VHF
COMMUNICATIONS

28 to 432 MHz
Transverter

Wolfgang Schneider
DJ8ES

VHF
COMMUNICATIONS
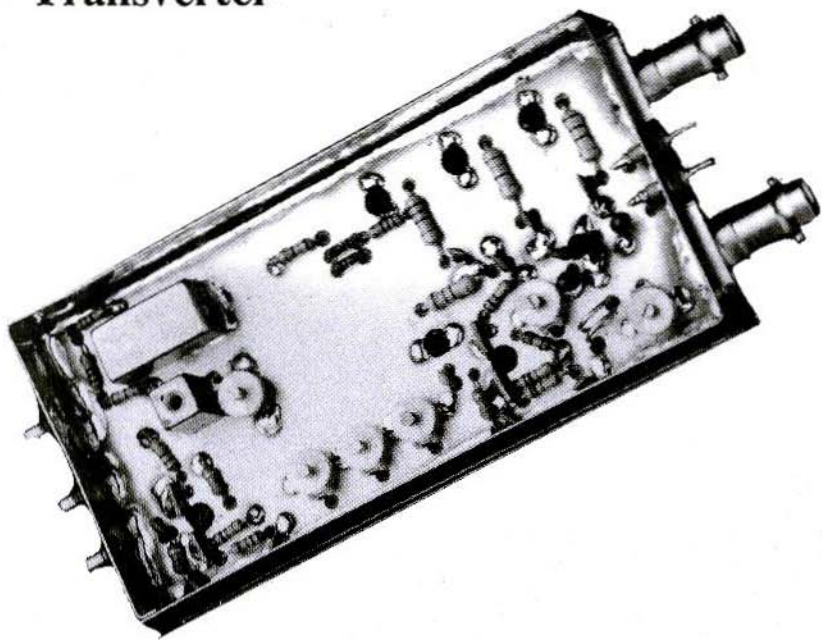
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Contents

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A Sweep Tuner for the VCO

There have been many VCO's described in VHF Communications, the latest ten appear in the last years volumes, (1) to (10). Obviously some of them are intended for use in conjunction with phase detectors in PLL's and are not to be tuned manually.This does not prevent them from use in some other, manually tuned applications.

Note the LMC6482/4 and LMC6572/4 amplifiers that have rail-to-rail input common-mode range and outputs range, operating down to 3V! Bias currents around 20pA.

I used the slightly older AD549J. The J-grade has a typical input bias current

1. DESCRIPTION

The component count for this project is low: One integrator and one DC motor!

Select an OP amp with very low bias current such as the AD 549 electrometer amplifier from Analog Devices, or something from the new LMC 6XXX-family from National Semiconductor.
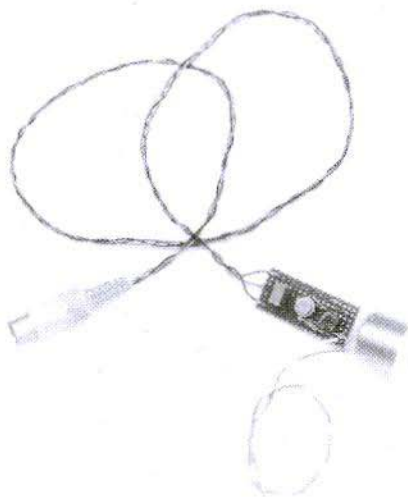
of 150pA, whereas the LMC 6XXX family has plenty of members with less than 10pA of bias current. The AD 549 has an offset trim option (10k pot to -Vs) which is useful here.

The amplifier is connected as an integrator with the lowest leakage capacitor you can find. Do not even dream of putting the virtual ground point on the circuit board! Bend the IC pin up in the air and connect the capacitor to it with air as an insulator!

The permanent magnet motor generates a voltage when the shaft is rotated. The faster rotation, the higher voltage. Polarity depends on direction of rotation.. Any motor can be used., even the cheapest piece of junk from some of the kids broken toys will give a nice smooth output from the integrator!

The preferred choice is a quality motor though, since, when it is fitted with a knob, one can spin it and it keep rotating for a while. The Swiss Portescap and the American Micromotor are to my knowledge the best.

Broken, non-working, Japanese home video cameras can sometimes be found on flea-markets. The zoom and distance servos on the lens are usually equipped with fairly nice motors as well. Tape recorders and turn tables can be worth looking into.

I have a Plexiglas disk, 6mm thick and 60mm diameter on my motor, giving it a good spin with a finger keeps it rotating for half a minute and a results in a shift of 3.5V. With the components indicated, the Escap SR 601 B1 motor gives a shift of 15mV per rotation. It is possible to set a voltage within some $10\mu V$. Considering the output range of ± 12 V this leads to the dial length of $\pm$ 1.2 million parts. Since the extremes thereof can be reached in less than a minute using a trick described later, it is a very versatile "potentiometer" for setting the frequencies of VCO's, levels of voltage for speed control, light, temperature, pressure, etc.

We live in an age when everything "is going digital" at a mindless pace. "You see, it can be controlled better that
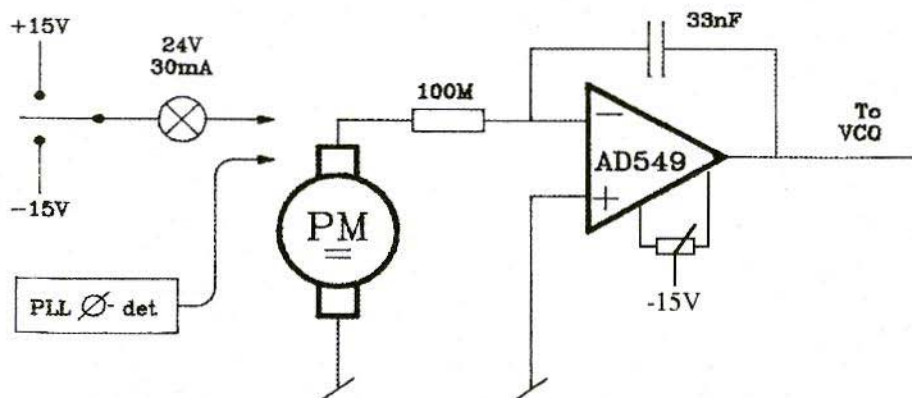
Fig.1: Circuit Diagram of the VCO Sweeper

way!" May it be said that the device described here corresponds to a 21-bit resolution D/A, is continuous, linear and has the right "finger feel" in operation. The resolution can be extended to 24-bits (a few μV settability) simply with a larger capacitor.

2.
CONSTRUCTION

The construction barely requires any description as it is very simple. If the voltage goes "the wrong way" switch the wires on the motor. If the voltage vary too fast, increase the resistor from the motor to the integrator or the capacitor value therein.

If the range desired is, say 5 to 25V for a VCO, feed the amplifier with +30/0V instead of +15/-15V. Note that at halfway, or thereabouts, ground must then be created for the other end of the motor. The most convenient method is to put this at the centre, or most used, voltage level in the application. The capacitor will then hold no charge (= no leakage) and only the offset trim of the amplifier becomes an issue.

3.
THE QUICK SPIN

To move fast over a longer distance, a voltage can momentarily be applied direct to the motor. The voltage is higher than would be the case if the motor was spinning at the speed it is going, so it hastens the move. The applied voltage will make the motor spin faster than can be achieved by hand. As the voltage is getting closer to the desired, one can just use a finger as a break and slow down the flywheel/ knob with a finger for the final adjustment.

$\pm$ 15V can be a bit much for some small, fine motors. Use a 24V/30-50 mA light bulb in series to limit the current. (A sudden reverse at high speed may blow a 12V lamp.) Use an ON-OFF-ON biased switch with spring return to middle OFF for a control.

Circuit AFC can be arranged by nudging the motor voltage very lightly via a resistor from the detector circuit. More juice here may provide the ultimate happiness, watching the knob rotate back and forth as the frequency is kept on target!

4.
LITERATURE

(1) 2/94 p. 120. R. Erping DB9JC and W. Schneider DJ8ES: 13 cm FM-ATV Exciter.

(2) 3/93 pp. 144-149. Dr.Ing. J. Jirmann, DB1NV: Theory and Practice of the Frequency Synthesiser, part 2.

(3) 2/93 p. 78. Dr F. M. Schimmel: A Simple Panorama add-on for Weather Satellite Receivers.

(4) 4/92 pp. 194-203. Dr.Ing. J. Jirmann, DB1NV: Broadband VCO's using Microstrip Techniques.

(5) 2/91 p. 108. W. Brueckner DL6MDA: RF Sweeping with the PC.

(6) 1/91 pp. 4-5. W. Schneider DJ8ES: FM-ATV Receiver for the 23 cm Band.

(7) 3/90 p. 142. G. Borchert, DF5FC: Universal Synthesiser for Frequencies up to and above 1000 MHz. Part 2.

(8) 2/90 p. 71. Dipl. Eng. D. Burchard: Shortwave Reception - Based on the Thirties Principles. Part 2.

(9) 1/90 p. 12. M. Salewski DC9DO: Sat-X Receiver for the Satellite IF band 900 to 1700 MHz.

(10) 2/89 p. 109. Dr. Eng. J. Jirman, DB1NV: A Spectrum Analyser for the Radio Amateur. Part 3.

Carl G. Lodström, SM6MOM/W6

A Grid-Dip Meter for VHF and UHF

This design was entered in the first design contest held by Radio & Television in Sweden (1). in March 1975. It won second place behind a calculator programmable with a cassette recorder.

With twenty years of hindsight this meter still seems to me like a good and useful instrument.

The commercial Grid-Dip (GD) meters, if at all available anymore, have not improved in sensitivity or internal resonances, or cost). Whereas programmable calculators now can be had for a very reasonable cost).

The GD meter is as useful an instrument as ever, so it seems that this article can fill a need.

1. INTRODUCTION

The concept used is the same as the classical "Bontoon" model 59 GD meter (Measurements Corporation, Division of McGraw-
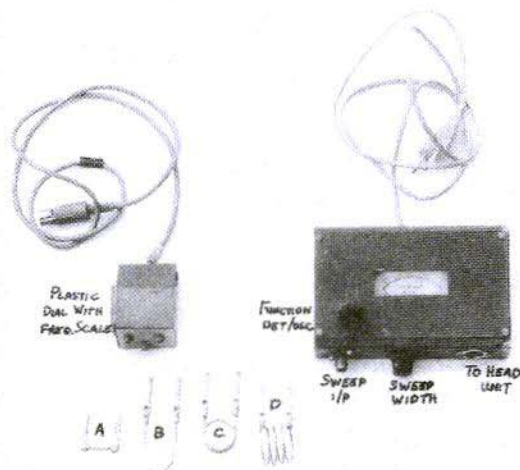
Fig.1: The Complete Grid-Dip Meter and Coils
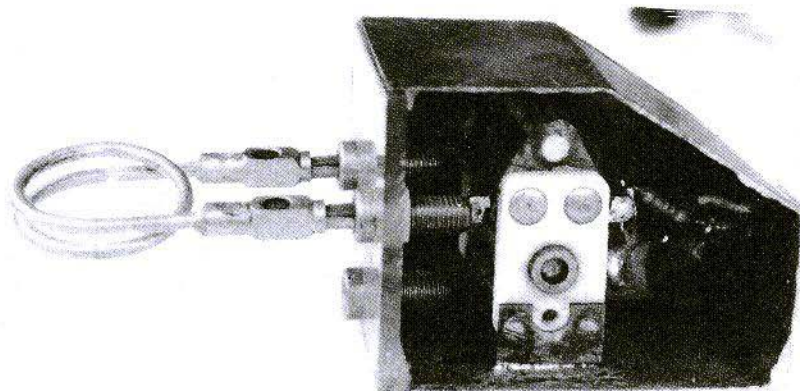
Fig.2: The Head Unit with a side plate removed

Edison, located in Boonton, New Jersey.) (Fig.4). The head is separate from the unit containing power supply and indicating meter.

This is by no means necessary, it can all reside in one box, as in the Millen, (Fig. 5) but the advantage is that different heads, for very different frequency ranges, can be built and connected to the same one power supply and meter.

At the time I built mine it never dawned on me to use two 9V batteries instead of line power! The consumption is so low that they should last "forever." Besides, better isolation, the absence of a mains lead and hum is worth a battery now and then.

The unit is shown complete with all the coils in Fig.1; whilst in Fig.2 the internals of the head with one side removed are shown and in Fig.3 is the complete circuit diagram.

2. CONSTRUCTION

Let me first recount the construction details and performance, as I remember them. I will also present suggestions that I have found in the fullness of time for improving upon the original weaknesses.

The size of the RF head depends on your needs (frequency) and what 2-gang variable capacitor is available. One can even visualise use of capacitance diodes, which can make a very small head possible. The frequency read-out could be on another (or the same) meter, or on a dial on a potentiometer in the power supply box. The frequency range would be less for each coil due to the lesser capacitance range of Varicap diodes, but more coils are easy to make.
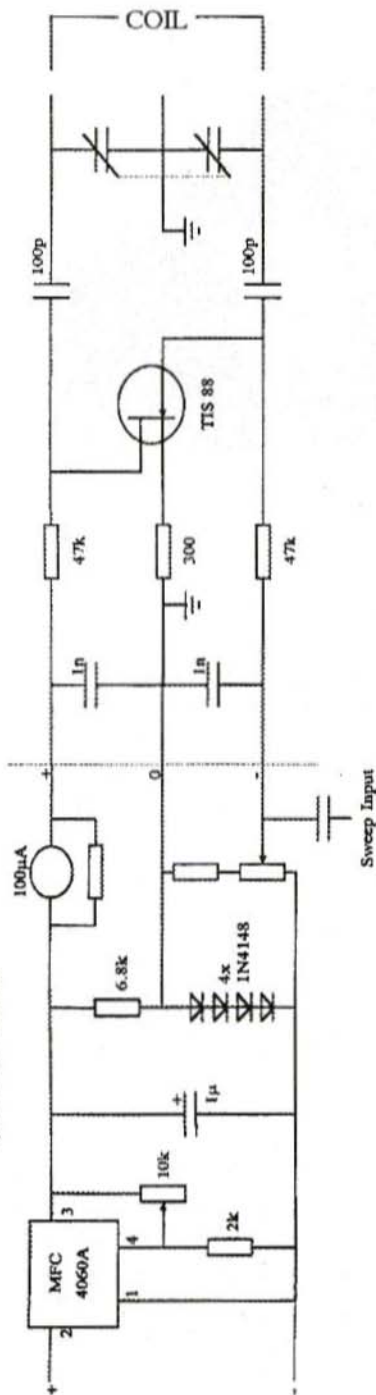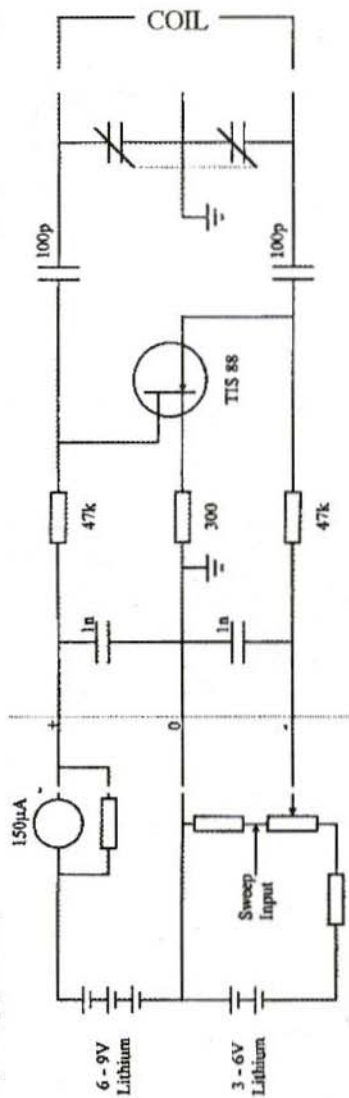
Fig.3: The Circuit Diagram of the Grid-Dip Meter

I have never seen a GD meter using Varicap diodes! Who will build the first one, and live to write an article in VHF Communications about it?!

I started "from the top", with the shortest coil I could make. "A" in Fig.1. The banana plugs connected by a straight brass tube. At first it did not want to oscillate at the higher end, so I had it silver plated, and that made all the difference. It got all the way up to approximately 420 MHz.

As I see it now, getting so close to the 70cm band, and yet not covering it, should have resulted in the capacitor catching a mild case of paradontitis and loosing a few plates! 450 MHz should be no problem to reach.

A few more coils followed. "C" followed "A", but since the 2m band ended up just about at the end of each, I also made "B" to centre it on the dial. The low end of "D" got down to 50 MHz where the transition from DET to OSC function no was longer nice and smooth, no doubt due to there being more gain in the transistor at lower frequency and the Q of the coils getting higher.

This combination does not allow for an "almost oscillating" state. Smaller coupling capacitors than the 100pF ones should help, values between 1 to 10pF may be enough.

A small capacitance in parallel with the source resistor would also give a more pronounced gain at higher frequencies, and maybe they can be balanced to provide a smooth operation over a larger range. If everything else fails, the lower frequency coils can be fitted with a resistor in parallel, or be made smaller, leading to a lower Q. A value for Q of 50-100 is about all that is needed anyway.

The shaft from the variable capacitor got "hot" even though it was grounded to the frame of the capacitor. Therefore it was necessary to use a plastic dial and knob. At first I had a metal dial, but the frequency changed when I touched it.

One key in reaching high frequencies without problems is the efficient decoupling after the 47kΩ resistors. I used disc capacitors, filing down one side of each to reach the metal of the wire and then soldering them both direct to each side of a piece of sheet copper that is also soldered to the centre divider/ ground of the variable capacitor. As we have seen in the article I wrote recently for VHF Communications (issue 4/94 pp. 249-253), capacitors become resonant, and are then not worth much for decoupling at higher frequencies.

