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Especially Covering VHF, UHF and Microwaves

VHF COMMUNICATIONS

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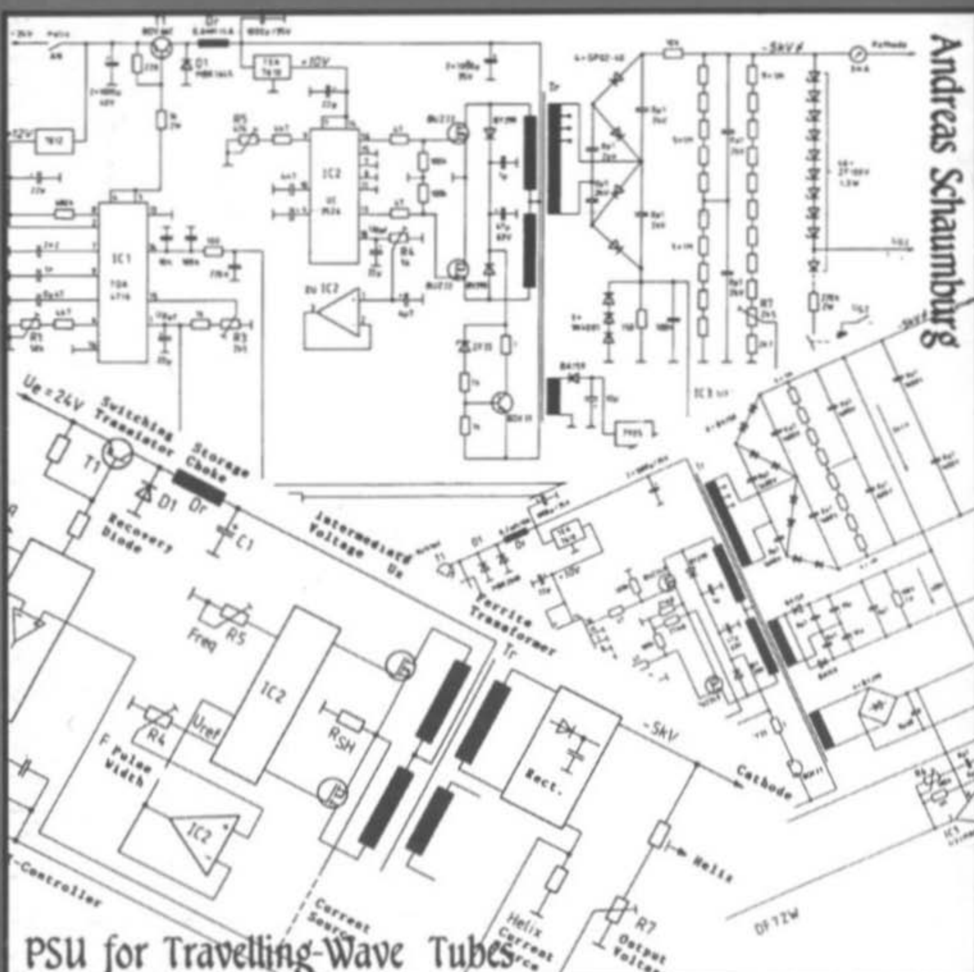

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Andreas Schaumburg, DF 7 ZW

Power Supply Unit for Travelling-Wave Tubes

There are two possible ways in which radio amateurs can increase their transmission power in the SHF bands. First, by using an end stage with expensive high-powered micro-wave GaAsFETs, and secondly by means of a travelling-wave tube as a power amplifier. By comparison with GaAsFETs, a travelling-wave tube, in a used but fully functional condition, can easily be obtained in amateur radio flea markets. Unfortunately, suitable power supply units which can deliver the special operating voltages for the tube are difficult to find.

1. GENERAL COMMENTS ON OPERATING A TRAVELLING-WAVE TUBE

The Siemens RW 1127 travelling-wave tube is normally used in the 11.7 - 13.25 GHz frequency range to feed television programmes to broadcast main stations through a microwave radio link. A 3-tone inter-modulation interval of 60dB is obtained with an initial picture-synchronised output of 3.5W. This travelling-wave tube was developed by Siemens in 1984. Such units have amassed a total of several thousand operating hours. Because people replace their tubes when competing in competitions, tubes in adequately functioning condition now find their way onto the surplus market and can come in useful for radio amateurs.

With regard to the circuit power supply (CPS) described below, this tube can briefly deliver a saturation power level of up to 50W in the 10 GHz amateur band. The input power of the circuit power supply here is up to 300W for an input voltage which can be between 20V and 35V. An input voltage of 12V would certainly be more favourable for portable use, but then currents of up to 25A would have to be connected up with low losses, which is very problematical at this low voltage.

In practical terms, a travelling-wave tube is a black box with SMA input and output sockets and a multiple plug for the power supply. Problems of low-loss wide-band tuning of the input and output impedances at 50, such as arise with GaAsFET amplifiers, have already been solved by the tube manufacturers. Putting a travelling-wave tube into operation has little to do with micro-wave technology, but is rather a problem of circuit power supply development. The following voltages and current levels are required for the operation of RW 1127 travelling-wave tubes:

Collector-1 voltage:	2400V, max. app. 60 mA
Collector-2 voltage:	400V, max. app. 130 mA
Helix (helix) voltage:	5000V, max. 3 mA
Grid-2 voltage:	3600V - 4600V, typ. 4300V
Heating voltage:	6.3V ±0.2V

The circuit power supply can also be used for other travelling-wave tubes if converted to other voltages and current levels. It is merely necessary to alter the winding ratios of the ferrite transformer.

Important tips:

The circuit power supply output voltages and current levels are highly dangerous and you must be careful to prevent accidental contact!

Never look into tubular conductors carrying transmitting power and open at the other end! Always keep a safe distance from the transmitter aerial!

The radiation protection regulations for micro-wave irradiation should be adhered to.

A high-frequency output power of 40 to 50W requires thick high-frequency cables and tight coax plugs.

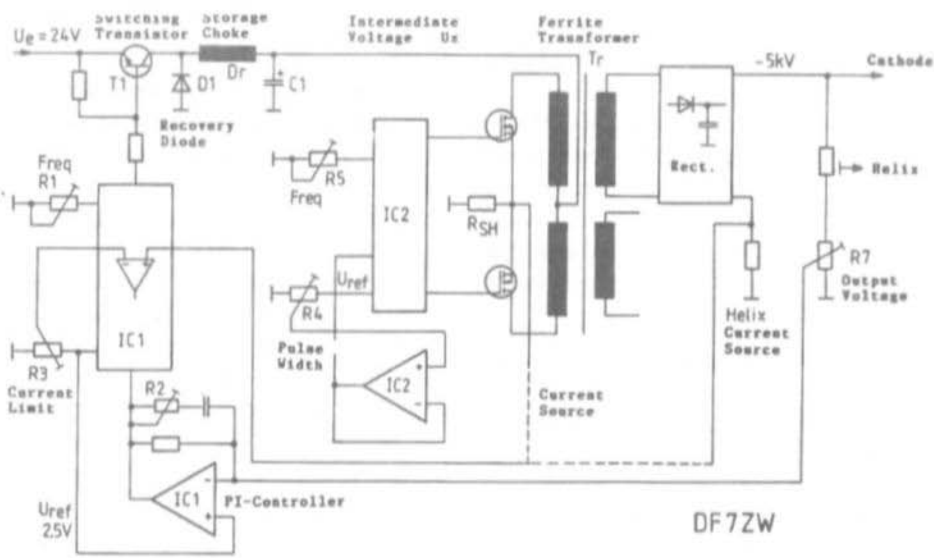

Fig.1: Block Diagram of the Power Supply

2. DESCRIPTION OF CIRCUIT POWER SUPPLY

The function of the circuit power supply is first explained in the text and then clarified through the block wiring diagram in Fig.1, which gives a much simplified view of the rectifier unit.

The circuit power supply is really two circuit power supplies (CPS) operating independently of one another. The first CPS (Fig.3) generates only the helix (Helix) voltage of 5kV, plus the grid-2 voltage derived from this through a chain of Zener diodes. The grid-2 voltage can be switched on and off independently of the Helix voltage, using a high-voltage vacuum relay, Re1, which is very important for building up the beam focusing in the travelling-wave tube.

The Helix voltage is applied with plus at the earth, so that acquiring the Helix current and the output voltage require no potential separation. The Helix current is acquired at the output side of the CPS, the output current being limited to a maximum of 3mA. The current limitation is very important, since if the current were excessive the Helix, which is made of thin tungsten wire, would soon be destroyed.

The other CPS (Fig.2), potentially separated from the first, generates the residual voltages for collector 1, collector 2 and the heating voltage. A large part of the power for operating the tube is converted in this CPS. The cathode potential of the tube amounts to -5kV to earth, so that all voltages and the optocoupler must be well insulated for voltage acquisition. The collector-2 voltage is regulated by means of the optocoupler. And the collector-1 voltage is also stable through the winding ratio of the ferrite transformer if both outputs are loaded approximately equally. The heating voltage must be stabilised using a 7806 stabiliser integrated circuit, for with a closed tube the circuit power supply will travel down and the input voltage of the stabiliser integrated circuit thus drops back sharply.

The current limitation circuit is intended to protect the tube and the circuit power supply itself. It is set up through an R_{sh} shunt resistance in the common earth wire of the MOSFETs on the input side of the push-pull circuit power supply.

The block wiring diagram, Fig.1, is somewhat similar for the two circuit power supplies. The differences arise in the number of output voltages and in acquiring the current and voltage for regulation. In practice, each CPS consists of two functional units: the down-converter CPS and the push-pull converter with a control circuit.

The power pack of the down-converter CPS generates the intermediate voltage, U_Z, and consists of the switching transistor, T1, the storage choke, Dr, and a recovery diode, D1. A PNP Darlington transistor is deliberately used as a switch here, since it is very easy to control. An N-channel MOSFET would switch with even less loss, but would need more expensive control here, despite the fluctuating potentials. The intermediate voltage, U_Z, can fluctuate between 10 and 15V, depending on the load and the high-voltage level set.

The push-pull converter is powered from the intermediate voltage, U_Z, and generates the high voltages. It consists of a ferrite transformer, two power MOSFETs and an IC 2 control circuit. The output signal of the ferrite transformer is rectified and filtered.

The 5kV or 2.4kV high voltage is obtained by quadrupling the voltage. The 400V required are generated by doubling the voltage. The heating voltage is simply rectified and stabilised by means of an IC 7806 (1.5A type).

The block wiring diagram (Fig.1) gives only a rough idea of the rectification. The two MOSFETs in the push-pull are controlled by the IC 2 control circuit (Unitrode UC 3526) of the push-pull converter, with a short dead time. Low switching frequencies have deliberately been selected, because of the very high internal capacitances with high-voltage turns: 16 kHz or 10 kHz for a Helix circuit power supply.

The internal capacitances in connection with the leakage inductances must be re-loaded for every switching procedure, which generates resonance currents and thus losses.

In contrast to the better known SG 3524, the UC 3526 has two integrated push-pull output drivers for direct control of the MOSFETs. If it is desired to use the SG 3524, which is easier to obtain, but has only open collector outputs, then two inverting drivers must always be wired up subsequently. At low switching frequencies, three parallel-wired 4049 CMOS grids can always be used.

The voltage regulation control circuit is closed by the push-pull converter in the down-converter direction. The controller, together with the current limitation comparator, is already integrated in the control circuit of the IC 1 down-converter. The voltage controller is wired up as a PI controller. The I behaviour is set using the R2 trimmer. The amplification of the controller at high frequencies is reduced, depending on the value of the individual trimmer, to the point at which the control circuit operates in a stable manner. The reference voltage, U_{ref} = 2.5V, generated in IC 1, is fed to the controller as the rated value.

The down-converter operates at a switching frequency of 25 kHz which is set through R1. The operation of the current limitation is set using the R3 trimmer. As soon as the current exceeds a specific value, the switching transistor, T1, is blocked by the comparator in IC 1.

The transition from voltage control to current limitation is critical for any circuit power supply and can cause oscillations in control. However, in practise this has no effect at all, since in any case the current limitation engages only in the event of a malfunction.

3. THE MAGNETIC STRUCTURAL COMPONENTS

Both storage chokes and high-voltage transformers are used as energy stores in the circuit power supply. Because of the DC polarisation, the core material used for the storage choke is a ring core with a low permeability of app. 100 for a maximum induction of 0.8 Tesla. Ferrite with a permeability of about 2000 and a saturation induction level of 0.3 Tesla is used for the transformer. You can wind the storage choke, Dr, yourself or buy it ready-made.

The choke for the Helix CPS has an inductivity, L, of 0.6mH and a current-carrying capacity amounting to 4 Amperes. 0.2mH and a current-carrying capacity of 10A are needed for the collector CPS. The author uses type 77552-A7 "Kool Mu Power Core" ring cores for both chokes. The core is wound with 47 turns of 1.5 mm. enamel copper wire for a 10A current, and 80 turns of 1 mm. wire are needed for a 4A current. The number of turns, and thus the inductivity of the storage chokes, is not critical. With a storage choke of unlimited size, a flat DC would be present. In practise, the storage choke is dimensioned in such a way that an AC with app. 20 to 40% of the DC amplitude is superimposed on the DC. These current ripples are smoothed out again by the subsequent filter capacitor, C1, so that the intermediate voltage, Uz, is approximately flat.

For the 10A storage choke in particular, the next largest core could also be used, since the specified core can actually easily be driven to saturation.

Preparing the two ferrite transformers for generating the high voltage is far and away more difficult. Here the insulation of the turns must be carried out very carefully, as the voltage is still 5kV. The ferrite cores used are type ETD 49, without an air gap. With the low switching frequency, the more favourably priced N 27 Siemens equipment can be used. The better N 67 ferrite material offers hardly any advantage here. The ETD 49 is a standard core and is also supplied by a number of other manufacturers.

The following elements are required to construct the transformer:

Ferrite cores, spools and two clamp clips each, together with enamelled wires, flexes, insulation tubing and insulation foil feathered on both sides, with plastic spray to soak the turns (e.g. Kontakt Chemie).

The transformer winding data for the collector CPS are as follows:

ETD 49 core:	Afe = 210 mm.2; f = 15 kHz; t = 33 micro-S; B = 0.2 T; n = 0.4 Wdg/V,
Primary:	2 x 0 - 15V, 2 x 6 turns of 60 x 0.2 mm. flex; 5 kV insulation
Secondary: -1-	0 - 540 - 600 - 660V, 0 - 216 - 240 - 264 turns, 0.3mm CuL, 5kV insulation
Secondary: -2-	0 - 160 - 200 - 230 -260V, 0 - 64 - 80 - 92 - 104 turns , 0.35mm CuL, 5 kV insulation
Secondary: -3-	0 - 15V, 0 - 6 turns of 0.7 mm. CuL

Winding data for transformer for Helix CPS:

ETD 49 core;	Afe = 210 mm.2; f = 10 kHz; t = 50 micro-S; B = 0.25 T; n = 0.4666 coil/V
Primary:	2 x 0 - 15V, 2 x 0 - 7 turns of 20 x 0.2 mm. flex
Auxiliary:	0 - 15V, 0 - 7 turns of 0.4 mm. CuL, 5 kV insulation
Secondary:	0 - 750V, 1000V , 1250V 1500V, 0 - 350 - 476 - 538 - 700, 0.17 mm. CuL

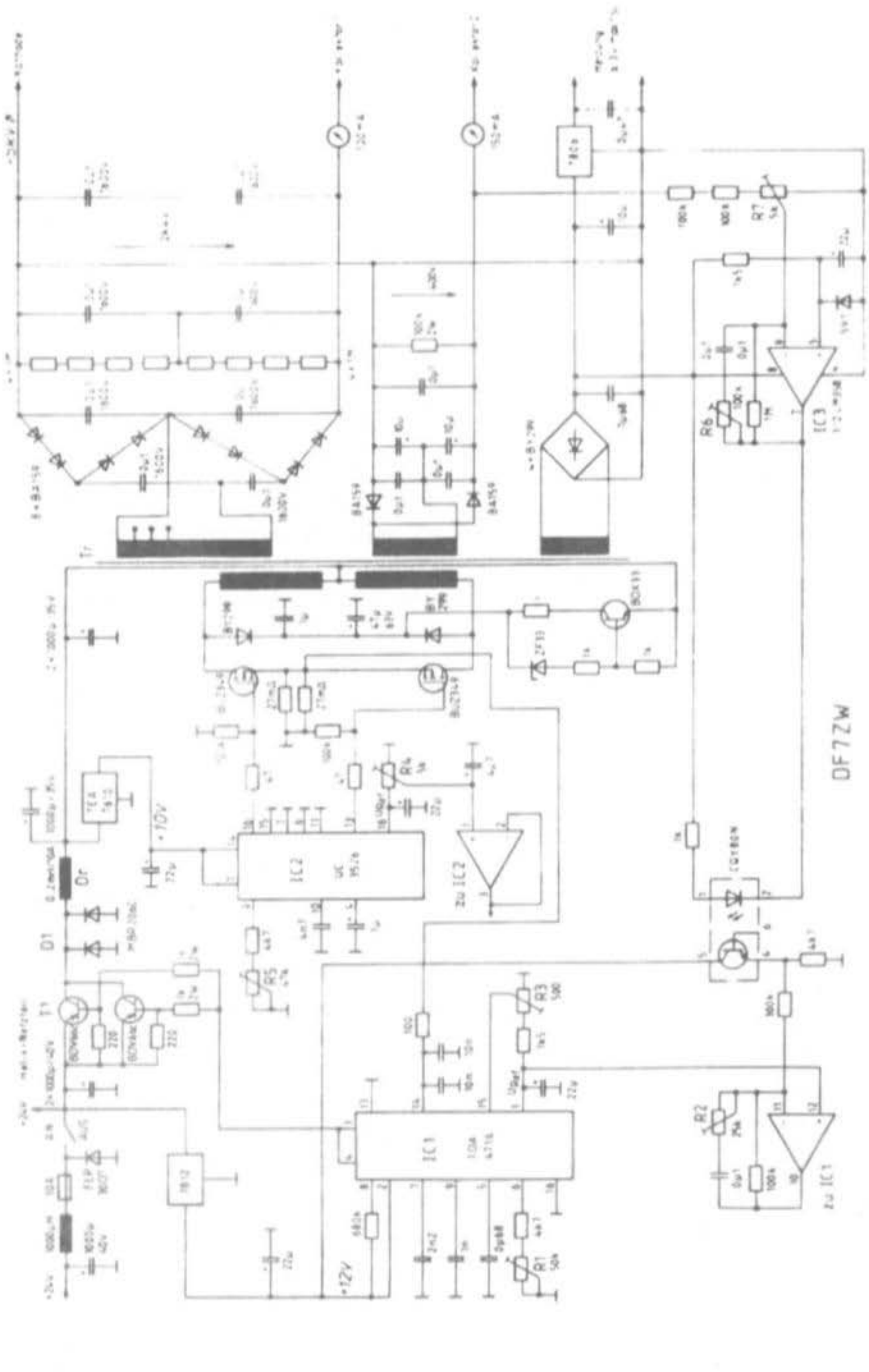

Fig.2: Circuit Diagram of the Collector 1, Collector 2 and Heater supplies

The specimens were wound manually by the author, using a hand drilling machine with a crank handle (clamped in a vice). The spool was held in the boring socket by a threaded rod.

