, therefore can not tell much about the microwave performance of the circuit. The attenuator includes four identical "PI" networks and four DPDT switches. This design allows a nominal attenuation of 0dB to -40dB in 10dB steps.

Standard SMD resistors of the size 0805 or smaller are certainly good enough for a 2 GHz spectrum analyser. It is much more difficult to find suitable DPDT switches with low parasitics. In the prototypes standard miniature DPDT toggle switches were used. The resulting frequency response was compensated with 1pF SMD capacitors soldered across the 68ohm resistors. Finally, the toggle switches were installed between two small printed-circuit boards to keep their impedance closer to 50Ω (to be described later in this article).

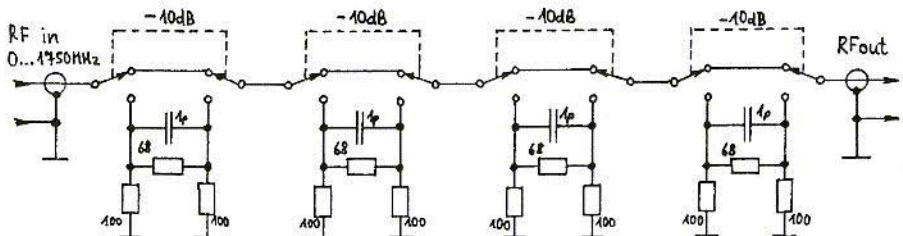


Fig.2: Input Attenuator



A careful construction together with some compensation capacitors allows to keep the frequency response of the attenuator within  $\pm 1$ dB for frequencies up to 1 GHz and within  $\pm 2$ dB for frequencies up to 2 GHz from the nominal attenuation value. Since the maximum power dissipation of SMD resistors is  $1/8$ W, the maximum input power to the described attenuator is about 250mW (+24dBm). At the maximum attenuation value of -40dB this means -16dBm at the first mixer input.

Of course additional (external) high-power attenuators or couplers are required for transmitter measurements. The input attenuator of a spectrum analyser is only intended for fine adjustments of the input signal level and/or finding the source of some signals: real signals or unwanted mixing products in the spectrum analyser?

Finally, the described attenuator does not include any protection against DC voltages that may be present in some RF circuits!

### 3. FIRST MIXER

The dynamic range of any receiver depends mainly on the performance of the first mixer. The low end of the dynamic range is defined by the mixer noise figure (insertion loss) while the high end of the dynamic range is defined by the mixer distortion (intermodulation). Further, the first mixer of a spectrum analyser may be easily destroyed by high RF levels, DC voltages and even static discharges on the input connector.

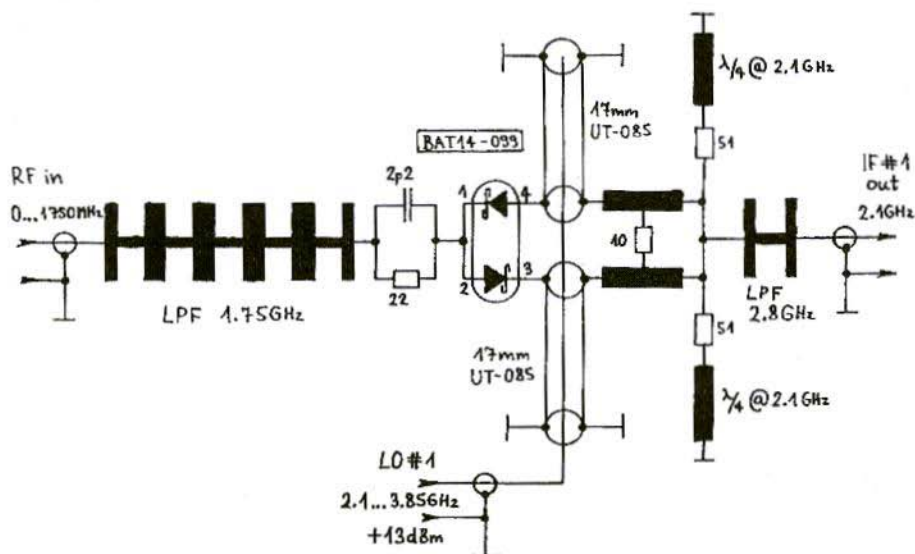


Fig.3: First Mixer

The first mixer of a spectrum analyser should therefore both provide the best possible dynamic range and at the same time allow quick and inexpensive repair in the case of damage. The latter requirement makes commercially-available, doubly-balanced mixers with Schottky quads and ferrite balancing transformers unpractical.

The described spectrum analyser is therefore using an inexpensive double Schottky diode BAT14-099 in the first mixer. The balun for the local oscillator is built from UT-085 semi-rigid cable, as shown on Fig.3. A completely symmetrical construction allows a balancing of at least 30dB without any tuning and more than 45dB by adding small drops of solder in the circuit. Of course the symmetry of commercially-built mixers can not be improved since the latter are built in hermetically-sealed packages.

A good mixer symmetry is required for many reasons. A balanced mixer prevents the oscillator noise from getting directly in the first IF. Further, some unwanted mixing products are suppressed in a balanced mixer. In the case of a spectrum analyser it is especially important to suppress second-order distortion products, so that the spectrum analyser can be used to accurately measure the suppression of the second harmonic in radio transmitters.

The first mixer module includes a microstrip lowpass filter with the cutoff frequency of about 1.75 GHz. The latter should suppress unwanted responses of the spectrum analyser in the microwave frequency range. Its insertion loss amounts to about 45dB at the first IF around 2.1 GHz and becomes even

higher at higher frequencies. The low-pass is followed by a RC network to slightly compensate the slow decay of the sensitivity at frequencies above 1GHz.

The mixer diode BAT14-099 is followed by several components to provide impedance matching and suppression of unwanted resonances. For example, the open end of the UT-085 balun is also used as a coupling capacitor. The two 51Ω resistors and corresponding quarter-wavelength lines provide a termination for the image frequency and other unwanted mixing products, reflected by the 2.8 GHz lowpass and the following cavity bandpass filter.

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#### 4.

### CAVITY BANDPASS FILTER

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In all wideband receivers, the first mixer should be followed by the best possible bandpass filter. In HF receivers covering 0...30 MHz, a 15 kHz wide crystal filter is usually used in the first IF around 45 MHz or 70 MHz. Considering the first IF the described spectrum analyser at 2.1 GHz, a cavity bandpass filter is the only technology that provides both high selectivity and low insertion loss.

Microwave cavities are electrically simple to describe, but usually require lots of precision mechanical work and special tools for manufacturing. Several efforts were spent in finding a reproducible cavity-filter design, made from standard materials using only simple mechanical tools.



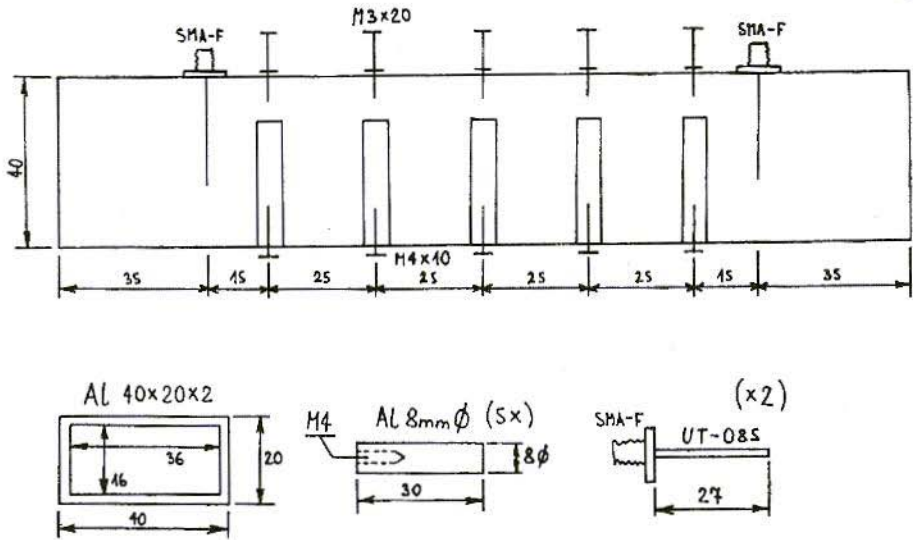


Fig.4: Cavity Bandpass Filter

The described bandpass filter is built inside a piece of standard aluminium tube of rectangular cross section with the external dimensions of 40mm x 20mm and 2mm wall thickness. Such rectangular aluminium tube can be found elsewhere in Europe. Of course, its internal dimensions of 16mm x 36mm are the most important parameter while building a cavity filter.

The construction of the cavity bandpass filter is shown on Fig.4. The filter includes five quarter-wavelength resonators made from 8mm diameter aluminium rod. All five resonators are oriented in the same direction ("comb" filter) to decrease the coupling between adjacent resonators. In this way the overall dimensions of the filter are smaller than in the case of an "interdigital" arrangement of the resonators.

The input and output couplings are made by two small rod antennas, supported by the corresponding SMA con-

nectors. The coupling is adjusted by the length (around 27mm) of the two antennas made of thin copper tube (UT-085 shield). The coupling between resonators is defined by the distance 25mm between resonator centres and sets the filter bandwidth to about 25 MHz.

Five M3 x 20mm tuning screws are used to bring all five resonators to the desired operating frequency. The tuning screws are inserted from the opposite narrow side of the cavity and secured with a lock-nut after tuning.

The cross section of the cavity is small enough that the electromagnetic field exhibits a very fast exponential decay at both ends of the rectangular aluminium tube. Covers are therefore not required for the electrical performance of the filter. On the other hand, covers are useful to keep dust and dirt outside. Covers may extend up to 10mm inside the cavity or stay at least 25mm away

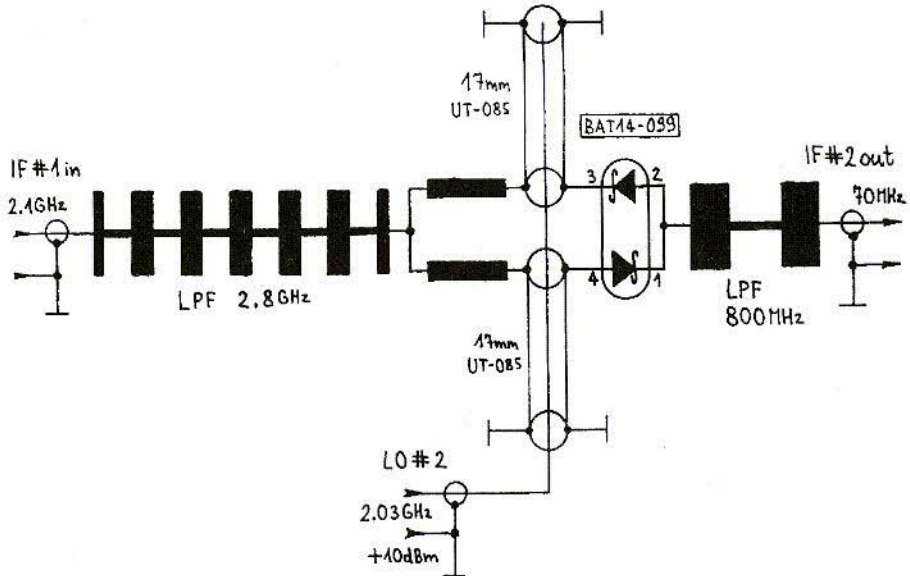


Fig.5: Second Mixer

from the coupling antennas without having any influence on the filter performance.

The described cavity bandpass filter provides over -100dB of suppression for the second-conversion image frequency around 1.94 GHz. The insertion loss is only around -2dB at the nominal first IF frequency of 2.1 GHz and over -100dB outside the passband anywhere between 0 and 4 GHz. Spurious higher-order resonances appear above 4 GHz, when the aluminium tube starts operating as a waveguide. A cavity filter alone is therefore not sufficient. Additional microstrip lowpass filters are therefore included in both the first and second mixer modules to suppress the spurious cavity responses above 4GHz.

The cavity design allows narrowing the passband down to just a few MHz. A bandwidth of 25 MHz was selected to

allow a narrow sweep of the second LO, to avoid some spurious responses of the first mixer and finally to allow for some frequency drift of the second VCO.

## 5. SECOND MIXER

The requirements for the second mixer are not as far as severe as for the first mixer, since most unwanted signals have already been removed by the cavity bandpass filter. Also the signal levels are about -10dB weaker due to the conversion loss of the first mixer and cavity bandpass insertion loss. Calculations and experiments show that no amplifier stages are required between the two mixers if the maximum dynamic range is desired.

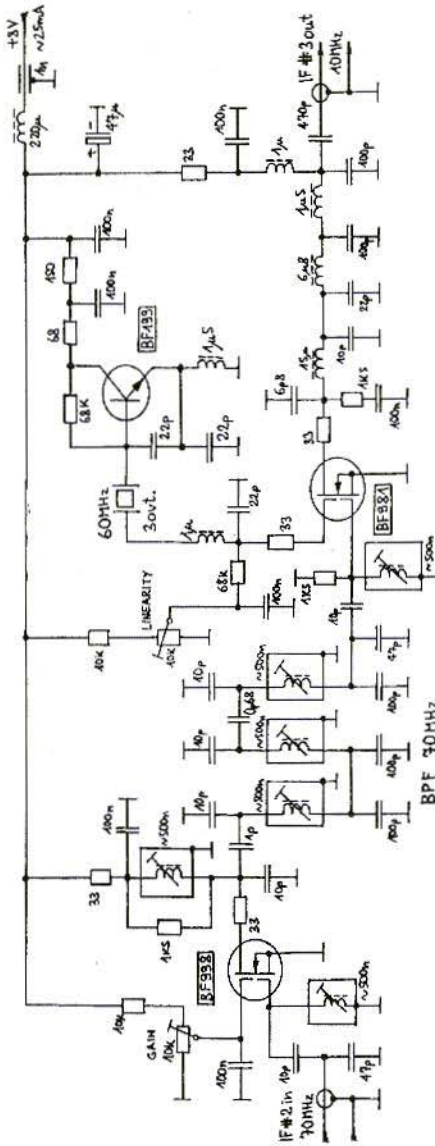


Fig.6: Third Mixer

The design of the second mixer is very similar to the first mixer, except that the input and output are interchanged, as shown in Fig.5. The second mixer is also using a BAT14-099 double Schottky diode and a balun made from UT-085 semi-rigid cable. Since the second mixer operates in a narrow frequency band, damping resistors and other compensation components are not required.

The second mixer module includes low-pass filters both at the input and output. The input lowpass filter cuts above 2.8 GHz to suppress the spurious cavity responses above 4 GHz. The output lowpass cuts above about 800 MHz to suppress unwanted mixing products and feedthrough of the LO signal.

## 6. THIRD MIXER

The first IF of the spectrum analyser around 2.1 GHz is far too high for the different IF filters and logarithmic detector. A more suitable choice for the final IF is 10 MHz. The latter can be conveniently reached from 2.1 GHz in two down conversion steps.

Due to the relatively low frequencies and low signal levels, the requirements for the third mixer are not particularly severe. An additional requirement is the maximum bandwidth  $B = 4$  MHz that requires a carefully designed bandpass filter at 70 MHz and wideband impedance matching at the final IF of 10 MHz.