For 450 MHz, the inductance in a 1nF will have to be on the order of 0.1nH. That is not possible unless the leads are completely eliminated, or kept to 0.15mm total length! 72pF surface mount capacitors of standard size, 3.2 x 1.6 mm, should be resonant at 450 MHz and may help if they can be fitted.

The J-FET transistor has parametric capacitances that make it possible to "VCO" the GD meter. The gate bias can be varied a little with the sawtooth sweep from an oscilloscope or a function generator. The stability was quite sufficient to sweep a regular narrow band FM radio and the linearity was
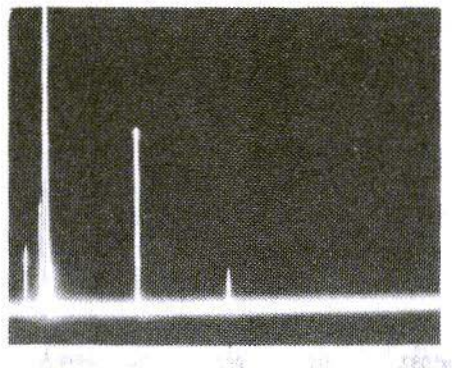
The Meter in OSC mode at 145 MHz showing the Harmonic content; 2nd harmonic -30dBc

The Meter in OSC mode at 145 MHz using Coil B with a 50 Hz Triangular Wave signal at the Sweep Input

good enough to sweep a broadcast FM radio. Markers can be obtained by letting a signal generator or a source rich in harmonics (2) interfere with the swept signal. It is valuable to add a potentiometer so the width of the sweep can be adjusted. An audio logarithmic potentiometer would be a good choice here.

If you base a design on capacitance diodes, the way to sweep the oscillator is obvious. Consider a little audio transformer in the Drain current lead in series with the μA meter. It can even be swept in the detector mode!

3.
OPERATION

An unusual feature of this GD meter is the transition from DET to OSC function. It is usually a "hard" switched function, either or. In this unit the transition portion is even more sensitive

to external fields of the same frequency, improving the sensitivity of the DET function. It even works as a detector while oscillating slightly.

Another unusual function, is that at the bottom of the DET position the instrument is shut off! Once the J-FET is cut off, the current consumption is zero! The only thing consuming current is the potentiometer dividing the negative bias for the gate. This can very well be a 2-5-10 MΩ pot, and will probably have to be put in series with top and bottom resistors anyway to provide the -2 to -4V bias, or whatever just makes your transistor happy and behave correctly.

Assume a partially discharged battery at 8V for this and a 4M7 potentiometer. Assume also a desired range of gate bias from -2 to -4V. The potentiometer will have to be put in series with 2M4 on top and 4M7 at the bottom for a total of 11M8 and a consumption of 678nA from 8V; 5.94mAh after a year...!

The other battery will be drained 50 - 100μA, but only when you use the meter in oscillator mode. So it will also last "forever."

Now there are some very nice Lithium batteries (3) available with a capacity of about 1Ah and no leakage. The terminal voltage soon drops to 8.5V, but then drops very slowly over its life to 8.0V. Don't forget to change it after some 168 years! With a 750kΩ load they should last 10 years!

We have truly reached the age when electronic equipment just can be left on for years between change of batteries! Over 8V at 57μA can be supplied for two years by one battery like this.

So you have an instrument with the three functions in one potentiometer; DET/OSC switch, sensitivity control and the ON/OFF switch!

The nicest operational range was found to occur with the drain current in the range 0 - 150μA, so a 100μA meter system was shunted a little. You may have to experiment, but start out with a 50 or 100μA meter, it is easy to shunt them, but not so easy to make them more sensitive!

The readers of VHF Communications and UKW Berichte are no doubt very familiar with the uses of GD Meters, so I will not try to "preach to the choir." The purpose of this article is to inspire the building of them by those of you who have less than well working ones. If so, it can probably be re-built along these lines. Half the job: coils, dials and meter will already be completed!

As I write this, it strikes me that one possibly can tap into, and combine this, with the technology described in (4) for a sensitive UHF detector with a large dynamic range.
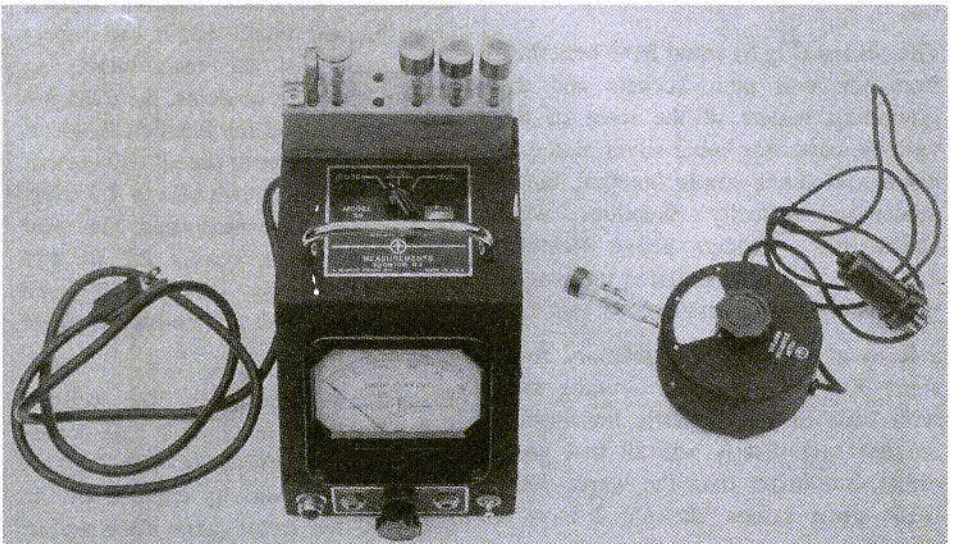
Fig.4: The classic Boonton Model 59 Grid-Dip Meter
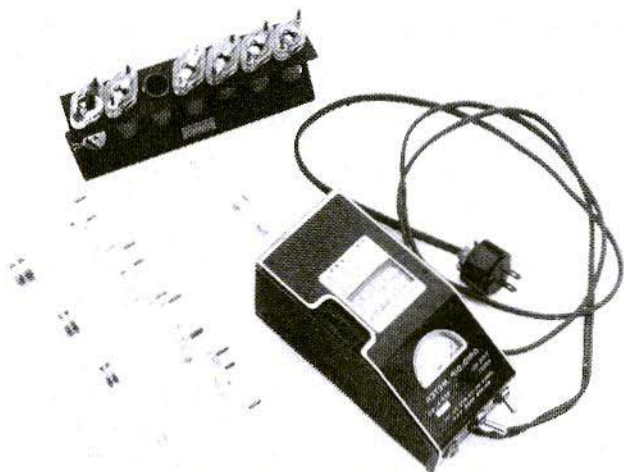
Fig.5: The Millen 90651 Grid-Dip meter

4
HISTORICAL REFERENCES

For reference, Figs.4 and 5 show two classic grid-dip meters and a lesser known one, outside the USA at least, in Fig. 6.

The Millen (Fig.5) could have benefited from the coil pins, sockets and the grounding wipers of the rotor in the variable capacitor being silver plated. It has a sometimes erratic function, but is otherwise basically functions well. Heads and shoulders above Heathkit and Eico, who have polluted the market with barely usable instruments.

I have a vague school memory from some 30 years ago, of a Grundig unit, with coils in green plastic housings. I am not sure that it was all that great either, but better than the worst. They made great radios, though! I have a couple.

The Measurements Corporation, "Boonton", Model 59 is the classic GD meter of all times! (Fig.4). The set pictured is a sight rarely seen, a complete set of 59 LF, 59 and 59 UHF, with manuals! I found it some 4 years ago at the annual Ampex Fleamarket in Redwood City, California, for a few hundred dollars! Once home I found that it had barely been used. All the seals intact, no bumps, scratches or dents, the dials are very accurate and the function flawless. It is the GD meter to top all GD meters. The 59 LF covers 100 kHz to 4.5 MHz with 4 coils. The next head, the most commonly found, covers 2.2 to 420 MHz with 7 coils. The 59 UHF head covers 420 to 940 MHz with one fixed coil.

Fig.5 shows the Millen. The wooden box is probably home-built. The coverage is 1.7 to 300 MHz with 7 coils. There is no sensitivity control, all the action of the meter takes place within the limits of the dial. It is a bit "lame"

on the 140-300 range, and silver plating the entire little loop coil would most certainly cure this. Not commonly found are the four extra coils on the right in Fig. 4. They cover 225 kHz to 2.05 MHz. Like the others, the Millen GD meter contains a triode, 6C4 if I remember right. The "Acorn" tube 955 is used in the two other GD meters.

Fig.6 shows a military GD meter, built by Stamford Electronics Co., Connecticut, USA. AN/PRM-10. It covers 2 to 430 MHz with 7 coils. It is very accurate and has good sensitivity and is flat over each band. Contributing to this is probably a fairly complicated system of resistors in each coil and a potentiometer coupled to the tuning capacitor. The coil plugs are silver plated, as are the entire coils for the highest frequencies. This meter can be found at the sales here for $60-100 and is well worth its price.

5.
LITERATURE

(1) C. G. Lodström, Transistoriserad Resonansmeter;Radio & television (Sweden) 3/1975 pp. 23-24.

(2) N. Rohde, EMC and its Consequences;VHF Communications 3/93 pp. 185-187.

(3) Ultralife U9VL-BP, 10 years guaranteed storage life. Ultralife Batteries, Inc., P.O.B 622, Newark NY 14513;or from the author at US $ 6.50 each. Postage may have to be added, they weigh 40g each in "bubble-pack."

(4) D. Burchard, Shortwave Reception Based on the Thirties Principles, VHF Communications 1/90 pp. 23-30 and 2/90 pp. 70-76.
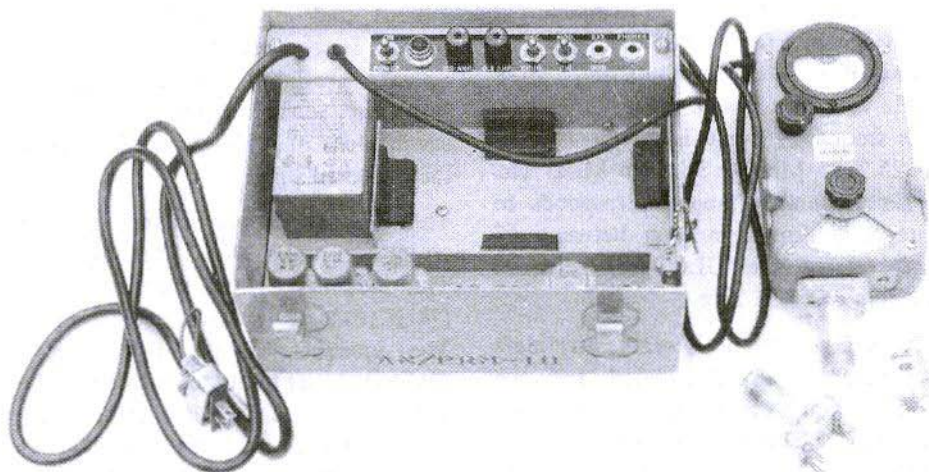
Fig.6: The AN/PRM-10 ex Military Grid-Dip Meter

Matjaz Vidmar, S53MV

A DIY Receiver for GPS and GLONASS Satellites Part-5

This part of the series continues with the construction details for the GLONASS PLL Synthesiser, Second IF Amplifier and the GPS/GLONASS DSP Hardware.

4.8 GLONASS PLL Synthesiser Logic

To convert the frequency range 1602 MHz to 1615.5 MHz down to 118.675 MHz the VCO must operate in the frequency range 1483.3125 MHz to 1496.8125 MHz. Subtracting 1458 MHz in the PLL synthesiser converter, this frequency range is downconverted to 25.3125 MHz to 38.8125 MHz. The latter frequency range corresponds to integer multiples ranging between 45 and 69 of the GLONASS channel spacing of 562.5 kHz.

The design of the PLL synthesiser logic is therefore straightforward and the corresponding circuit diagram is shown in Fig.36. The PLL synthesiser logic includes a PLL IF, a fixed divider by 32 to obtain the 562.5 kHz reference and a frequency/phase comparator.

The PLL IF signal in the range from 25 MHz to 39 MHz is first amplified to a TTL level in a two stage amplifier. The gain of this amplifier is set higher than required to have a considerable safety margin. The base bias resistor of the second stage may need some trimming although the suggested value of 2.2kΩ will usually work.

The programmable modulus counter is built from two synchronous counters: a 74F161 and a 74HC161. A 74Fxxx counter is required in the first stage to operate reliably at the highest frequency, since the PLL IF may be several tens of MHz above 39 MHz in the unlocked state!

The counter feedback network includes an inverter and a 74HC157 multiplexer. The modulus of the second counter (74HC161) is programmed directly. The modulus of the first counter (74F161) is set to 10, except during the last state of the second counter, when the 74HC157 switches the 74F161 preset inputs to the 4094 outputs. In this way setting the modulus of the

Fig.35: GLONASS Synthesiser Logic

second counter changes the whole divider modulus in steps of 10, while setting the modulus of the first counter during the last cycle only provides the single-count steps.

The 4094 is an 8-bit shift register with output latches. It is used as a serial-to-parallel interface driven by the computer. Of the eight available output lines, four are used to control the modulus of the 74HC161 counter and the other four to control the modulus of the 74F161 during the last cycle of the 74HC161. One should be especially careful when programming the modulus of the divider: the data is inverted and the first divider modulus should never be set too low to allow for the delays in the slower 74Hcxxx logic!

The 18 MHz master reference frequency is also amplified to TTL level in a single-stage amplifier (2N2369). A 74HC393 counter divides

this frequency by four to obtain the 4500 kHz sampling frequency and by 32 to obtain the PLL reference frequency. The 4500 kHz signal is attenuated with a resistor network to limit spurious radiations. It is then amplified back to TTL level in the dedicated DSP hardware module.

The frequency/phase comparator is a charge-pump circuit, including two D-type flip-flops (74HC74), a feedback network with a NAND gate and charge-pump switches with Schottky diodes. The backlash problem is solved by making the charge-pump circuit faster than the feedback network. This does not make the phase detector linear yet, but provides a stable locking point for the PLL with no dead zone, and produces a very clean synthesiser signal spectrum. An additional NAND gate provides a LOCK signal for test purposes.

Fig.36:
PCB Layout for
the GLONASS
Synthesiser Logic

The GLONASS PLL receiver logic is built on a single-sided printed circuit board as shown in Fig.36. The location of the components is shown in Fig.37. There are three wire jumpers on this board and two of them are installed below the 74HC157 multiplexer. A complex single-sided board also places some constraints on the installation of the resistors: those with a 10mm hole spacing are installed horizontally, whilst the others are

installed vertically to save space. All of the capacitors are ceramic with a 5mm pin spacing, including the 1uF (multilayer).

L1 is a commercial 100µH moulded choke. The module has three connectors with 3, 4 and 7 contacts, made from pieces of good quality IC sockets. The integrated circuits should be soldered directly to the board except for the 4094. It is recommended to install this IC on a socket, so that it can

Fig.37: Component Layout for the GLONASS Synthesiser Logic

Fig.38: GPS/GLONASS Second IF Amplifier

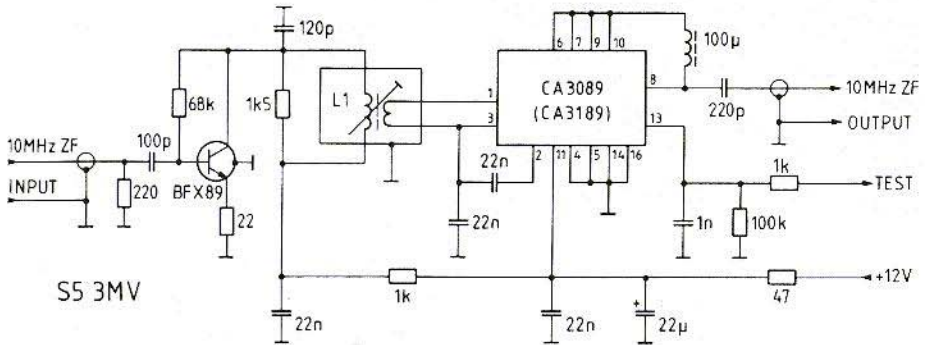be removed from the circuit and replaced by wire jumpers, to be able to test the synthesiser without the computer running.

The GLONASS synthesiser logic may require a single adjustment: the bias resistor for the second PLL IF amplifier stage. This may be adjusted if the output DC voltage deviates much from 1.3V (with no input signal) or if the programmable counter does not operate reliably.

4.9 GPS/GLONASS Second IF Amplifier

Both GPS and GLONASS receivers require a limiting IF amplifier at the final IF frequency around 10 MHz. The circuit diagram of this amplifier is shown in Fig.38. The GPS/GLONASS second IF amplifier includes a first stage with a bipolar transistor, BFX89, and a second stage using the integrated circuit CA3089.

The gain of the first stage is limited by the 22Ω resistor in the emitter circuit. The first stage is followed by a tuned circuit (L1) to limit broadband noise and avoid amplifying various spurious signals from the many oscillators inside a GPS or GLONASS receiver. The damping resistor in parallel with L1 sets the bandwidth of this tuned circuit to be comparable with GPS or GLONASS signal bandwidth.

The second stage uses a popular FM IF strip integrated circuit. The latter provides wideband amplification and limiting, while the discriminator section of this integrated circuit is not used here. The limited IF output is available on pin-8 and the signal level amount to a few hundred millivolts at 10 MHz. This is not enough to drive the following TTL logic directly and the remaining gain is built inside the dedicated DSP hardware module.