First, the secondary coil is wound with flex. The beginning and end of the flex are also insulated, using insulating tubing, and soldered to one side of the row of brads on the spool.

For the thick flex (for the primary winding), only insulated ends are brought through. Four-ply insulating foil is used for insulation between the primary and secondary windings. The first layer of the high-voltage turns is then wound.

The layers of enamelled wire must be neatly coiled on one another and there must be a space between the wire and the edge of the spool.

Every layer is methodically impregnated with plastic spray (this is particularly important) in order to avoid any air pockets in the coil. Otherwise a high-voltage, high-frequency corona can arise in air pockets, which destroys the insulation. Immediately after spraying, two or three layers of insulating foil are tightly wound on. Now you can start the next layer.

(Be careful to ventilate the work-room properly, as the enamel vapours can damage your health.)

The coil ends are soldered to the row of brads on the spool still free.

Spray the fully wound and completely assembled transformer with plastic spray again and leave it to dry for at least a day before first attempting to use it.

4. POWER SUPPLY CIRCUIT ASSEMBLY

The circuits are put together as sub-assemblies, e.g. on breadboards, and mounted in tinplate housings. There has been no attempt to draft a layout, since this is only a single unit.

The down-converter (Fig.3) of the Helix CPS was mounted in a 74 x 111 x 30 (mm.) housing, and the subsequent high-voltage generation took place within a 162 x 102 x 50 (mm.) space.

The output voltage of -5kV from the Helix CPS must be carefully filtered, since disturbances to this voltage could lead to an interference modulation of the electron beam of the travelling-wave tube and thus generate spurious emissions on the output signal. All voltages, even the connections between the two housings, go through feed-through capacitors in each case.

From this point of view, the collector CPS (Fig.2) is not critical. It was also mounted in a housing with dimensions of 162 x 102 x 50 mm, which is cut open at one end and thus offers space for high-voltage rectification and filter capacitors.

The specimen CPS components were mounted in a type 302 Schubert housing. All tinplate housings were mounted with the front face to the back wall of the main housing, so that the lost heat from the power semi-conductors can be transmitted to the heat sinks attached to the rear wall.

The three meters, for Helix current and collector currents 1 and 2, are mounted on the front plate. These three units absolutely must be mounted flush and insulated, particularly the Helix current meter, which is connected to a 5kV circuit. The scales of the built-in gauges are not insulated from the moving spool!

A meter with an current range of 3mA is not commercially available. You can make one yourself from a meter with three-part scaling (e.g. 30V) by removing the compensating resistance and incorporating a shunt resistance.

Three switches are mounted below the meters - the main power switch, the Helix power supply switch and the grid-2 voltage switch.

For cooling purposes, the power semi-conductors are insulated and screwed to an aluminium bracket in the tinplate housings. The control electronics was assembled on small Vero boards. The controls for the optocoupler in the CPS, together with the stabiliser integrated circuit of the tube, must be well insulated before mounting, as the assemblies carry a voltage of -5kV. The storage chokes can be soldered so as to be self-standing. The two ferrite transformers and the high-voltage capacitors are mounted on a larger Pertinax board, as a support, insulated and mounted in the tinplate housing.

5. PUTTING THE POWER SUPPLY CIRCUIT INTO OPERATION

When the complete Helix circuit power supply has been assembled, the push-pull circuit power supply is put into operation first.

To drive the push-pull converter (with voltage control and current limitation), the intermediate voltage, U_Z, approximately 10 Volts, is provided by an external power supply at the diode, D1.

Now the IC 2 circuit power supply must emit pulses to the two MOSFETs in the push-pull. The frequency is set to 10 kHz using trimmer R5. Trimmer R4 is responsible for the width of the pulses and is set in such a way that a dead time of app. 5% is obtained. The control pulse for a MOSFET then becomes approximately 45 micro-seconds long. A few micro-seconds are left for switching from one MOSFET to another.

There is already a high voltage at the output while these setting operations are being carried out. The voltage level can be altered through the intermediate voltage, U_Z, and is measured by means of a high-voltage scanning head. The author briefly increased the high voltage in the specimen to 7kV to test the insulation of the ferrite transformer. Anyone who has a short-wave receiver available can use it to determine whether corona discharges are occurring in the insulation. If the high voltage is increased above a specific value, the well-known high-voltage frying effect arises abruptly.

The voltage should not be raised above 7kV, since the electric strength of the Wima capacitors is given as only 2kV. The test voltage of 3.6kV for the capacitors corresponds to a maximum high voltage of 7.2kV.

As a preliminary calibration for the voltage regulation, the output voltage of the CPS is now set to 5kV precisely. The output voltage of the operational amplifier, IC 3 (LM 358), is set to 2.5V using trimmer R7.

This concludes the calibration of the

push-pull converter, and the entire CPS can be put into operation

The power supply voltage of the CPS is set max. 4V higher than the level of the intermediate voltage, U_Z, of the last step in calibration. If the CPS is working properly, the output voltage is now app. 5kV. The switching frequency of the down-converter is set to 25 - 30 kHz by means of trimmer R1.

The I behaviour of the controller is altered through trimmer R2. The value of R2 is reduced until the control operates in a stable manner. To check this, an oscilloscope is connected to the pin-10 output of the controller, IC 1. It must be possible to measure a clean DC level without any superimposed AC. Now the supply voltage can be increased to 24V and the current limitation can be set using trimmer R3.

As load resistors at 1.666 M, several 2 Watt resistances are wired up in series. At 5kV, an output current of 3mA is flowing. The current limitation is set using trimmer R3. When current limitation is used, the voltage controller goes to its maximum value. This can be measured at pin-10, IC 1, as above. The Helix CPS is now fully calibrated.

To calibrate the collector CPS (Fig.2), the Helix CPS is switched off and the cathode of the travelling-wave tube is earthed, which corresponds to a short-circuit of the Helix CPS.

This measure makes it possible to connect an oscilloscope to the potential-separated voltage acquisition of the collector CPS, which would otherwise have a potential of -5kV.

The collector CPS is calibrated in an analogous manner to the calibration of the Helix CPS. The intermediate voltage, U_Z, is applied again and a collector-2 voltage of 400V is set. Trimmers R4 and R5 are subsequently set, as already described above. R5 is used to set a switching frequency of 15 kHz. There should now be a heating voltage of 6.3V at the output of the 7806 stabiliser IC. Should the voltage be too low, the stabiliser IC's earth can be set slightly high through a voltage divider.

The level of the collector-2 voltage is set using trimmer R7. For this purpose, an oscilloscope is connected to the output of operational amplifier IC 3 (LM 358), and trimmer R7 is altered. As soon as there is a big change in the output voltage of IC 3, it normally jumps, and the output voltage of the collector CPS is initially calibrated.

To put the down-converter into operation, the supply voltage of the collector CPS is increased by about 4 Volts. R1 is used to set a switching frequency of 25 - 30 kHz. The I behaviour of the voltage circuit is set for stable operation using trimmers R2 and R6. The collector CPS is now fully loaded with load resistors.

The collector-1 output is loaded with 60 to 70 mA at 2,400V and the collector-2 output with 130 mA at 400V. If necessary, the output voltage must be corrected again slightly, using trimmer R7. The current limitation is set, by means of trimmer R3, in such a manner that it just fails to cut in.

Finally, check again that the CPS is operating in a stable manner under full load. A slight secondary adjustment of trimmers R6 and R2 may be called for.

Now the travelling-wave tube can be connected and put into operation.

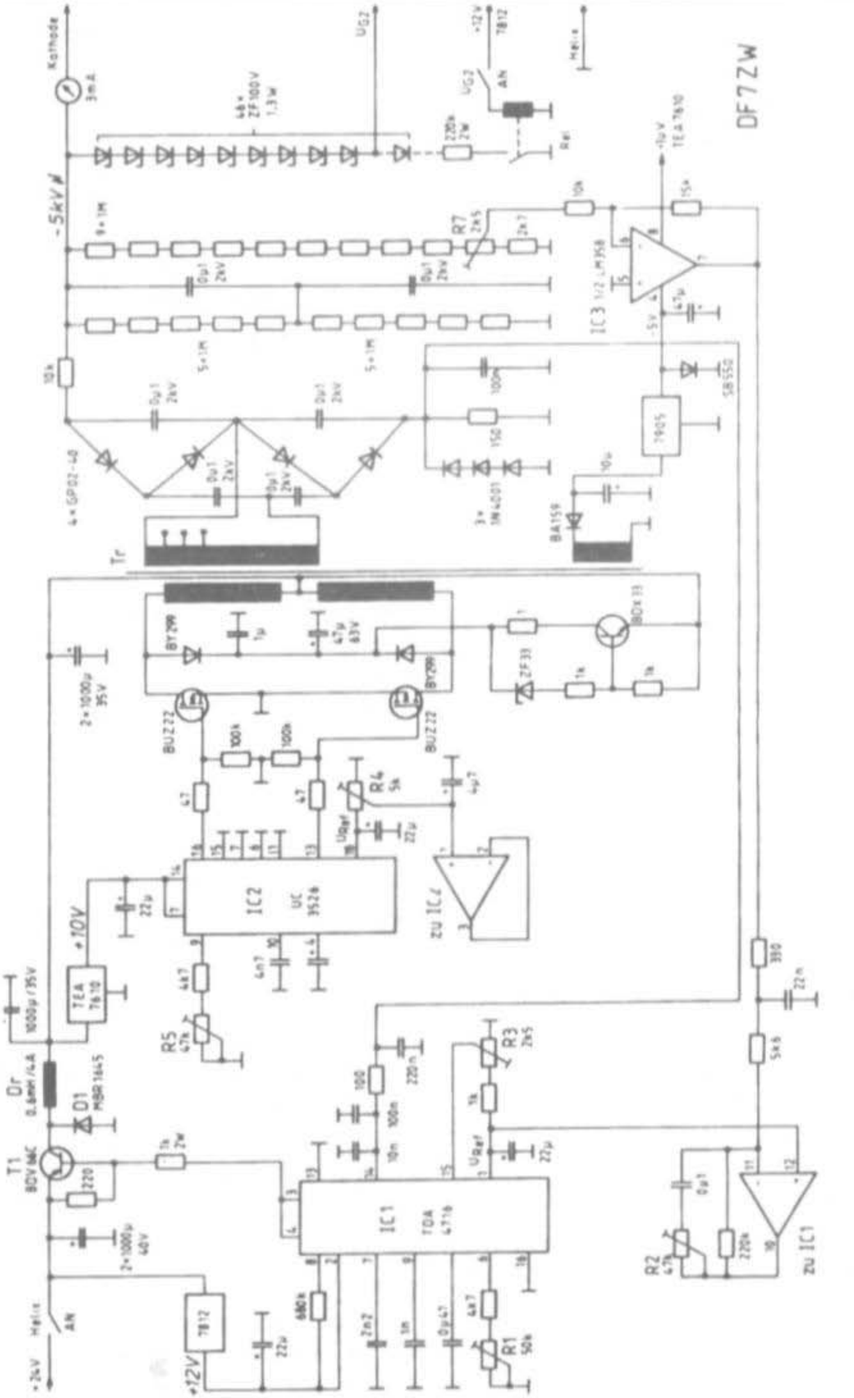

Fig.3: Circuit Diagram of the Helix and Grid 2 supplies

6. PUTTING THE TRAVELLING-WAVE TUBE INTO OPERATION

The travelling-wave tube should be mounted on a cooler sufficiently large to remove the lost heat. The manufacturer specifies a maximum flange temperature of 90 degrees. However, for test purposes it is even possible to operate the tube without a cooler for short periods. The housing temperature must be monitored, of course.

When the travelling-wave tube has been connected to the circuit power supply, the collector CPS is switched on. The tube itself is now in a closed condition, and no electrode currents are flowing. The tube is merely being heated up. The first time it is used it is advantageous to pre-heat the tube for several hours to re-activate the dispenser cathode. In the intervening period, the HF measurement system can be set up. You need a signal source for 10 GHz, the output power of which should be variable, as infinitely as possible, up to 50mW. At a driving power of 50mW, the tube is controlled right up to saturation.

The author used an X-band test single generator, variable between 8 GHz and 12.4 GHz, with a pin-diode attenuator, for the infinite variable adjustment of the output power.

An amateur radio transverter, the output power of which can be regulated through the control transceiver, is naturally also suitable. The travelling-wave tube output is connected to a power attenuator, to the output of which a thermistor head can be connected, or a spectrum analyser to analyse the signal.

Once the tube has been warmed up, the Helix current supply is brought into operation. Without any G2 voltage, the travelling-wave tube remains closed.

The G2 voltage is now applied, without HF control power. A current of app. 130 mA flows to the tube through collector 2. Without HF control, the Helix current and the collector-1 current are almost zero.

The sequence in which the voltages are applied absolutely must be followed. If the Helix voltage and the G2 voltage are switched on simultaneously, then the electron beam can not focus in the tube, and the Helix current limitation responds immediately. The level of the grid-2 voltage can be altered by the selection of the tap on the 46 series-connected Zener diodes. The G2 voltage is typically 4000 - 4300 Volts.

The control power is now switched on and slowly increased. The collector-2 current falls slightly and the collector-1 current increases as the control power and output power increase. If the control power is increased further, then the output power also increases, until the Helix current rises sharply. At this point, the maximum possible output power is obtained. If the control power is increased further, the electron beam is de-focused and the current limitation responds.

In this event, the G2 voltage must be switched off, and possibly the Helix voltage too, and the tube can then be put back into operation in reverse order.

The author was able to measure a maximum saturation power amounting to 50W in an experimental travelling-wave tube, shortly before the Helix current limitation set in. An output power of 15 to 20W can be obtained in continuous operation, using SSB and with a reasonable inter-modulation interval, with an amplification of more than 30dB. The manufacturer specifies an amplification of 37dB in the original frequency range between 11.7 and 13.25 GHz. The travelling-wave tube has proved itself to be very robust and to have a long working life in practise.

The author obtained some tubes from an amateur radio flea market, and was very surprised to discover that all the tubes were fully operational. Continuous-wave steady load trials were carried out for 30 minutes at power levels of 25W with the two worst examples.

The circuit power supply was capable of providing the power to operate the tube for longer periods without any modifications. Both tubes are still functioning today, with severe limitations, in spite of being considerably older and suffering from cathode consumption. The amplification band of the tube is very wide. Even at a frequency of 9 GHz, no great reduction could be detected in the output power and the amplification.

As the travelling-wave tube amplifier is wide-band equipment, it is especially important to use a spectrally pure control transmitter. The author's spectrum analyser can be used at up to 18 GHz, so that at 9 GHz the attenuation of the overtones can also be measured.

At an output power level close to saturation power, there was a reduction of app. 20dB. A typical value for wide-band amplifiers! There should be no problems in operating in the 10 GHz band with a slightly lower output power.

Unfortunately no measurements could be carried out at the time, since neither the spectrum analyser, nor the attenuators used, nor the adapter are specified by SMA for N standards greater than 18 GHz. Even at 18 GHz, measurements can already be very imprecise.

Crystal-stable frequency separation was used to control the tube for the measurement of the interference products on the output signal. The area around the carrier was analysed using the spectrum analyser. At the interval of the switching frequency of the circuit power supply, spurious emissions are attenuated by app. 50 dB, by comparison with the carrier, measured at an output power of 10W.

7. MEASURING EQUIPMENT USED

The measuring equipment used in the author's laboratory to develop the circuit power supply is listed here.

You can usually copy my experiments successfully with less measuring equipment.

Marconi 8-12, 4 GHz Signal Source 6158.

HP Spectrum Analyser System 141, high-frequency plug-in unit 8555A, intermediate-frequency plug-in unit 8552A.

HP 432A Power meter, 8478B measuring head.

Tektronix Current Probe P 6042; DC to 50 MHz.

Tektronix Oscilloscope 465.

Weinschel Attenuator Model 47-20-43, 50 W, 20 dB, DC-18 GHz.

Supply source: BFI IBEXSA, Dietzenbach

Fluke 80K-6HV 6kV high-voltage scanning head.