The circuit diagram of the third mixer and related components is shown in Fig. 6. The circuit includes a low-noise amplifier at 70 MHz, followed by a LC bandpass filter for 70 MHz and a dual-gate MOSFET mixer. The 70 MHz low-noise amplifier (BF998) is the only true amplifier stage in the whole receiving chain of the spectrum analyser. The only purpose of this stage is to compensate for the conversion loss in the mixers. Any gain increase or additional amplifier stages would just impair the dynamic range of the spectrum analyser.

The LC bandpass filter at 70 MHz has two functions. First, the image response of the third mixer at 50 MHz has to be suppressed. Second, the widest IF bandwidth of the spectrum analyser is defined mainly by the 70 MHz bandpass filter. The bandwidth of the 70 MHz LC bandpass filter itself is around 5 MHz, limiting the overall bandwidth of the complete receiving chain to about 4 MHz. The 70 MHz LC filter is built with adjustable coils (about 500nH) wound on shielded supports for IF transformers. The input and output are terminated with 1.5k resistors.

The third mixer is built with a dual-gate MOSFET BF981. The input 70 MHz signal is fed to the first gate while the 60 MHz LO is applied to the second gate. The mixer is followed by a lowpass impedance-matching network. The latter should both remove the 60 MHz LO signal and other unwanted mixing products as well as provide a wideband transformation of the MOSFET high output impedance down to 50Ω.

The design of a suitable lowpass/matching network is complicated, since the required bandwidth is comparable to the centre frequency 10 MHz. Impedance matching is therefore performed in several steps with lowpass LC networks.

The circuit shown in Fig.6 allows reasonable impedance matching in the frequency band 6...15 MHz and a high suppression of the 60 MHz LO at the same time. The circuit is built with fixed inductors of the size and shape of 1/4W or 1/2W resistors. The third LO includes an overtone crystal for 60 MHz and a BF199 transistor. The 1.5uH inductor in the emitter of the BF199 forces the crystal to oscillate on the third overtone. The crystal is also used at the same time as a filter for the output signal fed to the mixer.

The 1uH inductor in series with the crystal further reduces the amount of harmonics fed to the mixer. Finally, unwanted mixing products can be further suppressed by carefully setting the MOSFET bias with the trimmer "LINEARITY".

The trimmer "GAIN" is set for the lowest practical gain of the BF998 that does not impair the noise figure of the whole spectrum analyser.

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## 7. LC FILTERS

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The output of the third mixer can be fed directly to the input of the logarithmic detector, setting the IF bandwidth to about 4 MHz. If a narrower IF band-



width is desired, additional bandpass filters are required between the third mixer and the logarithmic detector. LC filters can be used for bandwidths above 100 kHz, while crystal filters are required for even narrower bandwidths.

A spectrum analyser should include several different IF filters, to be included as required in the IF chain. Filter switching can hardly be performed by standard mechanical switches only, since a crosstalk better than -100dB is required between the input and output of a filter. A better solution is electronic filter switching with PIN diodes. Each filter should also include an amplifier to compensate for the insertion loss of the filter. In this way the measured signal strength will remain the same while switching among different IF filters.

Professional spectrum analysers usually allow the selection of the IF bandwidth in steps of 1/3/10 etc. On the other hand, practical requirements show that just a few different bandwidths are required above 100 kHz. Many different bandwidths are only required below 100 kHz, where the IF bandwidth also defines the sweep time and limits the display update frequency.

The described spectrum analyser includes two different LC filters with the bandwidths set to 700 kHz and 150 kHz. In addition, the crystal filter bandwidth can be adjusted in smaller steps to 50 kHz, 20 kHz or 10 kHz. Finally, with no additional filters an IF bandwidth of 4 MHz is obtained, providing a total selection of six possible IF bandwidths.

The two additional LC filters are shown in Fig.7. Each filter includes a switching network with four diodes BA423. The filter is inserted in the IF chain by simply applying the corresponding +8V supply. At the same time the BF199 transistor cuts the direct signal path and further improves the crosstalk attenuation between the input and output of the filter.

The 700 kHz wide LC filter includes four adjustable coils (about 10uH), wound on shielded supports for IF transformers. The BFR96 amplifier compensates the insertion loss of this filter. Of course the exact amount of gain depends on the loss in the coils and can be adjusted with the 220Ω trimmer.

The 150 kHz wide LC filter includes two separate filters, each including two tuned circuits. The required inductors are in the range of 2.2uH. All four coils are wound on slightly larger shielded supports for IF transformers to achieve an unloaded Q of about 100. The insertion loss of both narrow LC filters is compensated by the BF981 amplifier. The gain of the latter is adjusted by the bias voltage (10kohm trimmer) on the second gate.

The LC filter module requires three different supply voltages. The +8V supply should be present at all times to bypass the filters while the latter are turned off. The +8V/700 kHz supply inserts the 700 kHz wide filter while the +8V/150 kHz supply inserts the 150 kHz wide filter. Both the input and output of the module should remain matched to an impedance of 50Ω at all times.

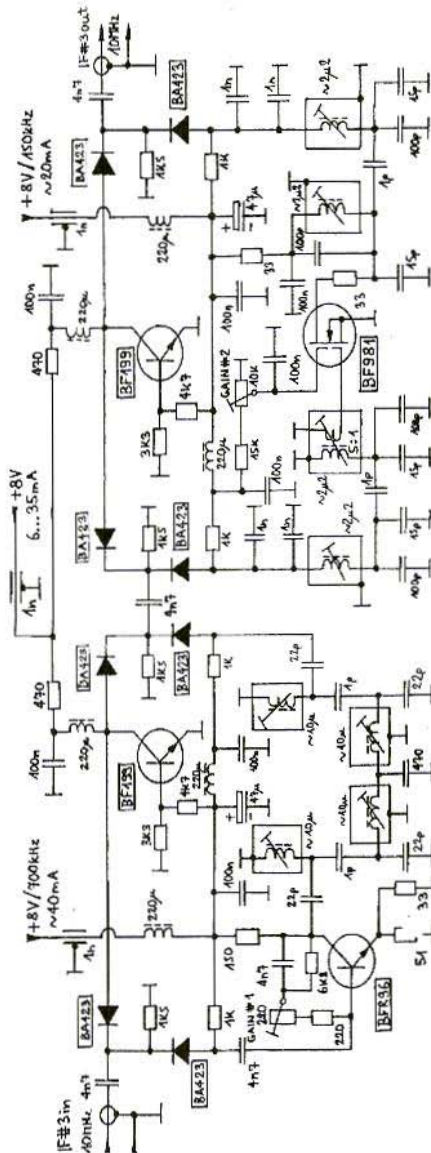


Fig. 7: LC Filters

## 8. CRYSTAL FILTER

Spectrum analysers require somewhat different IF filters than those installed in communication receivers. Communication receivers usually require filters with a flat passband, to avoid modulation distortion, and a very steep increase of the insertion loss immediately outside the useful passband, to reject adjacent channels. Such filters are not suitable for spectrum analysers, since their time response is rather slow (ringing!) compared to the filter bandwidth.

A slow filter response and/or ringing is especially harmful at small bandwidths, where the time response of the filter defines the sweep time and display update period. Commercial crystal and ceramic filters are therefore almost useless in spectrum analysers. A suitable crystal filter or set of different filters has to be specially built for a spectrum analyser.

A spectrum-analyser IF filter should have a "triangular" frequency response with a sharp peak and smoothly and symmetrically increasing attenuation outside the passband. In practice this requires under critically-coupled resonators or better a series connection of several single-resonator filters and buffer amplifiers, to avoid any interaction among the resonators.

The crystal filter shown in Fig.8 includes a series connection of four independent, single resonator filters. BF199 emitter followers are used to avoid unwanted coupling among the crystals. Each individual filter includes a

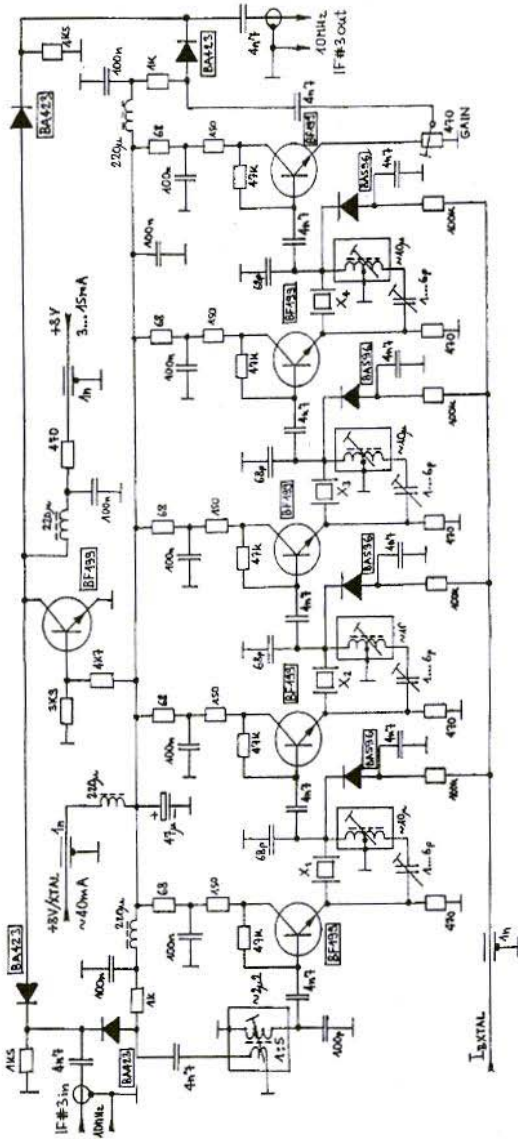


Fig.8: Crystal Filter

single crystal in a balanced network. The capacitive trimmer is used to compensate the capacitance of the crystal. The 10uH centre-tap coil resonates with the 68pF capacitor and compensates the remaining parasitic capacitors of the circuit.

The bandwidth of a single-crystal filter depends mainly on the source and load impedances. The source impedance is kept low by the previous emitter follower. The load impedance is adjustable, since a PIN diode BA596 is connected in parallel to the input of the following emitter follower. The filter bandwidth is therefore adjustable with the DC current IBXTAL fed to the four BA596 PIN diodes.

The crystal filter module includes a switch with four diodes BA423 to insert the filter in the IF chain. The supply voltage +8V should be present at all times to bypass the crystal filter while the latter is turned off. The supply voltage +8V/XTAL inserts the crystal filter. The insertion loss of the latter is mainly compensated by the emitter followers.

Some additional gain is provided by the 1:5 step-up transformer at the input, followed by the first emitter follower. The overall gain is adjusted by the 470Ω trimmer on the output.

## 9. LOGARITHMIC DETECTOR AND VIDEO AMPLIFIER

A spectrum analyser differs from a communication receiver also in the type of detector used. Spectrum analysers use a logarithmic amplitude display with a very wide span of 80...100dB while communication receivers use linear detectors. A spectrum-analyser display is therefore more similar to the operation of the S-meter of a communication receiver. Some inexpensive spectrum analysers (Hame ) and many amateur designs in fact use the S-meter output of popular FM-demodulator chips as a logarithmic detector. This is a very poor technical solution, since the S-meter output of most FM-demodulator chips is very inaccurate and has wide deviations from the ideal logarithmic curve. Further, the dynamic range of most single-chip logarithmic detectors hardly exceeds 70dB. Chips with a wider dynamic range require interstage bandpass filters to limit the wideband noise.

A logarithmic detector for a serious instrument therefore has to be built from discrete transistors, resistors and capacitors as shown in Fig.9. The circuit includes a series connection of ten identical stages, operating as limiting amplifiers and as linear detectors at the same time. Each stage includes a balanced amplifier, allowing a very high overall gain without instabilities and a precisely-defined saturation mechanism. The logarithmic response is obtained as a sum of linear responses that saturate above a certain level, provided that the

gain of each single stage does not exceed 10...12dB.

Similar balanced amplifier chains are used as FM limiters in all known integrated circuits. The main difference between the FM chips and the discrete circuit in Fig.9 are the 470pF emitter-coupling capacitors. The latter can not be built inside a monolithic integrated circuit. The emitter-coupling capacitors allow a simple setting of the bias point and centre frequency of the logarithmic detector.

Since the described logarithmic detector does not amplify DC or very low frequencies, "1/f" noise can be avoided resulting in an up to 20dB increase of the dynamic range. The noise level of the described detector is around -105dBm, while the logarithmic response is acceptable up to -10dBm, resulting in a dynamic range of at least 95dB.

Another advantage of the described circuit is that the output does not saturate but becomes linear for an additional increase of the input signal level of 10...15dB beyond the nominal logarithmic response. The latter is very useful to clearly indicate excessive signal levels on a spectrum-analyser display.

A drawback of the described circuit is a rather low output voltage, only around 100mV difference for a dynamic range of almost 100dB.

The output video signal is therefore related to a reference voltage provided by a dummy BF199 amplifier, operated at the same current as the remaining twenty BF199s used in real amplifier









The centre frequencies of both VCOs are set with 10-turn Helipot. The span width is set with a selector in steps 1/2/5/10 etc. At 500 kHz/div and narrower spans the second VCO is swept, while the frequency of the first VCO can be additionally stabilised with the STAB VCO#1 switch.

Finally, the zero-span function is activated by yet another switch that removes the sawtooth and connects 47 $\mu$ F capacitors in parallel with the control inputs of both VCOs.

As already described in [1] and [2], the frequency response of the wideband VCO with six BB833 varactors is quite non-linear. The tuning slope is the steepest around the centre frequency of the VCO coverage, where the tuning slope reaches 120 MHz/V at a tuning voltage around 7V.

At lower voltages the tuning slope decays slowly to about 90 MHz/V on the lower end. The tuning slope is much more non-linear in the upper part, where the tuning slope may decay below 10 MHz/V at the maximum frequency.

The lineariser circuit includes two operational amplifiers. A negative feedback defines the minimum gain at an output voltage of about 7V. At lower or higher output voltages some positive feedback is switched in through resistive dividers and diodes. The positive feedback increases the gain of the amplifier to compensate for the decay of the tuning slope of the wideband VCO.

Linearisation of the narrowband VCO is not required, since the latter is only swept over a small fraction of its frequency coverage. A single opera-

tional amplifier is required to drive the narrowband VCO, since the polarity of the sawtooth has to be inverted considering the frequency conversions used in the spectrum analyser. For the same reason it makes sense to interchange the connections to the Helipot that sets the centre frequency of the narrowband VCO (+30V on the beginning and ground on the end of the Helipot scale).

The blanking generator includes a quad voltage comparator LM339. Two comparators are used just as amplifiers for the trigger pulses to enable blanking during retrace.

The other two comparators check the control voltage of the wideband VCO, turning the display off when the VCO is tuned out of range.

The LM339 comparators have open-collector outputs and the final blanking signal is obtained from a wired-or connection of three LM339 outputs.

The lineariser (another MC33074) and blanking (LM339) circuits receive the same +30V supply as the sawtooth generator. The +7V reference voltage is obtained directly from the uA723 regulator on the same printed-circuit board.

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## 11. POWER SUPPLY

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The spectrum analyser is designed for a nominal DC supply voltage of 12V DC, negative grounded, with reasonable tolerances (10...15V). Of course, internally the spectrum analyser requires many





different supply voltages. There are also differences in the stability requirements. Some modules require very stable and well-filtered supply voltages while others are more tolerant.