The CA3089 integrated circuit includes an S-meter output. This is of little use during actual receiver operation, since the satellite signal levels are comparable to noise in the wideband IF. In the case of a GPS receiver, the S-meter output can only show the sum of all the signals present. On the other
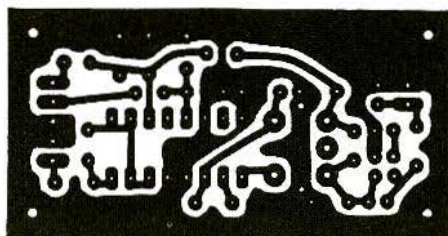
Fig.39:
PCB Layout for the GPS/GLONASS
Second IF Amplifier

hand, the S-meter output is very useful during receiver testing and alignment of the RF and the first and second IF tuned circuits.

Both GPS and GLONASS receivers include an S-meter function inside the narrowband IF processing. Since the latter is done in software, the real receiver S-meter as displayed on the LCD is just another software function and is NOT related to the hardware S-meter output of the IF strip.

The GPS/GLONASS second IF amplifier is built on a single-sided printed circuit board as shown in Fig.39, with the component location layout shown in Fig.40. Due to limited space all the resistors are mounted vertically. The capacitors are conventional disc ceramics (except for the 22µF) with a pin spacing of 5mm.

L1 is wound on a 10.7 MHz IF transformer set, including a fixed central ferrite core, an adjustable ferrite cup, various plastic support parts and a metal shielding can. The primary (resonant) winding of L1 has 10 turns of 0.15mm diameter enamelled copper wire, corresponding to an inductivity of about 2µH. The secondary (link) winding has two turns of the same wire. L2 is a 100µH moulded choke.

The second IF amplifier module has two connectors: a 2-pin connector for the input and a 5-pin for the output and supply voltage, both obtained from pieces of good quality IC socket with round contacts.

The tuned circuit with L1 is best adjusted after the GPS or GLONASS receiver is completely assembled, finding the maximum DC voltage on the test S-meter output in the same IF module. The final adjustment of all signal circuits is best performed on a real satellite signal obtained from a

Fig.40: Component Layout for the GPS/GLONASS Second IF Amplifier

Fig.41: GPS/GLONASS DSP Hardware (part-1)

Fig.21: GPS/GLONASS DSP Hardware (part-2)

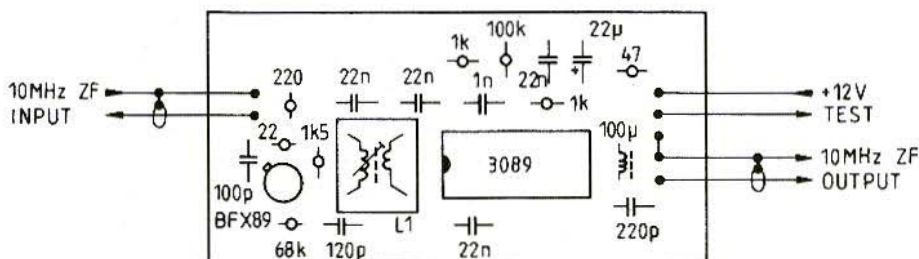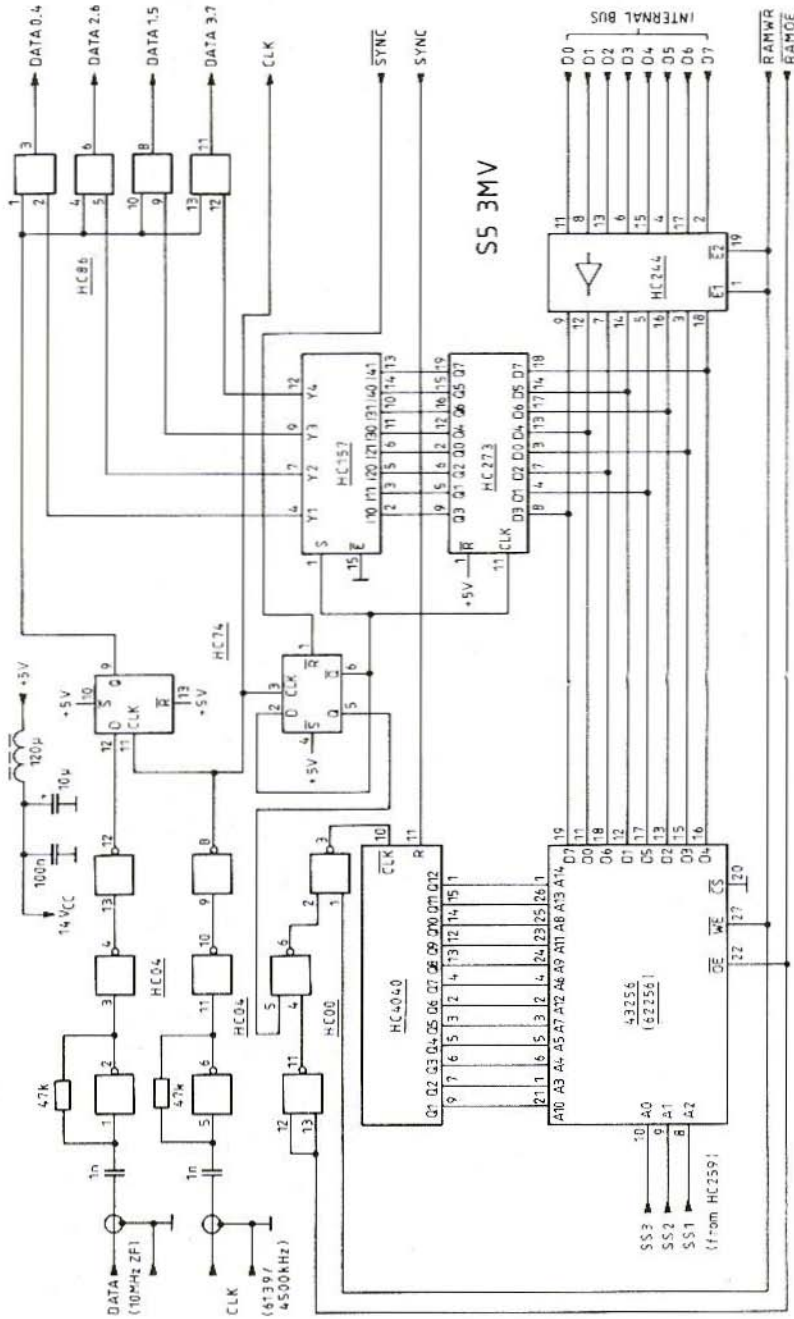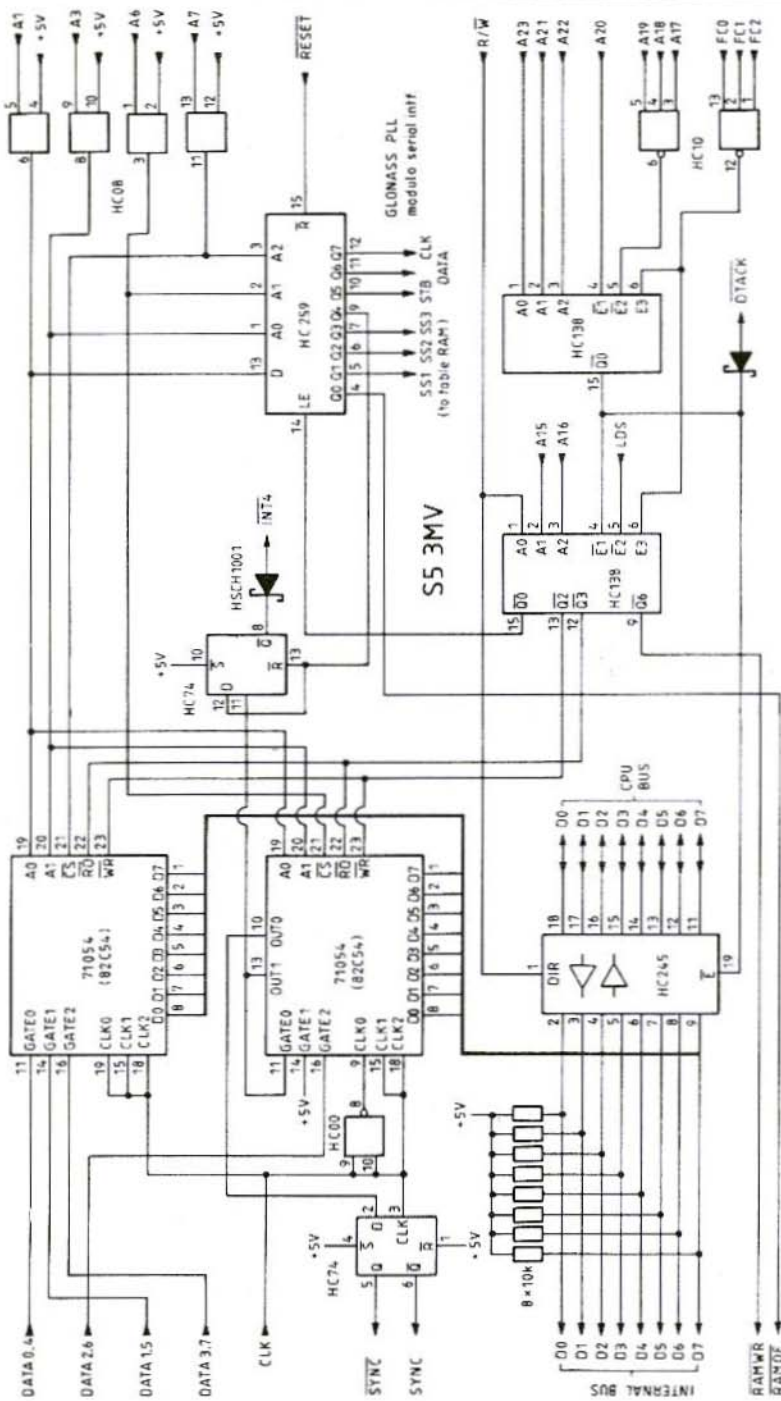directional antenna (a 15-turn helix or a small dish) pointed at an operating navigational satellite. A directional antenna should provide a higher than usual signal-to-noise (SNR) ratio of more than 10dB already available in the wideband IF. A high SNR is required to tune the circuits to the signal peak and not to noise or some spurious signals while observing the voltage on the test S-meter output.

4.10 GPS/GLONASS DSP Hardware

The theory of operation of the dedicated GPS/GLONASS DSP hardware has already been discussed in 3.6. Therefore the practical implementation only will be described in the following section.

The dedicated GPS/GLONASS DSP hardware is built as a peripheral plug-in module for the DSP computer (1), (2). The circuit diagram of the GPS/ GLONASS hardware is shown in Fig's.41 and 42. The DSP hardware module includes two amplifiers for the limited IF signal and master clock coming from the analogue part of the GPS or GLONASS receiver, a look-up table RAM, four signal despreading mixers, four correlator counters, all of the timing logic to scan the look-up table and generate interrupts to the CPU and all of the interface circuits necessary for the DSP computer bus.

The input signal amplifiers are built with 74HC04 inverters to amplify the input signals of a few hundred millivolts up to TTL levels. In this way the signal levels in the analogue part of

the GPS or GLONASS receiver can be kept more than 20dB lower, reducing the shielding requirements. The supply voltage for the 74HC04 is additionally filtered to reject the spurious rubbish usually present on the +5V computer bus power supply. The limited input signal is then sampled by a D-type flip-flop (½ 74HC74). Considering the relationship between the IF and the sampling frequency, the sampling operation also provides a third downconversion.

The look-up tables are stored in a 32k x 8 static RAM. The RAM area is divided into 8 separate areas of 4 kbytes each, selectable through a microprocessor output port. In this way the look-up table does not need to be rewritten when switching to another satellite. The receiver is usually time-multiplexed among four different satellites and all four different look-up tables are stored in the RAM. When switching to another satellite, the hardware is simply switched to another look-up table and this only requires executing a few instructions instead of rewriting the whole 4 kbyte table.

The 4 kbytes of each look-up table are written as bytes by the microprocessor. The microprocessor writes all of the bytes to the same location, since the address counter is incremented automatically after each write operation. In read mode the look-up table is scanned by the same hardware counter (74HC4040) clocked at half the sampling frequency (3069.5 kHz for GPS or 2250 kHz for GLONASS). The byte data is latched (74HC273) and then multiplexed to 4 bits (74HC157)
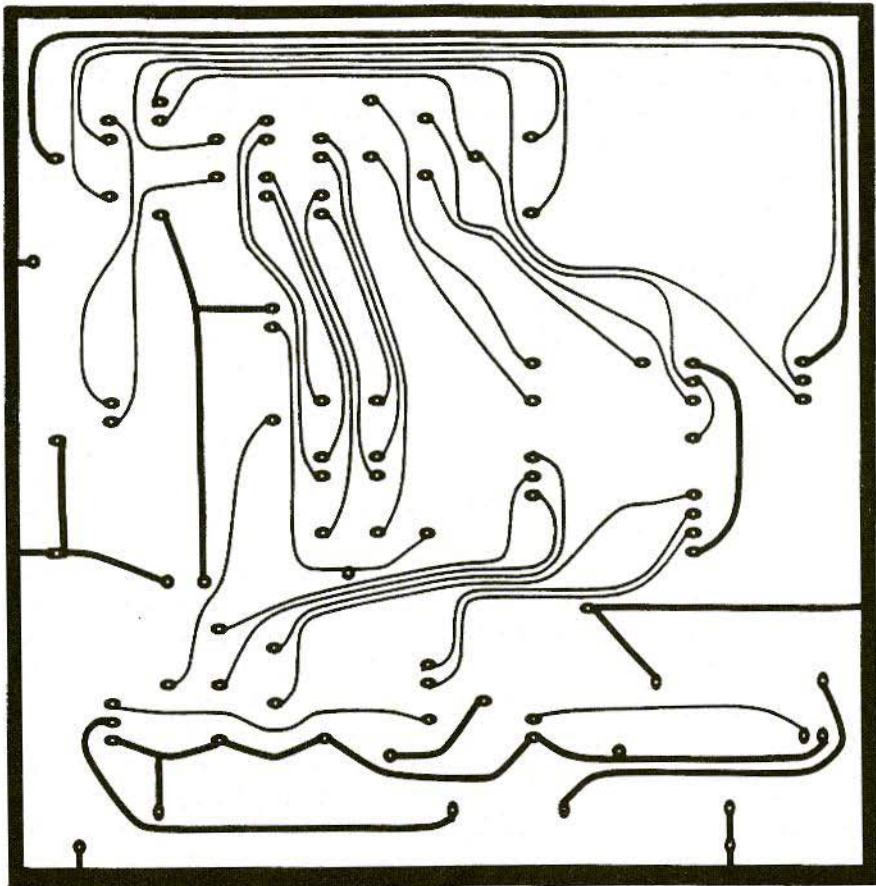
Fig.44: PCB Layout for the GPS/GLONASS DSP Hardware (top view)

to get all of the four required local signal replicas to be multiplied (EXOR 74HC86) with each input signal sample. Each look-up table may be thus up to 8192 samples long. The unused samples need not be written since they are not used by the hardware.

The dedicated DSP hardware requires six programmable counters: four correlation accumulators, a sampling frequency divider and a variable delay counter, all contained in two PD71054

(82C54) integrated circuits. Each PD71054 contains three almost independent 16-bit counters that can be programmed in different ways. For example, the four correlation counters are clocked with the same signal sampling frequency and the signals are fed to the GATE inputs, which are programmed as clock enables.

The signal sampling frequency (6139 or 4500 kHz) is divided down to 1 kHz to match the C/A-code period (1ms).
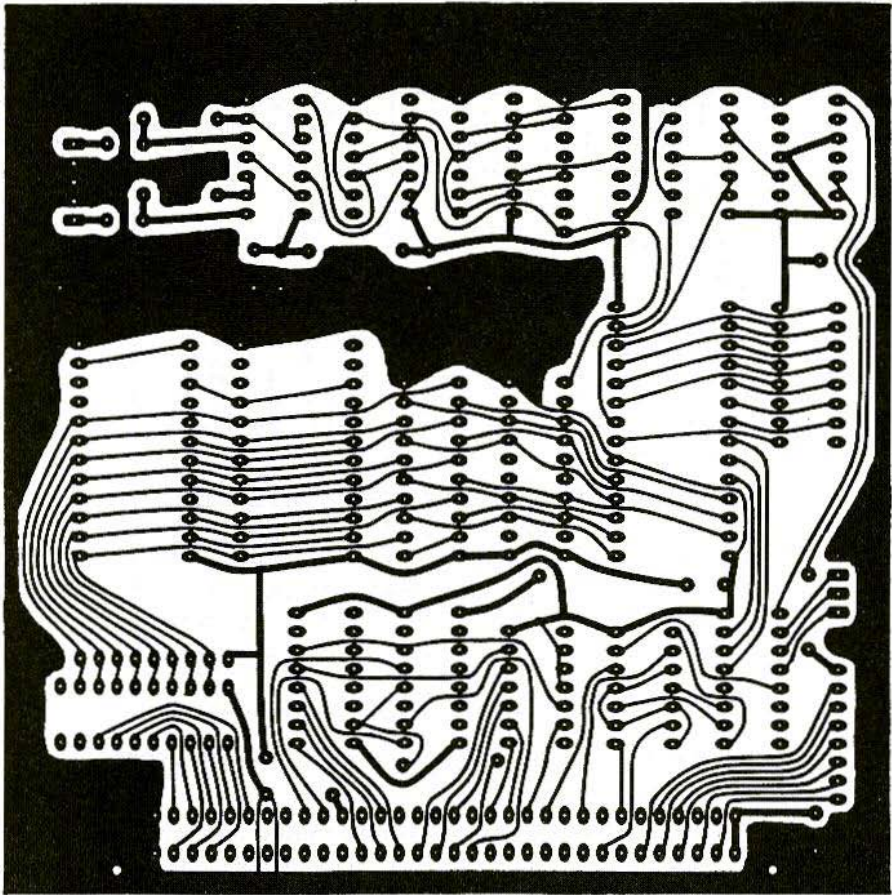
Fig.44: PCB Layout for the GPS/GLONASS DSP Hardware (bottom view)

This signal is also used to request interrupts from the CPU, since the correlation counters need to be read each millisecond. After receiving an interrupt request, the CPU will latch the contents of all counters in a single bus operation and then read the latched content of every single counter in separate bus operations. The sampling clock divider is also latched and read and its content is used as an accurate timing reference.

