VHF
COMMUNICATIONS

AGC test switch
Henning Christof Weddig, DK5LV

Contents

The first two articles that have been entered into the article writing competition are in this issue. I am hoping to have at least two such articles in each issue this year but at the time of writing I only have one more article promised. If you have an article that you want to submit for entry into the competition please send it to me and you could win the first prize of €150 worth of components from R F Elettronica or the second prize of €75 worth of components from R F Elettronica. To enter an article it should be something that will appeal to readers of the magazine on a subject covering the VHF, UHF or Microwave spectrum. I prefer articles with text in Word format and pictures/diagrams as jpeg image files but I can cope with other media. If your article is not quite ready for publication I can help to get it into shape. Send your articles by email or post to the addresses shown below.

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73s - Andy

Michael Klerkx, PA0MKX

ATV Transverter for conversion of the 23cm band to the 3cm band

Entrant in the 2013 article writing contest: Article Number 1

1.

Introduction

- If I have a 23cm ATV transmitter how can I turn it into a 3cm ATV transmitter in a simple and affordable way and at the same time minimise signal losses and minimise use of expensive cables?
- How can I keep the housing as small as possible and maintain an output level on 3cm of at least 10mW or more?
- How do I keep control of frequency, frequency stability of both exciter and local oscillator, as well as the modulation and output power, indoors?

This article deals with all these questions and is based on experiments that I performed over a period of 2 years. Every part of this transverter was built and tested separately during this period. The resulting design shows a compact transverter mounted on a double sided copper-clad PTFE substrate printed circuit board in a standard format of 74mm x 110mm that can be soldered into a standard tin plate housing.

I want to express my special thanks to Jos Disselhorst, PA3ACJ, for his technical advice and for the construction of the waveguide filter and the cavity filter (and silver plating them by electroplating as well!!)

2.

Considerations in choosing a prototype

Because of relative high cable losses at X band frequencies it is not desirable to transmit 3cm signals over a long coax cable to the antenna because power on 3cm is expensive (approximately €0.5-/mW from GaAsFET amplifiers). This means that the transverter should be operated as a remote masthead frontend that can be coupled directly to the antenna. All exciting signals are transmitted at a much lower frequency through the coax cable with low losses.

2.1 The DRO

We can achieve this goal by using a so-called DRO transmitters (Dielectric Resonator Oscillators), but they have the disadvantage that they cannot be tuned from the ground and the tuning range is limited to about 100MHz. Also, modulation with an ATV baseband signal is very critical and only possible in the middle of the tuning range. The bad phase noise characteristics of a DRO can result in a noisy ATV picture. The optimum tuning
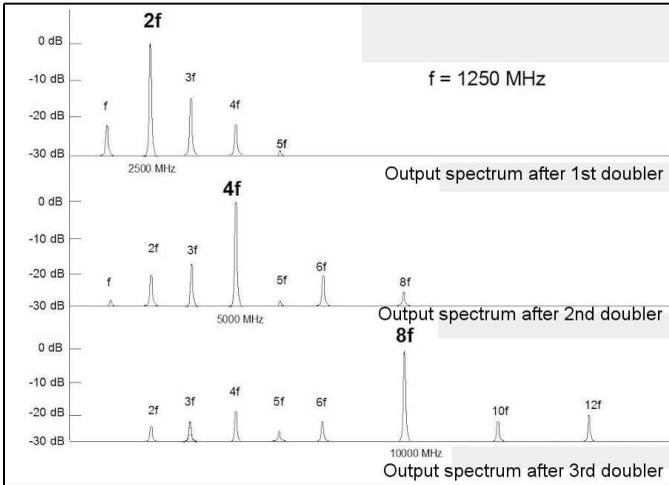
Fig 1: Unwanted harmonics after wideband frequency multiplication by 8 times.

frequency is not consistent when concerning both video carrier and audio subcarrier. The stability is poor and makes the transmitter frequency very temperature dependent. I dare to say that the DRO is not really meant for transmitting signals that are modulated. The reason that you see them used often is money. The DRO is by far the cheapest solution to produce 3cm signals. DROs have their main use in local oscillators for broadband receivers (LNB) or replacement of Gunn diode oscillators in e.g. Doppler detectors.

2.2 Frequency multipliers

Frequency multipliers are a good alternative, as they circumvent most of the disadvantages of the DRO. They can be tuned from the ground and to a specific frequency that is not influenced too much by temperature differences. The main disadvantage of a multiplier for ATV lies in the fact that every multiplication stage needs a dedicated amplifier followed by a dedicated filter that needs to have a specific bandwidth. The following multiplication stages need filters with multiplied bandwidth as well. The first stage that doubles 1250MHz to 2500MHz needs 120MHz bandwidth, whilst the last stage that doubles 5000MHz to 10GHz needs a 500MHz bandwidth filter. These individual bandwidths are critical because other

harmonics are also produced by tripling. We have to suppress the unwanted harmonics as strongly as possible. Using printed microstrip filters is an affordable solution for this problem. It is a compromise between minimal bandwidth, which is somewhat too narrow for broadband use, and selectivity, which is not optimal for this type of filter. However, it can just work well because the attenuation at octave and half octave distance from the centre frequency (sub harmonic and 3 x sub harmonic respectively) of -10 to -20dB is just enough to make the multiplier work effectively (see Fig 1).

The bandwidth of the multiplier is about 100MHz but it will work with limited power output over a much wider frequency spectrum. When the design is optimal, it can cover the entire 3cm band. A major disadvantage of multipliers is their nonlinear behaviour. They will not work with very low input power. Also, the noise figure is not very impressive, as it is not only the frequency that is multiplied but also the phase noise of the ATV VCO.

2.3 Linear transverters

Linear transverters are very suitable for use as a 3cm frontend. They make use of a frequency multiplier as well, but only

for the local oscillator signal. The modulated ATV signal is mixed with this local oscillator signal in a double balanced mixer (DBM) to produce the 3cm signal. Multiplication of a clear crystal generated un-modulated signal is a very different story because we can improve the selectivity by using narrower filters, like helical or High Q cavity filters.

Experiments show that only one High Q cavity is sufficient to perform frequency multiplication to 8x input frequency while suppressing all unwanted sub harmonics and harmonics, because of the extreme high quality factor of these cavities. The noise figure of the linear transverter is far better than the noise figures of both DRO and a multiplier and is mainly limited by the phase noise characteristics of the VCO used in the 23cm exciting ATV transmitter. This phase noise varies depending of the tuning range of the VCO. This means that a transverter of this type could also be used for narrow band applications e.g. SSB, NFM and FM stereo multiplex. That is virtually impossible with DROs and multipliers. Last, but not least, all controls can be done indoors i.e. transmitter frequency, local oscillator frequency (within the cavity limits ±2.5MHz) and output power from zero to maximum.

A very important and necessary part of the linear transverter is the filter that is placed at the output of the double balanced mixer. This filter has to be capable of suppressing all unwanted signals from the mixer RF port, of which the local oscillator frequency (9GHz) and the difference frequency (approximately 7.5 – 8GHz) are the strongest. Many other spurious signals are generated in the DBM, due to the nonlinear behaviour of the 4 schottky diodes in the ring mixer. These frequencies are so close to the output frequency, that they can saturate the final 3cm PA very easily, because they are also amplified. The filter must have a steep cut-off as close to the lowest transmitting frequency and still have enough bandwidth to permit the use of

the whole 3cm band. Normally, the only filter that is suitable for this purpose is the interdigital bandpass filter that is hard to construct by yourself and very expensive when you have to buy it. It also takes a lot of space. A filter of this type with acceptable dimensions is made by the Danish manufacturer PROCOM and has a price of approximately €250. However, the filter steepness is only necessary on the lower side of the passband because spurious and harmonics above 10.5GHz are negligible. Therefore, we can use a highpass filter instead of the interdigital bandpass filter.

In an article, written by WA1MBA, that I found on the web described a method of using a waveguide as a highpass filter. Quote: "Waveguide is the world's best highpass filter". It appears that every type of waveguide can be used on frequencies that are slightly higher than half of the maximum frequency specified for that waveguide. Let's take WR90 waveguide as an example:

WR90 is the best waveguide for 3cm because its frequency range in standard TE_{10} mode is: 8.2 – 12.4GHz. This is the range of negligible attenuation. (<0.2dB/cm). This waveguide shows a slightly higher attenuation (<1dB/cm) on frequencies below the specified 8.2GHz. This attenuation remains fairly constant upon lowering the frequency until about 6.8GHz. If we go a little lower in frequency, the attenuation suddenly increases dramatically to more than 10dB/cm. We have passed the so called cut-off frequency that is specified for WR90 at 6.557GHz.