Varactor diodes inside both VCO's require very stable voltages up to about +30V. Therefore the corresponding drivers require a very stable +30V supply. Considering the voltage drop in the uA723 regulator and BD139, a DC/DC converter with an output voltage of about +37V is required, as shown in Fig.13. The DC/DC converter includes a power oscillator with the transistors BC308 and BD139. Voltage spikes from the 100uH inductor are rectified by the 1N4148 diode to charge the 47uF electrolytic capacitor. When the voltage on the latter exceeds +37V, the two 18V Zener diodes turn on the BC548 to stop the oscillator.

Of course, a DC/DC converter is a potential noise source to be built inside a sensitive piece of test equipment. Both input and output of the DC/DC converter are well filtered with the 39uH and 1mH chokes as well as several electrolytic capacitors. The DC/DC converter is further built on its own printed-circuit board to be installed far away from the more sensitive circuits of the spectrum analyser.

The power supply of the remaining circuits of the spectrum analyser is shown in Fig.14. The input voltage +12V is first filtered with the VK200 choke and then fed to the DC/DC converter, to the video amplifier and to the 7808 regulator. The latter provides a stabilised +8V supply for most circuits of the spectrum analyser. The 7808

regulator uses the baseplate of the spectrum analyser as a heatsink and is installed together with the VK200, 470uF electrolytic capacitors and protection diode close to the +12V supply connector.

Fig.14 also shows the IF-bandwidth switching. The latter requires a two-section, six-way selector. One section of the selector is used to switch the +8V supply voltage to different filters. The two 1N5818 Schottky diodes keep previous filters inserted while new filters are added in the IF chain. The second section of the selector sets the current in the PIN diodes that define the bandwidth of the crystal filter.

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## 12. CONSTRUCTION TIPS

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A spectrum analyser is a sensitive receiver operating over a wide frequency range. The circuits of a spectrum analyser therefore require considerably more and better shielding when compared to conventional communication equipment. Better shielding also requires a larger number of shielded modules that contain just a few electronic components each. Besides shielding, microwave absorbers are required outside and inside some shielded enclosures.

The spectrum-analyser modules can be roughly divided in three groups: (1) microwave modules built as microstrip circuits inside shielded boxes, (2) IF modules built on single-sided printed-circuit boards inside shielded boxes and

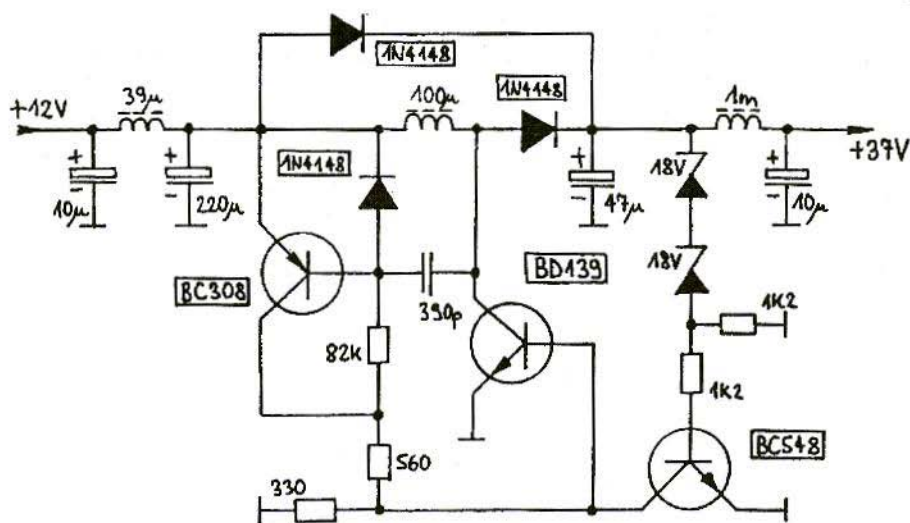


Fig.13: DC/DC Converter

(3) video/supply modules that do not require special shielding.

The microstrip printed-circuit boards (except the VCOs published in [1] and [2]) are shown in Fig.15. All microstrip circuits are etched on double-sided, 0.8mm-thick FR4 laminate. Only the upper side is shown on Fig.15, since the bottom side is not etched to act as a groundplane.

Only SMD components of the size 0805 or smaller should be installed on the microstrip circuit boards. The SMD components are grounded through 2.5mm diameter holes, covered with 0.1mm-thick copper foil on the ground-plane side and filled with solder before installing the SMD components. The completed microstrip boards are soldered in brass frames and covered with

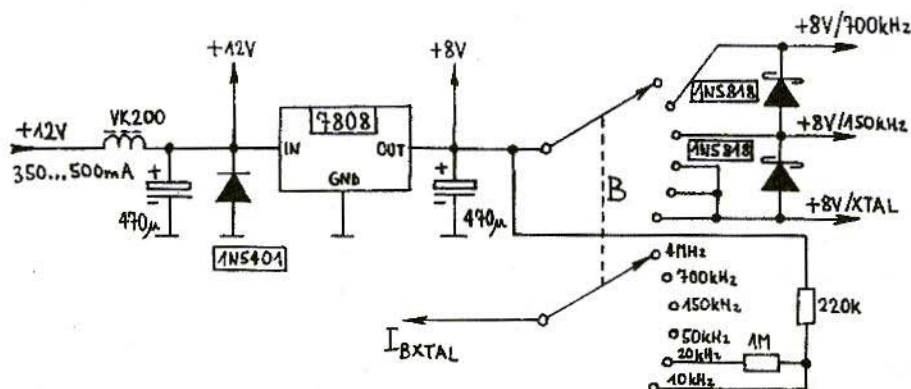


Fig.14: Power Supply

a shielding cover. A piece of 1cm-thick microwave absorber foam should be installed under the whole surface of the shielding cover to suppress parasitic resonances of the shielding enclosure.

The IF printed-circuit boards, shown in Fig.16, are all etched on single-sided, 0.8mm-thick FR4 laminate. SMD components should be soldered first: resistors, capacitors and semiconductors.

The printed-circuit boards are designed for SMD-component sizes 0805 or 1206. Inductors, IF transformers, trimmers, BF199 transistors, BA423 diodes, crystals and electrolytic capacitors are installed as conventional "through-hole" components from the other side.

The IF printed-circuit boards are also soldered in brass frames just like the microstrip boards. Of course, the IF modules do not require any microwave absorber foam under the cover nor any special shielding on the bottom. The supply voltages go through feedthrough capacitors soldered in the narrow sides of the brass frames while the IF cable shields are soldered directly to the brass frames.

The video/supply printed-circuit boards, shown on Fig.17, are all etched on single-sided, 1.6mm thick FR4 laminate. All of the components are conventional types with wire leads for "through-hole" installation. The resistors and diodes are installed vertically on the sawtooth and linearisation module to save some space. The video/supply modules do not require any shielding and are installed on the baseplate with four M3 screws in the corners. All connections go through simple connectors made from quality IC sockets.

The shielded enclosures of the RF/IF modules are shown in Fig.18. The input attenuator includes four miniature DPDT toggle switches installed in a sandwich between two microstrip boards. The microstrip boards are soldered to an "U"-shaped piece of 0.5mm-thick brass sheet. The four toggle switches are installed in 5mm diameter holes, while two 3.2mm diameter holes are provided for the input and output coaxial cable. The input attenuator does not require a shielding cover.

Both microwave mixers and all four IF modules are installed in a brass frame with the dimensions 120mmX30mm, while the two VCOs are installed in similar frames with the dimensions 80mm x 20mm. All frames are made from 0.5mm-thick brass sheet. Covers are also made from brass, although somewhat thinner brass sheet (0.4mm or even 0.3mm) is sufficient.

Both microwave mixers include baluns made from UT-085 semi-rigid cable. The shield of the latter is first tinned for a length of about 35mm to let expand the Teflon dielectric. Then the shield is cut 17mm from the open end and pulled away to obtain a 1mm-wide gap. Such a prepared cable end is then soldered on the microstrip circuit board and the board is then soldered in the brass frame. The balun cable pops out of the frame for about 2mm on one side, while a longer cable (20-25mm) is available on the other side to connect the VCO.

The overall spectrum-analyser module location is shown in Fig.19 inside an aluminium box with the following dimensions; width 220mm, depth 240mm and height 65mm. A central baseplate



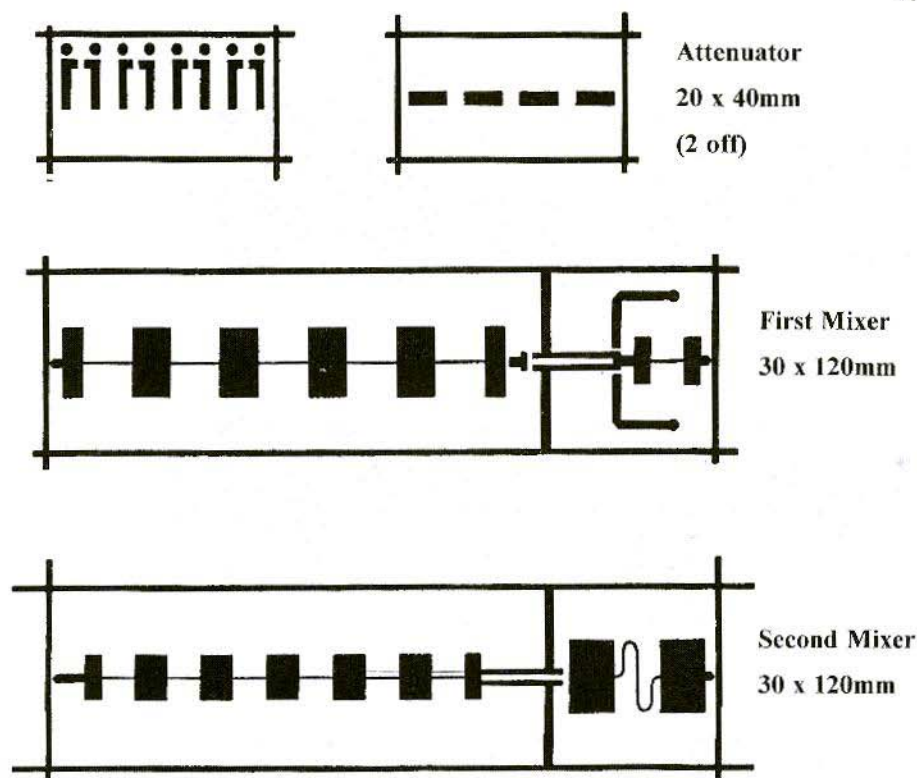


Fig.15: Microstrip Layouts

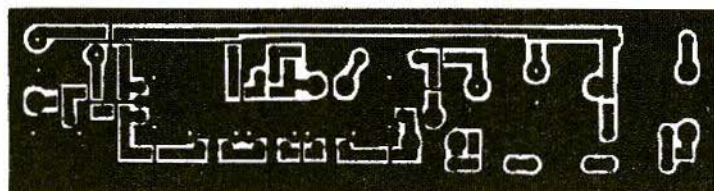
(1mm thick aluminium) holds all of the modules and divides the useful volume in two sections, each 32mm high.

The top section may require a piece of microwave absorber between the second and third mixers.

While positioning the different modules on the baseplate, care should be taken that the mounting screws do not interfere with the modules on the other side. The cavity is held in place by a pair of screws on each "cold" end, so that no metal parts enter the "hot" central part of the filter. The front panel is connected to the baseplate with the TNC

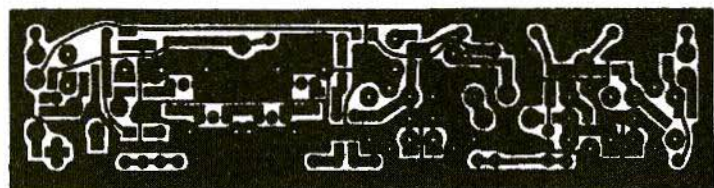
connector, toggle switches and Helipot, so no additional screws are necessary. On the other hand, the back panel is connected with four M3 screws in addition to the SMA connectors.

Finally, it is convenient that all display connections are brought to a single multipole connector on the back panel. The suggested 6-pole, 270° DIN connector is shown in Fig.20. Different displays therefore only require special cables. The same display connector also carries the +12V supply for a storage/normaliser unit or marker generator.



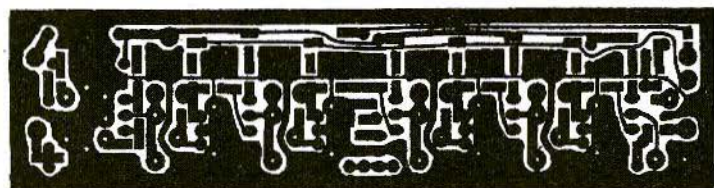
Third Mixer

30 x 120mm



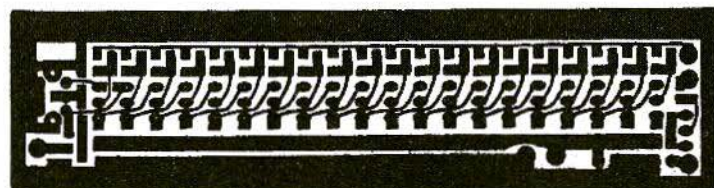
LC Filters

30 x 120mm



Xtal Filter

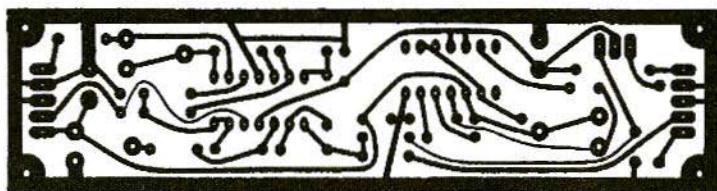
30 x 120mm



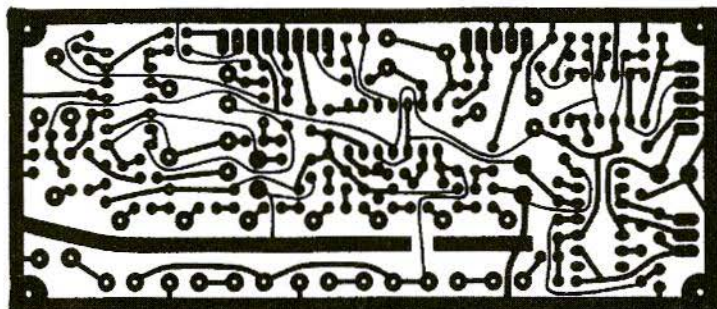
Log Detector

30 x 120mm

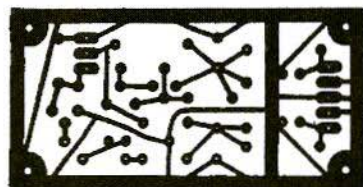
Fig.16: IF PCB Layouts



Video Amp  
30 x 120mm



Sawtooth &  
Linearisation  
50 x 120mm



DC/DC Converter  
30 x 60mm

Fig.17: Video & Power Supply PCB Layouts

### 13. ALIGNMENT AND TESTING

Before starting the construction of a complex piece of test equipment like a spectrum analyser, some simpler test equipment should be available as well as enough practical experience to efficiently use the latter. Besides a conventional analogue AVO-meter (digital

AVO-meters are useless for tuning), a grid-dip meter, a microwave frequency counter up to at least 4 GHz and a microwave power sensor should be available. Last but not least, the oscilloscope to be used as the display of the spectrum analyser should also be available.

The construction of the spectrum analyser should start with the most difficult part: the VCOs described in [1] and [2].