Every interrupt request sets a flip-flop that needs to be reset by the microprocessor after the interrupt has been serviced. Interrupt arm (enable) and reset (disable) is performed through one (Q4) of the eight output port bits provided by the 74HC259 addressable latch. Of the remaining 7 bits, three (Q1, Q2 and Q3) are used to select one of the eight look-up tables in the RAM and another bit (Q0) is used to select either write or read mode for the look-up table logic. The last three bits
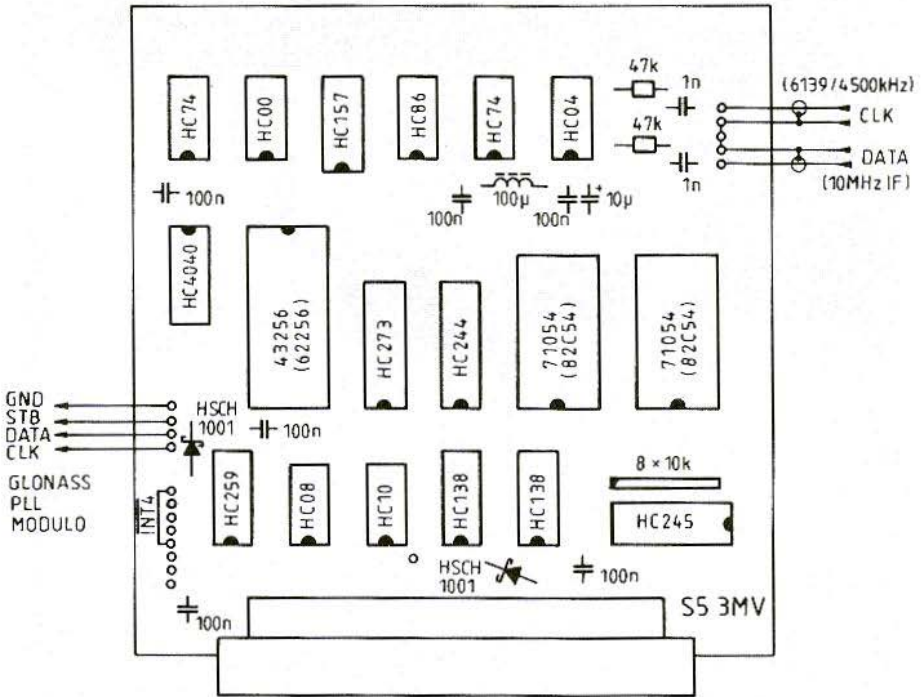
Fig.45: Component Location for the GPS/GLONASS DSP Hardware

(Q5, Q6 and Q7) are used to control the GLONASS PLL modulus or as spares in a GPS receiver.

The bus interface to the DSP computer includes a bi-directional data-bus buffer (74HC245) and an address selection logic (two 74HC138 and one 74HC245). The bus interface does not request any wait states from the MC68010 CPU. The address decoding for the PD71054 programmable counters must allow simultaneous write operations to both control registers of both peripherals, to be able to latch the contents of all of the counters at exactly the same time. Finally, the RESET signal is fed to the 74HC259 addressable latch essentially to prevent

any interrupts or other unintended operations before the whole DSP hardware is correctly initialised.

The bus addresses are assigned as shown in Table-1. However, one should notice that the remaining addresses in the range from $E0000 to $FFFFF are not fully or correctly decoded, although the module will acknowledge the access to the MC68010. Accessing other addresses in this range (either reading from, or writing to) will probably cause an erratic operation of the whole module. The 74HC259 addressable latch is programmed by writing to the specified locations. Since only the address is important and the data is ignored, CLR.B instructions are

The bus addresses are assigned as follows:

$E0001	Disable look-up table write mode
$E0003	Enable look-up table write mode
$E0009	Look-up table address A2 reset
$E000B	Look-up table address A2 set
$E0041	Look-up table address A1 reset
$E0043	Look-up table address A1 set
$E0049	Look-up table address A0 reset
$E004B	Look-up table address A0 set
$E0081	Reset and disable 1ms interrupt
$E0083	Enable 1ms interrupt
$E0082	Reset and enable 1ms interrupt, long transfer!
$E0089	GLONASS PLL modulus STROBE reset
$E008B	GLONASS PLL modulus STROBE set
$E00C1	GLONASS PLL modulus DATA reset
$E00C3	GLONASS PLL modulus DATA set
$E00C9	GLONASS PLL modulus CLOCK reset
$E00CB	GLONASS PLL modulus CLOCK set
$E800B	Common write to both 71054 command registers
$E8041	71054 #1 CTR0 - data 0,4 accumulator
$E8043	71054 #1 CTR1 - data 1,5 accumulator
$E8049	71054 #1 CTR2 - data 3,7 accumulator
$E804B	71054 #1 command register
$E8081	71054 #2 CTR0 - variable C/A-code delay
$E8083	71054 #2 CTR0 - GPS /6139 or GLONASS /4500 clock
$E8089	71054 #2 CTR0 - data 2,6 accumulator
$E808B	71054 #2 CTR0 - command register
$E8001	WRITE byte to look-up table

used to write to single bytes and a CLR.L instruction is used to reset and arm the interrupt flip-flop.

The dedicated DSP hardware board is built on a double-sided printed circuit board as shown in Fig's.43 and 44, with the component overlay shown in Fig.45. The single resistors, diodes and the 100μH choke are installed horizontally. The eight 10kΩ resistors are in a single 9-pin SIL package. The capacitors are ceramic except the 100μF tantalum, and all have a spacing of 5mm.

The 74HC4040 should NOT be replaced by the standard 4040 device, since the latter is too slow for correct operation in this circuit. To allow for troubleshooting it is recommended that at least the two 71054 counters and the

43256 RAM are installed in good quality sockets. The speed of the AM is unimportant, since even the slowest 150ns static Ram devices are fast enough for this project.

The dedicated DSP hardware module is inserted in the DSP computer bus with a 64-pole Eurocard A+C connector. The remaining connectors include a 5-pin socket for the IF input signal and clock, a 4-pin socket for the GLONASS PLL modulus control and an 8-pin socket for the interrupt selection. All of these are made from parts of good quality IC sockets.

The dedicated DSP hardware module requires no alignment or setting up.

(to be continued)

Details of kits and PCB's for this project appear on page 127 of this issue

VHF COMMUNICATIONS PROJECTS
KITS AVAILABILITY UPDATE

Kit	Issue	Description	Item No.	Price
DC8UG-PA	3/94	5W PA for 13cm Complete kit with LP's	06938	£266.00
DJ8ES-019	4/93	28/144 Transverter	06385	£139.00
DJ8ES-201	2/94	13cm FM ATV Exciter	06388	£ 70.00
DB6NT-001	4/93	Wideband Measuring Amplifier	06382	£ 65.00

KM Publications, 5 Ware Orchard, Barby, Rugby, CV23 8UF
Tel: (0)1788 890365   Fax: (0)1788 891883
Email: 100441,377@compuserve.com

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Frequency Doubling using Semi-Conductor Diodes

At low input voltages, every rectifier circuit has an area where there is a quadratic relationship between the output quantity and the input quantity. One use for this ratio is the measurement of the power or the root mean square value as carried out, for example, in (2). Another application is frequency doubling, which is the subject of this article.

1. FUNDAMENTAL PRINCIPLES

Semi-conductor diodes have in general exponential voltage-current characteristics. This can be used directly to derive a quadratic relationship for the rectified current at input voltages of below $2 \cdot U_T$ (U_T temperature voltage 25 to 50mV). The relationships are explained in more detail in (1). The following formula is also obtained from the same literature:

$$\frac{I}{I_S} = \frac{\hat{U}}{U_T} \sin\Omega t + \frac{1}{4} \frac{\hat{U}^2}{U_T^2} - \frac{1}{4} \frac{\hat{U}^2}{U_T^2} \cos2\Omega t \quad (1)$$

For a sinusoidal diode with the amplitude $\hat{U}$, the standardised diode current I/I_S has essentially three terms: the first, with the input frequency, which increases linearly with the input voltage; the second, which corresponds to the rectified current and increases quadratically with the input voltage; and the third, which is an alternating current with a doubled frequency (2Ω) and with an amplitude equal to the rectified current. It is this third term which is of interest to us here. It also increases quadratically with the input voltage.

The first term, the input frequency fraction, can theoretically be made to vanish by means of a push-pull arrangement using two diodes. In practise, of course, this is not completely successful. However, it is relatively easy to ensure that the doubled input frequency emerges from the output at least ten times stronger than any other fraction.
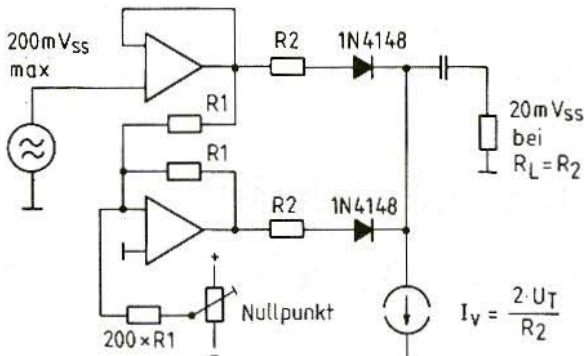
Fig.1:
Low-Frequency
Frequency Doubler

Nullpunk = Zero point
bei = at

Burchard

According to (1), the e.m.f. of the doubled frequency generated is independent of the type of diode and the point of operation. By contrast, the internal resistance at which this e.m.f. is available changes very considerably with the type of diode and the semiconductor material. Without a bias current, the internal resistance of Si diodes will be a few hundred megOhms, a few megOhms for Schottky diodes, and a few tens of kiloOhms for Ge diodes. The figures apply to small-signal diodes. In high-frequency circuits we would like lower Ohm values, right down to 50Ω, in order to match to a system impedance. This is possible if a suitable bias current is selected. In general, the AC resistance in the area of exponential voltage-current ratio is expressed by:

$$R_\sim = \frac{U_t}{I_v} \qquad (2)$$

For the desired value of 50Ω, then $I_v = 0.6mA$. If we now look at the curves contained in (1), we discover

that such bias currents are possible only using Schottky diodes. Ge diodes leave the exponential area above about 100μA. The border between Schottky diodes and Si diodes even earlier, at 10μA (1N4148). Other universal Si diodes go somewhat further: 1N4151 to 30μA and 1N4448 to 100μA. The border between Schottky diodes and Si diodes lies at the point where 50Ω technology begins, i.e. perhaps between 1 MHz and 10 MHz. Ge diodes are not required, unless we want to mange without a bias current.

2.
CIRCUITS

The frequency doubling circuit using semi-conductor diodes is in no way different from a rectifier circuit. Instead of the DC the AC is now coupled out. Fig.1 shows a circuit suitable from audio to radio frequencies, in which the input-side phase reversal and isolation of the source is carried out by operational amplifiers. The offset voltages of
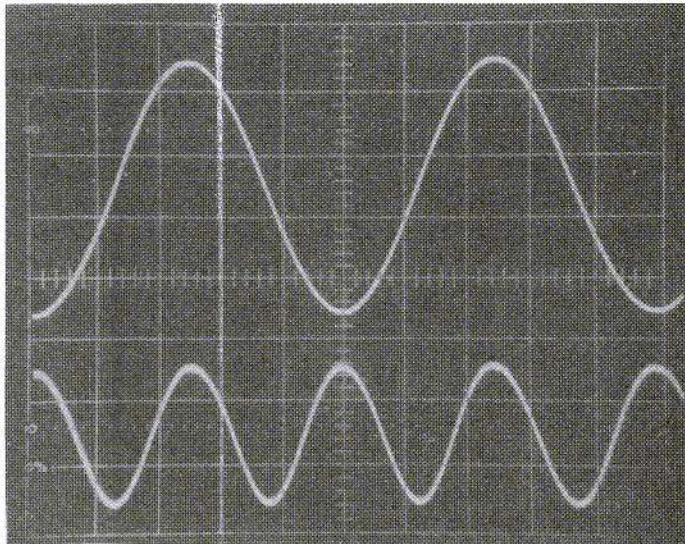
Fig.2: Input and Output Signals of Circuit in Fig.1
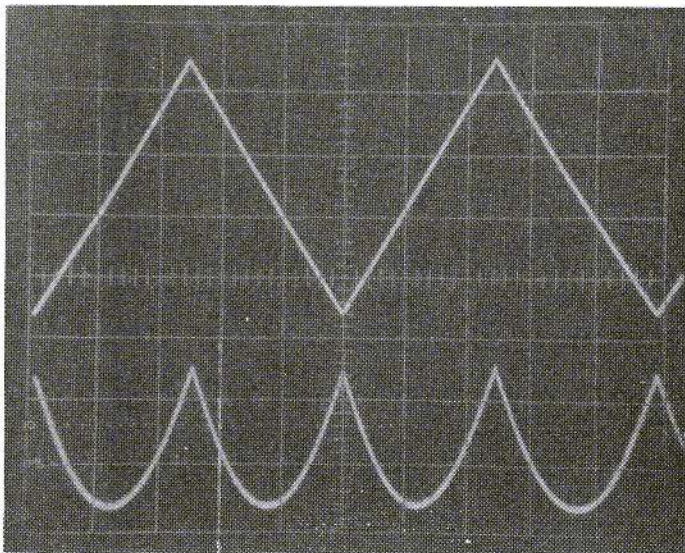X: 0.2 μs/div; Y1: 50mV/div; Y2: 10mV/div

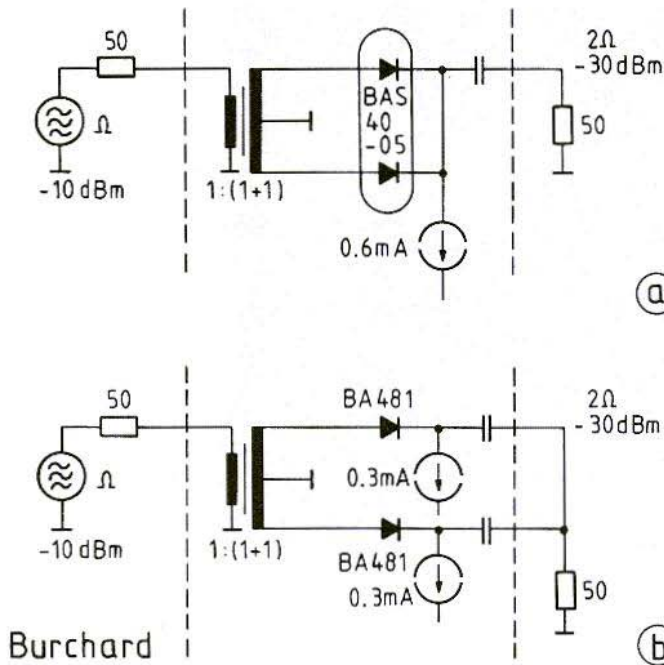Fig.3: As Fig.2, but with Delta-Form Input

Fig.4:
High-Frequency
Frequency Doubler
a. With a Diode
   Pair
b. With Individual
   Diodes which
   do not have
   Identical
   Characteristics

Burchard

the amplifiers and the variations in the diodes are balanced out by the zero preset. The resistances R2 prevent protect against low loads for the amplifiers. They would be selected to be between a few hundred Ohms and a few kiloOhms. If the load resistance $R_L = R_2$, the best power matching is obtained using the bias current marked up I_V. For idle operation, we naturally obtain double the output voltage, and then the bias current is not at all critical. Nor is any special current source required for the bias current. A resistor to the negative operating voltage of the operational amplifier is fully sufficient.

Fig.2 shows how well this doubler operates and how little distortion it produces. Here fast type LM361

opamps were used, so that even a signal of 1 MHz is still satisfactorily squared (Fig.3). Thus we have here a wide-band circuit, which functions from low frequencies right up to a higher limit determined by the opamp. For low distortion the input voltage of $200mV_{SS}$ should not be greatly exceeded and the bias current should be selected accordingly. We then obtain approximately $20mV_{SS}$ at the output, for power matching, and twice that for idle operation.

For frequencies higher than about 10 - 15 MHz there are no suitable operational amplifiers. A transformer is required for phase reversal now, so that finally a circuit like Fig.4a or b is obtained. It is also wide-band, and can be used in the HF, VHF and UHF ranges with a suitable transformer.
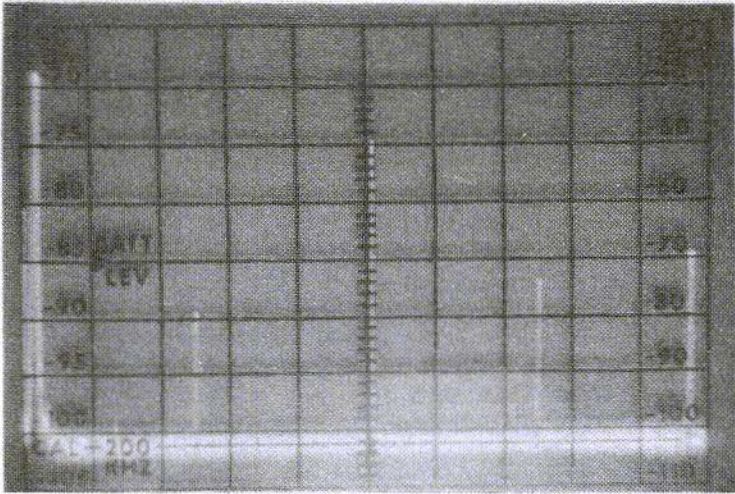
Fig.5: Output Spectrum of Fig.4b Circuit with 250 MHz/-10dBm at the Input
X: 100 MHz/div (0 to 1 GHz, zero marker on the left)
Y: 10dB/div (500 MHz line corresponds to -30dBm)

If the two diodes in a circuit are largely identical in their electrical properties, then the connections can be as in Fig.4a. However, a circuit like Fig.4b also functions with considerable discrepancies in the characteristics of the individual diodes.