Armed with this knowledge, we now consider WR62 waveguide with a specified frequency range of 12.4 - 18GHz. This waveguide has its cut-off frequency at 9.5GHz. A signal on 3cm (>10GHz) will pass fairly un-attenuated, while the local frequency (9GHz) is strongly attenuated. The difference frequency (7.5 – 8GHz) is even more attenuated because it is much lower than the local oscillator.
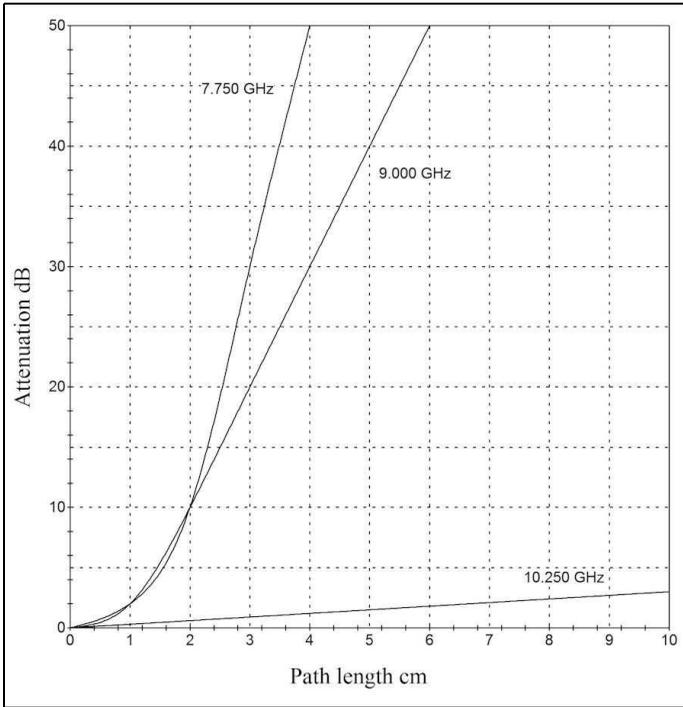
Fig 2: Attenuation properties of WR62 waveguide.

Therefore, WR62 is ideal to use for our filter (See Fig 2).

It should be noted, that the higher harmonics e.g. 20 and 40GHz will pass through with little attenuation, not only in TE_{10} mode, but also other modes. The effects are negligible. The cut-off frequency appears to be inversely proportional to the dimensions of the waveguide.

3.

The prototype

As a crystal oscillator chain takes considerable space and is very sensitive to temperature variations, it is not very attractive to place this part on the antenna mast. However, the last two doublers could be built very compactly into the outdoor mast-mounted device because they can be put together as a quadrupler.

Based on these facts an idea for the prototype was formed as follows:

Both local oscillator generation (signal A) and conversion of the ATV transmitting frequency (signal B) will be split into two parts:

- On the ground (indoors): Signals A and B on standard epoxy fibreglass board.(<2250MHz)
- On the mast (outdoors): processing of signals A and B on PTFE board. (>9GHz)

More specifically:

- On the ground, Signal A: Crystal chain (oscillator with 24x multiplication to 2250MHz)
- On the mast, Signal A: Quadrupler using a cavity filter and amplifier to 9GHz
- On the ground, Signal B: Modulated ATV signal 1000 – 1500MHz (approximately 10 - 30mW)
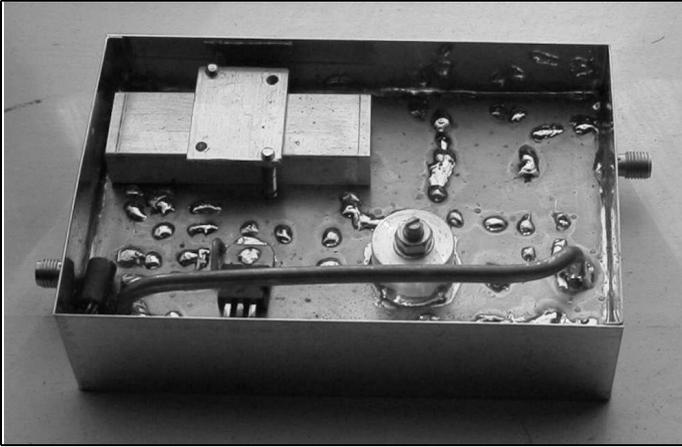
Fig 3: The soldered cylindrical cavity filter on 9GHz and the rectangular 9.5GHz highpass filter attached to the board by two M3 screws are clearly recognisable as well as the 2250MHz semirigid coax bridge.

- On the mast, Signal B: Mixing with 9GHz (signal A) in DBM to 10GHz, filtering and amplification to a minimum of 10mW and maximum of 200mW

The most attractive concept is to transmit both A and B signals together with the DC supply voltage over a single coax cable to the transverter frontend. In order to achieve proper merging of both signals on the ground for the coax feed and proper separation of both signals at the frontend, we have to make use of two identical duplex filters, fitted with a bias tee for the power supply. Recent technical developments (Minicircuits) yielded so-called LTCC filters with dimensions comparable to SMD resistors at a price of approximately €2 that are perfectly suited for our purposes. (LTCC is an abbreviation of Low Temperature Cofired Ceramics)

3.1 The mixer

Recently the American manufacturer Minicircuits developed a double balanced mixer based on LTCC technology. This mixer can be obtained at an affordable price (approximately €10) and has the type code MCA1-12G. This mixer can be used between 3.8 and 12GHz. The IF range is DC to 1800MHz with an average conversion loss is about 6.2dB. This type is the best choice for our transverter.

Soldering of this mixer is however a precision job and requires certain skills because of its SMD format, but it can be done safely with a normal soldering station.

3.2 The crystal oscillator chain

This article mainly deals with the mast-mounted parts on PTFE board. The crystal chain that is necessary to generate the local oscillator frequency of 2250MHz with an output power of at least 10mW is built on standard epoxy fibreglass board. It uses an old well known design consisting of a U310 FET and a 93.750MHz quartz crystal as exciter followed by; a tripler (BFR90), a doubler (BFR90) a doubler (BFR91A), a doubler (BFR91A) and amplifier (ERA-3). This makes multiplication of x 24. The print circuit board layout and parts are described chapter 7.

3.3 The ATV-transmitter

The ATV transmitter that drives the transverter should be capable of generating an ATV baseband modulated signal that can be regulated manually between 0mW and 30mW. The Comtech ATV transmitter that can be tuned from 1000-MHz to 1400MHz is well suited for this purpose giving an output frequency of 10,000 – 10,400MHz. There are many other options. I will not discuss the details of this part in this article.
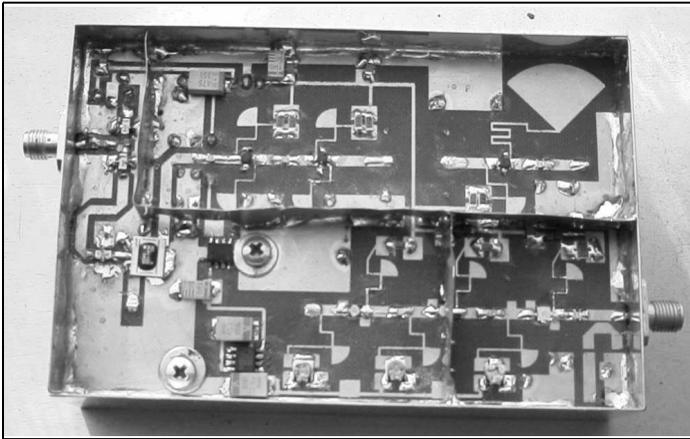
Fig 4: In this picture you can recognise the two M3 scews holding the 9.5GHz highpass filter and the several copper shields. These shields are discussed later on in this article. On the left the MCA1-12G mixer can be seen.

Photographs of the bottom side and the top side of the finished prototype are shown in Figs 3 and 4.

4.

Brief description of operation (signal path) and block diagram

On the ground, signals A and B are first
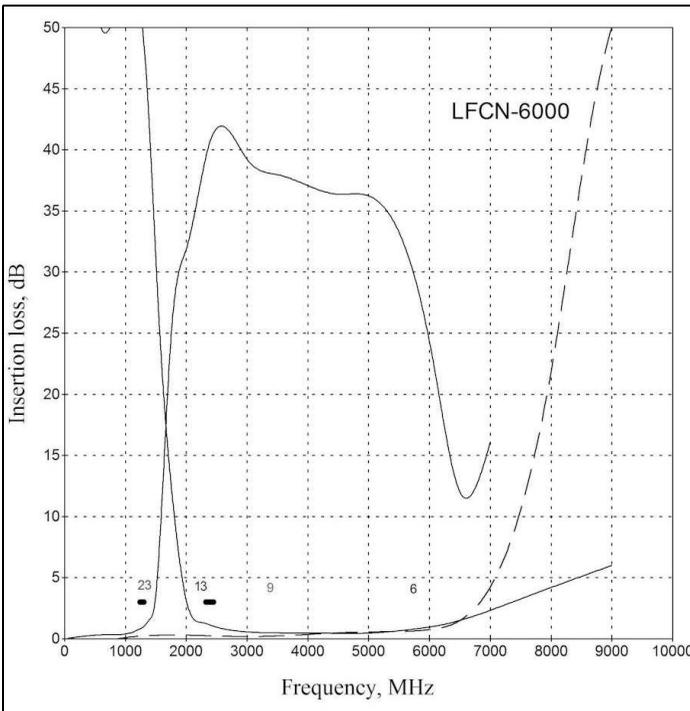
Fig 5: Frequency characteristics of the LFCN-1200/HCFCN-2000 duplex filter.

merged together by means of a 23cm/1-3cm duplex filter, consisting of a LFCN-1200 lowpass and a HFCN-2000 highpass filter (Minicircuits). The 12V DC supply is added through a bias tee directly after the duplex filter. These two signals with the added 12V DC are fed into a single coax cable. The far end of the cable at the mast frontend input the DC voltage is separated first by a bias tee. Then the two signals are separated by means of an identical duplex filter (Minicircuits LFCN-1200/HFCN-2000). As the 23cm lowpass filter (Minicircuits LFCN-1200) shows too little attenuation at frequencies of 9GHz and higher, an extra lowpass filter is added at the IF port of the mixer (Minicircuits LFCN-6000) that has enough attenuation for those frequencies and so prevents "leaking back" the local oscillator and output signals into the IF line. The passband characteristics of the duplex filter are shown graphically in Fig 5 together with the characteristics of the LFCN-6000 lowpass filter.