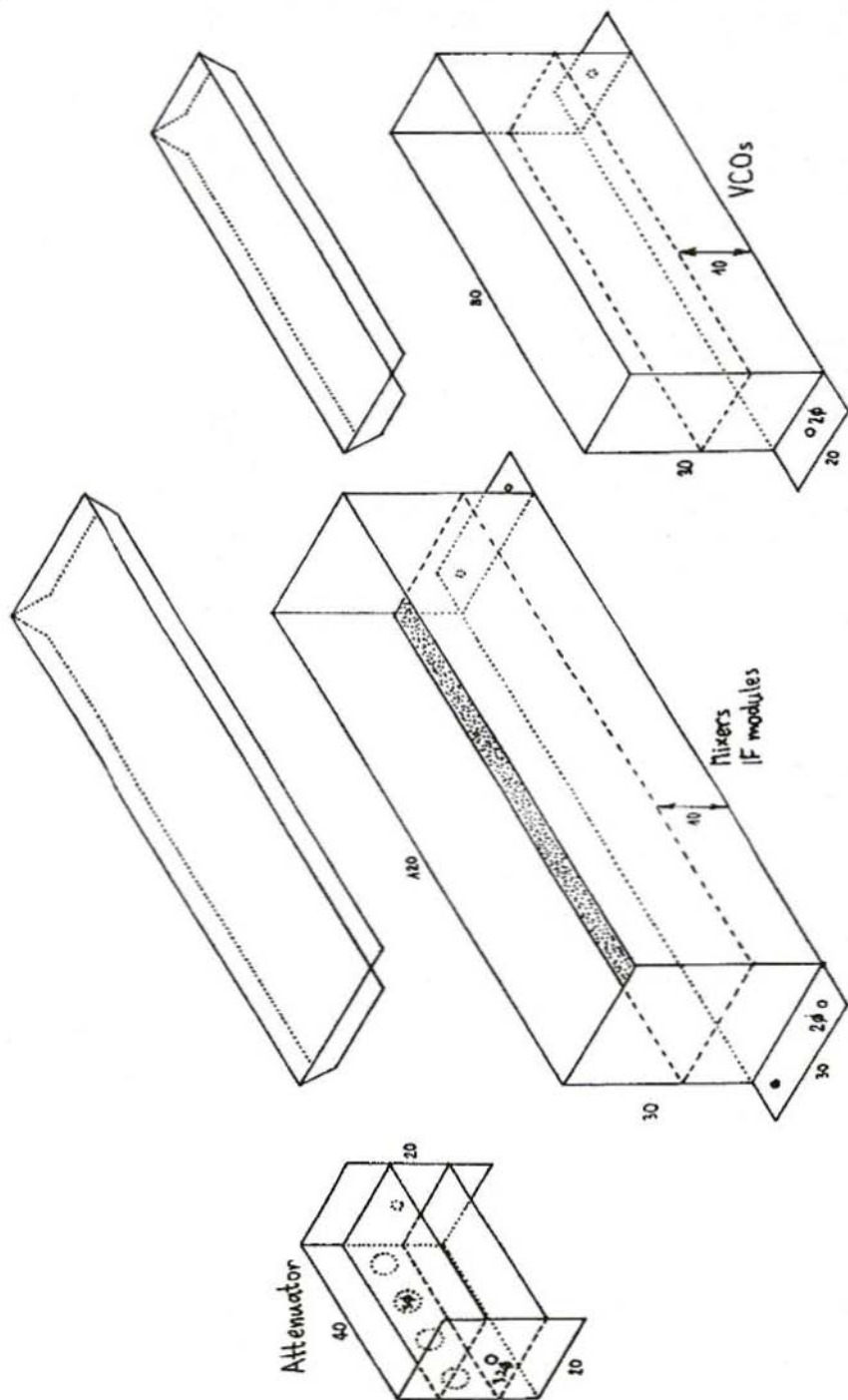


Fig.18: RF &amp; IF Module Shielding



Both VCOs have to be checked for the frequency coverage and output power. In fact it makes sense to build two wideband VCOs. The best VCO is installed in the spectrum analyser. The leftover VCO can still be used in other test equipment, like a harmonic converter that extends the coverage of a spectrum analyser, to be described in a later article.

When both VCOs are operating correctly, the remaining modules can be built except the LC and crystal filters. The third mixer and logarithmic detector can be aligned and checked with a grid-dip meter. The operation of the sawtooth generator, linearisation circuit and several selectors and potentiometers can be checked with the oscilloscope that will be used for display. All the trimmers in the linearisation should be set to their "cold" ends.

The cavity bandpass filter is the unit that is most difficult to align without complex test equipment. The tuning screws are initially set about 1...2mm above the "hot" ends of the resonators. The narrowband VCO is set to 2.03 GHz with a frequency counter while the wideband VCO is sweeping the whole frequency band. The cavity filter can then be tuned to the characteristic "DC peak" of all RF spectrum analysers.

When the "DC peak" is found and the cavity bandpass is roughly tuned to the correct frequency, an initial checkout of the spectrum analyser can be performed. A short antenna connected to the RF input will provide many peaks around 100 MHz (FM broadcast) and 900 MHz (GSM). Even better is a simple 100 MHz test source built with a 100 MHz

ECL crystal oscillator, as shown on Fig.21.

The adjustment of the lineariser requires several signals on known frequencies. The simplest solution is a comb generator (shown in Fig.22) that provides many equally-spaced harmonics of the same fundamental frequency. The usual image on the display is rather dense lines between 700 MHz and 1000 MHz, while the spacing of the lines gradually increases both towards lower frequencies and even more towards higher frequencies. The eight trimmers in the lineariser are then adjusted to obtain equally spaced lines across the whole display. If the adjustment range of the trimmers is not sufficient, other resistors in the circuit may be changed to match the response of the varactors in the wideband VCO.

When the spectrum analyser is operating correctly with a 4 MHz IF bandwidth, additional LC and crystal filters can be built. Finding suitable filter crystals is not simple. Cheap "clock" or "baud rate" crystals were found useless, since they have many spurious resonances close to the main resonant frequency. Such crystals may still provide some useful performance in narrow SSB filters, but their response is badly distorted in wide ( $B=50$  kHz) filters for spectrum analysers.

The only reliable crystals are those found in old 35 kHz-wide filters for 50 kHz FM-channel spacing. These crystal filters were usually built for a centre frequency of 10.7 MHz. One filter contains four different crystal pairs. Therefore two identical crystal filters have to be disassembled to obtain

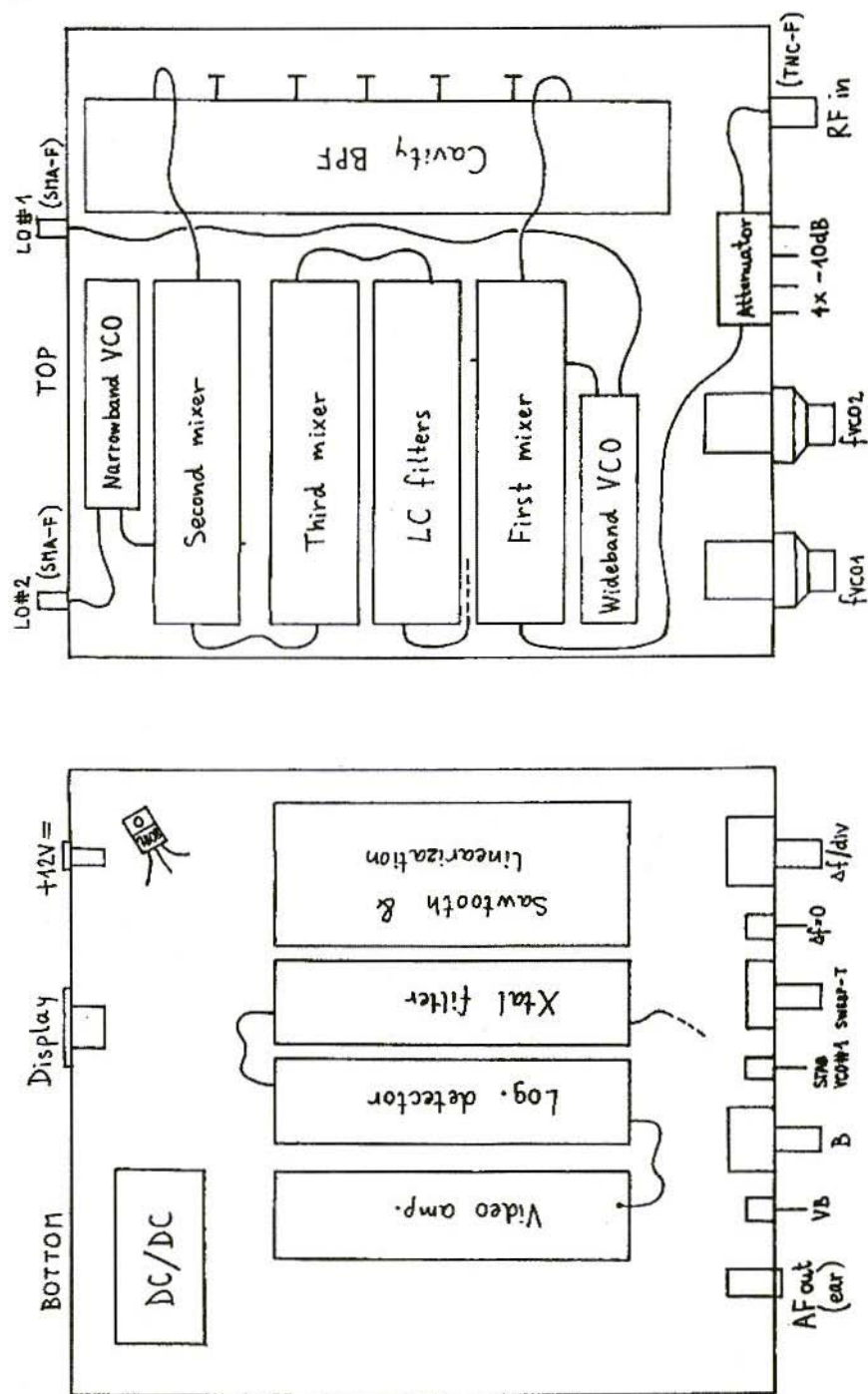
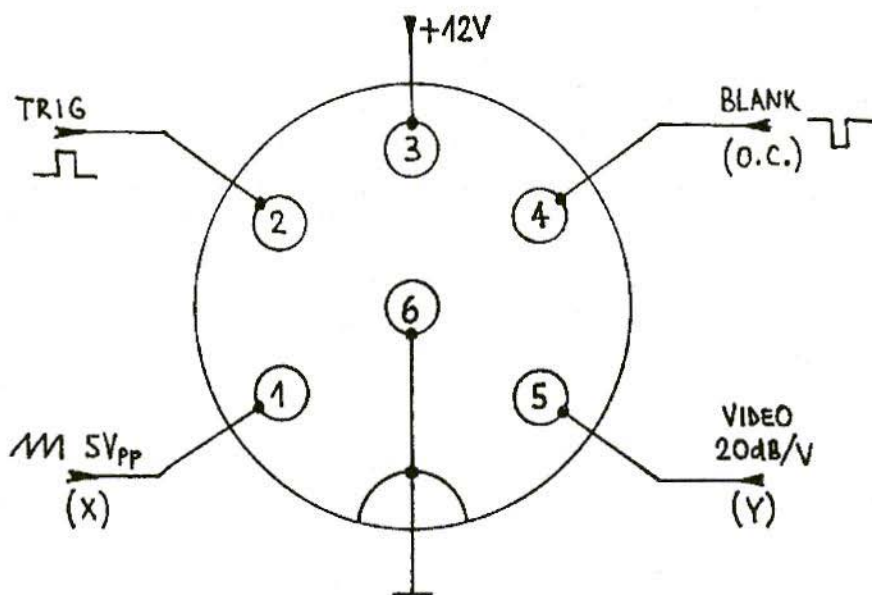


Fig.19: Spectrum Analyser Module Layout



**Fig.20: Display Connector**

enough crystals for four spectrum analysers.

Both crystal and LC filters are simply adjusted on the "DC peak", when the remaining parts of the spectrum analyser are already operating. The alignment of LC filters is straightforward and the desired response is obtained easily.

On the other hand, the alignment of the crystal filter is very demanding. Each filter stage has to be aligned independently while the remaining three crystals are shorted with  $100\Omega$  resistors. A symmetrical response of a single filter stage is sought with the capacitive trimmer. Next the coil is adjusted to obtain the broadest peak, of course with the current in the PIN diode turned off. This procedure has to be repeated several times for each stage, since the two adjustments interact.

After the LC and crystal filters are roughly adjusted, the two trimmers in the third mixer module can be adjusted. The mixer linearity is adjusted with a strong (around  $-10\text{dBm}$ ) signal on the first mixer input. Setting the resolution bandwidth to 150 kHz, some spurious responses may be visible  $\pm 2$  MHz around the desired signal. These spurious responses should be minimised by adjusting the mixer linearity. Finally, the gain of the 70 MHz amplifier is set so that the front-end noise just overrides the detector noise with the 150 kHz IF filter.

At the end, there are many fine adjustments to be performed. All IF filters should be aligned to the same frequency and same insertion loss. If the noise floor changes with frequency, then the symmetry of the first mixer should be improved by small drops of solder on

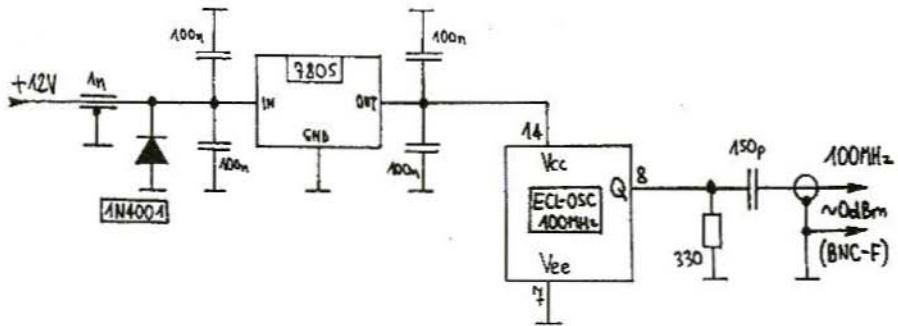


Fig.21: 100 MHz Test Source

the UT-085 balun. The remaining trimmers in the sawtooth and linearisation module have to be adjusted so that the display matches the selected span width.

Of course, a number of fine adjustments are only performed after spotting problems in the practical use of the spectrum analyser. Some adjustments, like the precise impedance matching of the wideband VCO to the first mixer, can only be performed with a tracking generator, to be described in a future article. In any case it is assumed that the builder is familiar with the theory of operation of a spectrum analyser, since

in this case most troubleshooting can be performed with the same spectrum analyser.