8. COMPONENTS AND SUPPLY SOURCES

Rel. high-voltage relay; Meder-type high-voltage reed relay H 06/24 (G 127.830) 12 Volts; Buerklin, Munich, order no. 30G395

Ferrite transformer: ETD 49 core, N 27 material; Siemens order no. B66367-G-X127; Spool order no. B66368-A1020-T1; Clip order no. B66368-A2000

Power supply Circuit IC's: TDA 4716C, Siemens order no. Q67000-A8313

UC 3526 Unitrode, P + H Electronic, Mainz-Kastell

High-voltage rectifier diodes: GP02-40 General Instrument; P + H Elektronik, Mainz-Kastell

High-voltage capacitors: 0.1 micro-F 2,000V and 0.1F 1600V; Type MKP 10, Wima

Storage chokes, complete or cores only: Kern Kool Mu Powder Cores; Order no. 77552-A7; BFI-IBEXSA, Dietzenbach

9. LITERATURE

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(2) Magnetics Kool Mu Powder Cores brochure (BFI-IBEXSA)

(3) Siemens Ferrite and Accessories Data Book 1990/91

(4) Siemens Travelling-Wave Tubes Data Book 1986/87

(5) Kilgenstein: Power Supply Circuits in Practice
Vogel Buchverlag Wuerzburg; ISBN 3-8023-0727-5
(Specimen circuits for TDA 4716 and SG 3524)

(6) Joachim Wuestehube: Power Supply Circuits expert
Verlag,; ISBN 3-88508-793-6

(7) Kaes/Pauli: Microwave Technology
FranzisVerlag Munich, ISBN 3-7723-5594-3

Michael Kuhne, DB 6 NT and Dr. J. Jirmann, DB 1 NV

Measurement Aids for the UHF Amateur

In the development of assemblies for the UHF and VHF bands, the amateur usually has to make do without expensive measuring equipment. To make the work easier, we have designed two simple pieces of auxiliary equipment, which have already solved many measurement problems, namely an adjustable broad-band measurement amplifier and a divider operating up to 5 GHz.

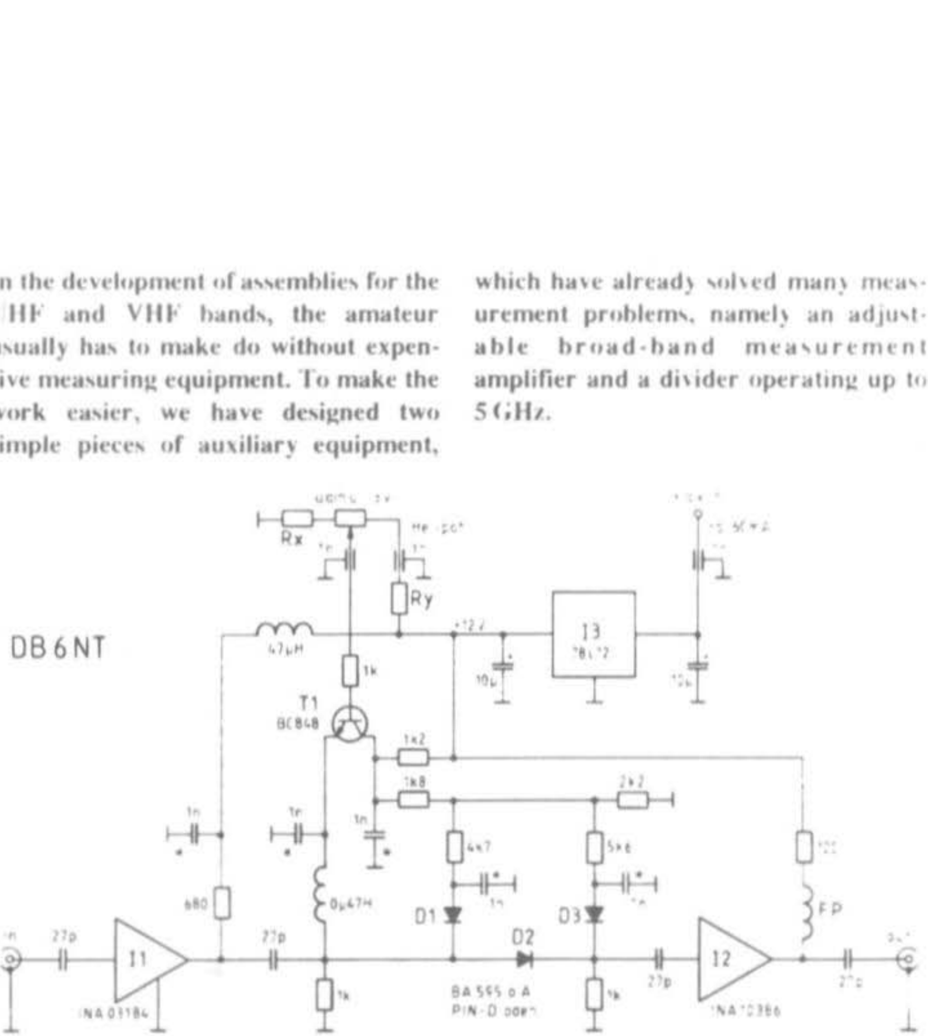


Fig.1: Broadband Measurement Amplifier: 100 MHz to 2.5 GHz

1. THE BROAD-BAND MEASUREMENT AMPLIFIER

1.1 Introduction

A broad-band and low-noise pre-amplifier with a band width going up to the 13cm. band can be used for many purposes, e.g. as:

- → Pre-amplifier for diode detectors in sweep measurements
- → Pre-amplifier for older spectrum analysers or microwave receivers
- → Makeshift aerial amplifiers

A few years ago, it would scarcely have been possible to produce such band widths on a repeatable basis using amateur resources, but nowadays integrated amplifiers are available at reasonable prices, which are a match, with regard to frequency response and noise factor with individually balanced discrete amplifiers.

1.2 Measurement amplifier circuit

Fig. 1 shows the result of the experiments: the first stage of the measurement amplifier is equipped with an INA 03184 MMIC (monolithic microwave integrated circuit) from Avantek/Hewlett-Packard. This module has an amplification of about 25 dB at band-widths between 0 and 2.5 GHz. The lower limiting frequency is determined only by the coupling capacitors at the input and output. The noise factor is below 3 dB.

The amplifier stage is followed by a PIN diode damping element in a PI circuit to adjust the amplification. This circuit will certainly be familiar to older readers from the television tuners of the pre-MOSFET age, and provides constant input and output impedances over the adjustment range. The BA595 SMD diode used, from Siemens, is especially suitable for damping elements in the frequency range between 1 MHz and 2 GHz.

The second amplifier stage is equipped with an INA 10386 from Avantek and

Fig.2: Measurement Amplifier frequency response

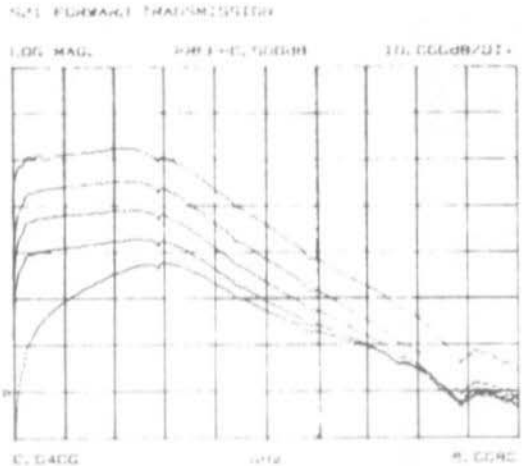


Fig.3: Control behaviour of the PIN Diode regulator

supplies about 25 dB amplification. The remaining circuit elements merely set the operating points of the amplifiers and PIN diodes. A voltage regulator stabilises the supply voltage, which should lie between about 14 and 20 V. Higher voltage levels of up to 35 V are possible if the 78L12 is replaced by a 78M12 or 7812 and appropriately cooled.

1.3 Measurement results

The amplifier described produces an amplification of 50 dB in the frequency range between 100 MHz and 2 GHz. As Fig. 2 shows, an amplification of 30 dB can still be obtained at 3.5 GHz, and 10 dB can still be obtained at 5 GHz. The lower frequency limit of 100 MHz can be pushed down if necessary by enlarging the coupling capacitors by 27pF.

Fig. 3 shows the effect of the PIN diode damping element. The adjustment range covers more than 20 dB at frequencies of less than 2.5 GHz, which should be enough for most applications. The additional frequency response is negligible at damping levels below 20 dB.

Measuring the noise factor using an Eaton 2075 noise factor meter yielded values of 2 dB at 100 MHz and 3 dB at 2 GHz.

1.4. Assembly and parts list

A DB6NT 001 printed circuit board made from 0.5 mm. thick RT Duroid 5870 forms the basis for the assembly of the measurement amplifier. Fig. 4 shows the layout. For assembly as per the components diagram (Fig. 5), proceed as follows:

First make the holes for the connectors (SMA or SMC flanged types), carefully producing a clearing hole on the earth side of the printed circuit board. The connectots can then be soldered in from the earth side. The flange should be soldered cleanly to the earth surface. Check carefully to make sure there is no copper swarf causing a short-circuit.

Then through-hole plate the earth connections of I1 and I2 in the usual way using thin brass foil and solder on the remaining assemblies (the order of operations is not critical here). But you should respect the basic rules when working with components where there is any danger of electros-

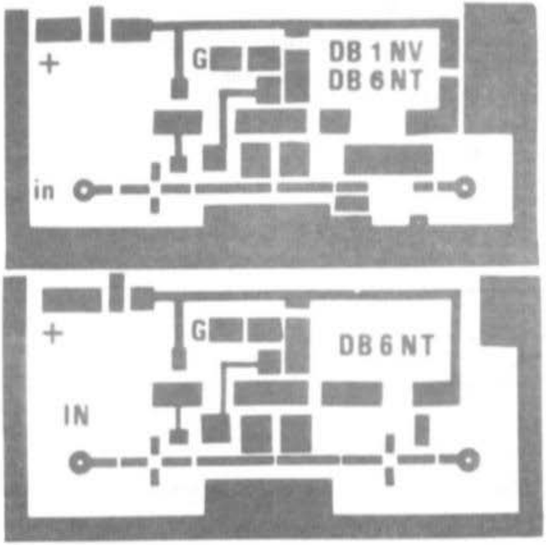
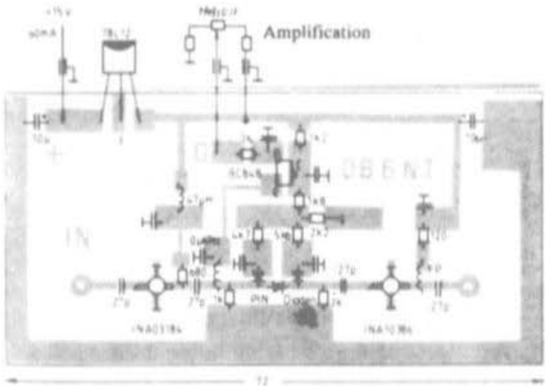


Fig.4:
Printed Circuit Board Layout of the Divider (top) and the Measurement Amplifier (bottom)



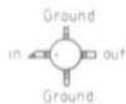


Fig.5:
Component Layout for the Measurement Amplifier

working with components where there is any danger of electrostatic phenomena and keep a steady hand when fitting the SMD components. Make sure you fit the integrated amplifiers I1 and I2 in the correct position - the slanting leg is the input! The blocking capacitors marked with a star on the circuit diagram, which are rated at app. 1nF (the value is not critical) must have the smallest possible format (SMD size 0603) and are soldered "vertically" into appropriate slots between the top and bottom sides of the PCB. This certainly requires precision work, but on the other hand it does provide extremely short earthing paths.

The entire assembly can be built into a suitable sheet metal housing. Feed in the operating voltage and the setting voltage for the PIN diodes through feed-through capacitors. The stop resistors Rx, Ry for the amplifier potentiometer are chosen in such a way that the setting voltage can be varied from about 0.5 V to 3 V.

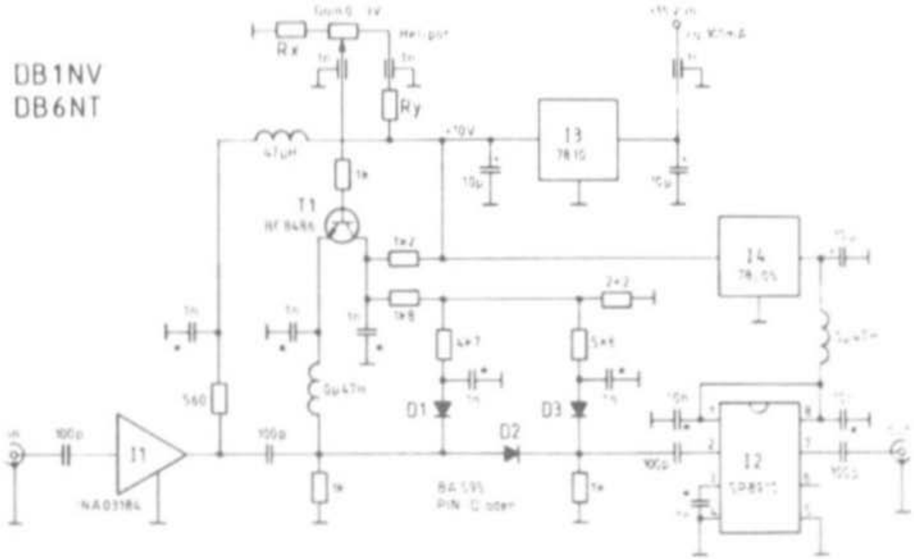


Fig.6: Frequency Divider circuit

The measurement amplifier is balance-free and should operate as soon as the supply voltage is applied if the damping element is set for minimum damping (3 V at adjustment input).

1.5. Parts list

I1:	INA 03184, Avantek/Hewlett-Packard Operates above BFI
I2:	INA 10386, Avantek/Hewlett-Packard Operates above BFI
I3:	78L12, or 78M12 (see text)
T1:	BC848b
D1-D3:	BA595, Siemens
FP:	Ferrite bead

All resistors: SMD, Minimelf or Chip - size 1206

All inductors: SMD

All capacitors: SMD, size not critical, blocking capacitors marked with a star on the circuit: size 0603

2. THE PRE-DIVIDER

2.1. Circuit Description

Until now, direct frequency measurement for the amateur at a reasonable cost has been possible only up to about 3 GHz (the frequency limit of the NEC dividers uPB 581 and uPB 582). Anyone who simply couldn't get hold of a microwave meter had to rely on elaborate transfer oscillator processes with harmonic mixing or expensive GaAs pre-dividers.

Over the last year or so, a family of pre-dividers from Plessey has become familiar which operate at up to 5 GHz for a price of app. DM 120. Types SP8902, SP8904, SP8908, SP8910 and SP8916 are available at present - the last two numbers give the divider factor. Thus any existing 500 MHz meter can be expanded for frequencies of up to 5 GHz.

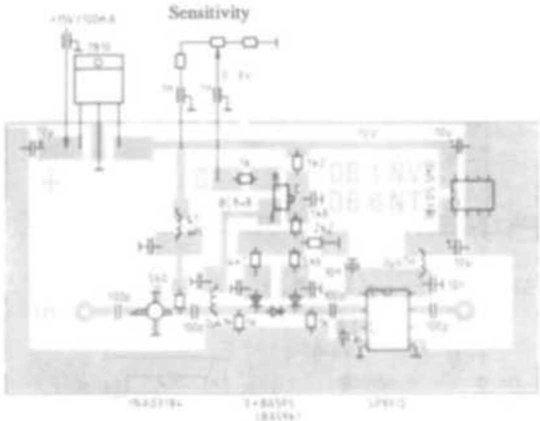
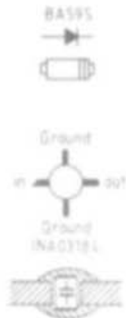


Fig.7:
Component Layout of the Frequency Divider



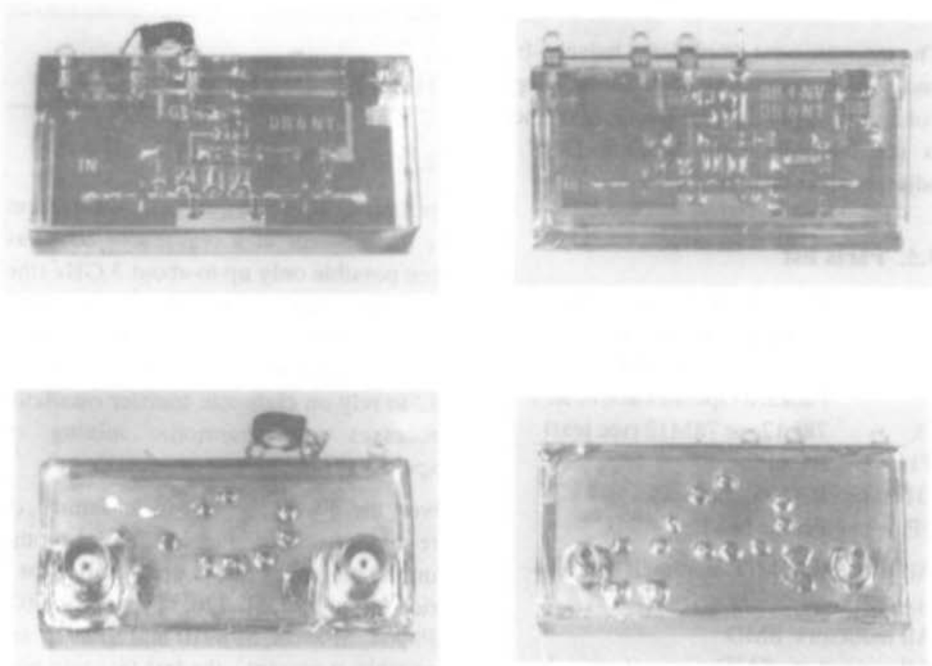


Fig.8: Prototype Measurement Amplifier and Divider units

For the frequency divider, type SP8910 was used, which operates in the normal DIL-8 housing with a scale running from 10 up to 5.5 GHz and needs a supply voltage of only 5 V.

As Fig. 6 shows, only the second amplifier, I1, was replaced by the pre-divider in the measurement amplifier circuit, and a permanent voltage controller was added for the 5 V power supply. The operating voltage of the first amplifier is still only 10 V, so that the entire circuit can be powered at 12 V. The pre-amplifier, I1, not only increases the input sensitivity of the divider, but also protects the divider module, I2, which is ten times more expensive, should the input level be too high. The pre-divider operates in the frequency range from app. 140 MHz to 6 GHz - with careful level app. 100 MHz to 6.5 GHz.

2.2. Assembly

A DB6NT printed circuit board made from RT Duroid 5870 0.5 mm. thick forms the basis for the assembly of the pre-divider. The layout has already been shown in Fig. 4, and the assembly as per the components diagram (Fig. 7) is carried out as for the measurement amplifier:

First make the holes for the connection bushes (SMA or SMC flanged bushes), producing a clearing hole on the earth side of the printed circuit board. The bushes can then be soldered in from the earth side.