Signal A (10mW 2250MHz L.O.) travels directly from the highpass port of the duplex filter at the shorter board edge, through the earth plane into a bridge of semirigid cable and back through the earth plane to the top side at the opposite shorter board edge. There it is fed directly into the quadrupler, consisting of an ERA21SM MMIC followed by a cylindrical cavity that is tuned to 9GHz. Signal A (now 9GHz) is then amplified by 12dB with two ERA21SM MMICs. At this point, the signal level is about 7dBm that is fed into the LOCAL port of the mixer (Minicircuits MCA1-12G) and mixed with signal B (23cm ATV signal, maximum 10mW!!). That comes directly from the lowpass port of the same duplex filter and fed into the IF port of the mixer via an extra lowpass filter (LFCN-6000). It should be noted that signal B has suffered some attenuation when it enters the mixer. This is because on its travel from the exciter on the ground to the mixer it has passed two bias tees and 3 lowpass

filters, resulting in an attenuation of about 3dB, depending on the cable length and quality. This means that maximum output power is reached when the ATV signal on 23cm is between 20 and 30mW.

The raw 10GHz signal (-17dBm) is generated at the RF port of the mixer. This signal is fed into the WR62 waveguide by means of a quarter wave monopole antenna positioned through the earth plane inside one end of the waveguide, just like a coax waveguide transition. The waveguide has a length of about 5cm and functions as a very sharp highpass filter with a cut-off frequency of 9.5GHz. So the local oscillator and difference frequencies (respectively 9 and 7/8GHz) are almost completely blocked. Attenuation by this filter on 10GHz is about 2 - 3dB and is very significant for the ultimate output power, so the filter must be constructed with as much precision as possible.

At the other end of the waveguide, an identical pickup monopole antenna is positioned and the signal (-18/-20dBm) travels through the earth plane back to the top side. It finally enters into a 3-stage 10GHz power amplifier, with high spectral purity, consisting of two MGF 1302 stages and a final stage with a power GaAsFET that can be chosen from different types: MGF 1302/1323, 10 - 30mW, or MGF1801, approximately 100 - 200mW. Total amplification depends on the choice of the final transistor and is 28 – 32dB. It should be noted that an MGF1801 GaAsFET is a pretty expensive device that operates best with an input signal of 50mW. If this level cannot be reached by the previous stages an MGF1323 is preferable because its gain is higher than the gain of an MGF1801 but its output is lower.

For a good overview everything discussed in the preceding story has been arranged in the block diagram shown in Fig 6. The parts that are operated on the ground or in the shack are shown as

Fig 6: Block diagram of the transverter system.

Fig 7: Circuit diagram of the transverter.
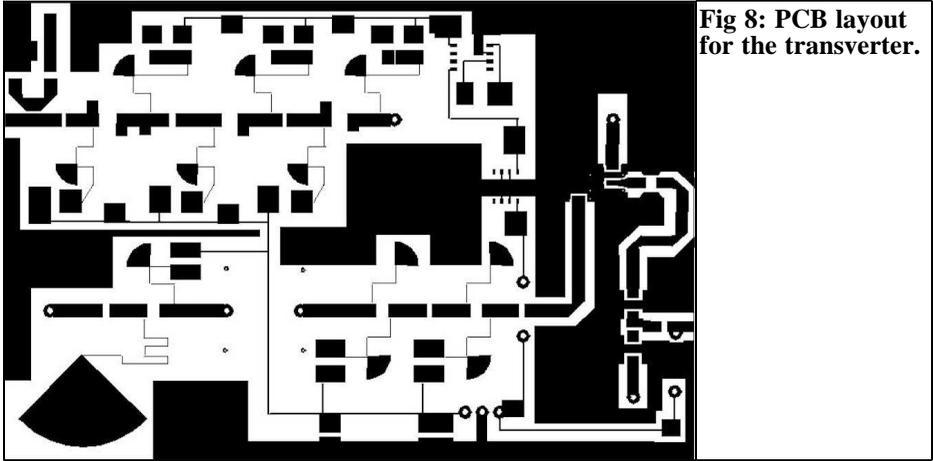
Fig 8: PCB layout for the transverter.

dotted.

The circuit diagram is shown in Fig 7. The PCB layout for the transverter is shown in Fig 8 it is made from PTFE laminate with $\varepsilon_0 = 2.33$ and thickness 0.75mm (Taconic Ultralam or Rogers RT-Duroid). The dimensions are: 10.85 x 7.10cm. The component layout for the transverter board is shown in Fig 9
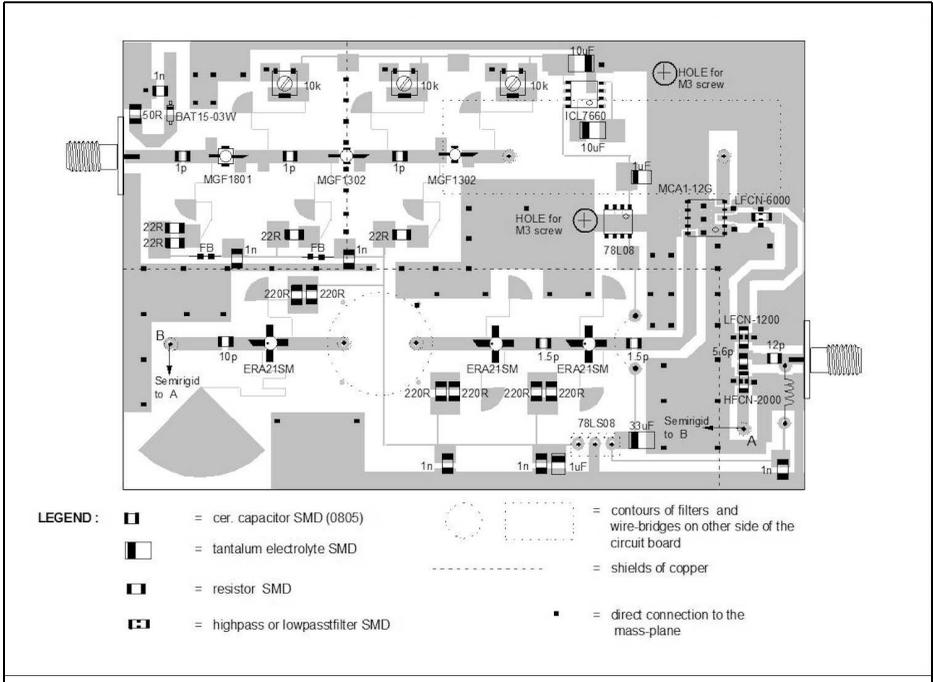

**Fig 9: Component layout for the transverter.**
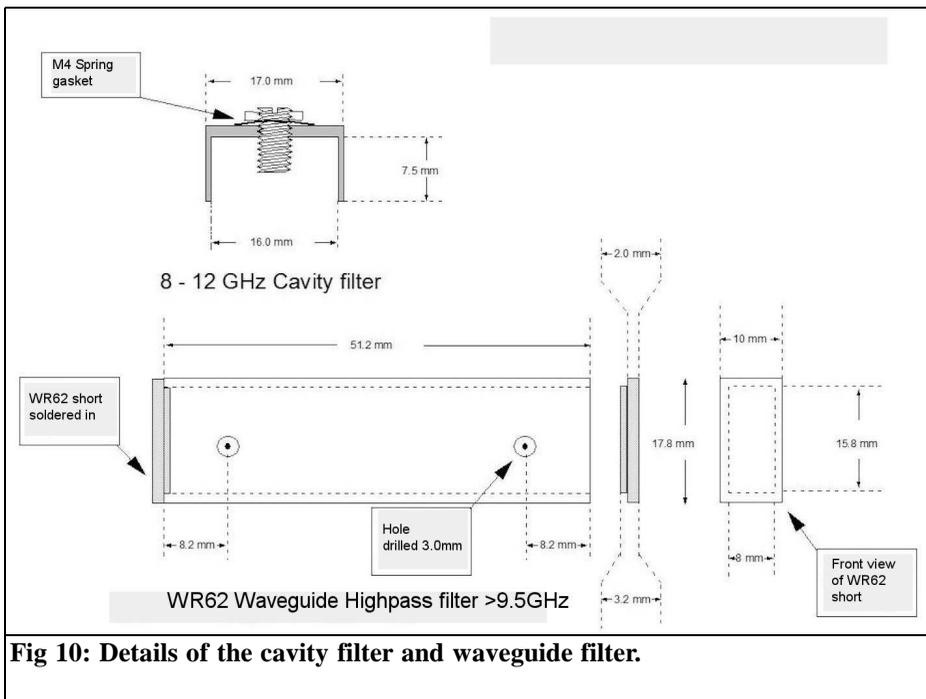
Fig 10: Details of the cavity filter and waveguide filter.