## 14. REFERENCES

- [1] Matjaz Vidmar: 'Rauscharmer, breitbandiger Mikrowellen VCO', pages 26-35/4-98, AMSAT-DL Journal, Marburg.
- [2] Matjaz Vidmar: 'Wideband & Low-Noise Microwave VCO', VHF Communications 4/98

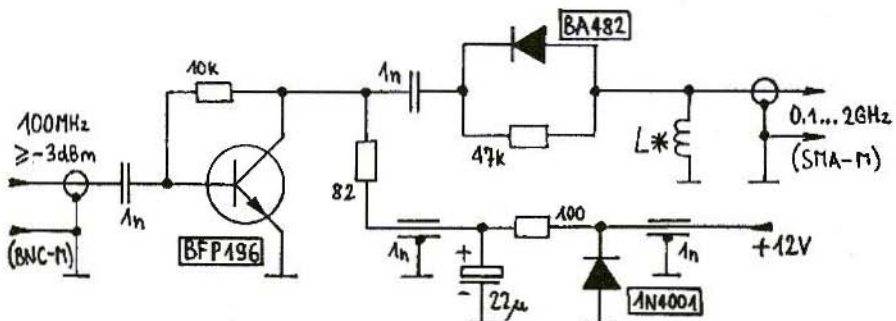


Fig.22: Comb Generator up to 2 GHz





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# Tips On and Improvements to: HF Synthesiser from Bernd Kaa, DG4RBF (Issues 2 and 3/1998)

---

## 1. WIRING DIAGRAM

---

The individual assemblies are described in detail in the above-mentioned article. Fig. 1 gives a clear representation of how they are wired up to one another. This makes the wiring easier to copy.

---

## 2. TIPS ON WIRING

---

### PLL / HF section:

The output (B) of the relay for the voltage change-over [K14] must also be connected to the two other points (B). These are [K15] the input of the 5-V stabiliser for the VNA25 and [K13] PIN diode switch, band B.

The output (A) of the relay for the voltage change-over [K10] must also be connected to the PIN diode switch, band A [K12].

The LO (1,404 MHz) should be powered through [K10] from the HF section, so that it is shut down when band (B) is in operation, and does not generate any undesirable signals.

If an attenuator (0-70 dB) is used as an option, the HF outputs of band (A) and band (B) should be connected together with a coaxial relay.

### Issue 3/1998, pages 165/166:

The output of the power control [K2] Fig. 23 is connected with the U rule, Fig. 25.

The output of the power detector Fig. 26 (to IN control circuit) is connected to the input of the power control [K1] Fig. 23.

### To Fig. 25:

The components in the frame are the new components. The components outside the frame are already on board DB1NV 012.

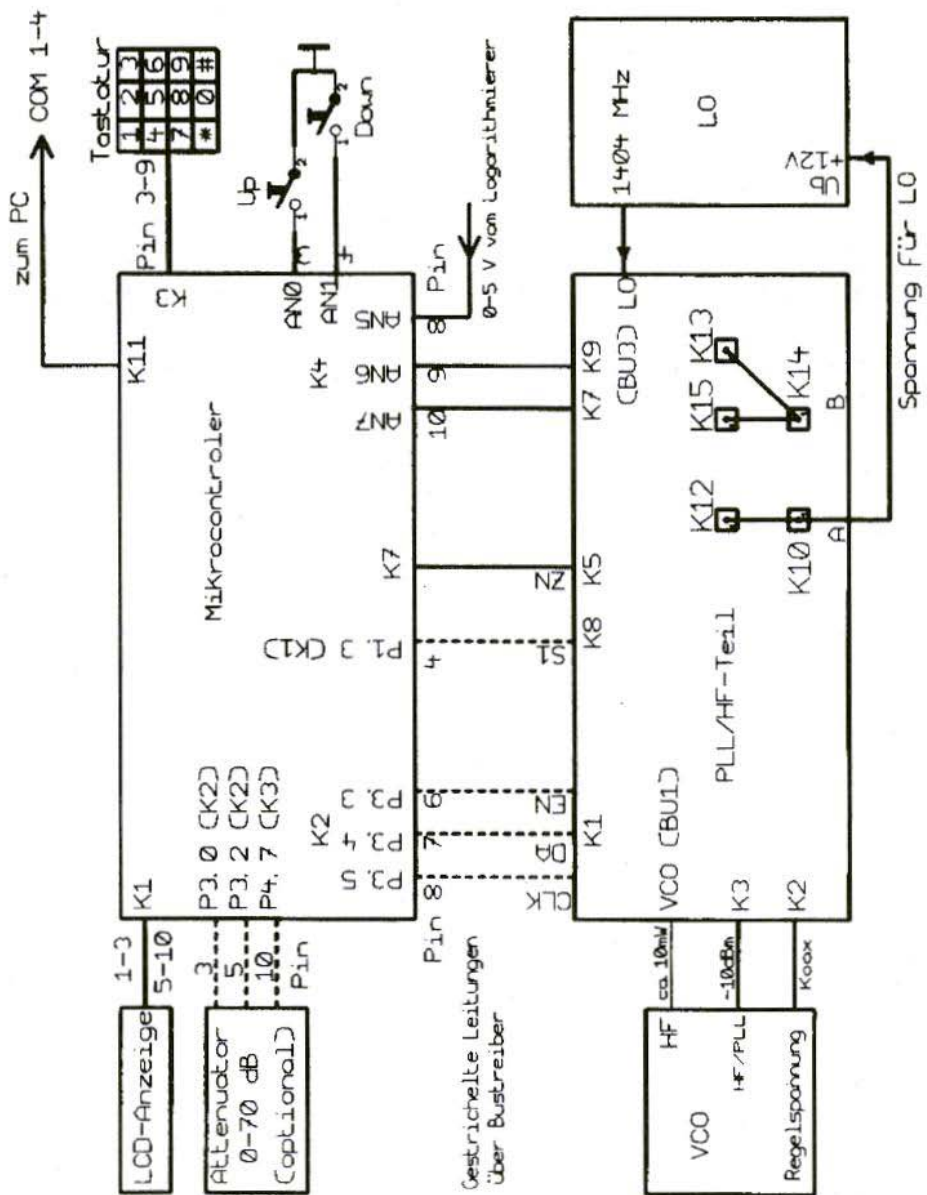


Fig. 1: Wiring Diagram for HF Synthesizer up to 1,450 MHz as per DG4RBF  
*Anzeige* = Display, *Gestrichelte Leitungen über Bustreiber* = Dotted circuits through bus driver, *Regelspannung* = Control voltage, *Mikrocontroller* = Micro-controller, *Spannung für* = Voltage for, *Zum* = To, *Tastatur* = Keyboard, *Vom Logarithmierer* = From logarithmiser



### 3. ADDITIONAL CIRCUIT TIPS

#### Issue 2/1998, page 116:

To sections 3.1.4.3. - Power recognition band B - and 3.1.4.5. - Power recognition band A: a voltage of 1.20 Volts should not be attained here. There are divergent values in the component tolerances for BAT 45. It may also help to adjust R54/R55.

A voltage of app. 1 Volt is not sufficient, so that complete resolution is not attained.

It is important to make sure that the voltages set for band A and band B are the same, since otherwise the power control for band A will diverge too far from that for band B.

### 4. ADDITIONAL TIPS FOR OPERATING SOFTWARE

#### Calibrating power:

Significance of numbers displayed during calibration procedure, e.g.:

Up/Down ok = #	
P: 10 mW	045/ 078

Voltage for PIN  
diode controller  
(0-255)  
(UP/Down)

Voltage detected  
on diode which is  
stored (0-255)

#### Frequency response:

Significance of number displayed during calibration procedure, e.g.:

Up / Down ok = # 097
P: 10 mW / 800 MHz

Voltage detected  
on diode which is  
stored (0-255)

### 5. CORRECTIONS

**On p. 107, last paragraph, the correct description is:**

... the pulse is direct (T1 and LED D9)... instead of (T4 and LED D1),... or inverted (T1, T2 and LED D10)... instead of (T4, T5 and LED D6).

Thus: at the PMB2306 we have T1, T2 and D10. Therefore only one LED is required, either D9 or D10!

**In addition, in Fig. 4 on p. 107:**

the value for R15 =  $\Omega$  has been omitted.

**On p. 165, second paragraph, right-hand column;**

the correct wording should be...AD converter (AN 5) ... and not AN7.

**In the components list on p. 113:**

4 x BC 548B are listed, which should be 3 x BC 548 B and 1 x BC 558 B.

Richard A. Formato, Ph.D., WW1RF

# Improving VHF Antenna System Performance with High Impedance Yagis

---

## 1. INTRODUCTION

---

One of the reasons that building good antennas is a challenge is that key parameters are often incompatible. This note considers the incompatibility between very efficient transmission lines, which have a high characteristic impedance, and typical Yagis, which have low input impedance. Antenna optimisation often ignores the transmission line. It is selected only after the Yagi design is complete, and it is almost always coax. The antenna/transmission line system is then made to work by adding a matching network. This approach is not necessarily the best. Designing the antenna and transmission line together may provide better overall performance, a point which is illustrated by the 12 element Yagi example discussed below.

---

## 2. FEED SYSTEM LOSS AT VHF

---

The problem at VHF is that coaxial cable loss increases dramatically with frequency, often robbing the antenna *system* of its full potential. On 2 meters, for example, the matched-line attenuation for widely used RG-8 foam dielectric coax is about 2.1dB/100 feet (1), which is quite substantial. This is the lowest possible value, because the attenuation increases as SWR goes up (which also reduces system bandwidth). Another source of attenuation is the required matching network and balun. Even the simplest network introduces losses in its electronic components and coax connectors, and these losses are usually much higher at VHF than at lower frequencies.





AWG	..... Zo ( $\Omega$ ) .....			
	300	350	400	450
10	0.622	0.945	1.434	2.176

**Table 1: Conductor Spacing (Inches) for Various Wires Size and Characteristic Impedance**

At VHF, where fractions of a dB can make a difference, using a transmission line with the lowest possible loss is obviously very important. The best transmission line is air-insulated open wire (not window line).

An open wire line made with #12 AWG conductors has an attenuation of only 0.25dB/100 feet on 2 meters (1). For comparison, 200 feet of open wire line delivers 89% of the input power to the antenna, while the same length of RG-8 delivers only 38%.

But there are problems with open wire line. Practical conductor spacings result in high characteristic impedance, which is difficult to match to a low impedance Yagi. Table 1 shows the centre-to-centre conductor spacing to achieve different line impedance with common wire sizes. The impedance is

computed from the formula  $Z_o=276 \log(2S/d)$ , where S is the centre-to-centre spacing and d is the conductor diameter, both in the same units (2).

### 3.

#### HIGH IMPEDANCE YAGIS

The easiest way to take advantage of the extremely low attenuation of open wire is to design a good Yagi with a *high* input impedance.

One approach is to use a half-wave folded dipole as the driven element(3). Another is to increase the input impedance of the usual centre-fed linear dipole driven element (DE) by proper placement of the arrays parasitic elements.

The feasibility of this design approach will be demonstrated by a 12-element Yagi that provides a nearly perfect match to 300-ohm open wire line and excellent overall performance. An added advantage of this design is that the balanced antenna is fed by a balanced line.

The balun required to match the unbalanced transmitter output to the line can therefore be placed at the transmitter output terminal instead of the antenna input.

#### 4. YGO2

The array was designed using the Yagi Genetic Optimiser version 2, which is a freeware program available on the web (4). YGO2 models Yagis using NEC-2D (Numerical Electromagnetics Code, Ver. 2, Double Precision), which is also on the web (4) or directly from ACES(5). Essential data from the YGO2 configuration file, and the NEC-2D input file for the final optimised array, appear in the Appendix.

The optimisation was done iteratively. The initial runs optimised only the input impedance by setting the coefficients d, e and f to zero, and by assigning a low value to coefficient a (see Appendix and YGO2 literature). Once a geometry was evolved that would provide a good match to 300Ω, the corresponding chromosome was used to seed subsequent optimisation runs in which the coefficients a, d, and e were gradually increased (f was always zero). This process was repeated until the desired balance between  $Z_{in}$ , gain, FB and FR was achieved. All runs were at one frequency (146 MHz). The element diameter of 0.0122 waves is equal to 1 inch divided by the wavelength at 144 MHz (299.8/frequency in MHz). Somewhat fat elements were chosen to broaden the Yagis response. Smaller diameter elements (but

not too small) should provide more gain at the expense of bandwidth, but this was not investigated in detail.

#### 5. 12 ELEMENT, 300Ω YAGI

The Yagis geometry appears in Table 2. All dimensions are in wavelengths at the design frequency  $F_0$ . *Length* is the end-to-end element length. *Spacing* is the separation along the boom from the previous element. *Position* is the distance along the boom from element #1 (the reflector). Element #2 is driven at its centre (centre-fed linear dipole). All elements have the same diameter (0.0122 wave at  $F_0$ ).

Fig.1 shows the YGO2 output screen, which plots the E-plane (azimuthal) radiation pattern and provides a scale representation of the array. The display is annotated with key performance data. The pattern is very clean, and its structure is typical of well-designed Yagis. One interesting aspect of this Yagi is its unusual geometry. Unlike standard designs (progressively shorter directors with increasing spacing away from the DE), the element lengths show no particular pattern, and some of them are quite out of the ordinary. The DE, for example, is the second longest

Genetically Optimized 12-el  
 Yagi E-Plane (Az) Pattern  
 Norm Freq, F/Fo = 1.00000  
 NEC File: YG0\_1.40  
 Run: 11-16-1998, 23:40:45  
 Chromo #1, Generation #40  
 Figure-of-Merit = 12.993  
 Boom Length = 2.21 wuln @ Fo

Gmax = 11.53 dBi @ 0°, HPBW = 38.4°  
 F/B = 19.2 dB, F/R = 19.2 dB  
 Zin = 299.3 + j2.9Ω SWR = 1.01 / 300Ω  
 SL1 = -15.4 dB, Az = 53°  
 SL2 = -19.2 dB, Az = 144°  
 SL3 = -19.9 dB, Az = 119°

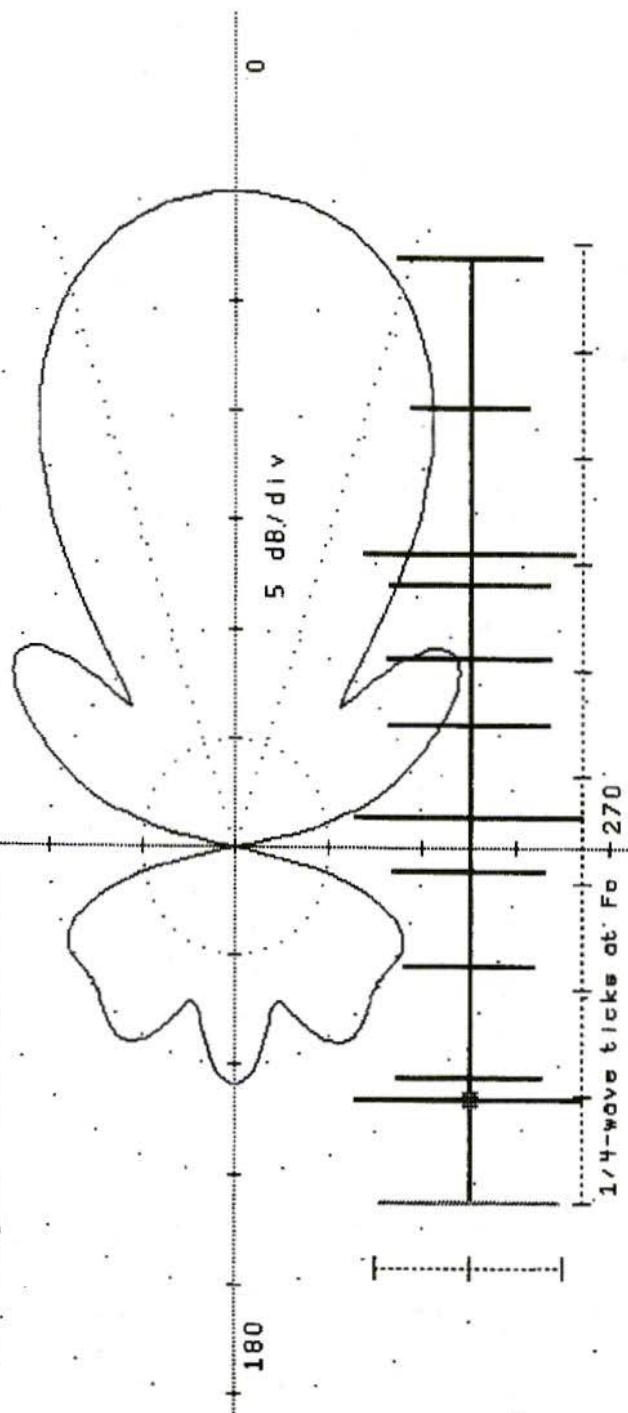


FIG.1: YG02 OUTPUT SCREEN



El #	Length	Spacing	Position
1	0.4839	0.0000	0.0000
2(DE)	0.5992	0.2397	0.2397
3	0.3865	0.0500	0.2897
4	0.3453	0.2635	0.5532
5	0.4094	0.2229	0.7761
6	0.6000	0.1224	0.8985
7	0.4306	0.2212	1.1197
8	0.4412	0.1559	1.2756
9	0.4271	0.1700	1.4456
10	0.5647	0.0729	1.5185
11	0.3141	0.3482	1.8667
12	0.3824	0.3447	2.2114

**Table 2: Yagi Geometry (dimensions in wavelengths at Fo)  
(note - all elements are 0.0122 wave diameter)**

element (0.5992 wave), the reflector is much shorter (0.4839 wave), and the longest element is director #6 (0.6 wave). Usually, the reflector is the longest element, the DE is close to half-wave, and the directors become progressively shorter.