The circuits are simple enough, and require only input power in the ranges which most signal generators can provide for them. A doubling circuit thus makes it possible to double the application tuning range of a measuring instrument. Many commercially available signal generators make use of this application for the top octave of their tuning range (3). Unfortunately, this also doubles the original noise deviation of the generator, and any amplitude modulation is distorted in accord-

ance with Fig.3, unless the modulator is positioned after the doubler.

Power matching is brought about by the selection of a suitable bias current. A 200Ω load should act at the output of the 1:(1+1) transformer. So each diode must have an AC resistance of 100Ω. This immediately results in output matching, for the diodes are parallel for the doubled frequency and the effective internal resistance is 50Ω. The bias current values given apply for $U_T=30mV$. Purists can vary the bias current until a VSWR optimum is obtained on the input and output sides, or until the optimum conversion efficiency is obtained. A degree of efficiency of 1% (conversion loss 20dB) is fairly typical for this circuit.

3.
FURTHER
CHARACTERISTICS

Things are not ideal in technical reality. The input curve form is not ideally sinusoidal, the diodes are not completely identical and the characteristics are not exactly exponential.

However, it is not very difficult to obtain an output signal from such a doubler, which is 20dB or more better than all other output products.

Fig.5 shows the spectrum obtained if circuit Fig.4b is used, with a balanced-to-unbalanced transformer from the UHF input of a clapped-out television and the output voltage of a Marconi TF2015 signal generator. A spurious suppression of better than 20dB is thoroughly consistent with most high-frequency measurements on selective objects. Indeed, signal generators which can be purchased are often no better than this.

Because of the quadratic relationship the output rises twice as steeply as the input. This relationship is shown in Fig.6.

It is thus possible, without any problem, to set very low defined output voltages by varying the input power level. Only two special features need to be taken into account:

– If the precision with which the input level can be set is, for example, ±1dB, then the output error can not be smaller than ±2dB.
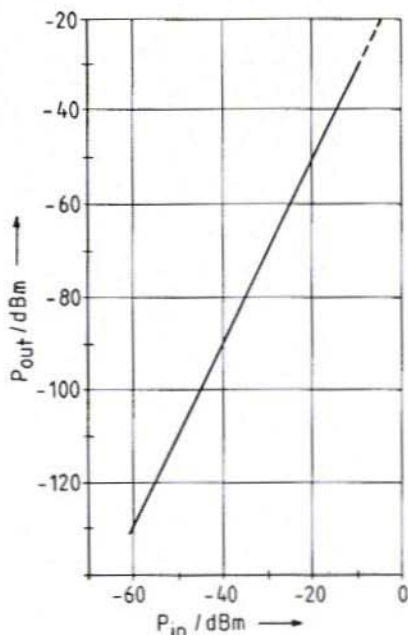
Fig.6: Relationship between Input and Output as per Fig.4

– Some spurious responses at the output do not decrease quadratically. So when the input power decreases, the suppression becomes worse. This can be remedied by an attenuator after the doubler.

4.
APPLICATIONS

I needed a doubler of this type to generate a defined input voltage for the meteorological satellite frequency (approximately 1.7 GHz). As I do not have a GHz signal generator this was the

only way to obtain a half-way exact measurement of the amplification of a converter I had purchased.

If we position a 20dB amplifier and another doubler after such a doubler, the frequency coverage can be even increased by a factor of 4. As the sweep of a frequency modulation is also multiplied accordingly, a circuit of this nature can be suitable for generating unusually large frequency sweeps. If necessary, we can mix back into the original frequency position using a fixed standby frequency.

Quadratic amplitude modulation might be of some interest for special purposes. I am happy to leave that for the future. If you want to avoid quadratic distortion and still modulate AM, a suitable modulator should be added to the doubler (e.g. PAS type from Mini-Circuits).

Because of the conversion losses of approximately 20dB and the differential frequencies at the input and output, the two gates of a doubler of this type are effectively decoupled. Matching to one is not spoilt by mismatching to the other. There is very little feedback to the drive generator.

If the generated signals are used to measure inter-modulation through a doubler, special care must be taken. The non-linearities can generate mixed products, which have nothing to do with the circuit to be tested.

5.
LITERATURE

(1) D. Burchard (1991): Basics of Rectification of Low AC Voltages using Semi-Conductor Diodes; VHF Communications 3/1991, pp.168-174.

(2) D. Burchard (1991): Absolute Calibration of a Noise Source; VHF Communications 2/1992, pp.76-89.

(3) N.N. (1987): Series 2022 Signal Generators Owners Manual, Marconi Instruments, Hertfordshire, U.K.
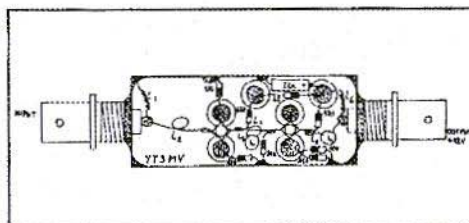
Wolfgang Schneider, DJ8ES

A 28/432 MHz Transverter in Modular Format

For some time, the author has been engaged on intensive work with transverters for frequencies from 28 MHz up to far above 2.5 GHz. Designs and detailed assembly instructions for transverters for various applications, resulting from his activities, have appeared in various publications and have been presented at the VHF Congress in Weinheim. The following article adds a specific assembly proposal to this occasional series: a 28/432 MHz transmitting-receiving converter.

The transverter described below is intended to fulfill certain basic restrictions. The most important factor here is the two-board technique - i.e., the oscillator and the transmitting/ receiving converter are to be separate assemblies. Moreover, it should be possible for the transverter to control a standard power amplifier directly.

Using wide-band amplifier IC's and a ring mixer makes the circuit universally applicable. By simply dimensioning the filters and the crystal oscillator, the tuning range can be matched to the requirements.

1. CIRCUIT DESCRIPTION

A well-known circuit which includes a U310 and has many applications is used as the oscillator to synthesise the frequency. The crystal frequency is 101 MHz.

The 404 MHz required is already available at the output of the subsequent quadrupler. An etched 2-pole filter provides the filtration.

The downstream integrated MSA0404 (IC1) and MSA1104 (IC2) wide-band amplifiers supply the desired output of 50mW. These MMIC's (Monolithic Microwave Integrated Circuit) are available at various amplifications and outputs. Their input and output impedance over a wide tuning range is 50Ω.

The correct dimensioning of the amplifier stages is important here. Thus only the amplification which is actually necessary should be used. Any excess increases the spurious transmissions present to an extent which is out of proportion.
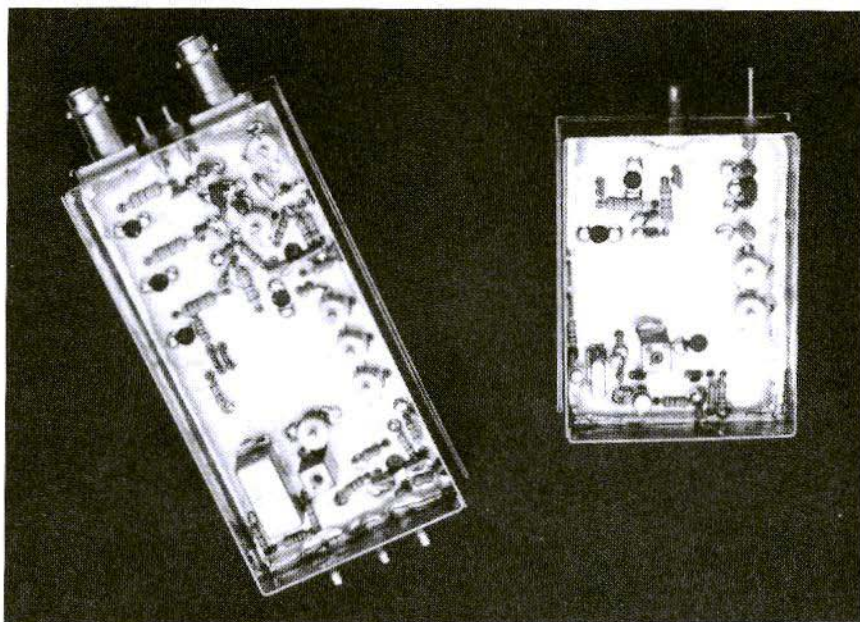
Fig.1: The two Assemblies of the 28/432 MHz Transverter

The SRA1H ring mixer used in the transmitting/receiving converter is suitable for use up to 500 MHz, and requires an oscillator level of 50mW.

The mixer is controlled through an attenuator, which should provide an intermediate-frequency level of no more than 1mW at the ring mixer. The attenuator must be dimensioned on the basis of the output available from the synthesiser.

Table 1 shows the resistance values required for the attenuator in relation to the synthesiser power level. All the values are based on the standard values from the E12 to E24 ranges.

The attenuator simultaneously serves as a wide-band 50Ω termination for the

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

Table 1: All values are in Ω and are
from the E12 or E24 range

ring mixer (SRA1H). Parallel to this, the received signal is matched at high impedance to the CF300 (T3) using L4 and C3. This low-noise transistor stage provides the necessary intermediate-frequency amplification.

Fig.2: Crystal Oscillator with Quadrupler for 404 MHz

The 70cm received signal is passed to the gate of the CF300 (T4) through a Pi filter (aerial impedance 50Ω). The pre-amplifier is directly followed by an MSA0304 amplifier module (IC5). When the receive +12V power supply is fed in, the PIN diode D1 (BA479) is biased on and the signal passed through.

The 3-pole filter for 70cm is an etched version and is used for both receive and transmit branches. The transmit signal initially passes through the filter and diode D2 is biased on. The subsequent amplifier is again constructed using integrated wide-band amplifiers (IC6, IC7, IC8). The combination of MSA0104, MSA0304 and MSA1104

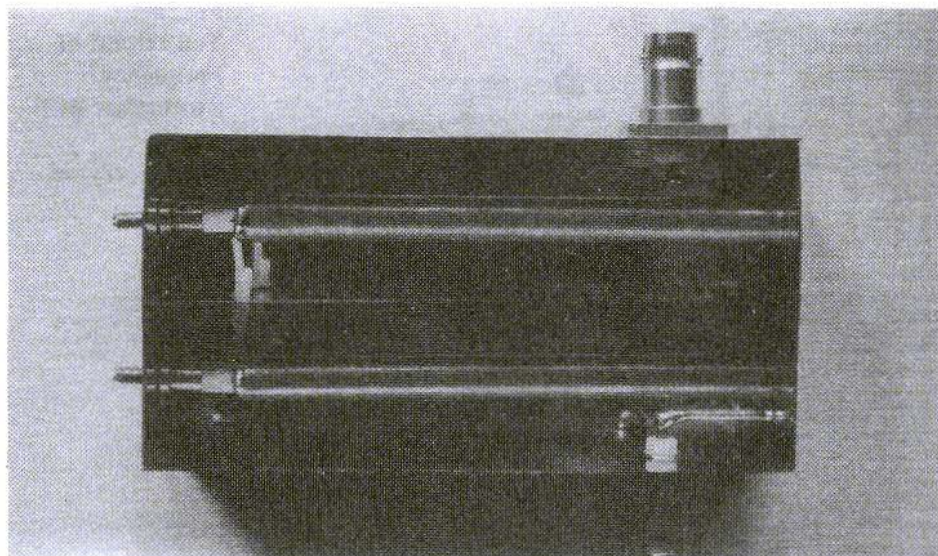Fig.3: Circuit Diagram of the Transmit/Receive Converter

Fig.4: Spurious Transmission/Harmonic Filter for 432 MHz

provides an output of 50mW (+17dBm) with a good 40dB amplification.

In practical operation, such transverters are used with the same driving unit; here an additional filter for harmonics and spurious transmissions is recommended.

Fig.4 shows a possible 2-pole bandpass filter. It can be assembled in air-core construction using a standard tinplate housing measuring 55.5 x 111 x 30 mm. Suitable constructions can also be found in the relevant literature for radio amateurs.

2.
ASSEMBLY INSTRUCTIONS

The 28/432 MHz transverter is divided into two independent assemblies: the oscillator frequency synthesiser and the transmitting-receiving converter. The dimensions of the boards (a double-sided coated epoxy board measuring 54mm x 72mm for the oscillator frequency synthesiser and a 54mm x 108mm printed circuit board for the transmission/reception section) allow for incorporation into a standard tinplate housing.

After being cut to size, the boards first undergo silvering and are then drilled. Suitable holes are drilled for the stripline transistors and the wide-band amplifiers; these components are thus mounted in the boards' surfaces.

The holes for the crystal, the trimmers, the Neosid coils, etc. are drilled on the earth side of the boards (fully-coated side) using a 2.5mm drill.

Suitable slots are to be sawn out in the printed circuit board for the SMC or

Fig.5a:
Top Layout of Frequency Synthesiser PCB

quarz = crystal

Fig.5b:
Bottom Layout of Frequency Synthesiser PCB

BNC bushes. The same applies to the pick-off capacitors with 1nF at the source connection of the amplifier transistors, T3 and T4.

Once these preliminary steps have been completed, the board can be sprayed with solderable lacquer.

The board is inserted in such a way that the connector pins are surface-mounted (cut off projecting Teflon collars with a knife first). When the "mechanically large" components (filter coils, trimmers, crystal and ring mixer) have been provisionally inserted in their positions,

it must still be possible for the housing cover to be placed on top without any obstruction.

When the individual boards have been soldered to the sides of the housing, the actual assembly can be undertaken.

2.1. Oscillator Frequency Synthesiser Component List

IC1	MSA0404 (Avantek)
IC2	MSA1104 (Avantek)
IC3	78L09 voltage regulator
T1	U310 (Siliconix)
T2	BFR90a (Valvo)
L1	Neosid BV5061 0.1μH blue/brown coil
L2, L3	λ/4 stripline, etched
C1, C2	6pF foil trimmer (grey), 7.5mm grid (Valvo)
Q	101 MHz crystal, HC18U or HC25U
1 x	Carbon film: 180Ω, 0.5 W
1 x	Carbon film: 120Ω, 0.5 W
1 x	SMC or BNC flanged socket (UG-290 A/U)
1 x	Teflon bushing
1 x	Tinplate housing: 55.5mm x 74mm x 30mm
2 x	1nF trapezoid capacitor
2 x	10μF/20 V tantalum capacitor

Ceramic Capacitors (2.5mm grid)	Resistors (1/8W/10mm)
1 x 2.7pF	1 x 100Ω
1 x 18nF	1 x 220Ω
1 x 82pF	1 x 2.2kΩ
6 x 1nF	1 x 22kΩ
1 x 100nF	

and in SMD format (model 1206 or 0805) 2 x 1nF

2.2. Transmitting/Receiving Converter Component List

IC4	78L09 voltage regulator
IC6	MSA0104 (Avantek)
IC5, IC7	MSA0304 (Avantek)
IC8	MSA1104 (Avantek)
T3, T4	CF300 (Telefunken)
D1, D2	PIN diode BA479
L4	BV5048 Neosid coil, 1μH, yellow/grey
L5, L6, L7	λ/4 stripline, etched
L8	1.5 turns, 1mm CuAg wire
C3	30pF foil trimmer (red) 7.5mm grid (Valvo)
C4, C5, C6	6pF foil trimmer (grey) 7.5mm grid (Valvo)
C7, C8	6pF foil trimmer (grey) 7.5mm grid (Valvo)
R1, R2, R3	Attenuator, see Table 1
1 x	Carbon film: 120Ω, 0.5W
1 x	Carbon film: 150Ω, 0.5W
1 x	Carbon film: 220Ω, 0.5W
1 x	Carbon film: 270Ω, 5W
5 x	SMC sockets (some of which may be BNC flanged: UG-290 A/U) (see photo of specimen assembly)
2 x	Teflon bushing
1 x	Tinplate housing 55.5 x 111 x 30mm
4 x	1nF trapezoid capacitor
2 x	0.1μH choke, 10mm grid, axial
2 x	10μH choke, 10mm grid, axial
1 x	10μF/20 V tantalum

Ceramic Capacitors (2.5mm grid)	Resistors (1/8W/10mm)
1 x 10pF	2 x 220Ω
12 x 1nF	1 x 270Ω
	1 x 680Ω
	4 x 10kΩ

and in SMD format (model 1206 or 0805) 6 x 1nF

3.
COMMISSIONING

When the equipment is used for the first time (and also calibrated), the following test equipment should be available: Multimeter, Frequency counter, Wattmeter and Received signal (e.g. beacon).

The assemblies are put into operation one after another.

Firstly, the oscillator is set to its operating frequency of 101 MHz with by adjusting self-inductive coil, L1. The onset of oscillation results in a slight increase in the collector current of T2 (monitoring voltage drop across 100Ω resistor). A frequency counter is loosely coupled and the oscillator frequency measured.

The 2-pole filter behind the quadrupler T2 (BFR90a) filters the useful frequency, 404 MHz. To this end, the two trimmers (C1, C2) are to be reciprocally set to the maximum level. The wiring diagram shows the approximate trimmer positions.

The oscillator frequency synthesiser assembly supplies an output of at least 50mW. The current consumption for an operating voltage of +12V is about 120mA.

The transmit branch of the transmitting/receiving converter is put into operation first. Only the 3-pole filter (C4, C5, C6) is to be calibrated here.

A current of approximately 130mA should be measured for an operating voltage of +12V. This is already an indication that the amplifier stages are operating satisfactorily. If the input attenuator is dimensioned as described in Table 1, an output greater than 50mW can be expected. Possible spurious transmissions (oscillator, image frequency, etc.) are suppressed better than 50dB here.

The receiver can be calibrated directly, using a strong received signal (e.g. a beacon). Because the same filter is used as in the transmit branch, the beacon signal should be audible immediately.

A further filter is mounted at the intermediate-frequency level (28 MHz) after the mixer. Here the trimmer, C3, should be adjusted to give the maximum signal. The directly connected parallel circuit naturally influences the transmit branch. However, with appropriate reserve capacity this influence is not brought to bear.