5.

Construction of the 9GHz cavity filter and the 9.5GHz highpass filter

The cavity filter is made of a standard solderable brass end cap for copper gas tubing with an inside diameter of 15mm and an outside diameter of about 17.5mm. The inside height is about 11mm and the outside height is 13mm. The open edge of the cap is milled off to an inside height of 7.5mm. The milled edge should be straight, sharp and as smooth as possible. Drill a 3.5mm hole in the centre of the top side and tap a M4 thread in the hole. This has to be done absolutely perpendicular to the surface. Now the cap is ready to use as a filter but for optimum performance and solderability it is better to mill off the inside diameter to 16mm and the outside diame-ter to 17mm, taking care that no material is milled off the top side. A silver plate finish is recommended but not imperat-ive. When the cap is finished, fit a headless silver plated screw (worm screw) with a length of 1 – 1.5cm. Screw it in until it has gone through the surface and is visible on the inside of the cavity. Then place a clamping ring over the screw and close it with an M4 nut. The cavity filter is now ready (See Fig 10).

For the construction of the 9.5GHz high-pass filter we start with a piece of WR62 waveguide that is sawn off and milled to an exact length of 5.10cm. Two holes with a precise diameter of 3mm are drilled on one broad side of the wavegu-ide (see Fig 10). Both ends are closed by soldering a brass block onto them accord-ing to the dimensions given in Fig 10. A silver plate finish internally as well as externally is recommended, but not impe-rative. All construction details are shown in Fig 10.
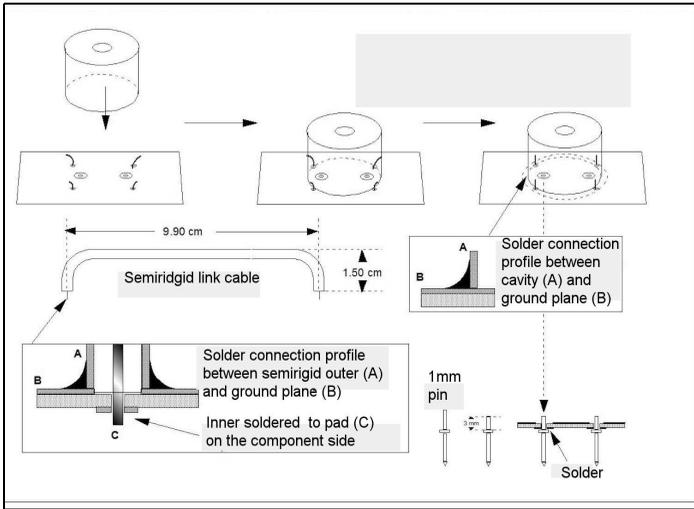
Fig 11: Attaching the cavity filter and semiridgid link cable to the ground side of the PCB.

6.

Detailed description of the transverter construction

Note: If you do not have a signal generator capable of producing UHF signals it is advisable to start with the construction of the crystal oscillator chain because a 2250MHz signal is necessary in the course of constructing the transverter.

We strongly recommend, that construction is done in order of the following steps:

Step 1: Adjusting the size of the printed circuit board to fit exactly in the housing and soldering the SMA connectors.

Cut one short edge of the board with a Stanley knife to the exact size using the housing as a guide. Repeat the same procedure for the opposite short edge. Cut two notches (3mm x 0.3mm) in opposite corners of the board and remove the sharp corners of the board. Fit the board into the two half sides of the housing. Adjust the board size until the board fits exactly. The short solder flanges of each half housing side should fit exactly in the notches at opposite corners. Use a marker pen to mark the positions of the SMA connector centres on both housing half sides. They should 1cm from the top edge and 2cm from the bottom edge. Drill the holes for the SMA connectors (only the centres and not the flange holes!!) Solder the SMA connectors into the housing as centred as possible. NOTE: Do not solder the board in the housing yet!!

Step 2: Drill 13 holes with a diameter of 1.0mm and isolating them from the ground plane.

These holes are for the waveguide highpass filter (2), the cavity filter (2), the choke (2) , the wire bridge for the 12V DC supply to the 78L05 regulator (2), both ends of the semirigid bridge (2) and the 78LS08 regulator (3) All are isolated from the ground plane except for the middle hole of the 78LS08 regulator. NOTE: The isolation distance for the cavity filter and the semirigid bridge from the ground plane should be at least 1.0mm but not more than 1.5mm from the centre hole.
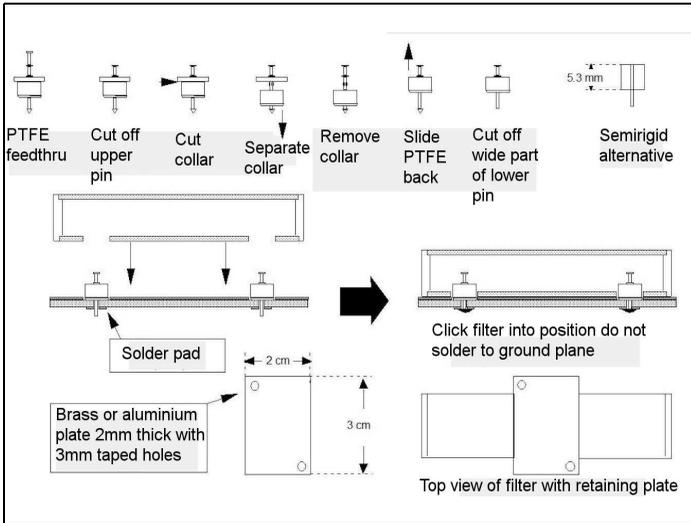
Fig 12: Making the waveguide filter.

Step 3: Drill 4 positioning holes for the cavity filter with a 0.6mm drill.

These 4 holes are indicated on the board (see Fig 9).

Step 4: Drill 3 holes for connections to the ground plane with a 0.6 mm drill

These 3 holes are located within the footprint of the MCA1-12G DBM and are indicated on the board (see Fig 9). Pull a bundle of 3 thin copper wires (<0.2mm diameter each) through all 3 holes and solder them to both sides of the board with excess solder. Remove excess solder with solder wick to keep the board as flat and smooth as possible.

Step 5: Soldering the cavity filter and the semirigid cable bridge on the board (Fig 11)

We start with attaching the cavity filter to the board. Position the board with its ground plane upwards on a good and stable table with a heat resistant top or perfectly a smooth wooden plate. Locate the 4 positioning holes and put a single 0.6mm wire (length about 0.5cm) into each hole. Bend the 4 wire tops slightly away from each other, so that they are arranged like a funnel. Check to see if the board is perfectly straight and smoothly resting on its support. Lower the cavity filter without its tuning screw and screw hole upwards between the 4 positioning wires. Apply pressure on the cavity and check that it is resting perfectly on the ground plane. If not, press it in place between the 4 wires. Bend the 4 wires back into an upright position touching the sidewall of the cavity. Keep a fairly high pressure on top of the cavity by means of a heat resistant tool e.g. screwdriver while soldering the cavity to the ground plane. This prevents solder leaking inside the cavity. Apply enough solder to cover the whole circular contact area with excess. (Caution, the cavity gets very hot!!). Apply heat until the soldering profile pictured in Fig 11 is reached all around the cavity. Keep the pressure on the cavity until the solder has cooled down. After reaching room temperature put the tuning screw, clamping ring and nut back in place.

Prepare two 1mm soldering pins according to the procedure shown in Fig 11 and solder them in place. The collar of the pin has to be firmly pressed against the board. The longer pointed ends on the other side of the collar are used to keep

the pin perfectly perpendicular and parallel while soldering. After cooling down, the long pointed ends are clipped off and flattened (see Fig 11).

Bend a piece of semirigid cable as shown in Fig 11. The two ends should be exactly parallel and perpendicular to the middle section of the cable. Shorten both ends to a total height of 18mm (see Fig 11). Remove the cable shielding and PTFE dielectric at both ends over a length of 3mm. The semirigid bridge is now fitted in the board on the ground plane side. Press the bridge until the copper cable shielding touches the copper of the ground plane. The height of the bridge is now 15mm. Solder the bridge at both ends to the ground plane using excess of solder around the cable shield until the soldering profile shown in Fig 11 is achieved. Solder the two protruding centre conductors to the solder eyes on the circuit side of the board (top side).

Step 6: Mounting the WR62 highpass filter on the board (Fig 12).