Then through-hole plate the earth connections from I1 and shorten all legs on I2 flush with the housing, except pins 4 and 5. Insert pins 4 and 5 through holes in the printed circuit board and solder them to top and bottom sides. All other connections are soldered only to the top side using the strip conductors. When installing the remaining components, once again insert the minimum size units (SMD size 0603) - app. 1nF and app. 10nF blocking capacitors - (values not critical, marked with a star in the wiring diagram) "vertically" into appropriate holes between the top and bottom sides of the printed circuit board. Like the measurement amplifier, the entire assembly can be incorporated into a suitable sheet metal housing. Fig. 8 shows two sample set-ups of the author's.

The pre-divider needs no balancing either. If there is no input signal, the SP8910 oscillates at about 5 GHz, and an unstable signal around 500 MHz can be measured at the output. This phenomenon should be well-known from experience with many other emitter-coupled logic dividers.

2.3. Parts list

I1:	INA 03184, Avantek/H.P.
I2:	SP8910 BDG, Plessey
I3:	7810
I4:	78L05
T1:	BC848b
D1-D3:	BA595, Siemens

All resistors: SMD, Minimelf or Chip, size 1206
All inductors: SMD
All capacitors: SMD, size not critical, blocking capacitors marked with star on circuit: size 0603

Other:	2 SMA or SMC flanged bushes
	3 feed-through capacitors, app. 1nF
	1 DB6NT 002 PCB
	1 tin-plate metal housing 74 x 37 x 30mm

Dipl-Phys. Nothart Rohde

EMC - and its Consequences Part-2

Unfortunately, the multiples of the divided down final frequency crop up most persistently in the oscillator as side-bands; this is especially true in cases when the distributor supplies high input voltages, the signal is intermediately amplified, or the outputs are carrying a leading load (here, one oscillator probe is already too many).

In the circuit shown in Fig.9, the pre-divider carries only a high-impedance load. So that the limiting frequency does not decrease too much due to the feed through capacitor and the cable, a resistance is wired up in parallel to the feed through capacitor. Two amplifier stages follow in the digital section, which prevent the feedback shown in Figs.5/6. The two stages can also be used separately.

Approximately speaking, it can be assumed that interference signals which are fed in at the output of an individual stage (a gate or a transistor) appear damped by 30 to 40 dB at the input of this stage. If an inverse-coupled gate is used as the sole buffer stage, you must ensure that the output is not driven into saturation.

7. TRANSMISSION OF ANALOGUE SIGNALS

You might think there are no serious errors to be made in this area; but far from it! It still seems to be a problem to terminate long coaxial cables with the correct impedance.

Urged on by some zealous IC manufacturers, the PC and video sectors are becoming enthusiastic about faster and faster drivers, which can convert still more nF/sec.. Incidentally, these are the same suppliers who wish to dazzle the user by offering more than a million shades of colour. Unfortunately, a circuit design of this nature messes up more than the power supply. Similar conclusions were reached years ago, even using the much simpler NE 555.

It is thus argued that the moving loads could ruin the 12-bit precision which - let's admit it strai ght away - has been bought so dear This is certainly not true in reality, for there are very precise 50/75Ω resistances with tolerances below 1%. Moreover, the signal is very easily distorted, due to the dependence on the frequency of the cable damping, irrespective of the wiring, as soon as you move away just a little from the DC voltage on the frequency scale.

So before you use the newest model of driver or receiver, you should first of all look for tried and tested ("good-natured") broad-band operational sections - for example, the LM 6361. These operational sections operate accurately enough, as long as they are operated as impedance converters (V = 1). The LF 356 also provides very respectable band widths, as long as the voltage dispersion remains in the voltage range. For zero-symmetrical signals, there is also the kind of circuit shown in Fig.8, with only a resistance leading to earth at the receiver.

8.
HOUSING RESONANCES

Screening measures can lead only to a damping of interferences coming in from outside, but do not eliminate the interference itself. Accordingly, it makes sense to keep fields as small as possible, even in closed housings, and to look for points with a low field when mounting the input and outputs.

From the EMC point of view, strip conductors and open resistance circuits have little significance, for the saving from the easily compressed strip conductors is only fictitious, since they result in greater expense on screening.

The diagram in Fig.10, on the right below, has very general application for feeds. It shows a case in which internal and external fields attack, and finally break through, the screening. It is true, especially for coaxial circuits, that the screen must be connected to the housing wall precisely at the point where it penetrates it. Otherwise sheathing waves occur, which lead to coupling. If you wish to save on expensive high-frequency plugs and sockets in high-frequency assemblies, you can use Teflon coaxial cables and solder the sheathing to the feed through point. This applies irrespective of scale, both for the smallest tinplate housings and for screened rooms.

In any case, it is foolish to solder the coaxial screen neatly to the printed circuit board and feed the cable neatly out of the housing, insulated by rubber rings! Incidentally, I didn't internalise this principle until a ZZF acceptance test had gone wrong for precisely this reason.

Apart from the interference from the outside world, high fields inside a housing

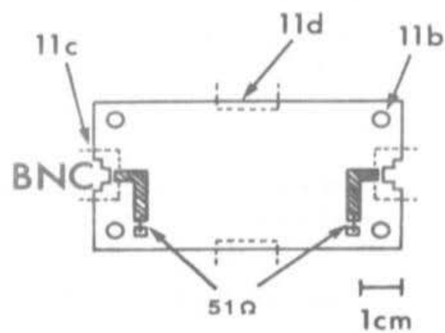


Fig.10: Foil side of a Printed Circuit Board for investigating Resonances.
- **The various earthing options for the earth plane refer to Fig.11.**

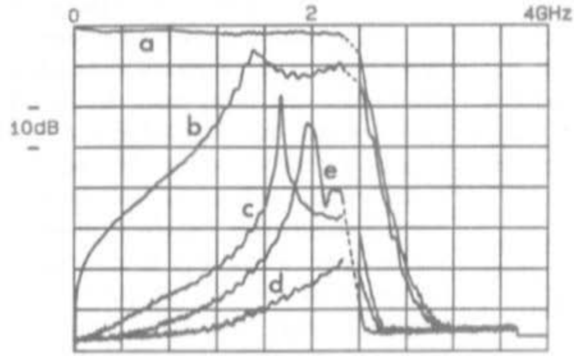


Fig.11:
Diecast Housing Simulation, with Printed Circuit Board
- a) Input power
- b) Earthing with 4 M3 x 10mm bolts at corners
- c) 2 additional Cu links on BNC sockets
- d) 2 additional earthing links in middle of PCB
- e) Earthing through 4 links only, without bolts

also have functional disadvantages: the oscillation of amplifiers, faulty selection of filters, etc. It is often very helpful here to lay thin coaxial cables (approximately RG 178) directly on the printed circuit board.

8.1 An example

Let's assume that two signals are fed into a housing through commercially available BNC flanged bushes, run a few centimetres in a compressed 50-Ohm circuit, and then arrive in a filter, an amplifier, or another component.

The question now arises of how to connect the two circuits.

Fig. 10 shows a test printed circuit board, looking at the foil side.

The continuous earth surfaces of the 1.5 mm. epoxy printed circuit board were connected to the housing in various ways.

The circuits were terminated using low-induction chip resistances, with a 3 mm. wide strip of copper foil leading directly from the resistance to the earth surface. Somewhere about half-way up, the printed circuit board was fastened in the tinplate housing (80 mm. x 40 mm. x 25 mm.).

Initially, an attempt was made to simulate a diecast housing. To this end, an intermediate floor was soldered in at a distance of 1 cm. from the foil side. A

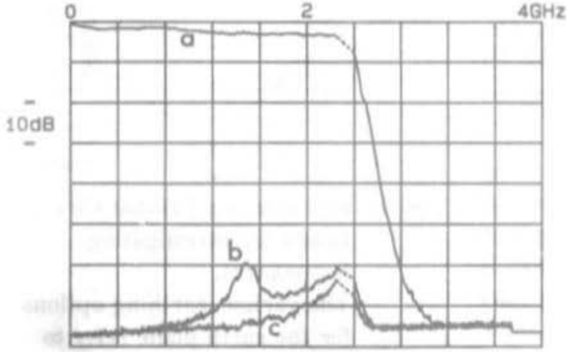


Fig.12:
Tinplate Housing Simulation with PCB as per Fig.10
- a) Input power
- b) Cover over conductor side, loose
- c) Cover soldered on

spectral analyser was connected to one socket of the test piece, and a locked-in oscillator was connected to the other. Fig.11 shows the result clearly. In connection with curve b, we see that the earthing through bolts, typical for this kind of housing, is already out of the question in the 70-cm. band, although such structures are to be found from time to time. Presumably, even the selection of a single high-quality filter, e.g. a helix filter, would ruin things with this rig. The signal generator, unfortunately, operates only up to 2.3 GHz. But the assumption is obvious that the coupling is also maintained, or even strengthened, above this level. The resonance behaviour must be brought about through the printed circuit board and not through the volume. This is indicated by the fact that the curves scarcely change if the cover is placed over the earth surface. Moreover, the resonance is strongly dampened in curve c as soon as you touch the earth surface with your finger.

Fig.12 shows another result: if the earth surface of the printed circuit board is soldered to the housing all round (as is usual with tinplate housings), the coupling is considerably reduced. Even the effect of the cover can then be recognised.

9. POWER PACKS

A neat power pack, as every reader knows, is always worth a good technical article. It is just as well known that switching power packs cause more or less pronounced interference, depending on their technology. But it would be tilting at windmills to try and abolish switching power packs. Moreover more IC's are being put on the market which generate other positive or negative voltages from +5 Volts, directly on the consumer, using rapid switching processes.

CMOS-based modules, especially those which operate only with switched capacities, are not so critical (e.g. type 7660 voltage inverter). But anyone who can manage it from the design point of view should stay with linear controllers. Inputs and outputs must be blocked on a large scale in order to prevent wild oscillation and thus further interference.

In any event, switched power packs generate secondary lines in the oscillators and thus secondary reception or inexplicable interference.

With powerful 5-Volt circuit power packs, it is already normal today to drop linear controllers completely and use the pulse-width repetition rate for control. This leaves a considerable high-frequency residual ripple which is difficult to combat.

The minimum equipment required here, from a metrological point of view, is a measurement receiver, or better a spectral analyser.

Examples of remedial measures are: separate feeds from the power pack for each assembly, so that chokes can be used with noticeable inductance and without saturation. With additional filtration using electrolytic capacitors, only types compatible with switching power packs should really be incorporated.

Experiments can also provide additional help with the earth lead. It can thus be expedient to treat the earth circuit of the power pack like a "hot" circuit, choke appropriately and only connect directly with the system earth at the consumer. These measures naturally cause the voltage at the consumer to fall to less than 5 Volts. Sensor circuits which measure the levels directly at the consumer are thus expedient.

Whereas in computers the power pack normally causes interference only in the outside world, with communications engineering equipment it causes internal interference as well. It is particularly irritating in this context that you have usually not thought about the magnetic sensitivity of many assemblies, whether they are coils in the intermediate-frequency section or chokes incorporated as a precaution.

In many pieces of equipment, we look in vain for a clean earth on which internal signals can be based, since they would be ruined through capacitance to source inductance. Strictly speaking, we have to live with these signals somehow and conceal them as well as possible. For example, the power pack cycle should not involve a whole fraction of the last intermediate frequency (e.g. 10.7 MHz).

10. MATERIAL QUESTIONS

If we pick up the stock list of a wholesaler, or the list of components released for constructors, in hand, we can look at a real wonderland of digital technology; any function you desire is there at once, in several variations.

In contrast, under the heading "Other, radio-shielding equipment", there is virtually nothing available. Neither the many network sockets with built-in filters nor the extensive range of chokes (e.g. from Siemens) can hide that fact.

The bold engineer who wants to do something in the EMC sector doesn't exactly have a simple task. For example, for experiments he or she needs a feed through capacitor with a specific value for de-coupling a low-frequency signal. He or she tests his or her reference sources and can then choose between:

The desired value, for soldering, 1 kg. bulk material direct from the manufacturer

A moderately-priced Unikat from a special transmitter, but unfortunately with exotic values

And a rather large, complete interference filter with an external inch thread for DM 15 - from a wholesaler

Even if, for design reasons, the choice has fallen on a housing which is difficult to solder, the bold engineer will still prefer to produce a construction which is suitable as regards EMC, from the point of view of the circumstances and costs, for the next development but one.

Some work has been going on for some time on 19-inch housings and their metric successors. A compelling design at a reasonable price is not even in sight, of course. Moreover, plug-in systems are in principle a pattern of slot aerials with corresponding problems; above all, the additional costs of EMC security.

Things look far worse for small housings! With the really wonderful plastic housings, e.g. for portable equipment, the manufacturers unfortunately have merely made a declaration of intent that they would like to metallise these units inside soon and produce them in suitable numbers. Thus there remains only the circumstantial route of incorporating a screening housing made of sheet metal into the beautiful plastic housing as well.

Nor is the situation in the market for screening housings much more convincing. In the lowest price range we find aluminium diecast housings, most of them British, together with tinplate housings in lightweight format with double covers, and "Teko boxes", with a cover on one side, as an intermediate form.

No milled aluminium housings are available at less than about ten times the price. These can certainly be bought ready-made, but it is usually worthwhile to get a quote from the nearest workshop with NC machines. I don't know of any other suitable housings.

The resultant compromises are familiar to everyone (see also

Fig. 11) - soldered tinplate boxes which you can't answer for in service conditions, or the combination of diecast and beginners' stock material, which rusts splendidly.

There simply isn't a favourably-priced housing with multiple printed circuit board earthing suitable for high frequencies.

Aluminium has broken through in the professional market due to its low weight and its easy machineability in metal housings. Anodic oxidation is out as corrosion protection, since it harms conductivity. The state of the art alternative is called chromating, a somewhat more moderate form of oxidation. This also forms a good base layer for lacquering.

The only real problem here is that there are scarcely any companies who provide chromating as a service. Thus a galvaniser in Freiburg is working virtually for no-one but me, and will shortly have to stop on allegedly environmental grounds.

Incidentally, there is a firm in Bayerischen (EM-Geraetebau, 8038 Mammendorf) which makes 19-inch standard housings from nickel-plated aluminium. This is still an interesting solution, even if nickel has been much talked of in connection with allergies.

Finally, something more on the topic of conductive foamed material. Conductive foamed materials filled with carbon have a field of application from IC packaging all the way to microwave absorption. I discovered the idea of using standard high-frequency damping panels in this journal one day. After a few measurements, I have come to the conclusion that the material must have rather low impedance, and can thus not be used for every format. I obtained good results from about 300 MHz upwards in damping larger housings which contained assemblies which were not completely sealed against the passage of high frequencies and had only single-screened cables. Inside smaller housings, the material is used rather to salve one's conscience. In experiments as per Figs. 11 and 12, covers with foamed material glued in place were also used selectively. But the changes were so slight that the preparation wasn't worth it. Here the printed circuit board had to be damped as well, not merely the cavity.

This fine material naturally also brings problems with it. Who guarantees that flexibility is retained for polyether with a low impedance base and that crumbs of the material won't be turning up in the circuit after only a year? What's the position about obtaining material? My preferred conductive foamed material, which functions really well, is obtained from packaging material for IC's from a well-known mail order firm. Naturally, they won't say where the material comes from and how it can be obtained.

All the problems listed here should indicate that in the achievement of EMC even the little things can develop into obstacles for the electronics engineer. This does not automatically apply to big companies in the car industry, or to communications or military technology, but it affects smaller companies and amateurs who like to experiment all the more. So it would be an important step to create a suitable forum for exchanging information and reference sources, for instance in the form of a permanent EMC section in the technical journals.

Editorial comment:

Attention should be drawn to the relevant DIN-VDE standards, for example:

- DIN VDE 0839 (Part 1 / Version 11.86): Electro-Magnetic Compatibility
- DIN VDE 0870 (Part 1, 07.84): Electro-Magnetic Influence
- DIN VDE 0872 (see Appendix 1/ 01.91): Radio Shielding of H.F. Eqpt. for Industrial, Scientific, Medical and Similar Purposes
- DIN VDE 0878 (Parts 1 & 2): Radio Shielding of Installations; (Part 3): Electro-Magnetic Compatibility of Information Processing and Telecommunications Engineering Installations

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Compiled by Hansgeorg Meyer:

Electro-Magnetic Compatibility of Automation Systems
VDE-Verlag 1992;
ISBN 3-8007-1511-2

Wolfgang Schneider, DJ 8 ES

28/144 MHz Transverter

Transverters for the 2m. band are always topical. In an attempt to match the current state of the art in amateur radio technology, a transverter was developed, with the help of modern components, which converts the 144 - 146 MHz range into the 10m. band.

Concepts such as high-level signal strength and test signal spectral purity have taken on increasing significance in recent times. The criterion that equipment can be copied with security is also of great importance for the author.

The transverter described below represents a circuit which corresponds to today's requirement profile.

1. CIRCUIT DESCRIPTION

Fig.1 shows a detailed circuit diagram of the 28/144 MHz transverter. In synthesising the frequency, we went back to a circuit proven many times. The crystal oscillator oscillates at 116 MHz with a U310 (T_1). This signal is amplified by the next stage. An MSA1104 (IC_1) integrated broad-band amplifier is used here. It supplies an output level of 50 mW.

The SRA1H high-level ring mixer requires an oscillator level of

+ 17 dBm (50 mW). This type can be used at up to 500 MHz.