Optimising the signal-to-noise ratio (Pi filter with C7, C8 and L8 at the receiver input) completes the calibration.

The current consumption of the receive branch is very low (only 50mA). The noise factor is app. 2dB, with a conversion gain of the order of 30dB.
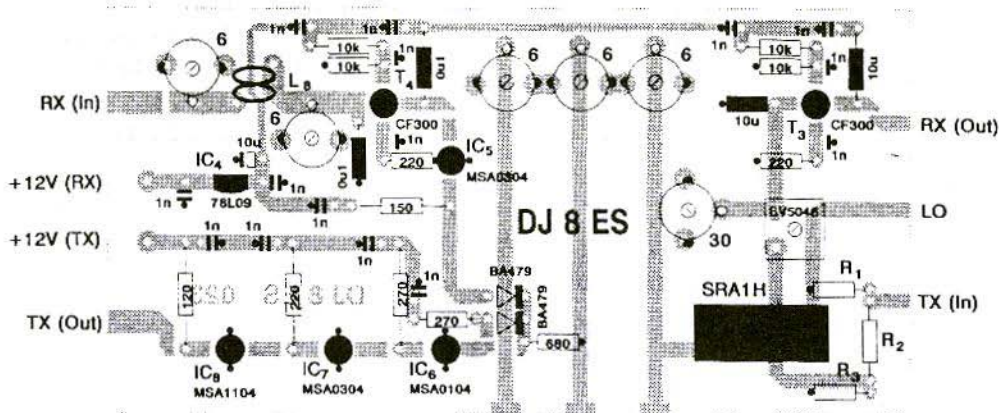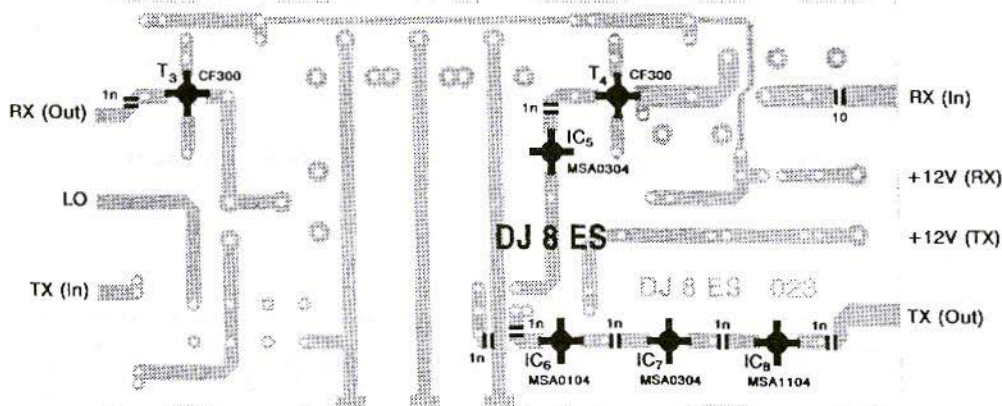
Fig.6a: Component Side of Transmit/Receive Unit

Fig.6b: Semiconductors and Coupling Capacitors on Copper Foil Side

4.
CONCLUSION

The author uses the transverter described in association with an external pre-amplifier and a power amplifier. Modern hybrid modules are just the thing for amplifier stages. The output signal can be increased from 50mW to 10 - 20W in one go, using such components.

Fig.7, for example, shows the circuit for such a module (type M55716 from Mitsubishi). A 2C39 type valve high-level PA can be fully driven by means of this 10W output.
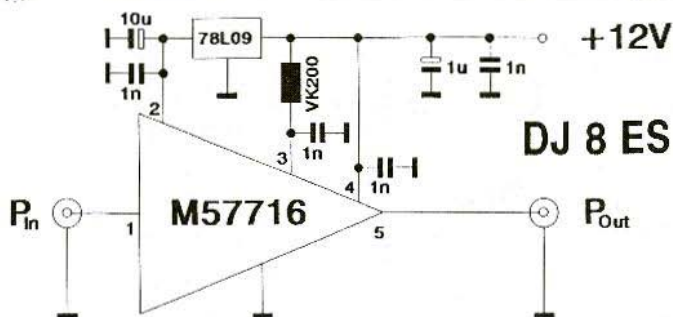
Fig.7:
Circuit Diagram
of the Hybrid
Amplifier Module
for 432 MHz

DJ 8 ES

5.
LITERATURE

(1) Wolfgang Schneider, DJ8ES:
Transverters for the 70, 23 or 13
cm. Tuning Areas.
1992 Weinheim Congress
Proceedings

(2) Wolfgang Schneider, DJ8ES:
28/432 MHz Transverter:
Instructions, Tips and Improvements: 1993 Weinheim Congress
Proceedings

Wolfgang Schneider, DJ8ES

A 28/50 MHz Transverter

With the de-controlling of the 6m band as part of a large-scale experiment, even local radio amateurs can now take an active part in events. In connection with this topical development, the present article describes a 28/50 MHz transverter.

A transverter for the 6m band can be produced on the basis of the tried and trusted concept of the 28/144 MHz transverter, which was described in VHF Communications 4/1993 (2). All that is required is modification of the components determining the frequency. Essentially, the oscillator and the filter need to be adapted.

1.
CLARIFICATIONS OF THE CIRCUIT

Fig.1 shows the complete circuit for the 28/50 MHz transverter. Its function has already been described in detail with regard to the 2m version (see issue 4/1993). The instructions for assembly and calibration given there also apply here, of course.

The entire circuit can be assembled on the DJ8ES 019 printed circuit board of the 2m transverter. Only the range of the pi filter at the receiver input needs to be altered. Fig.2 gives the details.

All the coils and some of the capacitors have new values due to the lower frequency range.

To make it easier to produce the 6m version of the transverter, the layout of the DJ8ES 019 printed circuit board with the appropriate components for the 50 MHz version is illustrated in Fig.3 and Fig.4.

1.1.   Transverter Component List

IC1	TA78L diameter 9F voltage regulator (SMD)
IC2, IC5	MSA1104 (Avantek)
IC3	MSA0104 (Avantek)
IC4	MSA0304 (Avantek)
T1	U310 (Siliconix)
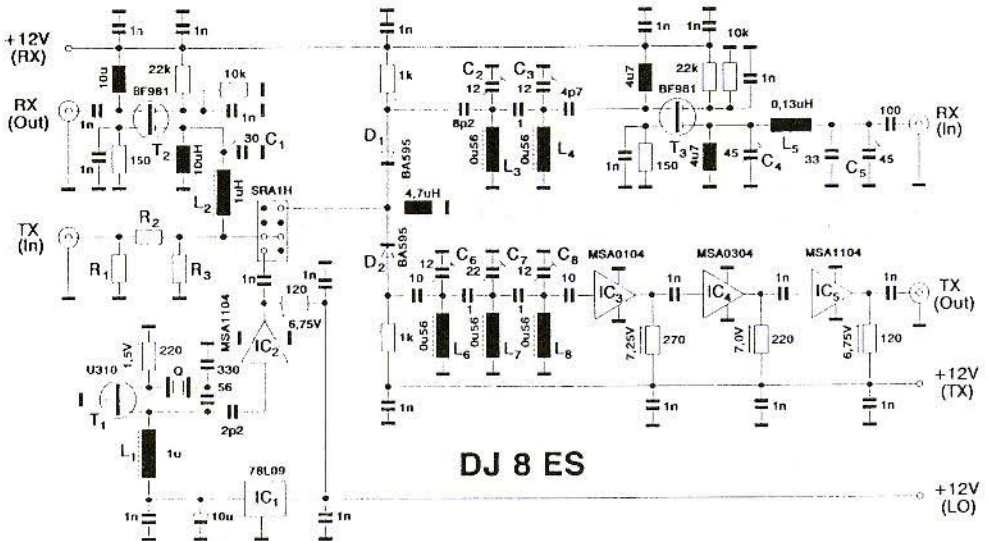
Fig.1: The Transverter Circuit Modified for 50 MHz

T2, T3	BF981 (Siemens)	1 x	220Ω / 0.5W Carbon film
D1, D2	BA595 PIN diode (SMD)	1 x	270Ω / 0.5 W Carbon layer
L1, L2	BV5048 Neosid coil,	4 x	BNC flanged connector
	1µH, yellow/grey		(UG-290 A/U)
L3, L4	BV5036 Neosid coil,	3 x	Teflon bushing
	0.58µH, orange/blue	1 x	Tinplate housing
L5	BV5063 Neosid coil,		55.5 x 111x 30 mm
	0.58µH, blue-orange	9 x	1.5mm dia. Copper rivets
L6, L7,	BV5063 Neosid coil,		
L8	0.58µH, orange/blue	All other components as SMD format:	
C1	30pF foil trimmer (red)	3 x	Choke, 4.7µH
	7.5 mm grid (Valvo)	2 x	Choke, 10µH
C2, C3	12pF foil trimmer (yellow)	1 x	10µF / 20VTantalum
	7.5 mm grid (Valvo)		

Ceramic capacitors   Resistors

C4, C5	45pF foil trimmer (violet)
	7.5 mm grid (Valvo)
C6, C8	12pF foil trimmer (yellow)
	7.5 mm grid (Valvo)
C7	22pF foil trimmer (green)
	7.5 mm grid (Valvo)
Q	22 MHz crystal,
	HC18U or HC25U
1 x	SRA1H ring mixer
2 x	120Ω / 0.5W Carbon film

3 x 1pF	1 x 150Ω
1 x 2.2pF	2 x 220Ω
1 x 4.7pF	2 x 1kΩ
1 x 8.2 pF	2 x 10kΩ
2 x 10nF	2 x 22κΩ
1 x 33pF	
1 x 56pF	
1 x 330pF	
17 x 1nF	

2.
HYBRID AMPLIFIER

An M57735 hybrid module is used in the separately assembled amplifier stage for the 6m band. This module has been specially developed for the tuning area around 50 MHz. About 10W can be expected at the output of the PA with the output pre-set from the transverter to 50mW.

The downstream low-pass filter provides the harmonic filtration required. Only components of appropriate quality (e.g. air-core coils and air-spaced trimmers) should be used here.

The 50 MHz amplifier can also be assembled on the printed circuit board of the 2m version of the DJ8ES 020.

2.1. Hybrid Amplifier Component List

IC1	M57735 (Mitsubishi)
IC2	TA78L09F
	voltage regulator (SMD)
L1, L2	0.5µH air-core coil
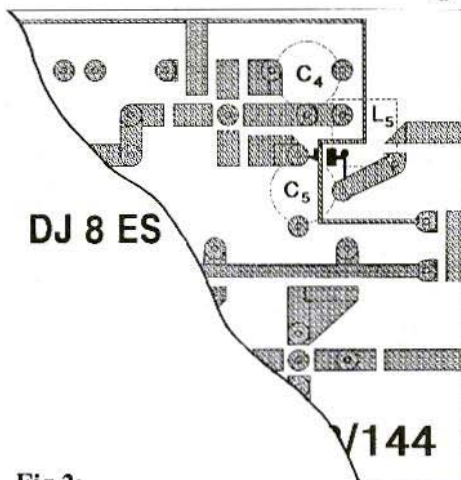
Fig.2:
Modification of the
Receiver Input

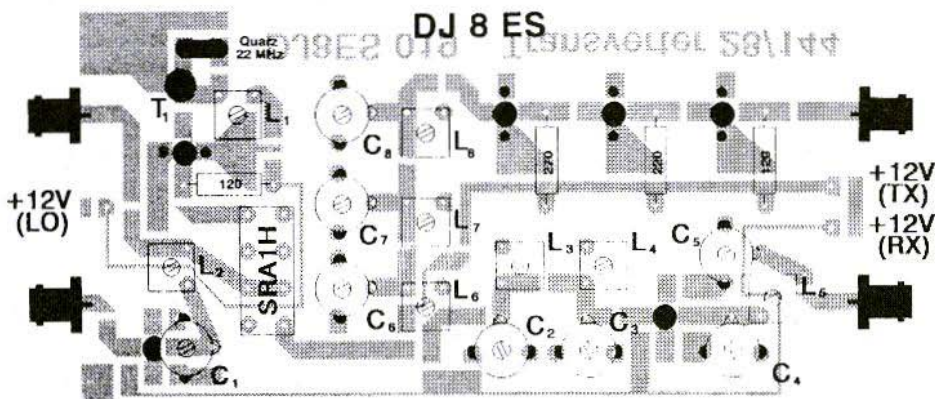|	
C1, C2	33pF trimmer
	with soldering lugs
1 x	VK200 UKW
	wide-band choke
1 x	1nF through-hole type
	capacitor, solderable
2 x	BNC flanged bush
	(UG-290 A/U)
1 x	Tinplate housing
	55.5 x 111 x 30 mm

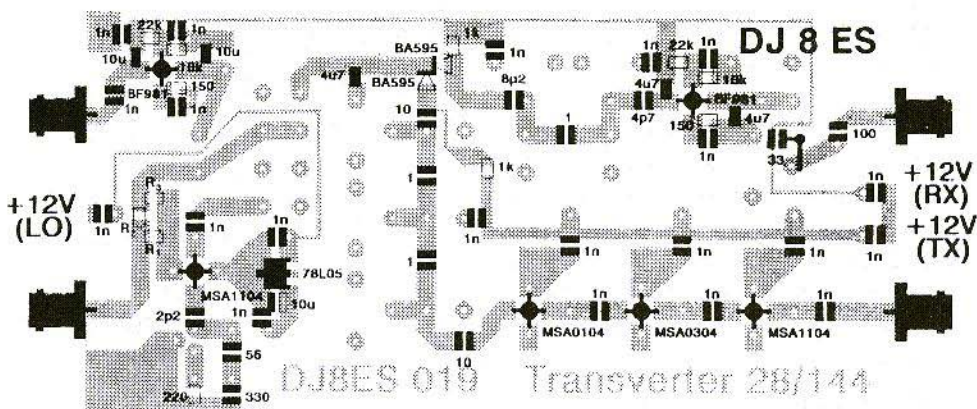Fig.3: The 50 MHz Transverter viewed from the Component Side

Fig.4: The 50 MHz Transverter viewed from the SMD Component Side

3. CONCLUSION

All other components in SMD format:

1 x	1µF / 20V Tantalum
1 x	10µF / 20V Tantalum
1 x	3.3pF, ATC chip
2 x	82pF, ATC chip
2 x	82pF, ATC chip
3 x	1nF, ceramic capacitor

The advantages of the 50Ω technology are clear and obvious. With only slight changes, an existing circuit can be used for a completely different tuning range. Only the components determining the frequency need to be changed for the new frequency band.

The 25W ERP permitted can certainly be attained with an output of around 10W and a standard 4-element or 5-element Yagi. We hope that the great experiment will end in positive results
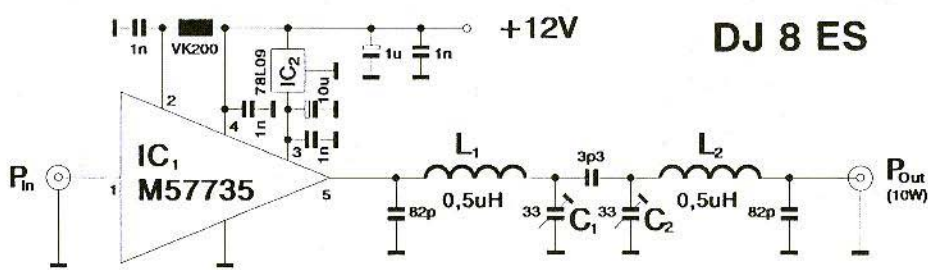

Fig.5: The 50 MHz Amplifier with Downstream Low-Pass Filter

and that, here in this interesting frequency band too, all options will remain open to us.

4.
LITERATURE

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(1) Wilhelm Schürings, DK4 TJ and Wolfgang Schneider, DJ8ES: Universal Transverter Concept for 28, 50 and 144 MHz VHF Communications 3/1991, pp. 175-187

(2) Wolfgang Schneider, DJ8ES: 28/144 MHz Transverter VHF Communications 4/1993, pp. 221-226

(3) Wolfgang Schneider, DJ8 ES: Hybrid Amplifier for 144 MHz VHF Communications 1/1994, pp. 56-61

(4) Mitsubishi: Data Sheet M57735 Mitsubishi RF Power Modules

Hermann Hagn, DK8CI

Receiving System Parameter Measurements using Radio Stars

The relationship between the antenna gain (G) and the system noise temperature (T_{sys}) is experimentally determined using a complete installation. This relationship (G/T_{sys}) includes all variables, such as the noise factor of the pre-amplifier, the gain, or more precisely the illuminated surface of the antenna, any possible errors in matching the exciter to the pre-amplifier, and the amounts of noise contributed by unavoidable resistive losses due to the antenna structure. The value which we obtain is determined from the measured increase in the signal from a known radio source. The value for the flux of this source can be taken from tables.

based on the researches of Kirchhoff, Stefan, Boltzmann, Wien and Rayleigh, and can be simply summarised as follows: any body which is at a finite temperature emits electromagnetic waves. The spectrum emitted is continuous and extends over an extremely wide frequency range. It has a
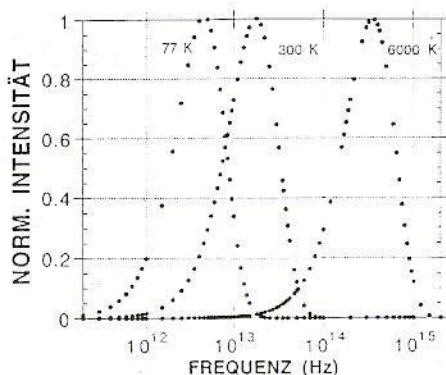

Fig1: **Position of Maximum Intensity Levels of Black Body Radiators with 77K, 300K and 600K. In reality the Intensity Levels are many orders of magnitude apart.**
norm. intensitat = **normal intensity**

1.
HISTORY & PRINCIPLES

In the year 1900, Max Planck (1858 - 1947) introduced his new radiation formula (today it is known as Planck's Radiation Law). Planck's findings were

maximum, which is dependent only on the temperature of the body (Fig.1). The frequency ranges which are a long way away from the maximum can be described by approximations to Planck's complicated radiation law. The "long-wave" frequency range which is of interest to us here is reflected in the Rayleigh-Jeans approximation. In this range, the intensity of the radiation increases, as a very good approximation, in a quadratic relationship with the frequency (Fig.2). If all the variables for the receiving equipment, such as the apex angle and polarisation of the antenna, are inserted into the Rayleigh-Jeans approximation, then for the power picked up by the receiving equipment we obtain the simple formula (1, 2, 3, 4):

$$P = k \cdot T \cdot B$$

where P is in W; $k = 1.38.10^{-23}$J/K; T in K; B in Hz (band width)

In 1928, Johnson published the results (5) of research into Ohmic resistances. In the same year, Nyquist published (6) the associated theory. The findings from both pieces of work state that charged particles execute movements due to inter-action with the thermally
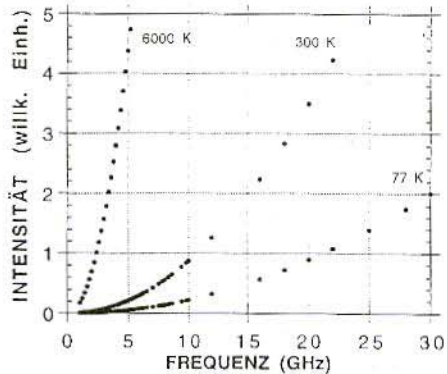

Fig.2: **Actual Intensity Ratios in Longwave Frequency Range of interest to us**
intensitat (willke. einh) = intensity (arbitrary units)

excited grid (in metals). These are irregular, come to a halt when the absolute zero point is reached, and increase as the temperature rises. But a charged particle movement indicates that a current is flowing. Since the charged particle movement is irregular, it occurs at no specific frequency. The frequency band encompasses the range from approximately 1 Hz right up to the highest frequencies.