Prepare two 1mm PTFE feedthroughs in 6 consecutive steps as indicated in Fig 12. Alternatively two pieces of 3.5mm semirigid cable can be used (see Fig 12). Press them both in the holes from the ground plane side of the board. The PTFE body of the pins should fit exactly in the isolation space in the ground plane around the hole. Check that they are perpendicular and solder them to the top side of the board. Mill off the soldering points to about 0.2mm above the board. Place the board on the table with its ground plane upwards and place the waveguide filter with its two holes directed downwards onto the two protruding feedthrough pins. Push and move the filter carefully in its place. If everything is OK the filter rests on the ground plane. If you have trouble with an exact fit it helps to taper the PTFE isolating collars. Remove the filter from the board and put it aside. Carefully drill two M3 holes in the board at the positions indicated on the

component layout. Prepare an aluminium or brass plate with dimensions 2cm x 3cm x 0.2cm. Position the plate on the ground plane over the two M3 holes in such a way that both holes are covered by two opposite corners of the plate and parallel to the board. When the right position is found attach the plate to the ground plane with Sellotape. Turn the board with its topside up and mark the plate through the M3 hole, with a marker pen. Then remove the Sellotape and the plate. Drill the two marked spots with a 2.5mm drill and tap with an M3 thread. Place the filter back in its position and put two M3 screws with washers through the M3 holes from the top side. The two screws protrude on the ground plane side exactly on both sides of the filter. Attach the plate above the filter with the screws until every part is touching the ground plane. CAUTION: Do not over tighten the screws because it will deform the board. Remove the filter and plate again and put them aside.

Step 7: Soldering the circuit board into the housing.

Draw a line on the inside of both half sides of the housing with a marker or a sharp knife. This line should be in line with the positions of both connector centre conductors. The distance from the top edge should be 1cm and the distance from the bottom edge consequently 2cm. (total height of the housing is 3cm without top and bottom covers.)

Solder both half sides of the housing together at their soldering flanges in the two opposite corners. Place the housing frame on the table with the connector centre conductors at 1cm above the table. The circuit board is now fitted in with its ground plane upward. Press the board down until it rests on both connector centre conductors and align exactly with the line drawn inside the housing. Keep the board as straight as possible. Secure the position of the board by soldering it to the housing at several points. Turn the frame to have its top side upward and

solder the centre conductors of the connectors to the circuit board. Turn again to the bottom side and solder the ground plane to the housing on all four edges without any gaps. Finally, the board is turned to the top side and soldered to the housing at the dedicated areas (see PCB layout).

Step 8: Soldering all components on the circuit board

Solder all SMD resistors, tantalum capacitors, the 3 SMD variable resistors, the 3 SMD filters and the SMD ceramic capacitors onto the board. Do not solder the Johansson trimmer capacitor and the MCA1-12G mixer. Solder the wire bridge, the choke and the 78LS08 regulator that are positioned with their connecting wires through the isolated holes in the ground plane. Finally solder the SMD regulator (78L05) and the SMD voltage inverter (ICL7660). Make a break at two places in the 8V DC feed to the GaAsFETs and bridge them each with a wire and two ferrite beads (see component layout).

Note: An optional 10GHz detector can be mounted on the circuit board consisting of a 50Ω SMD resistor, a 1pF ceramic capacitor and a schottky diode (BAT15-03W). This can be used for AGC of an eventual extra final PA stage.

Test all the supply before fitting the three ERA21SM MMICs in place and drill a 0.6mm hole near the ground pins on both sides of the MMICs and solder a through connection to the ground plane. Note: The through holes should be as close to the package of the MMICs as possible to prevent parasitic oscillations. Connect the power supply again via the bias-tee and measure the voltages on inputs and outputs of the MMICs. Inputs should be 2.5 - 3.0V. Outputs should be 3.5 - 3.8V. If this is not the case carefully inspect connections to the MMICs for short circuits or breaks.

Now fit the first MGF1302 GaAsFET. First bend both source connections 90°

downward as close to the package as possible. Temporarily fit the filter to the board. Cut two parallel slits 1.9mm apart through the ground plane and making a scratch in the filter housing as sown on the components layout. Remove the filter and push the GaAsFET source connections carefully through the board. On the other side of the board bend the connections as flat to the ground plane as possible and solder them leaving the ground plane as flat as possible by removing excess solder with solder wick. Now replace the waveguide filter with a small notch (0.5mm depth and 4mm diameter) milled in the filter to make room for the solder where the GaAsFET is fitted. This is marked on the filter housing by the two scratches made earlier. Secure the filter with its retaining plate.

Repeat the same procedure for the two remaining GaAsFETs (except for the filter manipulations). It is not necessary to bend the source connections flat against the ground plane for these FETs. If an MGF 1801is used the two parallel slits in the board should be 2.7mm apart instead of 1.9mm.

Finally, drill all through connections to the ground plane with a 0.6mm drill (as indicated on the component layout) and solder the connections to the ground plane, leaving ends of 3mm length protruding at the circuit side. Do not solder on the top side because they locate a the copper shields.

Step 9: Fitting the copper shields

Prepare a few strips of thin copper foil 15cm long and 1cm wide. Start with the shield that crosses over the second GaAsFET. As there are three GaAsFETs with a total gain of more than 20dB this shield is necessary to prevent spontaneous oscillation. Cut notches where the supply and bias feed the FET and for the FET. Note: The notch for the FET should fit exactly. Fit the strip between the protruding ground pins, check that the lower edge touches the board everywhere and the strip is perpendicular. Now solder all
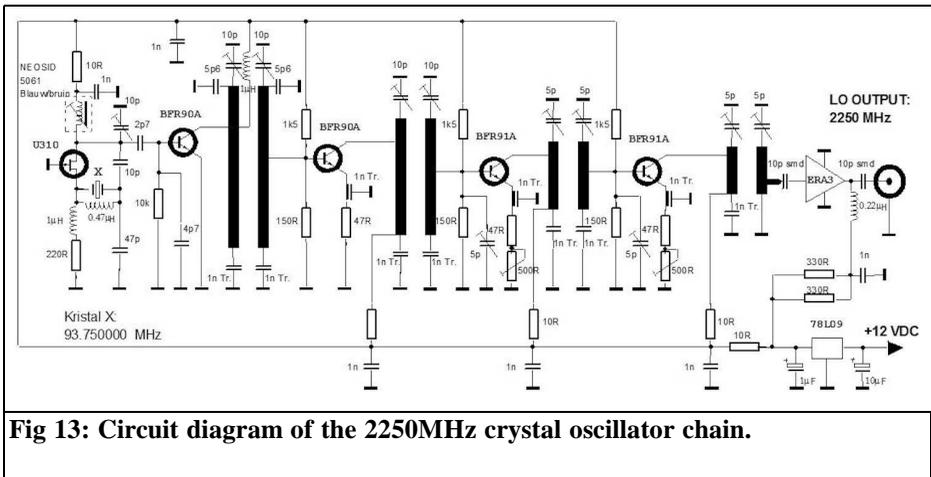
Fig 13: Circuit diagram of the 2250MHz crystal oscillator chain.

wires to the shield. Finally solder one end of the shield to the metal housing. The other end will be soldered to the second shield at a later time.

The second shield closes off the 10GHz signal completely and also prevents unwanted coupling of the 2250MHz LO signal. This is fitted in the same way as the first shield.

Step 10: Tuning the Local Oscillator to 9.0 GHz

Before the MCA1-12G mixer is fitted we have to tune and optimise the local oscillator.

Connect the power supply to the input connector via the bias tee. Connect a signal generator, or another signal source like the crystal oscillator chain capable of producing 10mW on 2250MHz, to the RF port of the bias tee. Solder the centre conductor of a coaxial test cable, which has been fitted with a connector on the other end onto the stripline that leads to the LO port of the mixer. Solder the cable shielding to a ground point on the board in the near vicinity of the LO stripline. Connect the cable to a spectrum analyser tuned at 9GHz. Tune the cavity with its tuning screw to a power maximum at 9GHz. (approximately 5mW). Next tune the analyser to 2250MHz. This

signal has to be significantly lower than the 9GHz signal!! (less than –10dBc) . When the signal on 2250MHz is almost equal to or stronger than the 9GHz signal a Johansson trimmer has to be soldered at the position indicated in the component layout. Tune the trimmer foe a maximum 9GHz signal output. Check the 2250MHz signal again. It should be considerably weaker and sufficiently low. Allow the circuit to warm up to a stable temperature and retune the cavity if necessary

Step 11 Fitting the Minicircuits MCA1-12G DBM

Finally we can solder in the MCA1-12G Double Balanced Mixer. It is recommended, to pre-solder the footprint (10 pads) and remove excess with solder wick. Position the chip and use thin soldering tip to solder all 10 contacts. This completes the construction.

Step 12: Biasing the GaAsFETs

For this procedure a laboratory power supply with a variable constant current limiter must be used set to 12V DC. Set the current limit to 500mA. It is recommended to use a multimeter in the +ve lead to measure the current correctly. Connect this supply to the transverter input connector via a bias tee. We should now read a current of about 150mA. Record this
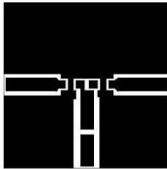
Fig 14: PCB layout the 2250MHz crystal oscillator chain. and duplex filter.