The Pi damping element, consisting of R_1 to R_3, is used for the control transmitter output adaptation. For a "clean" transmission signal (intermodulation products with multi-tone control

P_{in}	dB	R_1	R_2	R_3
1 mW	0 dB	---	0 Ω	51 Ω
2 mW	3 dB	300 Ω	18 Ω	300 Ω
5 mW	7 dB	120 Ω	47 Ω	120 Ω
10 mW	10 dB	100 Ω	68 Ω	100 Ω
20 mW	13 dB	82 Ω	100 Ω	82 Ω
50 mW	17 dB	68 Ω	180 Ω	68 Ω
100 mW	20 dB	62 Ω	240 Ω	62 Ω

All values are taken from the E12 or E24 ranges

Table 1: Resistance values for the damping element

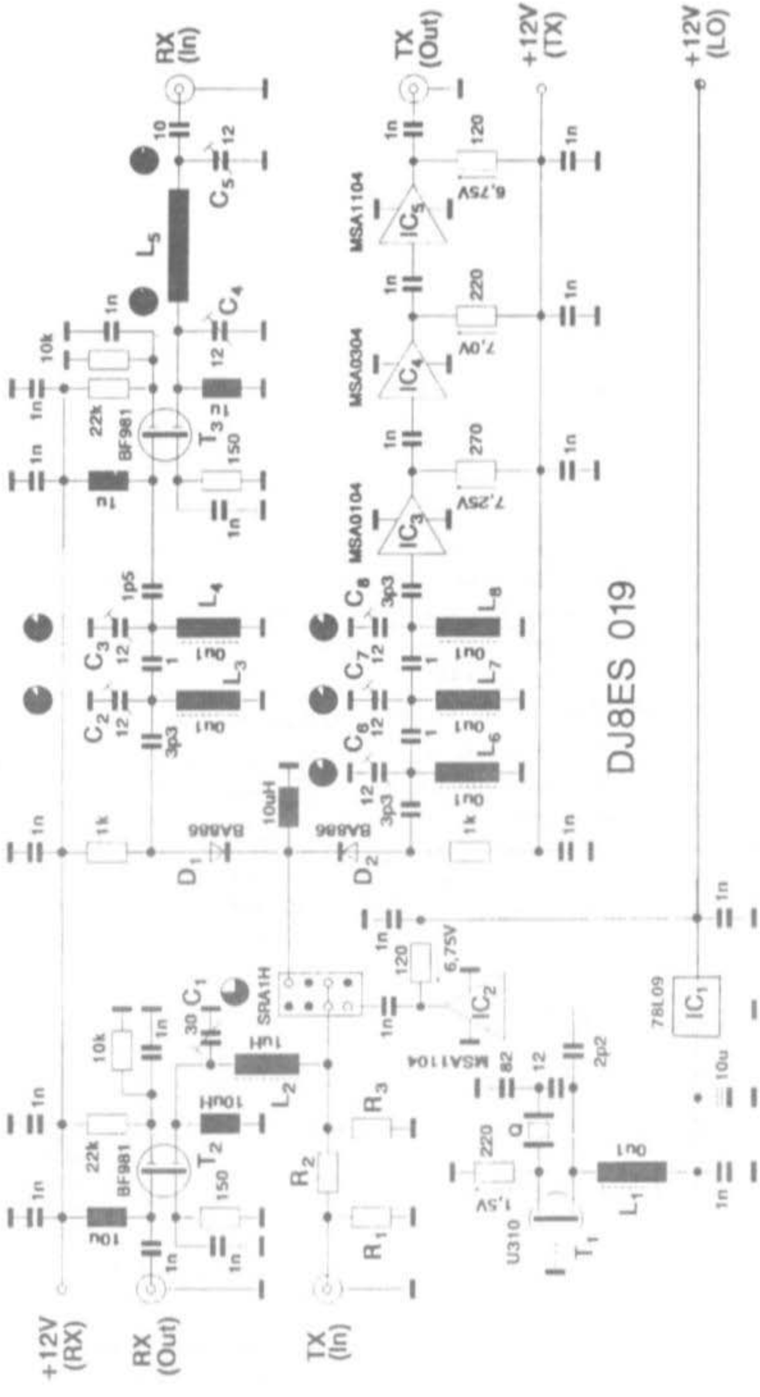

Fig.1: Transverter Circuit Diagram

< 50 dB), the ring mixer must be driven to full output at max. 1 mW (0 dBm). Table 1 shows the resistance values needed for the damping element plotted against the output selected. All specifications can be traced back to standard values.

The damping element simultaneously serves the ring mixer as a broad-band 50Ω cut-off. Parallel to this, the received signal is measured with a high impedance and matched to a BF981 (T_1) - a low-noise transistor stage which provides the intermediate-frequency amplification required - using L_s and C_1.

The 2m. received signal is transformed onto the gate of the BF981 (T_1) through a Pi filter (aerial impedance 50Ω). The pre-amplifier is followed by a 2-circuit filter. At the same time, the operating voltage supplied (+ 12 V) is switched through the PIN diode, D_1 (BA886).

In the transmission mode, diode D_1 (BA886) is activated. The signal first passes through a 3-circuit filter. The subsequent amplifier is assembled with integrated broad-band amplifiers (IC_1, IC_2, IC_3). The combination of MSA0104, MSA0304 and MSA1104 guarantees an output level of 50 mW (+ 17 dBm) here, with a good 40 dB amplification.

In practical operation, this transverter can be supplemented with any power

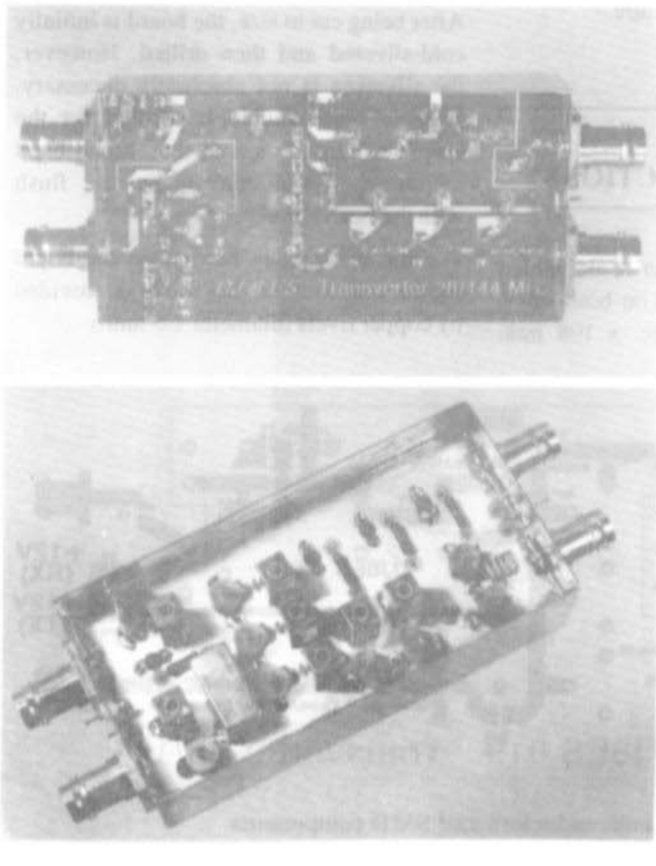

Fig.2:
Specimen Transverter

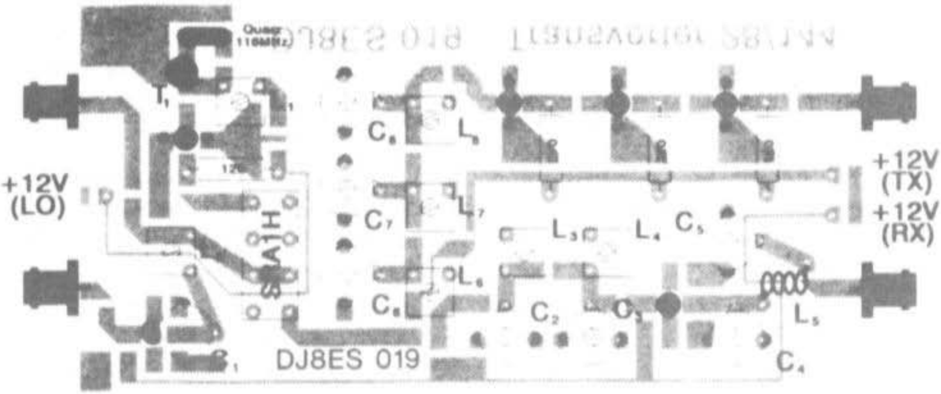

Fig.3: Transverter seen from the components side

amplifiers. In this case, additional harmonic filtration can be recommended. Suitable suggestions can be found in the relevant amateur radio literature.

2. ASSEMBLY INSTRUCTIONS

The 28/144 MHz transverter is assembled on an epoxy board, coated on both sides, with dimensions of 54 mm. x 108 mm. A board of this size can be incorporated into a standard tinplate housing (55.5 mm. x 111 mm. x 30 mm.).

After being cut to size, the board is initially cold-silvered and then drilled. However, the silvering is not absolutely necessary. Suitable holes should be drilled for the stripline transistors and the broad-band amplifiers. These components are flush with the board level.

The through-plating needed on the board for the coils and the ring mixer is provided by copper rivets (diameter 1.5 mm.).

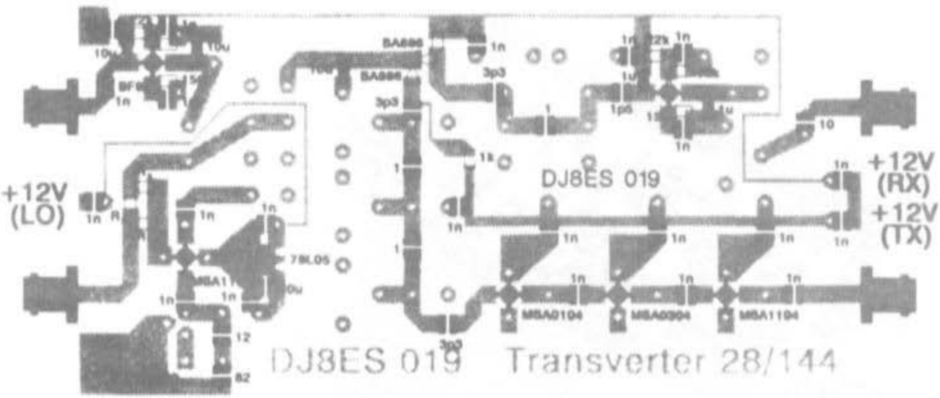

Fig.4: Track side with semiconductors and SMD components

After the drilling, the holes for the crystal, the trimmer, the Neosid coils, etc. can be reamed out on the earth side of the board (fully coated side) using a 2.5mm. drill. N.B.: earth connections must be left untouched! Recesses measuring app. 1 x 8 mm. should be sawn on the board edges concerned for the BNC bushes.

Once this preliminary work has been disposed of, the board can be sprayed with solderable lacquer.

The BNC flanged bushes must touch the cover edge with their flanges. If the board is now inserted in such a way that the bush pins are supported (cut off projecting Teflon collars with a knife first), it must still be possible to put the top cover on satisfactorily after a test insertion of the filter coils and the crystal.

When the board has been soldered to the side surfaces of the housing, the components are actually inserted.

3. COMMISSIONING

The following measuring equipment should be available for the initial commissioning and subsequent balancing:

- → Multimeter
- → Frequency counter
- → Diode probe
- → Wattmeter and
- → 2m received signal

First the oscillator is set using the tuned amplifier circuit, L. The power consumption of this operational unit is 65 mA, of which 55 mA are needed just for the MSA1104 broad-band amplifier (IC).

Only the 3-circuit filter (C to C) should be balanced in the transmission branch. The approximate trimmer locations are shown in the wiring diagram. It should be possible to measure a current of 130 mA with an operational voltage of + 12 V. This is already an indication that the amplifier stages are functioning satisfactorily. If the input damping element is dimensioned as described in Table 1, then an output exceeding 50 mW can be expected. Possible harmonics (oscillator, image frequency, etc.) are attenuated by better than 55 dB.

The initial receiver balancing can be done directly using a strong received signal (e.g. 2m. beacon). Here the 2-circuit filter (C, C) should be carefully set for the best signal strength. A further filter is positioned on the intermediate frequency level (28 MHz) after the mixer. Here the trimmer, C, is balanced to the maximum signal. Naturally, the direct coupling of the parallel circuit influences the transmission branch. But this influence does not have any effect, as appropriate reserve capacity is available. The optimising of the signal-noise ratio (PI filter with C, C and L at the receiver input) concludes the balancing.

At only 20 mA, the current consumption for the receiving branch is very low. The noise factor is app. 2 dB and the transmission amplification is app. 20 dB.

4. FINAL COMMENTS

The author is making successful use of the transverter described, with a masthead pre-amplifier and a power amplifier.

Modern hybrid modules are almost available as amplifier stages. With such components, the output signal of 50 mW can be increased to 20 W in one go. I intend to describe an amplifier of this type in a later article. Here too, the minimal external wiring of the hybrid modules provides for copying without any problems.

The results achieved with the 28/144 MHz transverter demonstrate once again that outstanding results can be obtained even with homemade equipment. Units like these can be assembled and successfully operated even without a wide range of measuring equipment.

5. PARTS LIST

IC_1	TA78L09F (SMD) regulator
IC_2, IC_5	MSA1104 (Avantek)
IC_3	MSA0104 (Avantek)
IC_4	MSA0304 (Avantek)
T_1	U310 (Siliconix)
T_2, T_3	BF981 (Siemens)
D_1, D_2	BA886 PIN diode (SMD)
L_1, L_2, L_3	BV5061 Neosid coil
L_4, L_5, L_6	0.1 µH, blue/brown
L_7	BV5048 Neosid coil, 1 µH, yellow/grey
L_8	4.5 winding, 1 mm. CuAg wire
C_1	30 pF foil trimmer (red), 7.5mm. grid (Valvo)
C_2 - C	12 pF foil trimmer (yellow), 7.5mm. grid (Valvo)
Q	HC18U or HC25U 116 MHz crystal
1 x	SRA1H high-level ring mixer
2 x	120Ω, 0.5 W carbon film
1 x	220Ω, 0.5 W carbon film
1 x	270Ω, 0.5 W carbon film
4 x	BNC flanged bush (UG-290 A/U)
3 x	Teflon bushes
1 x	tinplate housing: 55.5 x 111 x 30 mm.
9 x	copper rivets (1.5 mm dia.)

All other components in SMD format:

2 x	1 µH choke
3 x	10 µH choke
1 x	10 µF/20 V tantalum

Ceramic capacitors	Resistors
3 x 1 pF	1 x 150Ω
1 x 1.5 pF	2 x 220Ω
1 x 2.2 pF	2 x 1 kΩ
4 x 3.3 pF	2 x 10 kΩ
1 x 10 pF	2 x 22 kΩ
1 x 12 pF	1 x 82 pF
17 x 1 nF	

6. LITERATURE

(1) Rolf Albert, DK 8 DD: Compact 2m. transverter with low-noise preliminary stage and clean transmission signal; VHF Communications, 3/81

(2) Wilhelm Schürings, DK 4 TJ and Wolfgang Schneider, DJ 8 ES: Universal transverter concept for 28, 50 and 144 MHz; VHF Communications, 3/91

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An Antenna for all Meteors?

Meteor burst signals are reflected by meteor trails (ionisation columns) that are continuously formed in the atmosphere at about 100 km altitude. Because MB is an inherently weak-signal mode, similar to EME, MB links require good antennas. The conventional wisdom is that a good antenna for long links is one or more Yagis several wavelengths above the ground. This technical note describes an alternative antenna that should work well for any distance between the transmitter (TX) and receiver (RX). It should also be useful for EME work.

1. THE MONOPOLE ARRAY

Maximum gain at low take-off-angles is crucial for good link performance at long ranges. A typical antenna system might consist of two vertically oriented 3 to 3.5-wavelength Yagis stacked horizontally (side-by-side) at a height of 4 wavelengths above ground. Such an elaborate system is required to offset the "ground tuck" that occurs in the radiation pattern of any antenna over real earth. Ground tuck is a sharp drop in gain at shallow take-off angles (typically 10-30 dB). All antennas exhibit this effect, but it is less pronounced for vertical radiators.

Near-horizon gain can be maximised by using a vertically polarised antenna on a ground screen. If the screen, which acts as a reflector, is long enough in the direction of maximum radiation, it can all but eliminate ground tuck. This observation suggests the possibility of a single antenna designed for maximum near-horizon gain that is also suitable for intermediate MB ranges.

Fig.1 is a schematic representation (not to scale) of the type of antenna being suggested - a passive array of vertical

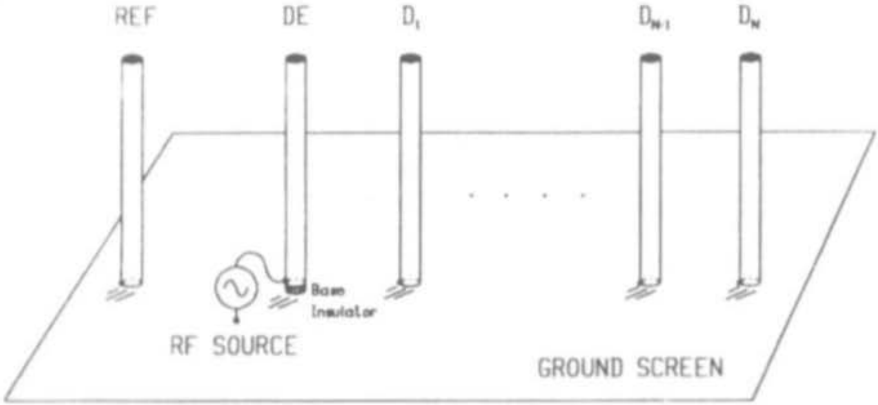


Fig.1: Monopole Array on Ground Screen

monopoles. The monopole array is essentially half a Yagi mounted on the ground screen. It consists of a single reflector (REF), a single driven element (DE), and N directors (D_i). All elements except DE are connected to the screen. DE is mounted on a base insulator and excited by the RF source (TX) which is connected between the base of the element and the screen. In practice, the feed is a coaxial cable witn its shield is connected to the screen and the centre conductor to an impedance matching device such as a transmission-line transformer (Unun) or a gamma match.