Another quite unusual feature is the position of director #1 (D1), which is very close to DE (only 0.05 wave separation). D1 appears primarily to function as an impedance matching parasitic, rather than as a true director. Even though it contains 12 elements, this Yagi seems to act more like an 11 element array.

The similarity to an 11 element array is even more apparent when the 12 element Yagi is compared to the 144 MHz family in *The ARRL Antenna Book*(6). Its 11-element array has a boom length of 2.2 waves, a gain of 14.15dBi

(12 dBi), a front-to-back ratio (FB) of 19dB, and an input resistance of  $38\Omega$  (reactance not specified). The 12-element Yagi in Fig.1 has the same boom length (2.21 waves), a gain of 11.53dBi, FB and front-to-rear (FR) ratios of 19.2dB, and it provides a nearly perfect SWR of 1.01 on  $300\Omega$  open wire line.

The 12 element Yagi thus provides nearly the same performance as the ARRL 11-element array, with two notable differences. First, it seems clear that the extra element increases the input impedance while contributing little to the gain. In fact, taken together, DE and D1 look much like a folded dipole. It is interesting that YGO2s solution places D1 as close as possible to DE (0.05 waves). Had the minimum spacing in YGO2.CFG (see Appendix) been lower, say 0.025 wave, YGO2 might have placed



D1 even closer, which would make DE/D1 look even more like a folded dipole.

The second (and more important) difference is that the 12-element Yagis gain is lower than the 11-elements by 2.62dB. But, while this gain reduction is significant, it is not realised in practice. If the two antenna systems are properly compared by taking into account transmission line loss, then the 12-element Yagi may well be better. It provides more overall gain than the 11 element array for any transmission line longer than 142 feet (assuming #12 AWG open wire and foam dielectric RG-8). And this result excludes matching network and connector losses. These are likely to be quite a bit higher for the coax than for the open wire line, which again gives an edge to the 12 element array.

Open-wire line is easily fed with a simple, very low loss air-core inductive circuit that impedance matches the unbalanced transmitter output to the balanced line(7). This circuit can be located at the transmitter, which is not possible with coax because the balun must then be at the antenna input. Placing the network in the shack reduces the weather-related losses and maintenance that are inevitable with outdoor networks. Another consideration favouring open wire line is the lines effect on signal-to-noise ratio. In a receiver noise limited system,

SNR is reduced by the amount of transmission line attenuation. At VHF, extremely low loss open wire line is therefore advantageous both for receiving and transmitting. It can provide better SNR than coax, and it delivers more power to the antenna.

---

## 6.

### Input Impedance & SWR

---

Figs. 2-7 provide more detailed performance data for 12-element, 300Ω Yagi. On each plot the abscissa is the normalised frequency,  $F/F_0$ , where  $F_0$  is the design frequency at which the array dimensions are computed. For convenience, the ratio  $F/F_0$  will be denoted by a lower case, italic  $f$ . Each parameter is plotted over a 10% bandwidth, 0.95  $f$  1.05.

Fig.2 shows the input resistance. The design value of 300Ω is achieved at 3 frequencies,  $f = 0.97, 1.00, \text{ and } 1.006$ . Maximum resistance is 350Ω at  $f = 0.957$ , with a secondary peak of about 330Ω at  $f = 1.003$ . The resistance is between 200 and 350Ω for frequencies from 0.95 to just below 1.01. Input reactance is plotted in Fig.3. Four resonances ( $X_{in} = 0$ ) occur at  $f = 0.968, 0.992, 1.00, \text{ and } 1.044$ . The maximum reactance of about +140Ω (inductive) occurs at  $f =$

## Input Resistance

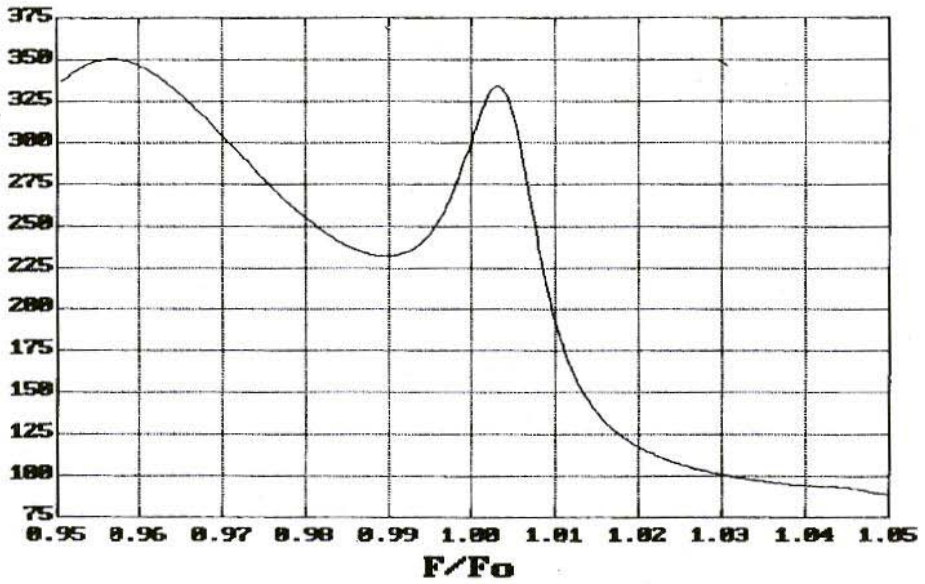


Fig.2

## Input Reactance

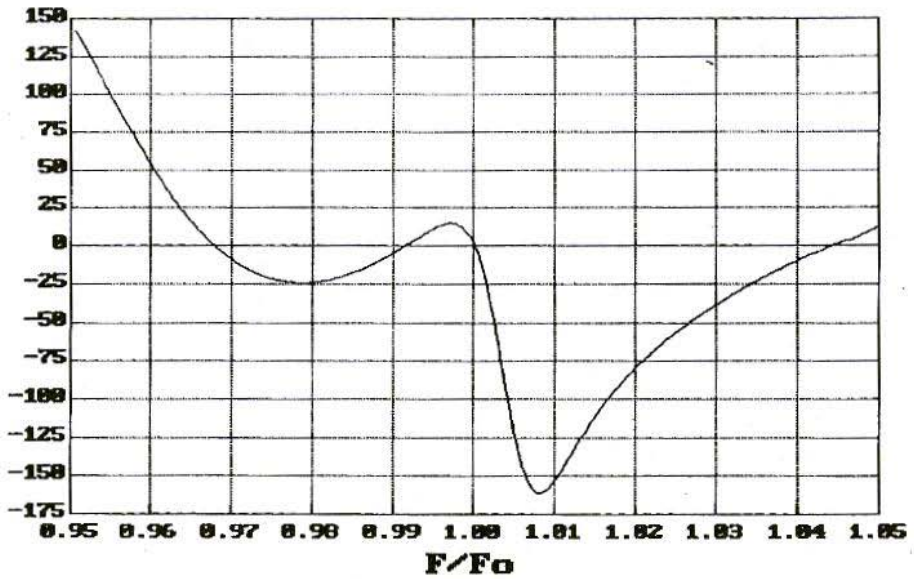


Fig.3



## Standing Wave Ratio

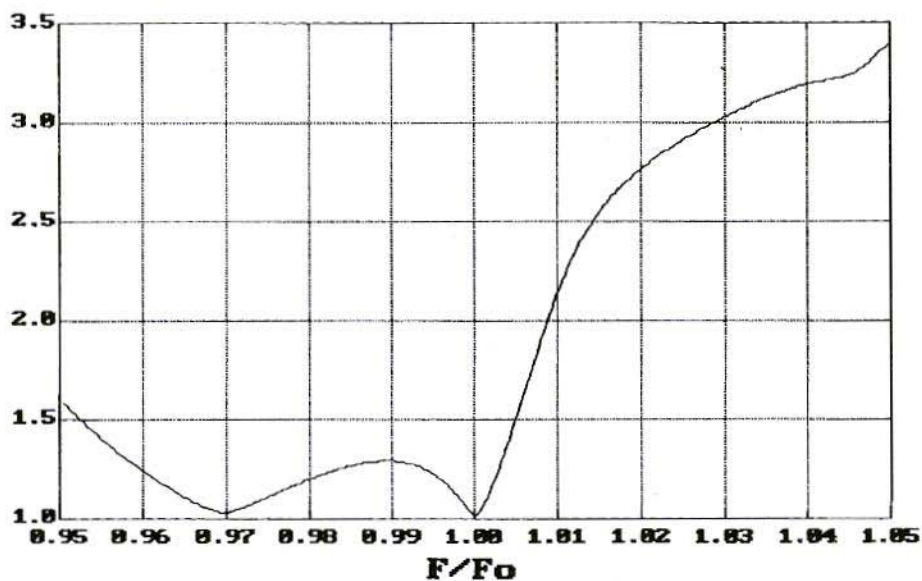


Fig.4

## Forward Gain

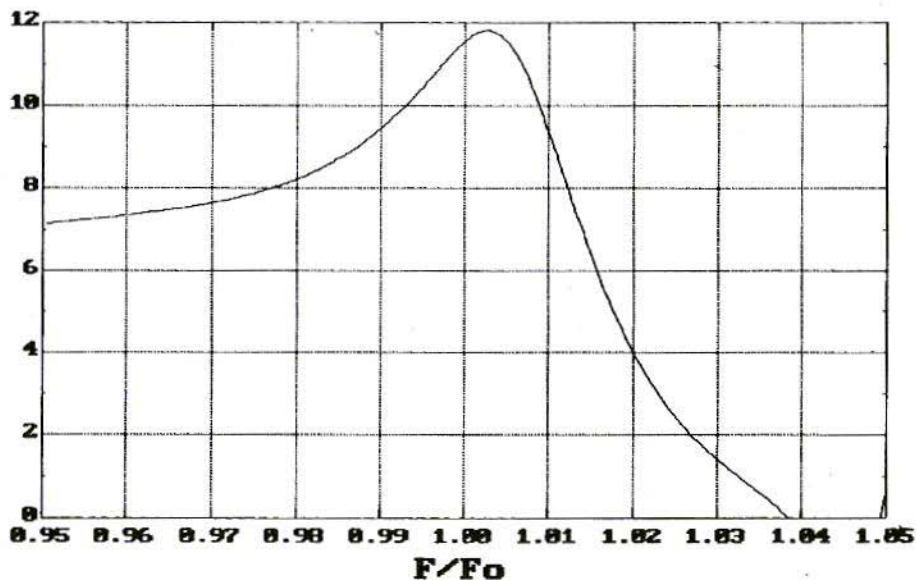


Fig.5

## Front-to-Back Ratio

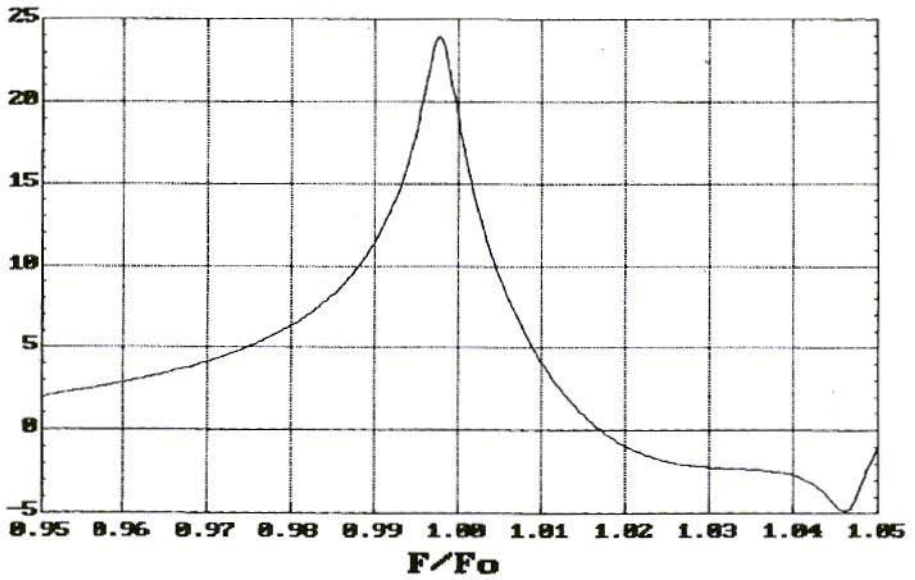


Fig.6

## Front-to-Rear Ratio

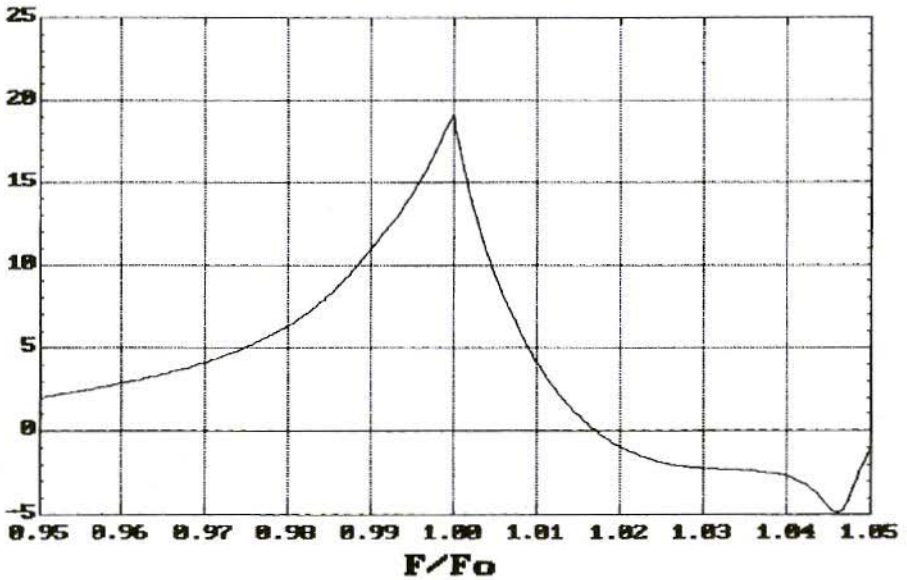


Fig.7





0.95, while the minimum of  $-161 \Omega$  (capacitive) is at  $f = 1.008$ . The reactance is less than 10% of the desired  $300 \Omega$  input resistance from  $f = 0.963$  to  $1.002$  (3.9%). For practical purposes, the Yagi is resonant over this entire range of frequencies. Of course, the most important antenna impedance parameter is SWR, which is plotted in Fig.4. At the lower band edge, the SWR is just over 1.5. It is below 1.5 from  $f = 0.953$  to  $1.005$ , yielding a 1.5:1 SWR bandwidth of 5.2%. The 2:1 SWR bandwidth is more than 5.9%, which is quite good.