Crystal oscillator chain PCB

Duplex filter PCB

value (accuracy to 1mA). This value we call "the starting value"

Notes:

- If the current is too high or rises to the 500mA limit, immediately disconnect the power supply!!

- Keep a close eye on the ammeter in the following procedure, as too high currents for too long time will damage the GaAsFET!!!

Adjust the variable biasing resistor for the first GaAsFET (MGF 1302) and set a current of 33mA above the starting value and record that value.

Adjust the variable biasing resistor of the second GaAsFET (MGF 1302 and set a current of 66mA above the starting value and record that value.

Repeat the same procedure for the last GaAsFET (MGF 1302/1323/1801). For the MGF 1302 or the MGF 1323 the current increase should be 110mA above the original starting value. For the MGF 1801 the current increase should be 166mA above the original starting
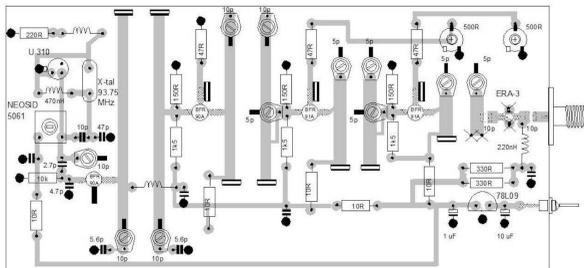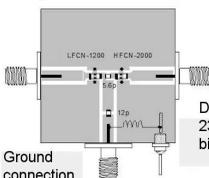
Fig 15: Component layout for the 2250MHz crystal oscillator chain.

17

7.

Construction of the 2250MHz Crystal Oscillator

Both crystal chain and duplex filter are built on standard 1.45 mm epoxy fibreglass board. The circuit diagram is shown in Fig 13.

The circuit diagram shows two variable resistors intended for controlling the output power. The schematic for the duplex filter can be deducted from the transverter input filter since they are identical. The crystal oscillator chain can be built in a standard housing of 111 x 55 x 30mm. The duplex filter can be built in a standard housing of 37 x 37 x 30mm using N connectors or 37 x 37 x 20mm when SMA connectors are used. The PCB layouts are shone in Fig 14 and the component layouts in Fig 15.

After finishing construction, tune the crystal chain to maximal output on 2250-MHz (approximately 15mW).

8.

Putting the transverter into action.

With a test configuration as shown in the block diagram, Set on the ATV transmitter to 1250MHz and about 2mW. The power meter or spectrum analyser will show an output between 1mW and 20mW. Switch off the ATV transmitter. Power output should now drop to 0mW. Repeat this procedure several times during the whole tuning procedure. If at any moment the power doesn't drop to 0mW when switching off a parasitic oscillation is occurring. The biasing of the GaAsFETs should then be checked and capacity should be added on the spot where the oscillation occurs. As the

mixer has a certain optimal LO input it has to be tuned to that power with the last variable resistor in the crystal chain circuit to yield maximum output power at 10.250GHz. Regulate the output power of the ATV transmitter to maximum output power on 10.250GHz. Then place the top cover on the housing. An increase in output power will be observed. Output power should be at least 30mW. If the power is too low it is recommended that ground continuity is checked throughout the whole circuit. In this way a power of 40mW was reached for this prototype

9.

Transverter specifications:

Output power on 10.25GHz:	10mW (with MGF 1302) 20mW (with MGF 1323) 40mW (with MGF 1801)
2250MHz LO input:	10mW (at transverter input)
1250MHz ATV signal input:	10mW (at transverter input)
Isolation LO/23cm signal:	>50dB
Output flatness between: 10.0 and 10.5GHz	<2dB (at 0dB flatness ATV transmitter)
Stability Local oscillator:	1ppm (between 0 and 30°C)
Stability LO with PTC: temperature regulation 30°C	0.5ppm (between 0 and 30°C)
Supply voltage:	12.0 ±1.5V DC
DC current consumption:	approximately 250mA (with MGF 1302/1323) approximately 350mA (with MGF 1801)
Spectral purity at 10.250GHz output and 125 MHz drive:	
9GHz LO suppression:	approximately -30dBc
7.75GHz image suppression:	approximately -37dBc
2250MHz spurious suppression:	approximately -30dBc
1250MHz drive suppression:	>-50dBc
20.5GHz 2nd harmonic:	approximately -60dBc

9.

Parts List - Transverter and External Duplex filter

Semiconductors:

3 x	ERA-21SM
3 x	MGF 1302
1 x	MGF 1323 or MGF 1801
1 x	78LS08
1 x	78L05 (SMD)
1 x	ICL7660 (SMD)

SMD Resistors:

4 x	22Ω
6 x	220Ω
3 x	10kΩ Trimpot

SMD Capacitors:

3 x	1pF
2 x	1,5pF
1 x	2.2pF
2 x	5.6pF (incl. ext. duplex filter)
1 x	10pF
2 x	12pF (incl. ext. duplex filter)
6 x	1nF

SMD 25V Tantalum electrolytics:

2 x	1µF
2 X	10µF

Special Minicircuits SMD parts:

1 x	MCA1-12G DBM
2 x	LFCN-1200
1 x	LFCN-6000
2 x	HFCN-2000

Special filters and other parts:

1 x	Cavity bypass filter 9GHz
1 x	Waveguide WR62 highpass filter
1 x	PTFE PCB
2 x	1nF ceramic feedthrough
1 x	Fiberglass PCB
1 x	Housing 74 x 111 x 30mm
1 x	Housing 37 x 37 x 20(30)mm
5 x	SMA connector
2 x	1µH high current choke
1 x	Jonansson trimer 0.5 - 2.5pF
1 x	Copper strip 20 x 1cm
2 x	Solder pin
2 x	PTFE feedthrough
1 x	3.5mm semiridgid 15cn long
4 x	Ferrite bead

10.

Parts List – Crystal Oscillator-Chain   2250MHz

Semiconductors:

2 x	BFR90A
1 x	BFR 91A
1 x	U310
1 x	ERA-3
1 x	78L09

Resistors:

5 x	10Ω
3 x	47Ω
3 x	150Ω
1 x	220Ω
2 x	330Ω
2 x	500Ω trimpot
2 x	1.5kΩ
1 x	10kΩ

Capacitors:

1 x	2.7pF
1 x	4.7pF
6 x	5pF SKY trimmer
2 x	5.6pF
1 x	10pF
5 x	10pF SKY trimmer
2 x	10pF (SMD)
1 x	47pF
6 x	1nF
10 x	1nF ceramic trapezoidal
1 x	1nF ceramic feedthrough
1 x	1µF electrolytic
1 x	10µF electrolytic

Inductors:

1 x	Neosid 5061
1 x	0.22µH
1 x	0.46µH
2 x	1µH

Other parts:

1 x	93.750000MHz crystal
1 x	Housing 55 x 111 x 30mm
1 x	Epoxy PCB
1 x	SMA connector

Henning Christof Weddig, DK5LV

AGC test switch

Continuation from 1/2012
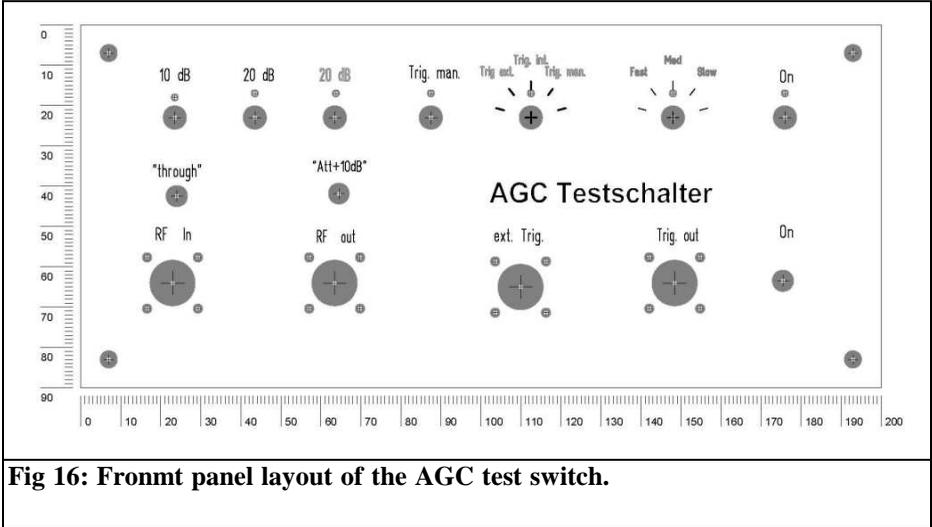
Fig 16: Fronmt panel layout of the AGC test switch.

7.