Fig.2 shows the radiation pattern for a well-designed 18-element monopole array on an infinite ground plane. Maximum gain for this design is 18.6 dBi, which is consistent with MB at the longest ranges. On a finite plane, the gain falls off very close to the horizon, but the ground tuck is substantially reduced by the presence of the ground screen. Two opposing effects occur as the take-off angle increases: (1) the antenna gain drops rapidly because of the highly directive main lobe, and (2) the MB propagation loss decreases rapidly because the TX-RX range is reduced. Lower antenna gain reduces available RX power, while lower path loss increases it.

Comparing MB link performance with varying TX-RX range to the performance at maximum range provides a simple measure of how well a particular antenna works. If the link performance is acceptable at the maximum range, then all the antenna has to do at any other range is meet or exceed that performance to provide a solid MB link at the shorter TX-RX range. Fig.3 is a typical comparison. It plots the ratio of available RX power vs. range to the RX power at maximum range (zero take-off angle). This curve applies to mid-point reflections along the TX-RX great circle path (GCP), and it assumes a spherical earth of radius 6371 km with a 100 km trail height.

Fig.3 shows that the 18-element monopole array on a ground screen, which is designed to provide maximum gain near the horizon for the best possible maximum-range performance, also provides a positive system margin at almost all intermediate ranges. Near 600 km the margin is about 13 dB! The effects of the 3 secondary lobes in the pattern are clearly evident between approximately 70 and 350 km range, where the margin varies considerably but is always positive. Only at the shortest ranges does it fall below zero.

Even though the monopole array is designed with the single objective of maximising near-horizon gain, it works well at all TX-RX ranges except the very shortest. This behaviour is explained heuristically as follows. For mid-path MB reflections, the received power varies as the cube of the distance from TX to the meteor trail. As the range decreases (increasing take-off angle), the RX power increases very quickly because of this cubic behaviour. By comparison, the antenna gain, which falls off as the range decreases, drops much more slowly than a cubic variation. The decrease in antenna gain at shorter link ranges is more than offset by the increase in received power on the shorter propagation path.

Of course, MB reflections do not occur only at mid-path. Trails on either side of the GCP, trails at different heights, and even trails behind the TX or RX can support a link. Nevertheless, the behaviour for GCP mid-point reflections appears to illustrate a generally valid effect: when highly directive antennas are used to maximise near-horizon gain, the lower MB path loss at shorter ranges offsets reduced antenna gain and results in positive MB link margins (more RX power compared to maximum range) at all ranges except possibly the very shortest. The monopole array on a ground screen is a typical antenna system. Other configurations should exhibit similar behaviour.

On the practical side, the monopole array on a ground screen is mechanically simpler, almost certainly less expensive, and easier to install and maintain than Yagis on a tall tower. In addition, it eliminates unnecessary, but potentially substantial, feed system losses associated with long cable runs, power dividers, and

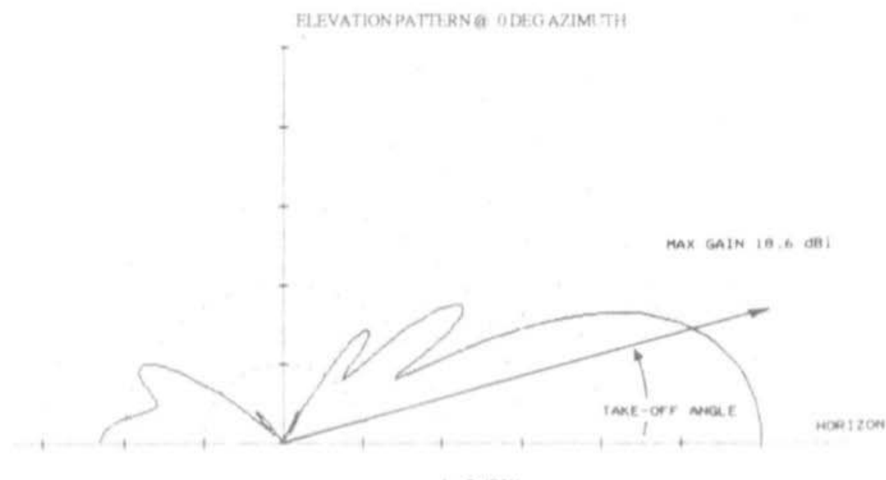
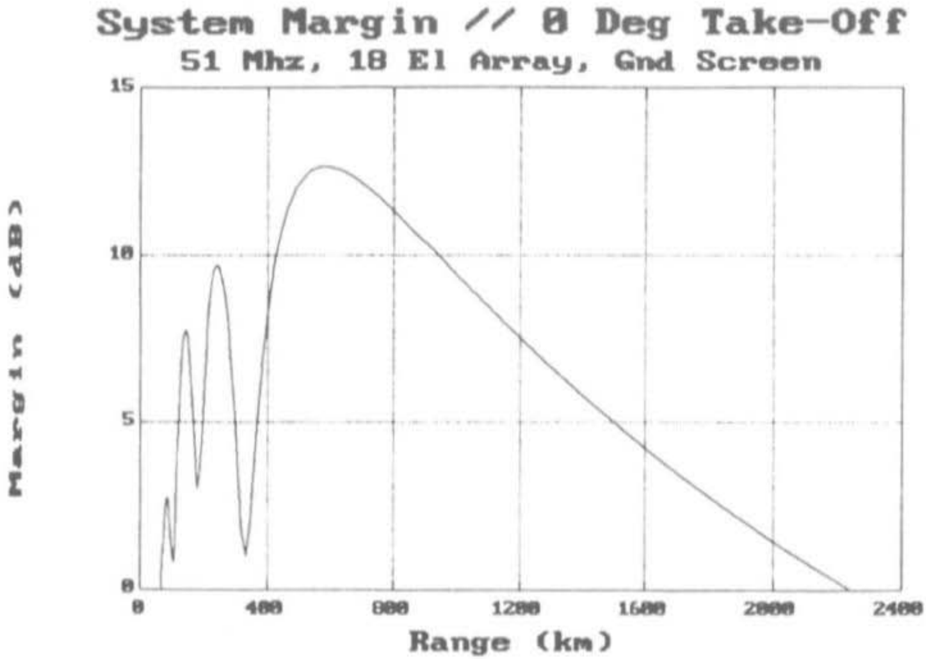


Fig.2: Radiation Pattern for 18-element Monopole Array



coax connectors. Of course, the monopole array suffers the disadvantages of not being rotatable and requiring a large ground screen.

Unfortunately, there is no simple way to specify just how big the screen should be. In theory, it should extend hundreds of wavelengths or more in the direction of the main lobe to essentially eliminate all ground tuck. But, in practice, a few wavelengths should work quite well. The screen should extend several wavelengths beyond the last director (the more the better), and as far as possible to the side and behind the array. Because FB for a long "half-Yagi" is high (in the example, about 18.5 dB), the screen size behind the reflector is probably the least critical.

As an indication of what should be achievable, computer simulations of MB performance at maximum range using a screen extending 10-wavelengths beyond the last director show a 6-fold improvement in average throughput, with nearly a factor of 10 improvement at certain times of day. The comparison antenna was essentially the stacked array system described above.

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Observation of Scintillations while Receiving Meteosat

"Scintillations are rapid oscillations in the field strength and/or direction of incidence around an average value, brought about by non-homogenities in break values (ITU Conference, Geneva, 1982, Rep. 881)." Quotation from (2).

1. INTRODUCTION

We find more on this topic in (2): "Scintillations due to non-homogenities in the electron density in the ionosphere are strongly dependent on solar activity" (AGARD Conf. Proc. 284, 1980).

"On propagation paths which penetrate the ionosphere in the vicinity of the magnetic equator, oscillations of 10 dB at 4 and 6 MHz were exceeded in 0.1% of a year at the sunspot maximum, whilst at the sunspot minimum they were never above 1 dB (AGARD Conf. Proc. 332, 1982). These scintillations occur regularly in the evening hours (18.00 - 3.00 OZ) during the Spring and Autumn equinoxes".

In (1) I described signal oscillations, which can also be referred to as scintillations - points marked S in Fig. 6 of (1). All in all, the word means nothing more than oscillation in intensity and was predominantly used to describe the twinkling or glimmering of stars before it was adopted by researchers into wave propagation. My call in (1) for reports of similar observations has gone unheard.

2. DESCRIPTION OF THE PHENOMENON

Some days after the recordings for (1) were completed, scintillations could be recorded for over three hours on the Meteosat frequency. Since then, the phenomenon

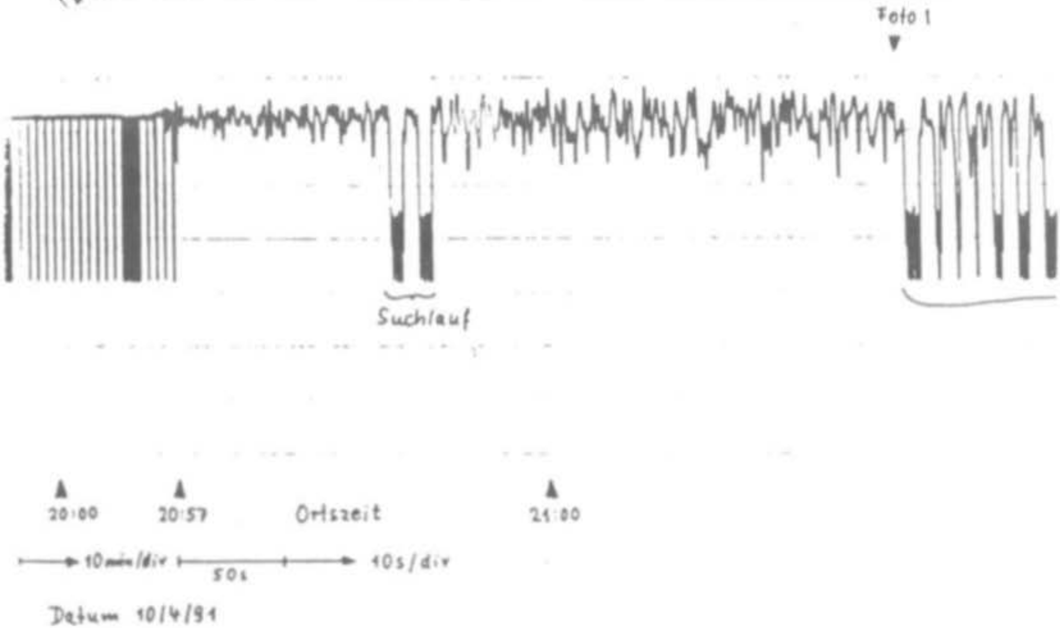


Fig.1: Recording of Scintillations at 21.38 on 04/10/92;

has been observed several dozen times on several evenings. Fig. 1 was obtained in exactly the same way as Fig. 6 in (1), but between 20:57 and 21:25 OZ the time-scale was switched to 10 s./div., so that the resolution was 60 times better than for the long-term recording.

The highest recorded increased input power is - 83 dBm, 6 dB above the normal level. The lowest recorded input power is - 97 dBm, 8 dB below the normal level. The FM threshold for my reception equipment was then undershot by 3 to 4 dB, which led to a break in vision. Less pronounced interruptions caused grainy spots to appear on the image.

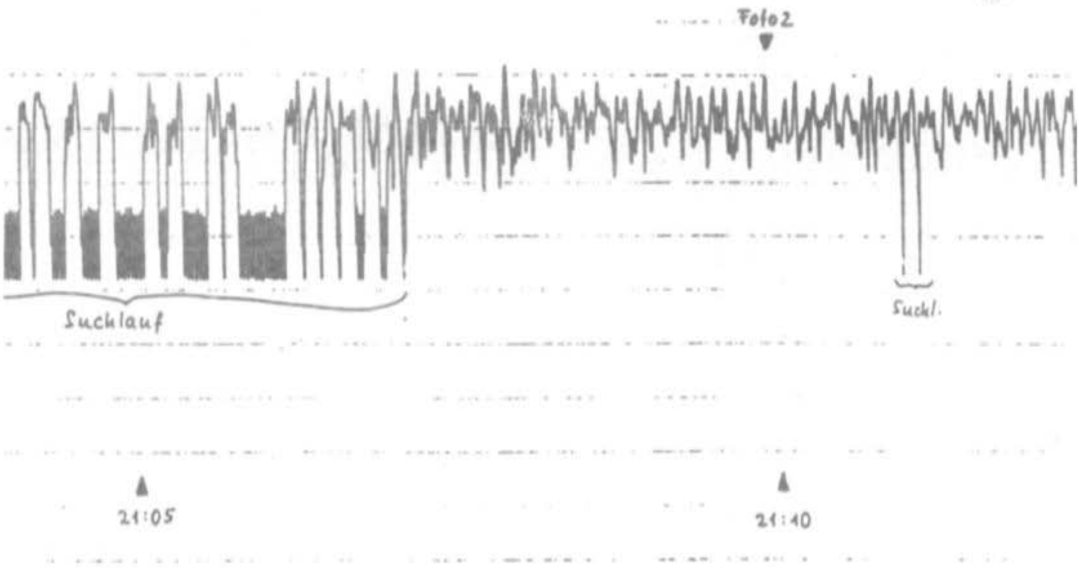
The frequency of the scintillations lies between 0.05 and 0.5 Hz. Undershooting of the FM threshold occurs every 5 to 50 s. The impairment of the image quality ranges from serious to insignificant. Fig.2, the quality of which is unsatisfactory due to the readings having been plotted by hand and to completely artificial lighting, can nevertheless convey an impression.

In day-to-day continuous operation, scintillations attract attention only if they repeatedly lead to the undershooting of the FM threshold during an image. But even then they are not infrequent and are not confined to the equinoxes.

Anyway, in the Tropics the days and nights are always more or less the same length as one another, so that the expression "equinoxes", which comes from the developed world, is really misleading. Nonetheless, I shall continue to use it.

Observations over more than a year can be summarised as follows:

→ Sporadic scintillations occur all through the year



Suchlauf/Suchl. = Station finding; Ortszeit = Local time; Datum = Date

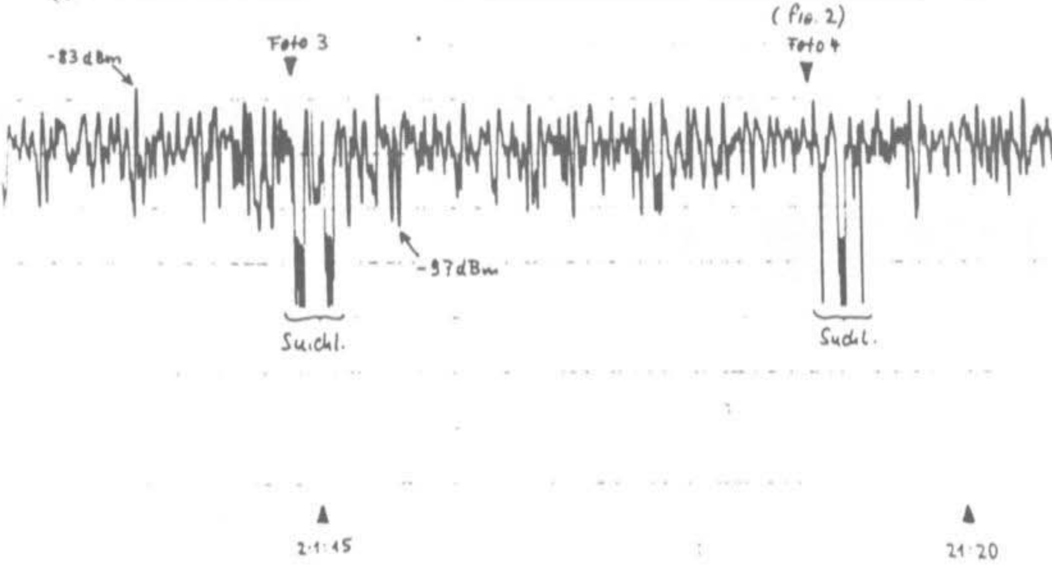
- → They are most frequent during the Spring and Autumn equinoxes
- → Even during the equinoxes, there are periods without interfering scintillations.

Scintillations were observed only between 18:00 and 24:00 OZ. No connection with the local weather could be established

3. AN ATTEMPT AT AN EXPLANATION

Fig. 3 provides an impression, to scale, of the path of the energy from Meteosat to my receiver aerial. You just have to imagine you're looking at the North Pole of our planet from rather a long way away. The Pole Star would make a suitable standpoint (Alpha Ursa Minor). In this representation, Nairobi is near enough on the horizon. Seen from there, Meteosat is at an angle of approximately 45 degrees. The point on the Earth's surface above which the energy received by me enters the ionosphere lies as far to the West as the ionosphere is thick

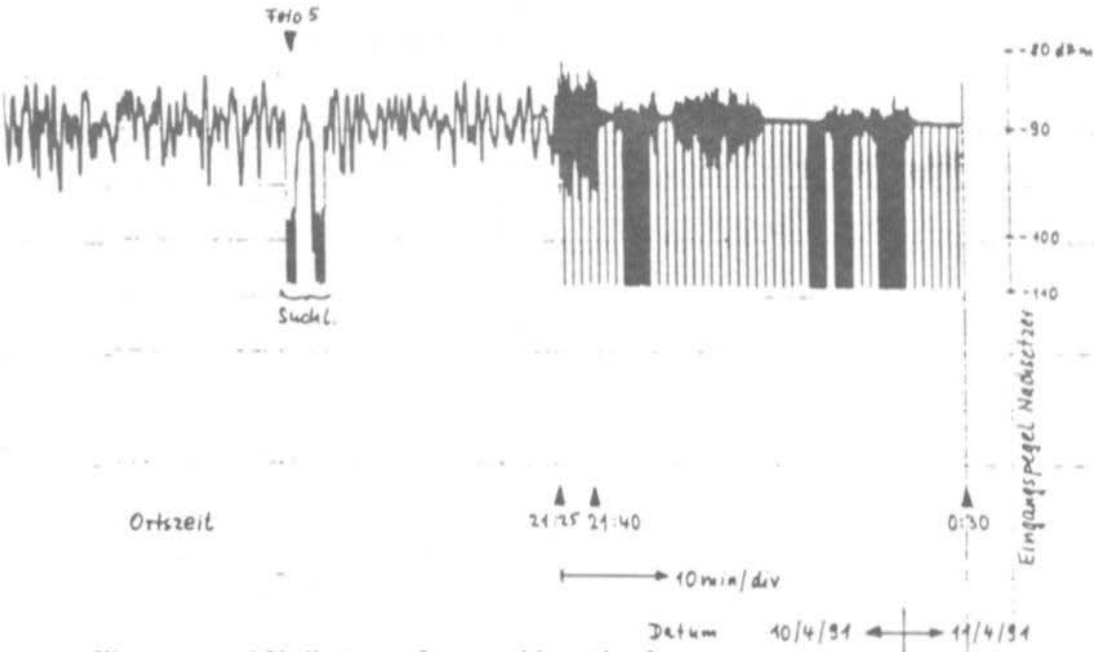
Fig.2: (Poor) weather image during scintillations



(1,000 km. = 1 Mm.). This thickness is still quite clearly displayed in Fig. 3, but not that of the troposphere (10 km.) or the average cloud height (3 km.). The weather behaviour takes place in the troposphere, which the energy of reception enters only 10 km. West of the reception location. That location is Dagoretti, which was already in existence when the explorer J.W.Gregory discovered the Great Rift Valley in 1897. Nairobi (i.e. "cold water") was then nothing but a rather large stream. Anyway, I can follow the changes in the weather up to Dagoretti with the naked eye. The scintillations observed were independent of whether the sky was clear or contained cumulus clouds or even threatened a severe storm.