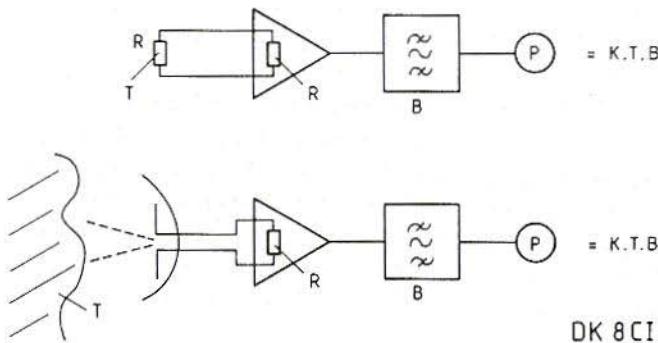Only in the 100 GHz range and at very low temperatures do certain corrections

Fig.3:
Calibration of Load Resistor using matched loaded Antenna during Power Matching
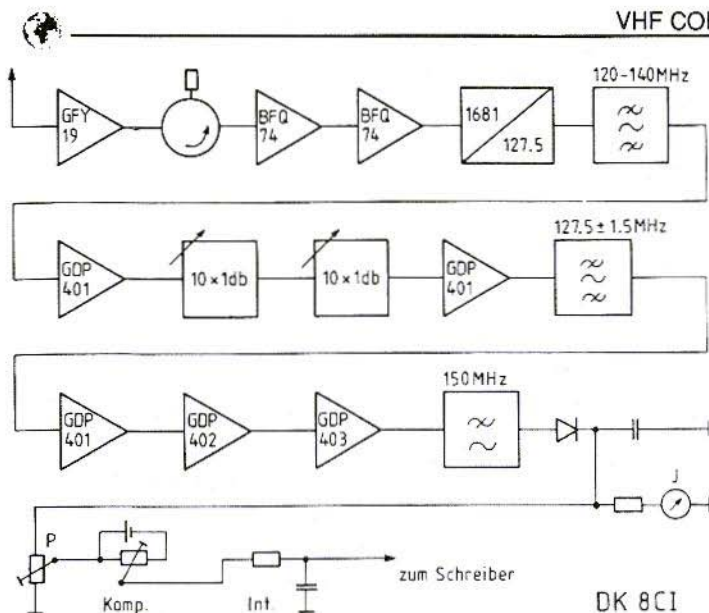($R_1 = R_2$) leads to Equal Noise Output)

DK 8 CI

Fig.4:
Block Diagram of
Receiver for
1681 MHz

komp = comparator
zum schreiber =
to recorder

DK 8CI

become necessary (7). The thermal noise power generated by the flowing of this current in the resistance was described by Nyquist as:

$$P = 4k \cdot T \cdot B$$

where P is in W; $k = 1.38.10^{-23}$J/K; T in K; B in Hz (band width)

From this, we obtain an effective noise voltage for the uncharged resistor of:

$$U_{eff} = \sqrt{(P \cdot R)} = \sqrt{(4kT \cdot B \cdot R)}$$

The maximum power which can be taken from this resistor, R, with the terminal voltage, U_{eff}, is $P = k.T.B$. (with a power match of $R_i = R_a$). The resistance, R_a, acting as a charge in our case is the input resistor of the pre-amplifier.

Result: Whether we connect up a resistance of the same size, with the temperature T, to our pre-amplifier with the input resistor, R, or "look at" a body with the temperature T using the

loaded, polarised antenna, this leads in both cases to the same noise output at the pre-amplifier output. The radiating body must "illuminate" the antenna completely (Fig.3).

2.
RECEIVING INSTALLATION FOR 1681 MHz

The antenna consists of a 3m parabolic type reflector with a 75cm focal distance. The exciter is a "beer can radiator" (8). A GaAsFET amplifier with a CFY19 is mounted directly on the exciter (Fig.4). A circulator wired up as an isolator is fitted at the output. This is followed by a two-stage amplifier with bi-polar transistors (BFQ 74). The total amplification for the two amplifiers amounts to 42dB. The cable leading to the receiver (RG 213) has

about 10dB attenuation. The receiver consists of a converter (9) with a broad-band output (120 - 140 MHz). at the output of this frequency converter, a filter suppresses all frequencies outside the 120 - 140 MHz range which arise in the frequency converter. This is followed by an Avantek GDP401 broad-band amplifier (14dB) and two switchable 10 x 1dB attenuators to set the measurement range and for calibration. There follow another GDP401, a 127.5 MHz ±1.5 MHz filter and a three-stage amplifier array - GDP401,

402, 403 (40dB amplification). Following these amplifiers, a bandpass filter (limiting frequency 150 MHz) suppresses any harmonics of 127.5 MHz which may arise. A low-barrier diode is used as a rectifier.

The rectifier is followed by the instrument, I, for monitoring the total amplification and the potentiometer, P, for setting the input DC voltage. The DC voltage generated by the receiving installation's internal noise is much greater than the DC voltage from the noise-form signal of a radio source. (The single exception is the Sun.) To be able to amplify these very small changes, it is advantageous to calibrate the DC voltage obtained from the installation's inherent noise with a counter-voltage of the same size to zero in the compensator and to continue working with only the small changes. In the subsequent integrator (R-C unit), the mean value for these small changes is formed. The larger the time constant, the more accurately this is done. But the intermediate-frequency band width and the receiver noise temperature, T_r, also have an effect. Dicke (2) gives the following relationship for the smallest measurable temperature change (radio meter formula):

$$T_{min} = \frac{T_r}{\sqrt{(B \cdot \tau)}} \; [K]$$

where T is in K; B in Hz; τ in seconds.

Fig.5: Flux of strongest source plotted against Frequency. The Vertical Line at 1.7 GHz corresponds to our Observation Frequency

The relationships given here are described in great detail in (1). The setting and calibration of the installation and the evaluation of the data obtained are described in the appendix.
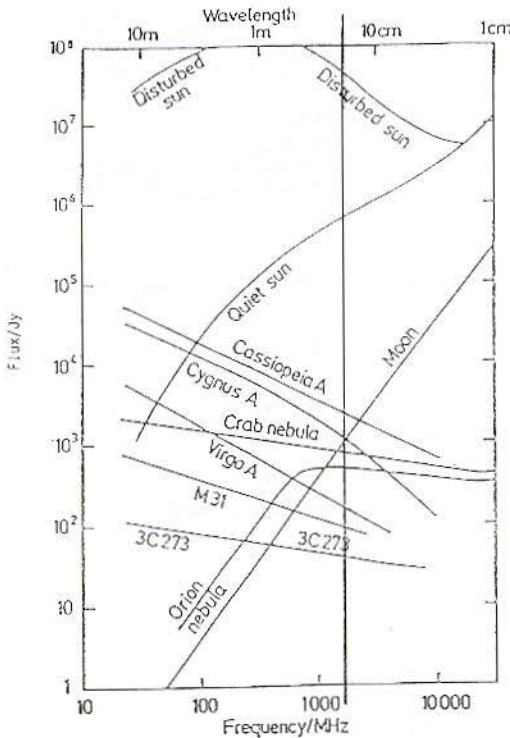
Name	Right Ascension [h]	[m]	Declination [°]	[']	Flux* [J_Y]	Identification - comments
(a) HII-Regions/Emission Nebula						
W3	02	22.7	61	51	170	IC 1795
Ori A	05	32.8	-05	27	520	NGC 1976, 3C145
Ori B	05	38.4	-01	54	95	NGC 2024, 3C147.1
W28	17	58.2	-23	22	360	M20
W29	18	01.0	-24	22	260	M8, NGC 6523
W33	18	10.4	-18	00	190	IC 4701
W37	18	16.3	-13	45	260	M16, NGC 6611
W38	18	17.8	-16	09	1060	M17, NGC 6618
W49A/B	19	08.2	-09	02	75	NRAQ 598 + 3C398
W51	19	20.8	+14	08	710	3C400
(b) Remains of Supernova(SNR)						
3C10	00	22.6	+63	52	44	Tychos SNR of 1572
3C58	02	01.9	+64	35	34	
Tau A	05	31.5	+21	59	875	M1, Crab Neb, SNR/1054
3C157	06	14.3	+22	36	190	1C443
3C358	17	27.7	-21	27	15	Keplers SNR of 1604
W41	18	31.6	-08	57	75	
3C391	18	46.8	-00	59	20	NRAQ 583
W44	18	53.6	+01	15	171	3C392
W78	20	48.2	+29	30	90	Cirrus Neb., Cygnus Loop
Cas A	23	21.1	+58	33	2480	3C461, SNR circa 1700
(c) Radio Galaxies						
3C84	03	16.5	+41	20	14	NGC1275, Seyfert Galaxy
3C123	04	33.9	+29	34	47	
3C218	09	15.7	-11	53	43	D Galaxy
3C270	12	16.8	+06	06	18	E Galaxy
Vir A	12	28.3	+12	40	198	M87, E Galaxy
3C295	14	09.6	+52	26	23	D Galaxy
Her A	16	48.7	+05	05	45	3C348, D Galaxy
3C353	17	17.9	-00	56	57	D Galaxy
3C390.3	18	45.9	+79	43	12	N Galaxy
Cyg A	19	57.5	+40	36	1495	3C405, D Galaxy
(d) Quasars						
3C48	01	34.8	+32	54	16	
3C138	05	18.3	+16	35	10	
3C147	05	38.7	+49	50	23	
3C196	08	10.0	+48	22	14	
3C273	12	26.6	+02	20	46	
3C279	12	53.6	-05	31	11	
3C286	13	28.8	+30	46	15	
3C309.1	14	59.0	+71	52	9	
3C380	18	28.2	+48	43	14	
3C454.3	22	51.5	+15	53	11	

* for 20cm Wavelength

3.
SELECTION OF RADIO SOURCES

The most powerful source in the sky, and for frequencies above 100 MHz, is the Sun. However, in the frequency range going up to 20 GHz, its flux is subject to very strong fluctuations, so that it can be ruled out as a calibration source. The second most powerful source is Cassiopeia (Cas A), the remains of a supernova. Its flux is known very precisely over a wide frequency range (20 MHz - 10 GHz), and it is thus very suitable for calibration purposes. But a slight disadvantage is the position of this source. It stands so far to the North that the equipment to be measured must be swung over one's head, or must be movable within the North-South azimuth. The third most powerful source in the frequency range up to 1000 MHz is Cygnus A (with X). Cygnus A is a radio galaxy and CyX is an emission cloud in our galaxy. The two objects can be separated only using very large antenna (diameter >10m) at below 1,000 MHz. But up to 13cm CyA + CyX is a very good calibration source, though it too stands very high in the sky. From 2000 MHz onwards, the Moon is also a strong radiator, but you have to take into account that its intensity fluctuates as its distance from the Earth varies. From 3000 MHz onwards, the Moon is the most powerful source after the Sun.

The fourth most powerful source, which is also well-known over a very wide frequency range (20 MHz - 30 GHz), is the Crab Nebula (M1). Its flux is almost constant over the entire frequency range. M1 also represents the remains of a supernova. Chinese astronomers observed the explosion in the year 1054. In addition, from 1000 MHz onwards, the Orion Nebula (M42) and Virgo A, a radio galaxy, are also suitable for measurements.

We selected the Crab Nebula and Virgo A for our measurements. 3C157 (remains of a supernova) was also measured with the Crab Nebula (Fig.5, Table).

One really good calibration source for smaller antenna (1 - 3m diameter) is the Milky Way, but make sure you always observe over the horizon at the same elevation, as the flux changes with the elevation. The flux also depends on the apex angle of the antenna, as we are dealing with an expanded source. For antenna of up to 3m diameter and for frequencies above 1000 MHz, the Milky Way is the most powerful source after the Sun.

4.
EVALUATION OF MEASURED DATA

The following relationships (formulae) are used: effective antenna area, gain over isotropic radiator (1)

$$A_{eff} = \frac{G_{iso} \cdot \lambda^2}{4\pi} \qquad (1)$$

This formula can also be used to calculate the "effective area" of a Yagi antenna. Antenna temperature, Flux:

$$Ta = \frac{S' \cdot A_{eff}}{k} \quad (2)$$

S is the total flux for an unpolarised radiation source. With a polarised antenna, only half the flux is measured (1)

$$S' = S/2; \; S = \frac{W}{m^2 Hz} \; ; \; 1 \, J_y = \frac{10^{-26}W}{m^2 Hz}$$

(Jy → Jansky)

antenna gain: log, lin $\quad G_{dB} = 10 \cdot \log G$

Signal rise in dB when passing through a source:

$$dB = 10 \log \frac{T_{sys} + T_A}{T_{sys}} \rightarrow$$

$$T_{sys} = \frac{T_A}{10^{} - 1} \quad (3)$$

By T_{sys}, we understand the noise temperature of the receiving system + the temperature of the source-free sky for the antenna setting in question.

From (1), we obtain $G_{iso} = \frac{A_{eff} \cdot 4\pi}{\lambda^2}$

We want to calculate G_{iso}/T_{sys} and we obtain:

$$G_{iso}/T_{sys} = \frac{A_{eff} \cdot 4\pi \, (10^{dB/10} - 1)}{\lambda^2 \cdot T_A}$$

T_A inserted from (2):

$$G_{iso}/T_{sys} = \frac{A_{eff} \cdot 4\pi \, (10^{dB/10} - 1) \cdot k}{\lambda^2 \cdot S' \cdot A_{eff}}$$

S' = S/2 inserted:

$$G_{iso}/T_{sys} = \frac{2.4\pi \, (10^{db/10} - 1) \cdot k}{\lambda^2 \cdot S}$$

dB: measured signal increase, without dimension

$$k = 1.38 \cdot 10^{-23} J/K, \; S = \frac{W}{m^2 Hz}$$

(from table)

λ = wavelength in m, G = without dimension

J = W . sec = W . 1/Hz,
T_{sys} = K (Kelvin)

The relationship G_{iso}/T_{sys} is determined from the measured signal increase (in dB over the system temperature, T_{sys}), the known flux, S (from the table) and the wavelength, λ (in m) with the help of the given formula. For the GHz and Jy (Jansky) values mostly required, the formula is:

$$G_{iso}/T_{sys} = 0.385 \cdot 10^6 \frac{(10^{dB/10} - 1) \cdot f^2_{(GHz)}}{S_{(Jy)}}$$

It is amazing how accurately the fluxes from other sources can be determined with the help of this formula - once G_{iso}/T_{sys} has been determined. (See evaluation of readings).
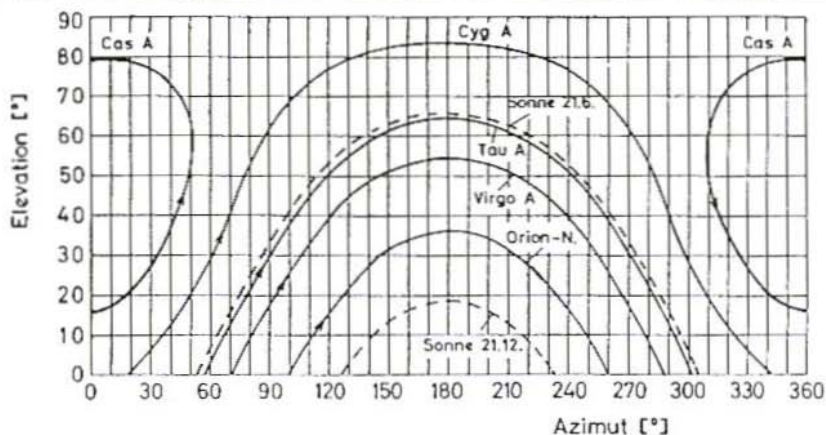
Fig.6:   Cosmic Noise Sources: Curves for Antenna Directions

5.
APPENDIX

Setting and calibration of our receiving equipment. The antenna is set to the height of the radio star to be measured (Fig.6). The second 10 x 1dB attenuator is set to 10dB and is adjusted, with the first attenuator, to a DC output voltage of approximately 200mV at the instrument, I. The potentiometer, P, is set to the top limit, and zero output voltage is set using the compensator.