---

## 7.

### Gain & FB/FR Ratios

---

Forward gain in dBi (dB relative to an isotropic radiator) appears in Fig.5. Maximum gain is 11.8dBi at  $f = 1.003$ . The gain is above 10dBi from  $f = 0.993$  to  $1.009$ , a 10dBi gain bandwidth of 1.6%. The half-power (-3 dB) frequencies are 0.986 and 1.011, yielding a -3dB bandwidth of 2.5%.

The FB ratio (Fig.6) peaks at 23.9dB at  $f = 0.998$ . FB is above 20 dB from  $f = 0.996$  to  $1.00$  (0.4%), above 15dB from 0.993 to  $1.002$  (0.9%), and greater than 10 dB from 0.988 to  $1.004$  (1.6%). FR (Fig.7) shows similar behaviour, but its peak value is

lower. Maximum FR is 19.2dB at  $f = 1.00$ . It is above 15dB from  $f = 0.996$  to  $1.002$  (0.6%), and greater than 10dB from 0.988 to  $1.004$  (1.6%).

---

## 8.

### Conclusion

---

The 12 element  $300 \Omega$  Yagi is indeed a very good antenna. It directly matches extremely low loss, balanced  $300 \Omega$  open wire line. As a system, the 12 element Yagi with open wire line outperforms an optimised 11 element array of the same boom length fed by RG-8 coax, as long as the transmission line is longer than 142 feet (ignoring matching network and balun losses). The 12 element system provides more overall gain and better SNR. To the extent that the coax matching network/balun has higher losses than the open-wire network, which is very likely, the 12 element Yagi will provide better performance even for lines shorter than 142 feet.

There is a substantial potential advantage in designing high input impedance Yagis that operate directly with high impedance open wire lines. The design example discussed here shows that this objective can be achieved with very robust antenna system performance.



---

**9.****Literature**

---

1. The ARRL Antenna Book, 17th ed., R. Dean Straw, editor, American Radio Relay League, Inc., Newington, CT 06111, USA, 1994, Fig.22, p. 24-16.
2. *ibid.*, Eq. (20), p. 24-14.
3. Antennas, 2nd ed., John D. Kraus, McGraw-Hill Inc., New York, 1988, p. 483.
4. Ray Anderson, WB6TPUs, NEC Archive web site, URL: <http://www.qsl.net/wb6tpu>
5. Applied Computational Electromagnetics Society (ACES) Attn: Dr. Richard W. Adler, ACES Executive Officer, ECE Dept., Code ECAB, Naval Postgraduate School, 833 Dyer Road, Room 437, Monterey, CA 93943-5121 USA.
6. The ARRL Antenna Book, 17th ed., R. Dean Straw, editor, American Radio Relay League, Inc., Newington, CT 06111, USA, 1994, Table 11, p. 18-25.
7. *ibid.*, Figs. 2(A), 2(C), p. 25-3.

---

**10.****Appendix - Yagi Modelling Data**

---

**Essential Data from File YGO2.CFG**

Name of NEC Executable File - NEC2D100.EXE

# NEC Input Files/Gen in Output - 5

starting with generation number - 20

-----  
Number of Elements in Array - 12

Number of Segments per Element - 7

Feed System Zo - 300 ohms resistive

Assume Feed Reactance Tuned Out? NO



-----

Population Size - 10 (# chromosomes)  
 Max # Generations - 1  
 Save Percentage - 2.0 (% best chromos/gen saved)  
 Crossover Probability - 0.8000  
 Mutation Probability - 0.0200  
 Max Mutation Rate - 2 bits/chromosome

-----

Selection Method # - 1  
 (1-Binary Tournament, 2-Proportionate)  
 Minimum Fitness [0-1] - 0.5 (proportionate only)

-----

#### FoM Terminology

-----

Gfwd - Forward Gain (dBi)  
 FB - Front-to-Back Ratio (dB)  
 FR - Front-to-Rear Ratio (dB)  
 Rin - Feed Point Input Resistance, ohms  
 Xin - Feed Point Input Reactance, ohms  
 MaxSLL- Maximum Sidelobe Level (dB//Gfwd)  
 ^ - Exponentiation  
 \* - Multiplication  
 / - Division  
 ABS - Absolute Value

-----

Figure-of-Merit (averaged over all frequencies):

$$\text{FoM} = \{a * \text{Gfwd} - b * \text{ABS}(Z_0 - R_{in}) - c * \text{ABS}(X_{in}) + d * \text{FB} + e * \text{FR} - f * \text{MaxSLL}\} / (a + b + c + d + e + f)$$



## FREQUENCY TABLE

## # FREQUENCIES USED - 1

Freq #	Freq(MHz)	DE	a	b	c	d	e	f
1	146.00	2	260.0	4.0	6.0	1.0	85.0	0.0
2	144.20	2	40.0	2.0	3.0	0.0	0.0	0.0
3	144.30	2	40.0	2.0	3.0	0.0	0.0	0.0
4	144.40	2	40.0	2.0	3.0	0.0	0.0	0.0
5	144.50	2	40.0	2.0	3.0	0.0	0.0	0.0
6	144.60	2	40.0	2.0	3.0	0.0	0.0	0.0
7	144.70	2	40.0	2.0	3.0	0.0	0.0	0.0
8	144.80	2	40.0	2.0	3.0	0.0	0.0	0.0
9	144.90	2	40.0	2.0	3.0	0.0	0.0	0.0
10	145.00	2	40.0	2.0	3.0	0.0	0.0	0.0
11	145.10	2	40.0	2.0	3.0	0.0	0.0	0.0
12	145.20	2	40.0	2.0	3.0	0.0	0.0	0.0
13	145.30	2	40.0	2.0	3.0	0.0	0.0	0.0
14	145.40	2	40.0	2.0	3.0	0.0	0.0	0.0
15	145.50	2	40.0	2.0	3.0	0.0	0.0	0.0

-----  
 Target FoM - 9999 (not normalised)  
 -----

Crossover allowed only at gene boundary? YES

Print Percent - 20 (% chromos/gen printed in output file)  
 -----

Use Seed Chromosome? YES

Seed Chromo:

```
01100001000000000000000001011011000000000000000111111100000000000000
00001000010000000000000001001101101000000000010001101111111100000000
10010100111101100000000010000110000111100000000000111100001101100000
00000010001010000111000000001011000000110000000000001001010101100010
0000000011100101
```

Print Chromo Sequences in YGO.DAT? NO



## GENE TABLE

Gene #	Name	Length (bits)	Min (wvln)	Max (wvln)
1,	"Refl_Length ",	8,	0.3000,	0.6500
2,	"Refl_Radius ",	8,	0.00610,	0.00610
3,	"Refl_Spacing ",	8,	0.0000,	0.0000
4,	"DE_Length ",	8,	0.3000,	0.6500
5,	"DE_Radius ",	8,	0.00610,	0.00610
6,	"DE_Spacing ",	8,	0.0500,	0.428
7,	"D1_Length ",	8,	0.3000,	0.6500
8,	"D1_Radius ",	8,	0.00610,	0.00610
9,	"D1_Spacing ",	8,	0.0500,	0.428
10,	"D2_Length ",	8,	0.3000,	0.6500
11,	"D2_Radius ",	8,	0.00610,	0.00610
12,	"D2_Spacing ",	8,	0.0500,	0.428
13,	"D3_Length ",	8,	0.3000,	0.6000
14,	"D3_Radius ",	8,	0.00610,	0.00610
15,	"D3_Spacing ",	8,	0.0500,	0.500
16,	"D4_Length ",	8,	0.3000,	0.6000
17,	"D4_Radius ",	8,	0.00610,	0.00610
18,	"D4_Spacing ",	8,	0.0500,	0.500
19,	"D5_Length ",	8,	0.3000,	0.6000
20,	"D5_Radius ",	8,	0.00610,	0.00610
21,	"D5_Spacing ",	8,	0.0500,	0.500
22,	"D6_Length ",	8,	0.3000,	0.6000
23,	"D6_Radius ",	8,	0.00610,	0.00610
24,	"D6_Spacing ",	8,	0.0500,	0.500
25,	"D7_Length ",	8,	0.3000,	0.6000
26,	"D7_Radius ",	8,	0.00610,	0.00610
27,	"D7_Spacing ",	8,	0.0500,	0.500
28,	"D8_Length ",	8,	0.3000,	0.6000
29,	"D8_Radius ",	8,	0.00610,	0.00610
30,	"D8_Spacing ",	8,	0.0500,	0.500
31,	"D9_Length ",	8,	0.3000,	0.6000
32,	"D9_Radius ",	8,	0.00610,	0.00610
33,	"D9_Spacing ",	8,	0.0500,	0.500
34,	"D10_Length ",	8,	0.3000,	0.6000
35,	"D10_Radius ",	8,	0.00610,	0.00610
36,	"D10_Spacing",	8,	0.0500,	0.500

----- End of Gene Table -----

\*\*\*\*\* End of File YGO2.CFG \*\*\*\*\*

## NEC-2D Input File for YGO2-Optimized Array

CM NEC File: YGO\_1.40 (Run ID: 11-16-1998, 23:40:45)  
 CM Chromosome #1, Generation #40  
 CM Figure-of-Merit = 12.993  
 CM Feed System Zo = 300 ohms resistive  
 CE  
 GW 1,7,0.,241961,0.,0.,-241961,0.,.0061  
 GW 2,7,239741.,299608,0.,239741,-299608,0.,.0061  
 GW 3,7,289741.,1932355,0.,289741,-1932355,0.,.0061  
 GW 4,7,.5532.,172647,0.,.5532,-172647,0.,.0061  
 GW 5,7,.776141.,204706,0.,.776141,-204706,0.,.0061  
 GW 6,7,.898494,3,0.,.898494,-3,0.,.0061  
 GW 7,7,1.11967,215294,0.,1.11967,-215294,0.,.0061  
 GW 8,7,1.275552.,220588,0.,1.275552,-220588,0.,.0061  
 GW 9,7,1.445552.,2135295,0.,1.445552,-2135295,0.,.0061  
 GW 10,7,1.518493.,282353,0.,1.518493,-282353,0.,.0061  
 GW 11,7,1.866728.,157059,0.,1.866728,-157059,0.,.0061  
 GW 12,7,2.211434.,1911765,0.,2.211434,-1911765,0.,.0061  
 GE  
 GN-1  
 FR 0,1,0,0,299.8,0.  
 EX 0,2,4,0,1.,0.  
 RP 0,1,181,1001,90.,0.,0.,1.,10000.  
 XQ  
 EN

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## Individual Indication of Reasons for Cut-Out by Protective Circuits of a High-Power

If part of a transmitter/receiver system breaks down, it usually happens in the most inconvenient situations. So it is important to obtain clear information on where the malfunction is located.

It has happened several times that one of the standard built-in protective circuits in the PA of our club station has responded, for reasons which can

no longer be traced, thus cutting out the PA and protecting it from destruction. So the protective circuits have certainly done their job reliably, but the actual reason for the cut-out is still not clear.

Things can be considerably simplified here by the kind of individual indication described in the following article.

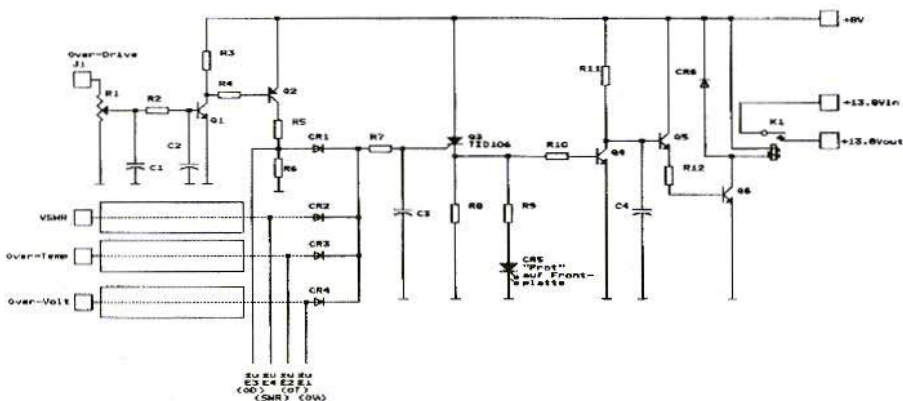


Fig. 1: Circuit Diagram Extract for BEKO Protective Circuits  
Auf Frontplatte = On front plate

## 1. INTRODUCTION

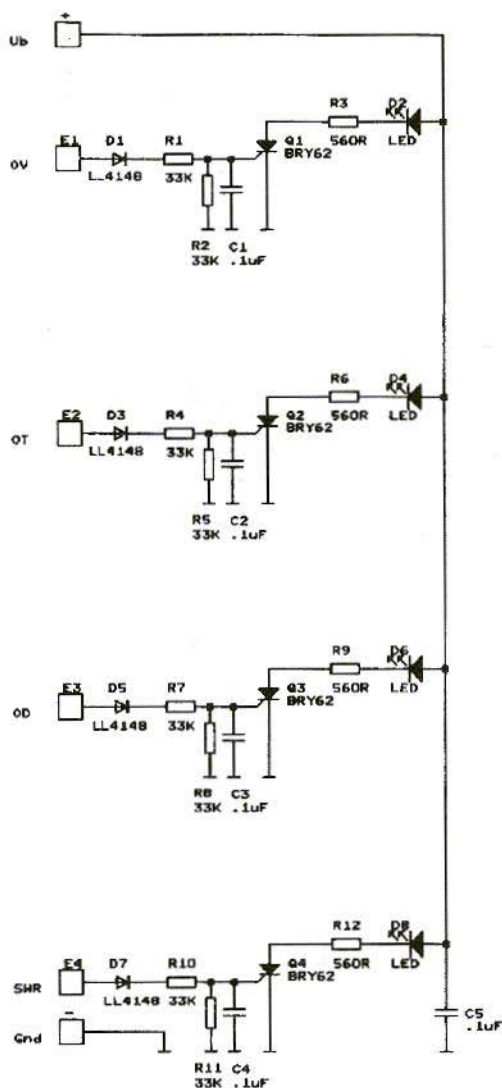


Fig. 2: Individual Fault Display Circuits

The function and assembly of equipment for the individual indication of faults is explained here, using the example of a type HLV120 BEKO transmission amplifier, with an output power of 100 Watts for the 70cm band.

This individual indication system can certainly be incorporated into other BEKO amplifiers which are fitted with similar protective circuits.