Results and measurement

Fig 19 shows the measurement result of the RF switch in the through position when St6 is fed with a DC voltage of +12V. The transmission loss (S21) is approximately 1dB; the reflection coefficient (S11) in the lower frequency range is better than –20dB and –16.9dB at 145MHz that is still sufficient.
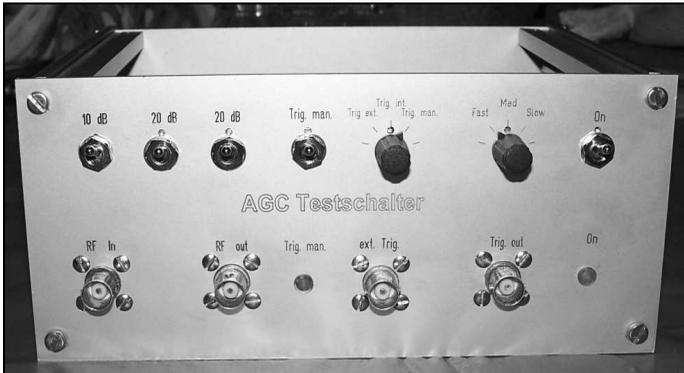
Fig 17: Picture of the prototype AGC test switch.
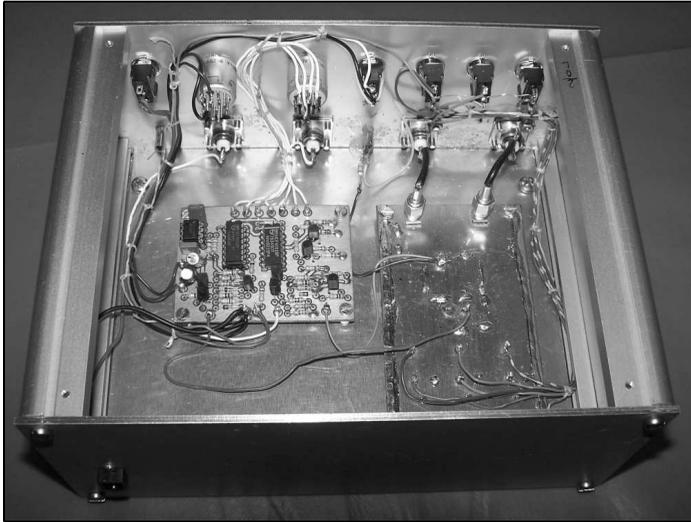
Fig 18: Picture of the prototype AGC test switch.

Fig 20 shows the RF switch in attenuation position, i.e. St3, St1, St4 and ST8 fed with +12V. The transmission loss (S21) is approximately 13dB, 3dB more than expected. The reflection coefficient (S11) in the lower frequency range is better than –20dB and still sufficient at 50MHz getting better at 145MHz. This indicates a periodical compensation of the attenuation.

Fig 21 shows the RF switch in the 10dB attenuation position, i.e. St3, St2, St4 and ST8 fed with +12V. The transmission loss (S21) is approximately 23dB, 3dB higher than expected; the reflection coefficient (S11) is better than –20dB in the lower frequency range. The notch in attenuation is pulled to a lower frequency
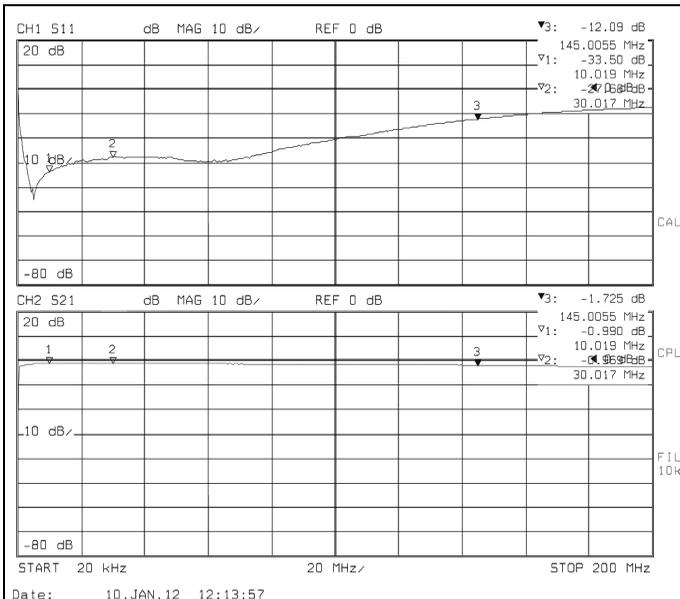
Fig 19: RF switch in "through" position.
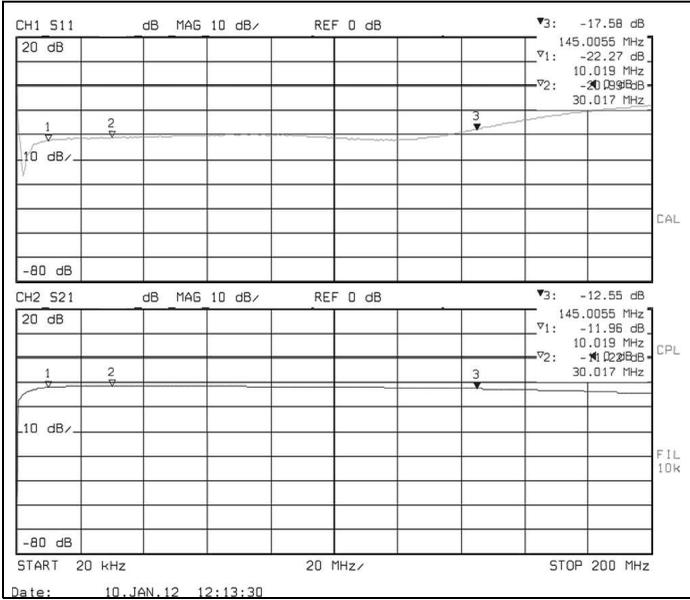
Fig 20: RF switch in "attenuation" position.

(110MHz), the attenuation is still good at 145MHz.

Fig 22 shows the RF switch in the 20dB attenuation position, i.e. St3, St1, St9 and ST8 are fed with +12V. The transmission loss (S21) is approximately 33dB, 3dB higher than expected; the reflection coefficient (S11) is better than –20dB in the lower frequency range. The notch has disappeared; the attenuation is still sufficient at 145MHz.

Fig 23 shows the RF switch in the 30dB
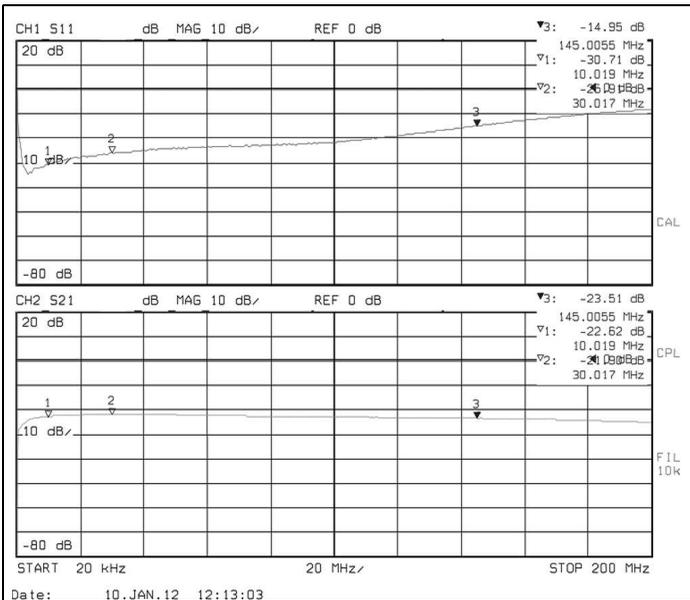
Fig 21: RF switch in the 10dB attenuation position.
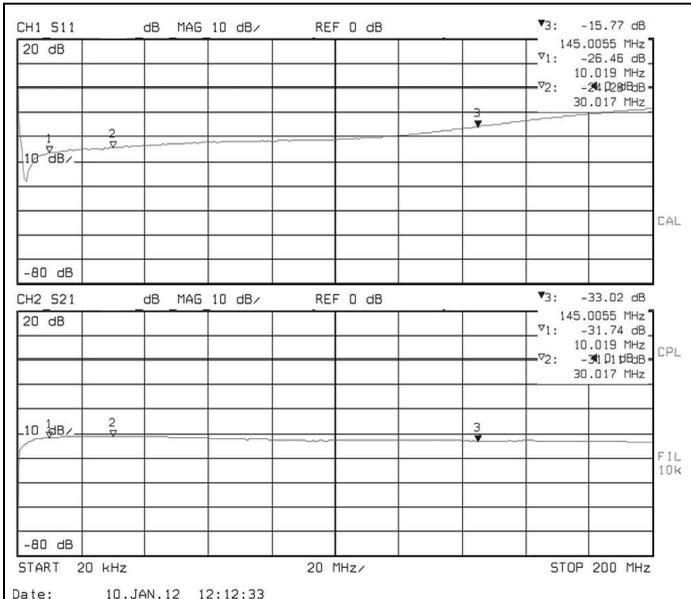
Fig 22: RF switch in the 20dB attenuation position.

attenuation position, i.e. St3, St1, St4 and ST9 are fed with +12 V. The transmission loss (S21) is approximately 43dB, 3dB higher than expected; the reflection coefficient (S11) is better than –20dB in the lower frequency range. The notch has

disappeared; the attenuation becomes good at 145MHz.