The weather pattern over the next 1,000 km. towards the West is shaped by the mountains at the edges of the rift, the big lake in the Rift Valley and the meeting of air masses from the Congo Basin with those from the Indian Ocean. On the shores of Lake Victoria, there is an area with very intense storm activity. On the floor of the Rift Valley (Ol Donyo Lengai) and on the edges (Virunga) there are several active volcanoes. Nevertheless, it should not be assumed that these "interference sources" can have an effect several hundred kilometres up into the ionosphere.

So finally, as put forward in (2), only solar activity is left as a source of parasitic induction. There was a sunspot activity peak in 1990 and the next trough is expected in 1996. High activity causes increased ion concentration in the ionosphere. There is also a daily cycle. At sundown, the concentration near the ground first diminishes and then, once it gets dark there too, actually increases.



Eingangspegel Nadisetzer = Increased input level

There is thus a concentration gradient, which compels a tangentially incident wave to travel in a curve. It thus approaches the aerial from the wrong direction. However, I am quite sure that I aligned my aerial optically correctly.

This type of thinking might certainly explain a reduction in the received power, but not a 6 dB increase!

A concentration effect must therefore be effective, at least in terms of time. This can be explained if we assume that streaks are being formed. Optical streaks can be observed, for example, in a glass of water, in which a lump of sugar is dissolving. A movement occurs automatically due to the difference in density between pure water and sugared water. The projection of the streaks in transmitted light shows lighter and darker strips alternating, in comparison to the uniformly mixed solution. The movement of ions in the ionosphere could be connected with the strong horizontal components of the earth's magnetic field near the Equator. Streaks could be formed under conditions which would in any case be favourable to this.

I leave it open whether such a model makes this phenomenon clearer! Moreover, I understand too little of the processes in the ionosphere. Unfortunately, there isn't another Meteosat receiver station in this country which could confirm my observations. True, there is another station belonging to the Government Meteorological Service, which picks up the D6 WEFAX image at 15:24 OZ for the evening television news, but because it only operates until 17:00 it observes nothing which happens at night.

From stations in Europe it is known that

Fig.3: Sketch of geometric relationship between Ground (Erde) , Nairobi and Meteosat

Meteosat reception is often interfered with by radar equipment. I think that can be ruled out in my case. Firstly, the long-range radar equipment at the international airport (15 km. away) has been out of action for many years, and secondly, such interference would have looked different in Fig.1.

4. CONSEQUENCES

The provision of food supplies in Africa, which is obviously crucial, is essentially dependent on improvements in weather forecasting. The conventional weather service supplies less reliable data today than 30 years ago, because many stations have ceased their activities (e.g. 49 out of 50 in Zaire!). So there have been continuous proposals from various aid organisations for the setting up of WEFAX receiving stations in the main farming areas. To ensure that such stations could operate smoothly even during sunspot activity peaks and after sunset (worst case design), their aerials would have to be over-dimensioned.

So for 8 dB fade reserve, at the same distance from the FM threshold, the reception aerial would have to have an aperture area of 6.4 m^2!

Perhaps, to identify the reason why this difference from the conditions in the temperate climatic zone exists, it might be quite a good idea if I operated my Meteosat receiver station (1) during a sunspot peak.

5. LITERATURE

(1) D.Burchard (1991): Cylindrical parabolic aerial with Meteosat compact converter; VHF Communications 4/91

(2) Meinke & Gundlach (1986): High Frequency Engineering Pocket Book; S.H 33, Springer-Verlag, Berlin

Guenter Sattler, DJ 4 LB

Addenda and Comments on the Article: A 10 GHz FM-ATV Transmitter with Dielectric Resonator (Issue 2/1992)

1. DISCUSSION ON THE CONCEPT

When TV amateurs experimenting at 10 GHz talk about the "pill", they mean the dielectric resonator - the DR.

Every satellite LNC (Low Noise Converter) contains one or more "pills", which stabilise the frequencies of the local oscillators at 10 GHz. Thus unmodulated test signals are already available in the 10 GHz amateur band.

Many TV amateurs conjectured that it must "somehow" be possible to modulate the frequency of such a DRO (oscillator with DR), so that it could also be used as a 10 GHz ATV transmitter. Denys Roussel, F6IWF, has now made this a reality and described how to assemble the equipment (VHF Communications 2/1992). 10 GHz broad band technology is here shown a path of its own, specially for FM ATV. The indisputable advantage of this concept is that a harmonics-free signal can be generated directly at 10 GHz in a single transistor stage, without any expenditure on filters.

The frequency stability which can be achieved is certainly poorer, by several orders of magnitude, than the "crystal stability" known through the 10 GHz SSB technique. But it should be taken into consideration here that the intermediate frequency band widths of the satellite receivers used for FM-ATV (approximately 27 MHz) are almost 10,000 times as big as those of SSB receivers.

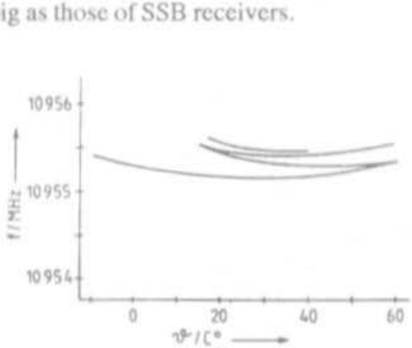


Fig.1: Temperature Graph of a DRO

The frequency changes from 100 to 250 MHz which are described in Section 3.2., and which occur when the cover with the tuning screw is put on, should not be understood as an indication of frequency instabilities in this order of magnitude. All DRO's in satellite LNC's behave in a similar way.

However, on the basis of a stable and well thought-out mechanical structure for the balanced and closed LNC's, the manufacturers can guarantee frequency deviations of less than 1 MHz in the temperature range of - 10 to + 60 degrees, as a corresponding measurement curve (Siemens) (Fig.1) indicates.

2. COMMERCIALLY AVAILABLE DR FREQUENCIES AT 10 GHz

Conventional satellite LNC's, which convert satellite programmes with either horizontal or vertical polarisation, usually have DRO's at 10,000 GHz.

There are now also LNC's being sold (Sat-LNB 4243, Conrad Electronic), which separate TV channels of both polarisation levels, which are interpenetrating as regards frequency, and mix them with various beat frequencies. The DRO frequencies used here are 9.70 GHz for the vertical components and 10.25 GHz for the horizontal ones.

DR's at 9.10, 10.00 and 10.50 GHz are now available from the specialists (e.g. Giga-Tech, Karl Himmler, DB 3 UU).

3. DR's MODIFIED FOR AMATEUR RADIO FREQUENCIES

The grinding process to increase the DR frequencies described in Section 3.2. should be carried out using wet grinding for safety reasons, since it can not be completely excluded that hard DR materials contain toxic constituents, which could be released during the grinding process.

As also explained in this section, ceramic discs are suitable for changing the DR frequencies to lower values. An optimum basis for this consists of round ceramic disc capacitors without connection wires, with capacitance values of between 1 and 6.8 pF. The metallisation on both sides is removed using a 600 wet emery paper, leaving no residue. To test the effect, place the individual discs on the DR pill one after another and determine the DRO frequency in each case.

The following list of measurements gives the reference values for changing the frequency by adding ceramic discs.

Original capacitance of disc capacitors	DRO frequency change
1.0 pF	- 70 MHz
1.5 pF	- 90 MHz
2.2 pF	- 140 MHz
3.3 pF	- 240 MHz
4.7 pF	- 300 MHz
6.8 pF	- 440 MHz

Should there still be enough space left free between the DR and the tuning screw, a second ceramic disc can be put on with (originally) identical capacitance values. This brings the frequency down again by 25 - 45% of the frequency change brought about by the first disc.

If the right ceramic disc has been found for the range of application in question, it is glued to the DR, using the minimum quantity of a two-component epoxy adhesive (e.g. UHU-plus endfest 300), to produce a component modified for amateur radio frequencies.

4. DRO FREQUENCY MODULATION

An additional AM component can not be avoided in the DRO frequency modulation described in Section 1, using the displacement of the operating point of the oscillator MESFET.

Circuits are known here from the professional manufacturers (e.g. 3.5 GHz DRO, Rhode & Schwarz) which, with the help of an additional strip line, make the electronic fine tuning of the DRO frequency possible. The electrical length of the strip line here is changed by means of voltage-controlled varactor diodes.

The conversion of 11 GHz LNC's to the 10 GHz amateur band yields further information on frequency changes in a DR caused by the influence of the strip lines connected.

If the 10-GHz DR built in is replaced by a 9.1 GHz DR, also commercially available, then the latter oscillates at about 9.4 GHz, due to the gate and drain strip lines dimensioned for 10 GHz. (By gluing a ceramic disc on which, as described above, is made from a 6.8 pF disc capacitor, you can pull the DRO down to the 9.0 GHz desired.)

A further DRO development for ATV applications could use the electrical lengths of strip circuits, which can be altered through varactors (or alternatively FET's) for frequency modulation or for AFC or PLL operation.

5. SIMULTANEOUS TRANSFER OF IMAGE AND SOUND

The additional components proposed in Section 6 for "simultaneous transfer of image and sound" represent considerable expense in practise.

Even if high-level audio signals are now available, for example from cameras or camcorders, these are suitable for modulation only to a restricted degree without further preparation.

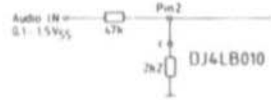
Widely varying breaks in discussions can occur in ATV operation, due to activities in front of or behind the camera or on the directional aerial. It has been shown in practise that the modulation index required can be maintained only by using an automatically controlled amplifier with a control range of app. 40 dB.

Moreover, at least one audio pre-emphasis, a sound sub-carrier oscillator, tuneable between app. 5.5 and 7.5 MHz, and a link between the sound sub-carrier and the video basic band signal are required for TV sound transmission.

5.1 Use of DJ4LB010 image-sound processing

All these circuits are included in the DJ4LB010 image-sound processing assembly (VHF Reports 1/1990). Since we are dealing with general image-sound processing for FM ATV transmitters, it is obviously also suitable for the 10 GHz ATV transmitter from F6IWF. The basic band signal can be fed from the DJ4LB010 assembly through a coax cable of any length to the "mod-input" socket of the DRO module. The R1 resistance could be changed from 100 to 75 Ohms in here for exact cable matching, or for level setting, using a 100-Ohm potentiometer with a 330-Ohm parallel resistance.

If a high-level audio signal from a modern video-audio unit is available for sound modulation, it is convenient in this case to wire up the DJ4LB010 microphone input as per the diagram below.



Joachim Danz, DL5UL

Assembly Instructions and Experiences with the DB1NV Spectrum Analyser Design

The spectrum analyser as per DB1NV is a relatively complex apparatus, and many problems could crop up for those who copy it.

Starting with a short systems comparison between two up-to-date publications, I'd like to pass on my experience, with some tips on copying and balancing the individual assemblies. The problems of harmonic mixing in ring mixers are then briefly discussed.

1. GENERAL COMMENTS ON SPECTRUM ANALYSERS IN THE AMATEUR LITERATURE

Before starting to copy a spectrum analyser, you have to have a clear idea of what requirements the spectrum analyser should fulfil and what the technical specifications are. At the moment, there are essentially two descriptions which it is safe to copy:

- → A Spectrum Analyser for Amateurs Dr. Ing. Jochen Jirmann, DB1NV(Published in Issues 3 & 4/1987, 2 & 3/1989, 1 & 3/1990, 3 & 4/1991, 1/1992 and 1/1993 of VHF Communications)
- → Spectrum Analyser 5...1,500 MHz Wolf-Henning Rech, DF9IC(Published in Proceedings of 13th and 14th GHz Congresses in Dorsten)

2. SYSTEMS COMPARISON

Since the purposes of the two designs are not identical, no direct comparison can be carried out here. Both designs have advantages and disadvantages. Nevertheless, I'd like briefly to point out the most essential differences.

2.1. Operating principle

Some of the properties have already been established through the choice of design.

DB1NV: Superheterodyne receiver with two intermediate frequencies. Level determination at second intermediate frequency at 10.7 MHz (dynamic: up to 80 dB)

DF9IC: Direct mixer or superheterodyne receiver, depending on frequency range. Level determination in base band (dynamic: up to 65 dB)

2.2. Frequency range

The analysis range is an important factor in selection, since all ranges in which you might wish to be active should be included. Naturally, the frequency range can be extended for both designs by means of external mixers and oscillators.

DB1NV: 0 - 500 MHz and the 1,000 - 1,500 MHz image frequency (the image frequency should be handled with great care) - in the meantime also continuous up to 1,900 MHz.

DF9IC: 5 - 1,500 MHz in two ranges without gaps (in the meantime expandable up to 4,500 MHz by means of an external mixer).

2.3. Filter band widths

The band width of the filters is determined by the spectrum lines on the screen. Broad filter band widths simultaneously mean that spectrum lines which lie close to one another can no longer be individually resolved. Narrow filters require a certain response time, so that the image display speed must be reduced for narrow filters.

DB1NV: Extensive band width control using filter crystals from 1.5 kHz to 50 kHz, which also makes it possible to measure, for example, transmissions on adjacent channels

DF9IC: Direct mixer with a band width of 4 MHz or 1.5 MHz (determined by low-pass)

2.4. Range of application

DB1NV: This spectrum analyser represents a very expensive apparatus, which, up to a frequency of 465 MHz, does simply everything a radio amateur could wish (even a tracking oscillator has been described!). However, in the image frequency range you need to know exactly what you're measuring since, due to harmonic mixing, a great many additional lines occur which can be attributed only with difficulty. The only thing which will really help here is a pre-filter running at the same time which suppresses these undesirable lines (the harmonics originate in the ring mixer!). Or else you can obtain some help from a simple pocket calculator, with a lot of patience (see also Section 4). But this effect can also be made use of: depending on the ring mixer, spectrum lines up to, for example, 4 GHz can be identified. However, only very limited findings can be deduced concerning the level of these lines.

DF9IC: This apparatus is designed as an aid to balancing up to 1.5 GHz. If you only want to know which frequency fractions are generated at which levels, this equipment can be warmly recommended. However, you can not obtain detailed information about the signals. And for the assembly of, for example, oscillators or transverters, this isn't so important, for here other test options are usually available such as beacons, QSO partners or the like. With this design, problems caused by harmonic mixing can effectively be avoided, i.e. the spectrum lines you see are actually there. Anyone wishing to build radio equipment which goes beyond 70 cm. is well advised to choose this design.

2.5. Comments

What I have tried to do here is to give a very brief idea of what I think are the most essential differences. It is in any case worth while to study the documentation more closely and weigh the characteristics.

Both pieces of apparatus can be assembled without additional complicated gauges. You just need a frequency counter, a high-frequency source with harmonics and a watt meter. The rest of the balancing procedure is carried out using the equipment itself.

In my opinion, it would also be possible to alter/expand the analyser as per DB1NV: 1. LO of 1 - 2 GHz; first intermediate frequency at 1 GHz; then as in original: second intermediate frequency at 465 MHz, third intermediate frequency at 10.7 MHz.

The range for which unambiguous measurements can be obtained can thus be expanded to 5 MHz to 1 GHz. Putting this into practise, of course, stands or falls by the availability of suitable ring mixers and oscillators.

3. ASSEMBLY INSTRUCTIONS FOR INDIVIDUAL UNITS

I am restricting myself to the assembly of the original version of the spectrum analyser, as described in Issues 2 and 3/1989 of VHF Communications. I have assembled, but have not yet integrated, an additional oscillator of 1 - 1.7 GHz, since an interesting suggestion has been published by DB1NV in Issue 1, 1992 in the meantime. I have so far not been able to conclude additional experiments using travelling pre-filters to my own satisfaction, although it is undoubtedly worth carrying on with this work. For the power supply, I used a cascaded stabilisation with a 78xx range IC and the well-known LM 723, which turned out well. The characteristics of the power pack from Volker Esper, DF9PL (1) are naturally not matched by this circuit.

The various units can be assembled in the following order. The run-off control is positioned first, followed by the VCO/PLL unit, next the high-frequency/intermediate unit and finally the crystal filter unit. It makes no sense to carry on assembling modules if the sections already finished are not fully functional. If you stick to this order of operations, stage-by-stage balancing, using a frequency counter and a watt meter, poses no problem.

3.1. DB1NV 006 high-frequency/ intermediate frequency unit (VHF Communications 2/1989)

If the external pre-selector is not firmly integrated into the entire concept, it is advantageous to include a high-quality ceramic capacitor in the radio-frequency input of the unit. Without this capacitor, it could happen that the ring mixer input transformer is destroyed by DC at the input.

The ring mixer, I1, should operate only to the extent that it can actually be measured. Using a broader-band type here only causes more lines on the screen, which are caused by harmonic mixing (see also Section 4). To reduce undesirable mixed products, the ring mixer, I1, should be cut off in the broad band at 50Ω. It is possible to provide for 3 - 6 dB damping elements at all gates. A broad-band intermediate amplifier with good input matching (e.g. MMIC's from the MSA or MAR ranges) at pins-3 and 4 is the most favourable solution here.