The protective circuits are:

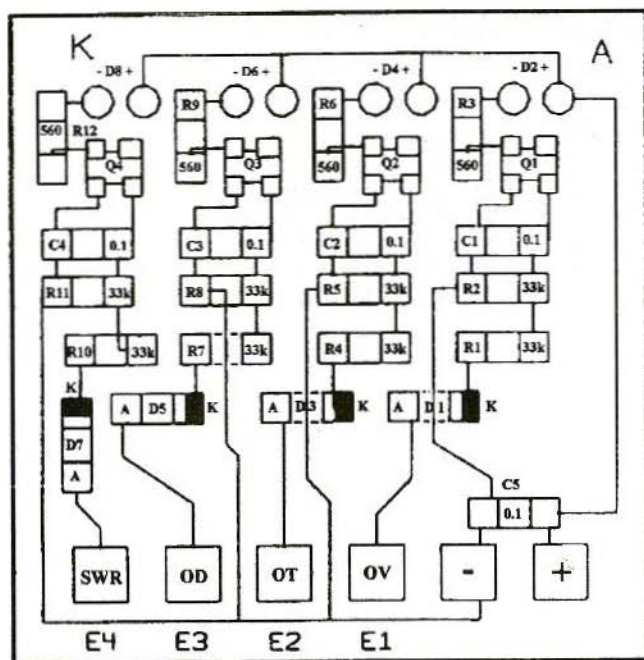
- Overvoltage (OV)
- Overtemperature (OT)
- Standing Wave Ratio (SWR)
- Overdrive (OD)

In the original format, the signals from the four monitoring devices, which protect the transmission amplifier from impermissible operating conditions during its operation, are connected to an alarm system totalling all the faults, by-pass the power section and cause the "Protection" LED to light.

The PA is also protected against any reversal of the operating voltage terminals, but no indication takes place.

To narrow down the reason for the cut-out more easily in the case of a fault, it makes sense to have the output signal from each individual protective circuit indicated





**Fig. 3:**  
SMD Board  
Component Overlay

separately. To bring this about, the four signals from the individual faults are measured before they are even combined, and are displayed, each with a separate LED, with the aid of the circuit described here.

The LED's can be mounted directly on circuit boards, behind or into the front plate. A suitable proposal for mounting follows below.

## 2.

### CIRCUIT DESCRIPTION

Starting from the standard protective circuits of the transmitter amplifier, the "status" signal is measured in each case, and is processed further for the individual fault display.

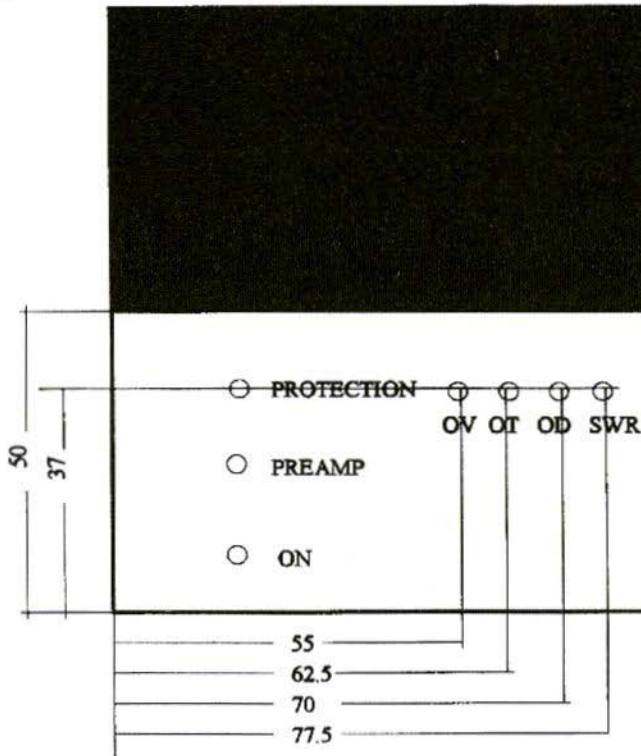
The different protective circuits all operate in a similar way: thus, for example, "Input Overdrive" works as follows (Fig.1):

If the limiting value is exceeded in the PA, this causes the output of the switching transistor to go to the logical "High". The thyristor Q3 is triggered through the diode CR1.

The cathode of Q3 goes to "H", the "Protection" LED CR5 on the front plate lights.

Q4 is also switched through, which makes the Darlington stage - consisting of Q5 and Q6 - currentless, and the relay K1 opens, and the end stage is by-passed.

The four new supplementary circuits also all operate the same way (Fig.2):



**Fig. 4:**  
Hole dimensions for  
Additional LED's

In each case, at the point where the signals are connected together through a diode, the corresponding status signal is now measured, and goes through a diode and a voltage divider to the gate of a small thyristor.

Parallel to the gate, a block-off capacitor is connected to earth. Then, in the anode circuit of the thyristor, there is the LED, in series with a resistor to limit the current.

If one of the protective circuits now responds, the corresponding small thyristor is switched through and the LED lights.

The thyristors continue to be switched through until the operating voltage is

switched off, which undershoots the holding current.

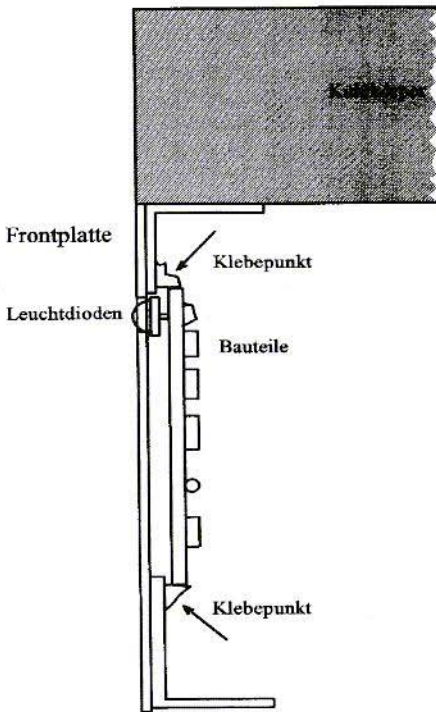
---

### 3. ASSEMBLY

---

So that as small a printed circuit board as possible could be used, it was decided to assemble the unit in SMD format. The display board can thus be positioned at a free location behind the front plate.

Only the standard tools for SMD placement are needed for assembly.



**Fig. 5: Side View of Fitted Additional Boards**

*Kühlkörper* = Heat sink, *Frontplatte* = Front plate, *Leuchtdioden* = LED's, *Klebpunkt* = Glue point, *Bauteile* = Components

### 3.1 Placement

The SMD components are positioned on the foil side in accordance with Fig. 3.

Parts list:

- 1 x printed circuit board DH6SBN001
- 4 x SMD diode LL4148 (Mini-Melf)
- 4 x LED 3 mm., red
- 4 x BRY 62, thyristor, SOT 143
- 8 x 33k $\Omega$ , SMD 1206
- 4 x 560 $\Omega$ , SMD 1206
- 5 x 0.1 $\mu$ F, 50V, SMD 1206

The "fitting" of the thyristors is not critical. First we tin-plate a speck of solder, then we lay the thyristor onto the printed circuit board, with the somewhat wider connection coming onto the unconnected soldering point, and then we solder up the previously tin-plated speck of solder. Then the remaining connections of the thyristor are soldered on. These are followed by the capacitors, resistors and diodes. The LED's are left to the very end, because these are inserted from the other end and are then soldered on the foil side.

### 3.2 Checking

Following the placement, the printed circuit board, the position of the components and their soldering joints are given a careful visual check with the aid of a magnifying glass.

If the assembly is correct, the operating voltage of + 8 Volts is applied. None of the LED's should light.

If an additional voltage of app. +3 to +8 volts is applied to each of inputs E1 to E4 in turn, then the corresponding LED should light in each case. It should not go out even when the control voltage is switched off again. The LED's should not go out until the operating voltage is taken out of the circuit.

If the circuit does not function as described, the following sources of error may be present:

LED terminals reversed; triggering voltage to thyristor interrupted or short-circuited; one of soldering joints not in order or - the least likely explanation - a faulty component has been fitted.

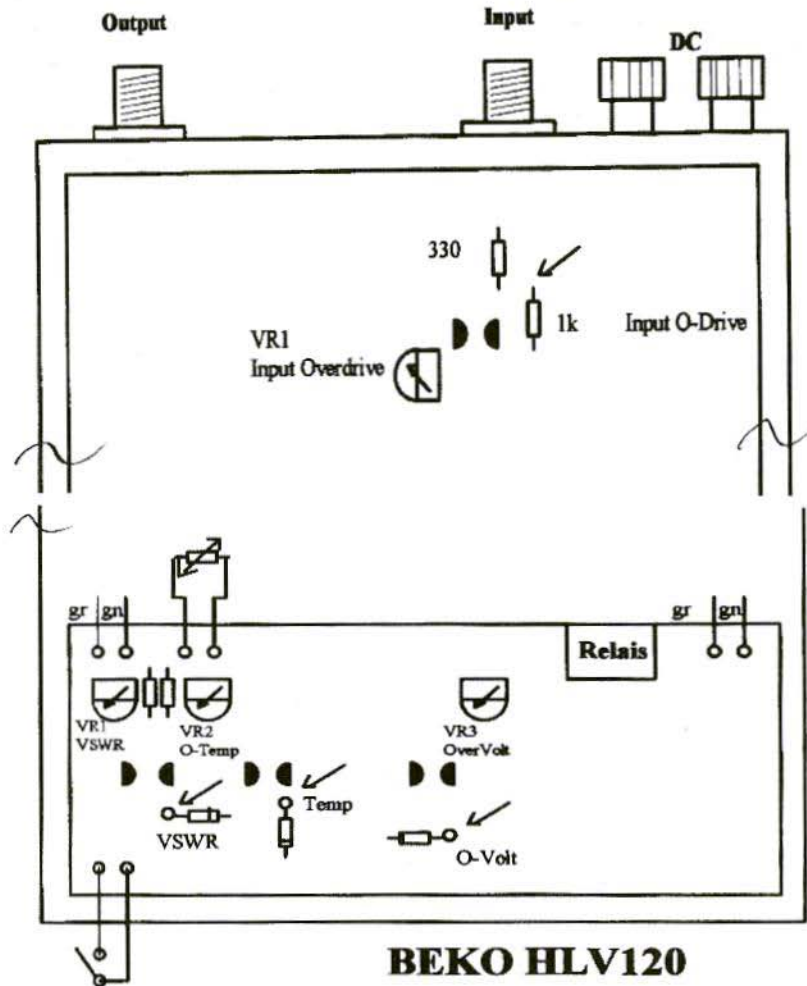


Fig. 6: Connecting Points for Individual Display in PA; View from below; *Relais* = Relay

### 3.3 Fitting the display unit

Attention! The display unit has to be fitted extremely carefully, to ensure that the PA operates just as well afterwards as it did before.

Any interference with new appliances during the guarantee period renders the manufacturer's guarantee null and void!

The equipment is unscrewed. To do this, the four screws in the baseplate, and one screw each in the centre and below the front and rear plates are removed.

First the lines leading to the two switches ("On/Off", W) are taken to the front plate, marked and unsoldered.

The feeds from the LED's, which are



**Fig. 7: Modified Standard Equipment HLV 120 from BEKO**

permanently connected to the front plate, are also unsoldered.

The front plate can now be screwed off, bored as per Fig.4, and then labelled.

The lettering can be done, for example, using transfer letters or laser-printed self-adhesive film.

The modified front plate is re-fitted, and the LED's for the display board are inserted into the front plate from behind, until they are visible and project uniformly through the front plate (Fig.5).

If the front plate holding angle is too great, file out at the appropriate places.

The display board is then mechanically fixed using hot melt or silicone.

### 3.4 Wiring

The operating voltage and control voltage for the display board are wired up as per Fig's.1, 2 and 6.

A screened line should be used to connect up the "Input Overdrive" signal, since this goes directly past the power unit of the PA.

Finally, the feed lines for the switches and the LED's are soldered on again.

The fitting is now complete. The high-level stage is re-assembled complete (Fig.7) and a functional test is carried out.

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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**

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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
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Frequency Range:	925 - 1800 MHz
Sensitivity:	-57dBm (300 $\mu$ V)
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 4 $\Omega$ minimum
Controls:	Main Tuning, Volume, Video Gain, Video Bandwidth, LNB Power ON/OFF
LNB / Preamp Powe:	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

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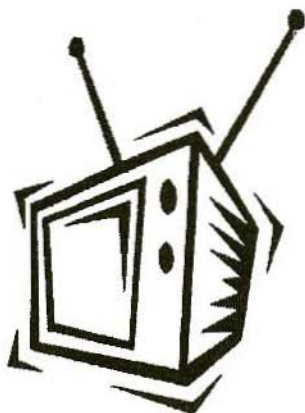
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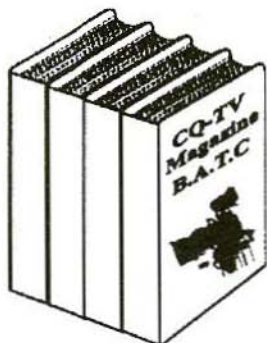
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DG4RBF-006	HF Synthesiser 5 to 1450 MHz, Regulator for VCO	2&3/98	-----	£ 12.00
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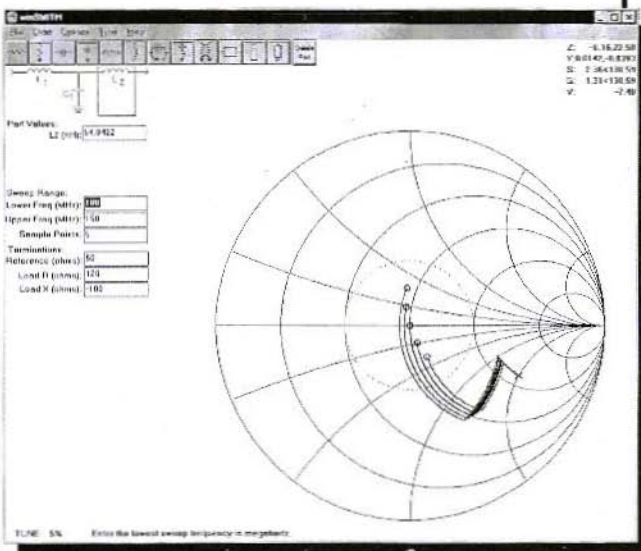
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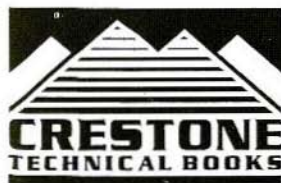


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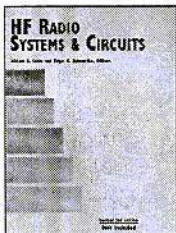
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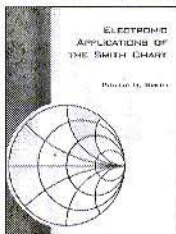
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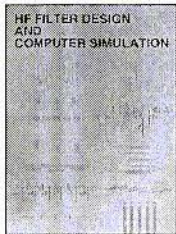
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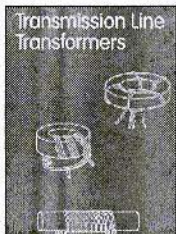
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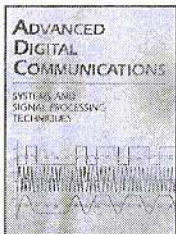
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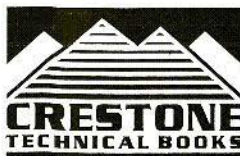
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