Naturally, the pre-condition here is that the antenna is orientated towards the cold, source-free sky. One very suitable compensation voltage source is a large mono-cell, which is charged at app. 100µA. If we now reduce the attenuation on one of the two switchable attenuators by 1 dB, then we obtain a DC output voltage of slightly greater than 20mV. A level of precisely 20mV is set, using the potentiometer, P, and the attenuators then reset. The two latter calibration procedures may have

to be repeated once or twice, until a 1dB attenuation change corresponds to exactly 20mV at the output. The recorder is set to a full-scale deflection of 2mV, which corresponds to a signal increase of app. 0.1dB above the system temperature of the installation. The equipment is ready for measurement.

6.
EVALUATION OF READINGS

Fig.7 shows a beautiful passage through the Crab Nebula and the nearby radio source 3 C 157. The second curve, at approximately 35 scale divisions, originates from a 1.8m antenna, which was orientated on a point near the Crab Nebula and served as a reference.

The signal increase during the passage through the Crab Nebula is 1mV, corresponding to 0.0528dB above the system temperature (at 20mV = 1dB). That gives a G/T_{sys} of 15.30.
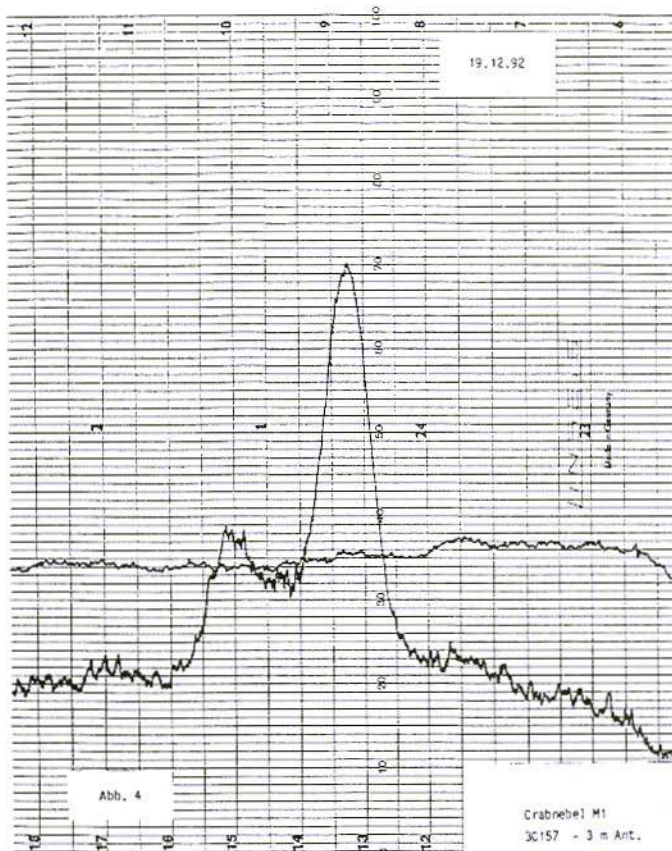
Fig.7: Transit of the Crab Nebula M1 and nearby Radio Source 3C157

If we insert this value, with the flux of Virgo A, into the formula G/T_{sys}, we obtain a signal increase of 0.012dB (corresponds to 0.027mV), which correlates very well with the measured value from Fig.8.

The manufacturers (Telefunken) specify the following mid-band gain values for the antenna, based on isotropic radiators:

1.7 - 1.9 GHz 32.7dB
1.9 - 2.1 GHz 33.7dB
2.1 - 2.3 GHz 34.9dB

so that no really big errors will be made if a gain of 31.7dB is assumed for the range between 1.5 and 1.7 GHz. For the measuring frequency 1681 MHz, we then obtain

$$32.2\text{dBi} \cong 1659.6$$

From this, we can calculate the effective antenna area:

$$A_{eff} = (1659.6 \cdot \lambda^2)/4\pi = 4.21\text{m}^2$$

If we compare this with the geometrical area of the antenna, which has a 3m diameter, then we obtain an illumina-
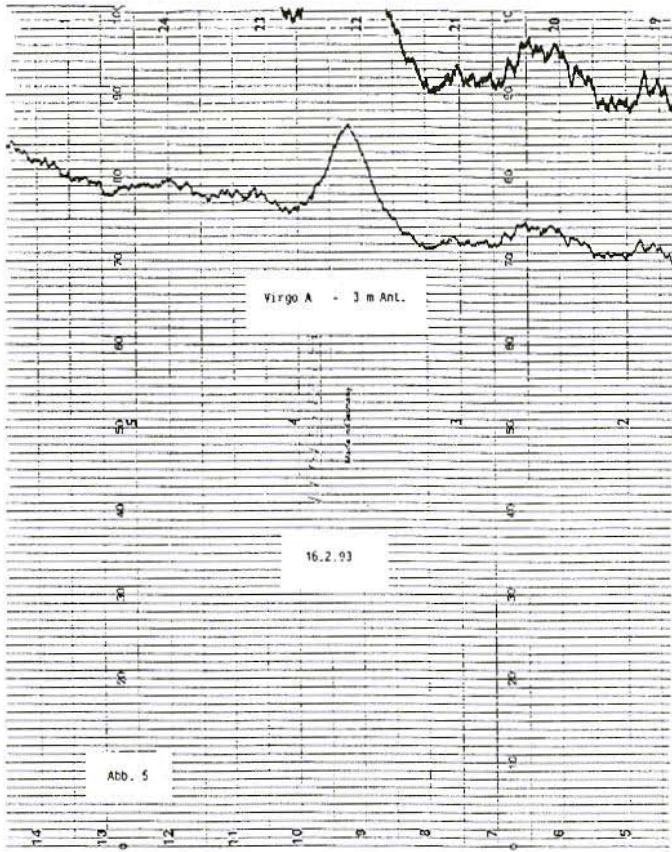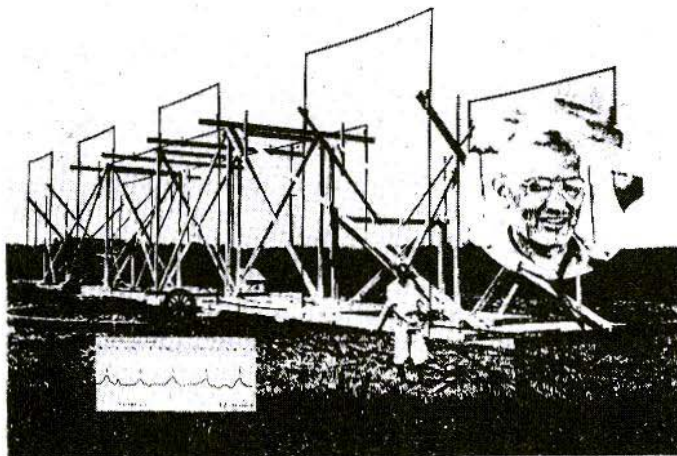
Fig.8:
Virgo A measured
using a 3m Antenna

Fig.9:
With this Antenna
Karl Guthe Jansky
'opened' the 'Radio
Window' to the
Universe in 1932

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tion of 59%. The antenna's minor lobes are a clear 20dB below the main lobe, so that is a reasonable value.

G/T_{sys} = 15.30 (from previous calculation) and G = 1659.6 give T_{sys} = 108.4 K. This value is derived from the receiver temperature, the temperature contribution of the minor lobes, the antenna temperature fraction (Ohmic loss), the atmosphere, radio sources unresolved by our antenna, and the background radiation (2.7 K, measured by Penzias and Wilson). All these fractions, minus the receiver temperature, give a value of about 40 ± 5 K for our antenna. This value was determined using "hot-cold" measurements. A T_0 moving load served as the hot source, with the source-free sky as the cold source. So we were left with approximately 68 K = 0.92dB for the receiver. In measurements of this nature, great emphasis must be laid on the constancy of the amplification (all operating voltages are at least double-stabilised) and on very good cable connections. As can be seen from Fig.7, the equipment's stability was better than 0.01dB over four hours.

I would like to express my sincere thanks to Dr. Hans J.Hartfuss, DL2MDQ, for looking through my manuscript and for many suggestions and comments. I would also like to thank Mr. Max Münich, DJ1CR, for "many long years" of support by word and deed in assembling the equipment (= providing me with the most exquisite radio-frequency components); and Mr. Knut Brenndorfer, DF8CA, for support with "low-noise components".

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Eugen Berberich, DL8ZX

A Big Wheel Antenna for the 70cm Band

As a horizontally polarised circular beam antenna, the big wheel antenna has proved to be outstandingly suitable for the observation of activities in contests and as a transmit antenna for beacons, and it has other applications as well.

This article describes a format for the 70cm band. It is constructed as an "indoor" antenna, but can also be set up in the open air, if weather-resistant materials are used and if the connectors are potted.

To keep the problems of materials procurement in the background, my construction proposal uses materials which can be found in any DIY supermarket, apart from the trimming capacitor.
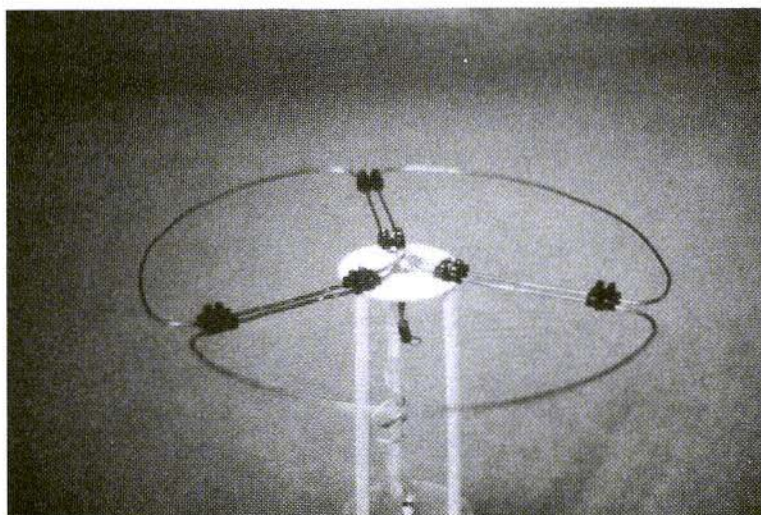
Fig.1: Prototype of a Big Wheel Antenna for the 70cm Band

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1.
OPERATING PRINCIPLE

The big wheel antenna consists of three horizontally mounted loops, each of which is one wavelength long, and which are wired up in parallel to the power splitter in the correct phase to give an impedance of approximately 50Ω. The inductive reactive component is compensated for by means of a trimming capacitor. Only the external arc, with a length of λ/2, acts as a radiator. The λ/4 circuits (spokes) are there only for power splitting purposes, as in them the currents run in opposite directions and theoretically cancel each other out (1).

2.
CONSTRUCTION

To give the antenna some stability, the spokes of the big wheel must be fixed to a base plate (Fig.1). The lid of a 70mm branch box, for example, is suitable for this purpose, with the holding clamps removed - or a disc made from epoxy material. Biscuit connectors are suitable for fastening the wires, with a hole in the middle for fastening using an M3 screw.

The "wheel" can be made from 2.8mm diameter bare Cu wire (e.g. 6mm² NYA or earth wire). The formation can be given additional stiffening through
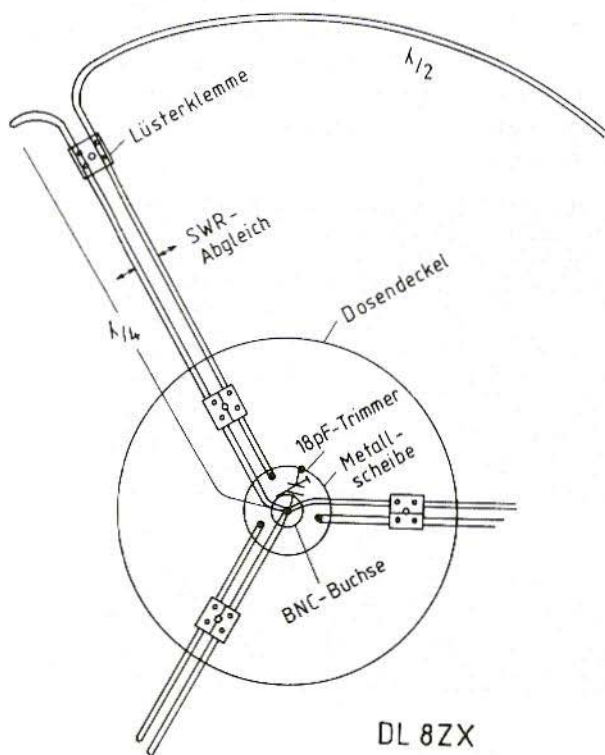

Fig.2:
Sketch demonstrating the Construction of a Big Wheel Antenna

Lusterklemme = Biscuit connector
SWR Abgleich = SWR adjustment
Dosendeckel = Box lid
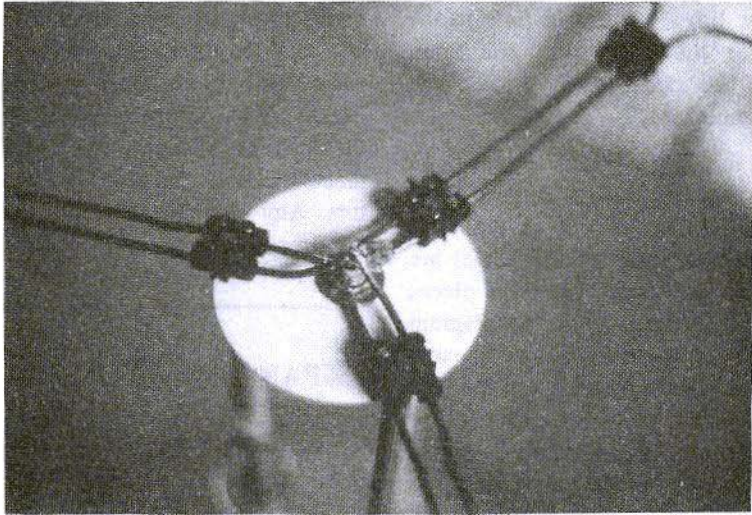*Metal-scheibe* = Metal disc
BNC-Buchse = BNC connector

Fig.3: Connecting the Wires Ends to the BNC jack and the Metal Disc
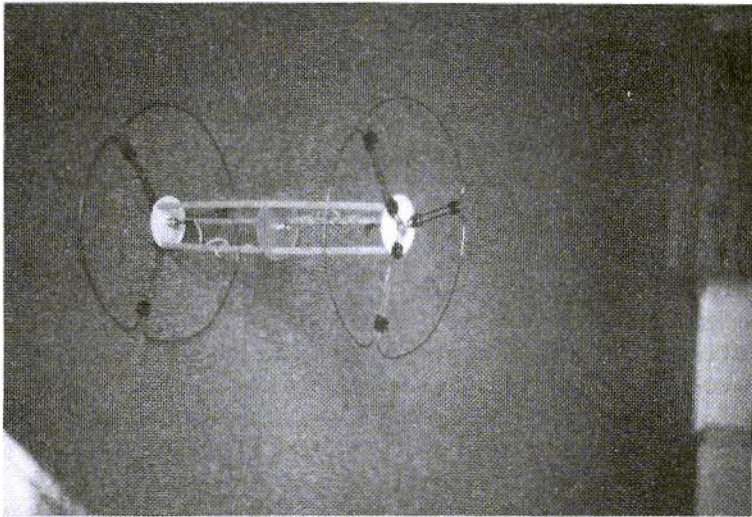

Fig.4: The Twin Antenna Assembly

three additional biscuit connectors at the points where the spokes meet the wheel. A large shim or a piece of brass plate with a suitable hole acts as an earthing point directly on the BNC jack.

3.
ASSEMBLY PROCEDURE

Measure and cut three pieces of wire, each 70cm long. Prepare the assembly

plate (box lid) - i.e. drill a hole in the middle for the BNC jack and 2mm holes at 120° intervals to attach the biscuit connectors.

A 10mm shim (or a brass plate) with an external diameter as large as possible is attached at the same time as the BNC jack. The earth ends of the wheel are soldered on here later. The three pieces of wire are now bent as in the diagram in Fig.2. Then push the biscuit connectors over the wires and prepare the wire ends to be soldered to the earthing disc and the internal conductor of the BNC jack (Fig.3). The spokes should run parallel as far as possible. The distance between them is fixed by the biscuit connectors.

As the box lid is thermoplastic, the wires should be soldered to the earthing disc before assembly takes place. An assembly disc made of epoxy material would be more practical here, as it is less sensitive to heat, and would thus give the formation rather more stability.

After assembly, the formation is screwed together and calibrated. The three circuits can then be soldered to the internal conductor. In the final stage of the assembly, the 18pF foil trimmer is soldered directly to the BNC jack between the centre connector and earth, so that the rotor of the trimming capacitor is lying against earth.

4.
SWR CALIBRATION

Set up the antenna at a sufficient distance from metallic components, and connect up a standing wave meter and a 70cm transmitter through a coaxial cable. Use the trimmer to set to the lowest SWR. Improvements can also be obtained by changing the "spoke distance" between the two biscuit connectors. Any changes should be made on all three spokes.

5.
EXPANSION OPTIONS

For my application (DF0ANN beacon), I stacked two of the antenna together to increase the gain by app. 3dB.

The stacking distance is λ . 0.8 and the two radiators are fed by 75Ω circuits with an electrical length of 0.75λ giving a transformation to 100Ω. With two antenna wired in parallel, this gives 50Ω.

The two planes must naturally be constructed to be equal and in phase. The stacking distance can be obtained using 3 hardwood rods or plastic rods between the assembly plates.

To stiffen the construction, and for the mounting of the BNC jack for the two antenna, an assembly platform can be provided in the middle between the two antenna (Fig.4).

6.
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