To improve the decoupling of the first and second intermediate frequencies, a partition can be soldered onto both sides of the printed circuit board, passing transversely through the housing, which must be tightly sealed to the covers. Problems with an additional amplifier between the first ring mixer, I1, and the helical filter, Fi1, are eliminated by these measures.

In the area of the ring mixer, I1, and the two filters, Fi1 and Fi2, highly conductive metal oxide semi-conductor foam should be inserted into the tinplate housing on the components side as well, in order to reduce the coupling between the individual stages further.

On the second ring mixer matching is uncritical, since undesirable mixed products can not fall into the second intermediate frequency. However, you must ensure that the 4.7 nF capacitor at the ring mixer output is of a type which acts as a choke above app. 400 MHz. If ceramic types are used, the 47Ω and 10Ω resistances are to be wired up in parallel, which causes a loss of sensitivity.

It is advisable to mount the oscillator in an external housing and connect it through an additional plug and socket connection. This makes it easier to carry out comparisons or to replace the unit by a less noisy type of assembly. A separate power supply also de-couples the oscillator. I structured the self-inductive coil to be self-contained, which makes it slightly more temperature-stable and improves its quality.

Be careful to install L7 correctly, as otherwise the consequence could be less sensitivity up to 18 dB.

The sensitivity of the TDA 1576 can be altered by app. 40 dB by a DC level of between 0 V and 3 V at pin-14, with the open pin, as per data sheet, being the most favourable solution.

The units can be balanced in the following way. The ring mixer, I2, is not installed until the second LO is functioning smoothly. First the oscillator output at pin-2 of I2 is measured using a watt meter. Here the ring mixer needs a level greater than 7 dBm (5 mW). As soon as the tuning range and the output of the oscillator match, the ring mixer can be soldered on.

The IF out and IF in sockets are then connected. The input gain is set at + 15 Volts. The filters, Fi1 and Fi2, are each bridged over by means of a 1 nF capacitor. If the first LO is now wobbled from app. 440 MHz to 480 MHz, the image of the 10.7 MHz filter curve can now be seen twice on the oscilloscope, with an intervening gap of 21.4 MHz. The TDA 1576 generates app. 0.4 V voltage per 10 dB of level. With this signal, coils L1 and L3 to L7 are balanced at a maximum video signal (the interval for the two maxima must be 21.4 MHz). Then the two 1 nF C's are replaced by lower values, until the two maxima can only just be seen. The screws of the two filters, Fi1 and Fi2, are turned equally from above until the image of the filter curve at the higher frequency is at its maximum. If we now remove the two parallel capacitors, the weaker filter curve disappears. Directly behind the ring mixer, I2, the output is de-coupled and displayed on the oscilloscope. If the balancing is carried out on a roof which is as flat as possible, this can be checked by manually wobbling the second LO at the video output.

3.2. DB1NV 007 VCO/PLL unit (VHF Communications 2/89)

In the meantime, DB1NV has described an oscillator with a frequency range of 450 - 1,450 MHz in (2), through which an analysis band width of up to 1.9 GHz can be obtained.

It is worth experimenting with the resistance values and the coils of the original VCO. With my oscillator, I was able to obtain a reliable oscillation of 430 - 1,065 MHz. It is important for the ring mixer that the output power over the entire frequency range is greater than 5 dBm. Slight changes in the tapping point on the self-inductive coil, together with variations in the capacities of the amplifier stage, have a significant influence here. With my set-up, the tuning voltage does not coincide with the voltage referred to by DB1NV, but this can easily be compensated for by suitable alteration of the resistances at the tuning input. A capacitor of a few nanofarads can be wired up in parallel to reduce the diode noise from D3.

Should the oscillator frequency be unstable, this may be due to the fact that the 5-Volt power supply is not sufficiently constant (particularly critical at I6 and for the 390-Ohm resistance connected to pin 13). This can be remedied only by wiring up blocking capacitors (e.g. 2.2 F electrolytic capacitor) to the power supply connections of all TTL IC's. The VCO can be powered by a separate power pack for better de-coupling from the rest of the analyser.

As mentioned by DF7ZW, I6 absolutely must be from Motorola. Other manufacturers' products are considerably poorer here. For an even better linearity, an NPN transistor in a common emitter circuit, which functions as a low-impedance switch, can be introduced between pin 13 of the 74LS221 and the 100 H coil, instead of the two resistances.

3.3. DB1NV 008 crystal filter assembly (VHF Communications 3/89)

Four cheap crystal units can also be used, instead of the four identical filter crystals. True, these produce spurious resonances, but they disappear relatively quickly as the band width is reduced.

According to DB4DY, the four 3.9 H chokes can be replaced by suitable Neosid coil kits. They can be balanced with considerably greater sensitivity than the foil trimmers.

Replacing the pin-diode commutation by relays brings no advantage here, since it produces only a slight increase in the overall amplification.

The crystal filter assembly can be balanced directly in the analyser, even without external oscillators and detectors. To this end, the first oscillator is set to the first intermediate frequency, and the signal thus obtained without mixing is used for balancing. If the second oscillator is wobbled, an exact image of the entire filter curve is obtained. This line (zero response) corresponds to the zero frequency or to double the first intermediate frequency at the input, and should also be used later in operation for test purposes.

3.4. Wiring/inter-connection of units

An infinitely variable Preh damping element (50Ω !) can also be introduced as an input damping element, and can, in some circumstances, be obtained at a very favourable price in the flea markets.

For an optimal setting range for the P5 potentiometer, it is advisable to measure the required voltages first and then calculate values for fixed resistors to limit the matching range.

In my case, the tuning range of the first oscillator is greater than 500 MHz. It therefore makes sense to leave S4 out and, in contrast, to expand switch S5 by one position, 50 MHz/div.. There is then no difficulty in locating the lines when switching to another resolution.

To make it easier for even an inexperienced operator to use the spectrum analyser, the sweep speed and the band width of the crystal filter can be firmly preset, depending on the band width represented.

4. IDENTIFICATION OF UNEXPECTED LINES WHEN WORKING WITH THE SPECTRUM ANALYSER

Once all the modules had been assembled and balanced, I was naturally eager to see how, for example, the output signal from my 23cm transverter would look. When it was connected up, so many spectrum lines appeared on the screen that I almost threw the transverter away. Luckily, it was soon possible to confirm the suspicion that not all the spectrum lines stemmed from the transverter. They were more likely to be harmonics from the first ring mixer.

4.1. Frequency multiplication and mixing with diodes

Diodes possess a non-linear characteristic, which displays the following relationship:

$$I = Is * \left[\exp\left(\frac{e * U}{k * T}\right) - 1\right]$$

where:

I = Current through diode
U = Voltage over diode
T = Temperature
Is, e, k = Constants

The exponential function can be developed into a power series:

$$\exp(x) - 1 = \frac{x}{1!} + \frac{x^2}{2!} + \frac{x^3}{3!} + \ldots + \frac{x^n}{n!} + \ldots$$

If we now specify a harmonic voltage curve ($U_0 * \cos(\Omega * t)$ and convert the powers of the cosine function, we obtain:

$$\begin{aligned} I = Is * (C * \cos(\Omega * t) \\ + C^2 * ½ * [1 + \cos(2 * \Omega * t)]/2 + \\ C^3 * ¼ * [3 * \cos(\Omega * t) + \cos(3 * \Omega * t)]/3! + .. \end{aligned}$$

Thus signal fractions arise at multiples of the frequency of the stimulating voltage.

If we now feed in the sum of two harmonics, in addition to the harmonics we also obtain mixed products. Using the binomial formula for the second-order term, for example, we obtain the following mixed product:

$$\begin{aligned} I = Is * [\ldots + 2 * C1 * C2 * [½ * \cos((\Omega 1 + \Omega 2) * t) + \\ ½ * \cos((\Omega 1 - \Omega 2) * t)] / 2! + \ldots] \end{aligned}$$

These oscillations are normally the desirable output signals of a mixer.

Terms of a higher order also create mixed products from the harmonic frequencies. As an example, let us consider the third order term here and solve it using the mixed products:

$$\begin{aligned} I = Is * [\ldots + ½ * C1^2 * C2 * [\cos(\Omega 2 * t) + ½ * \cos((2 * \Omega 1 + \Omega 2) * t) + \\ ½ * \cos((2 * \Omega 1 - \Omega 2) * t] / 3! + ½ * C1^2 * C2 * [\cos(\Omega 1 * t) + \\ ½ * \cos((\Omega 1 + 2 * \Omega 2) * t) + ½ \cos((\Omega 1 - 2 * \Omega 2) * t)] / 3! + \ldots \end{aligned}$$

These spectrum fractions are used, for example, in a sub-harmonic mixer. However, these fractions are undesirable for most mixer circuits.

4.2. Harmonic mixing in spectrum analyser

The mixing of two signals with diodes is used for frequency conversion in ring mixers. When four diodes are wired together in the ring mixer, some signal fractions occur in the transformers, as they in phase opposition. These fractions are then strongly suppressed at the outputs, although they definitely arise in the mixer.

If a suitable intermediate frequency is selected and if filtration is provided, in normal transceivers these effects play no part. But since no input filters are provided for in the spectrum analyser, undesirable mixed products with harmonics can occur, even at the intermediate frequency level:

As an example, let us look here at the case in which an input signal of 1 GHz, free from harmonics, is exclusively fed in.

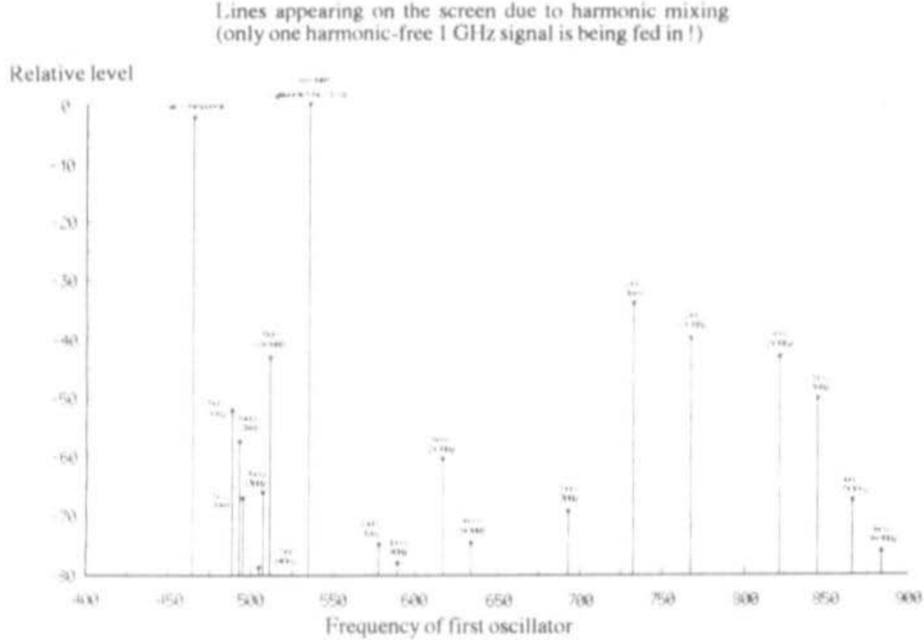


Fig.1 Ring mixer mixed products for an inputsignal of 1 GHz and various settings of the first LO; (gewünschte linie = Target line)

Harmonics arise in the ring mixer at 2 GHz, 3 GHz, etc. The first LO should be varied over a range between 450 and 950 MHz. If the first LO now oscillates at 530 MHz, an output signal occurs at 470 MHz (1 GHz - 530 MHz = 470 MHz) due to mixing with the 1 GHz input signal.

This signal appears on the screen as a line (Fig.1). Only this spectrum line is undesirable. If the first LO now oscillates at 735 MHz, the first harmonic, at 1,470 MHz, can also mix with 1 GHz (2 * 735 MHz - 1 GHz = 470 MHz)

470 MHz is again the output frequency, which causes additional lines to appear on the screen. If the first LO is oscillating at 490 MHz, among other things a third harmonic of 1,470 MHz also arises. This mixes in the same way with the input signal at an intermediate frequency of 470 MHz (3 * 490 MHz - 1 GHz = 470 MHz). Thus three lines have already appeared on the screen, without any harmonics from the input signal. But the ring mixer also generates harmonics of this input signal, and still more lines appear:

2 * 1 GHz - 2 * 765 MHz = 470 MHz

2 * 1 GHz - 3 * 510 MHz = 470 MHz

3 * 823.33 MHz - 2 * 1 GHz = 470 MHz

4 * 617.5 MHz - 2 * 1 GHz = 470 MHz, etc.

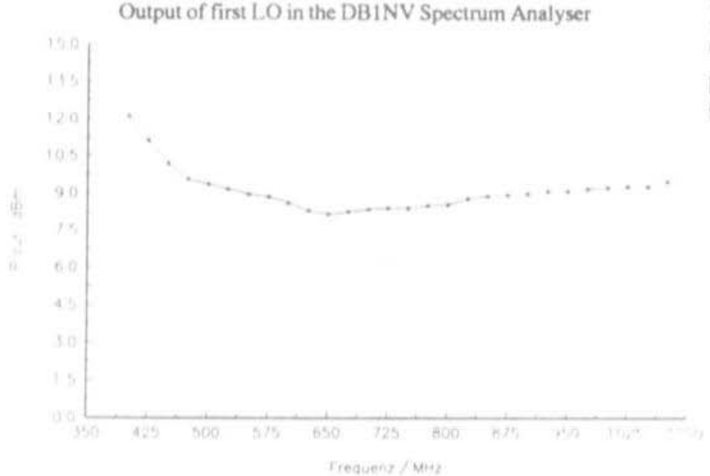


Fig.2: Output of first LO plotted against frequency

Values for additional lines can easily be worked out with a pocket calculator.

The X axis is the frequency of the first oscillator, the Y axis the relative level of the spectrum lines displayed, and the analyser must be selected in such a way that just the top line is obtained. The description of the lines can be taken from their origin:

"3 * fo - 2 * 1 GHz" means that the second harmonic of the oscillator mixes with the first harmonic of the test signal.

Since only poor bases can be obtained from undesirable mixed products in ring mixers, I have tuned these level values for an SRA1, using known data, and the results which I have measured. The levels can thus serve only as examples. The individual results are greatly dependent on the mixer used. I used a Datel M21L mixer, which is specified for use up to 3 GHz, and thus generates lines up to and beyond 4 GHz. The circumstances in my rig are actually still worse, due to harmonics which are generated by the first LO. Of course, anyone who knows which lines are expected can also use this effect, for example, to balance an oscillator at 2,556 MHz.

Harmonic mixing can be effectively combated only by means of a travelling pre-filter. A suitable control voltage is already available in the form of the tuning voltage of the VCO in the local oscillator assembly.

5. MY ANALYSER'S TECHNICAL DATA

Sensitivity: -10 dBm (to obtain top line)
- 90 dBm (bottom line)
(without additional IF amp)

Band width at narrowest filter position:
-45 dB → 10 kHz (with cheap oscillator crystals)

Level change if crystal filter band width is altered: < 1 dB

Linearity of frequency control loop:
Deviation < 2 MHz over entire frequency range (Motorola 74LS221 with buffer amplifier)

Ripple of 465 MHz helix filter:
< 1 dB

6. FINAL COMMENTS

After using the analyser for about two years, I wouldn't want to be without this piece of equipment. Balancing many assemblies would be more difficult, or even impossible, without it. Using a high-impedance key, you can carry out measurements between individual stages of fully assembled pieces of equipment and detect sources of error. I can particularly recommend such measuring equipment for beginners who are just starting to build their equipment themselves. With frequency counters, watt meters and a crystal oscillator which generates suitable harmonics, you can, for example, fully balance a 70cm or 23cm transverter.

7. LITERATURE

(1) Volker Esper, DF 9 PL: Highly stable, low-noise power pack; VHF Communications 1/93

(2) Dr. Ing. Jochen Jirmann, DB1NV: Wide-band VCO's using microstrip technology; VHF Communications 4/92

(3) Dipl.-Ing. Wilhelm-Hartmut Weishaupt, DB4DY: Assembly instructions for the DB1NV spectrum analyser (unpublished)

(4) Data sheet LM723/CA723 E.g. Valvo, Harris, etc.

MATERIAL PRICE LIST OF EQUIPMENT

described in VHF COMMUNICATIONS

DB1NV	**Broadband VCO's using Microstrip Techiniques**	**Art. No.**	**Ed. 4/1992**
PCB	DB1NV 012	6480	DM 33.00
PCB	DB1NV 013	6481	DM 33.00
Components	400 - 1250 MHz 3 x BB619; 1 x BB811; 1 x BFG96; 2 x AT42085; 1 x BFQ69; 2 x 2.2nF & 1 x 27pF Feed-through Cap.; SMC Connectors; 2 x 0.47 H SMD Choke; 1 x housing 74x55x30mm; 1 PCB DB1NV 012	6482	DM 81.00
Components	450 - 1450 MHz as above but: 1 x BFG65 instead of BFG96	6483	DM 81.00
Components	800 - 1900 MHz as above but: 4 x BB811 instead of BB619 and PCB DB1NV 013 instead of 012	6484	DM 85.00

DB6NT	**Broadband Measurement Amplifier**	**Art. No.**	**ED. 4/1993**
PCB	DB6NT 001	06379	DM 36.00
Kit	DB6NT 001	06382	DM 125.00

DB6NT	**Frequency Divider**	**Art. No.**	**ED. 4/1993**
PCB	DB6NT 002	06381	DM 36.00
Kit	DB6NT 002	06383	DM 208.00

DJ8ES	**28/144 MHz Transverter**	**Art. No.**	**ED. 4/1993**
PCB	DJ8ES 019	06384	DM 36.00
Kit	DJ8ES 019	06385	DM 229.00

Tuning Diode	BB619	10 off	10450	DM 20.00
Tuning Diode	BB811	10 off	10451	DM 31.50

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