Fig. 24 shows the RF switch in the 40dB attenuation position, i.e. St3, St2, St5 and ST8 are fed with +12V. The transmission loss (S21) is approximately 53dB, 3dB
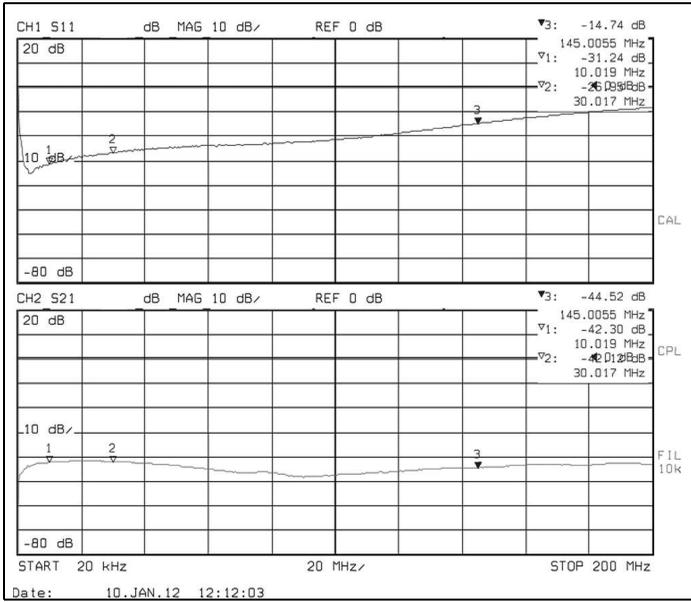
Fig 23: RF switch in the 30dB attenuation position.
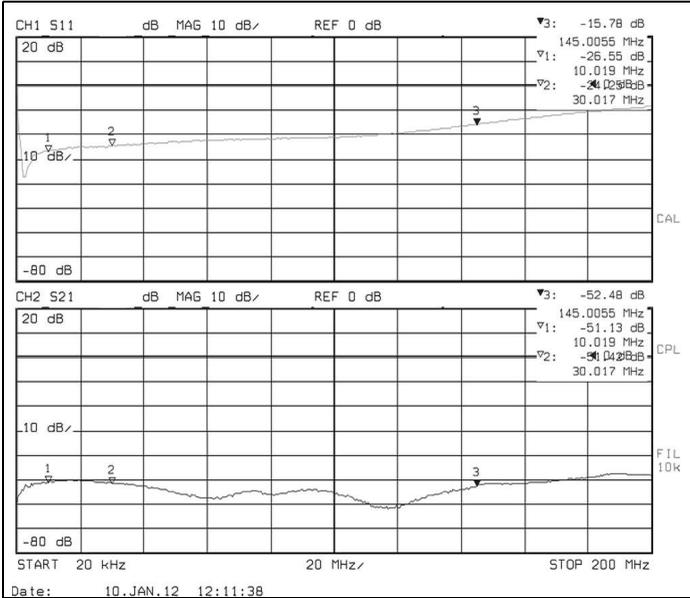
Fig 24: RF switch in the 40dB attenuation position.

higher than expected; the reflection coefficient (S11) is better than –20dB in the lower frequency range. The notch in attenuation is about 100MHz; the attenuation becomes good at 145MHz.

Fig 25 shows the RF switch in the 50dB

attenuation position, i.e. St3, St1, St5 and ST9 are fed with +12V. The transmission loss (S21) is approximately 63dB, 3dB higher than expected; the reflection coefficient (S11) is better than –20dB in the lower frequency range. There is a weak
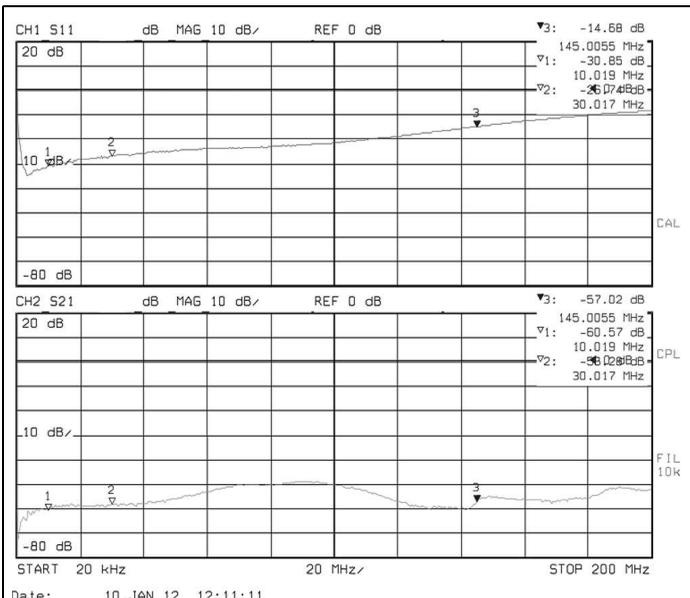
Fig 25: RF switch in the 50dB attenuation position.
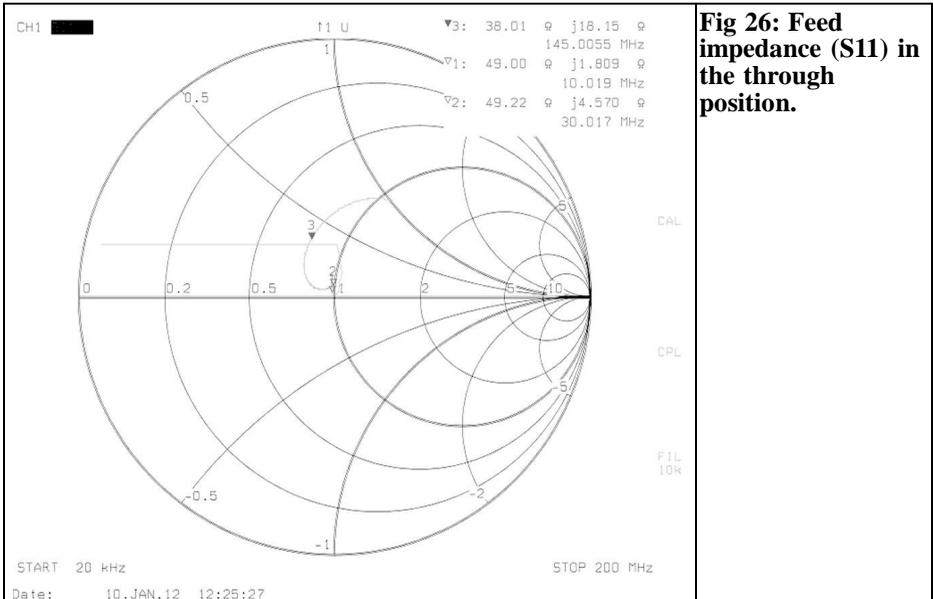
Fig 26: Feed impedance (S11) in the through position.

notch in the attenuation at about 70MHz, the attenuation is sufficient at 145MHz.

Fig 26 shows the feed impedance (S11) in the through position. The feed impedance falls with frequency to 38 +j18 ohms at 145MHz.

Fig 27 shows the feed impedance (S11) in the 50dB position. The feed impedance falls with the frequency to 44 +j16 ohms at 145MHz

It is noticeable that in both cases an inductive portion of the feed impedance
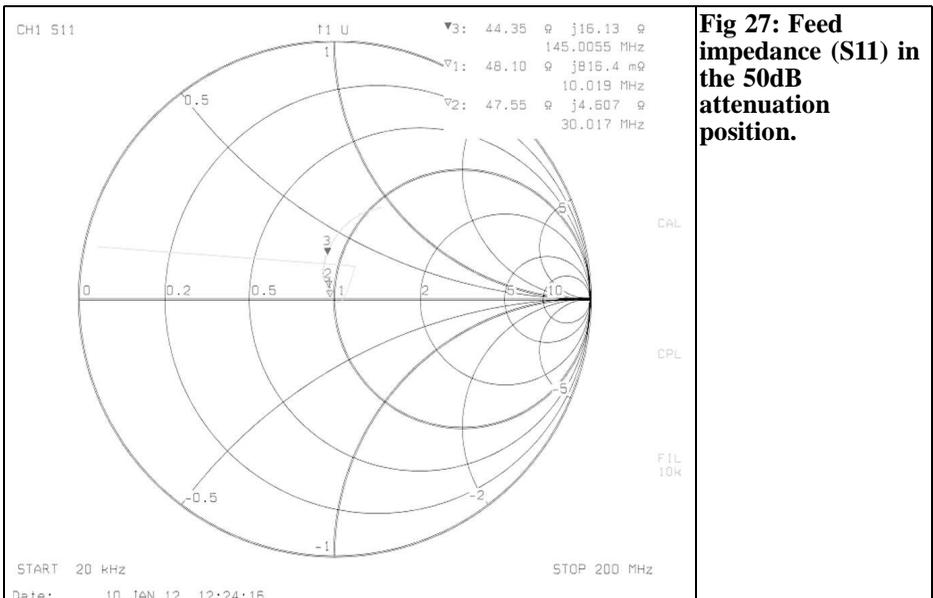
Fig 27: Feed impedance (S11) in the 50dB attenuation position.
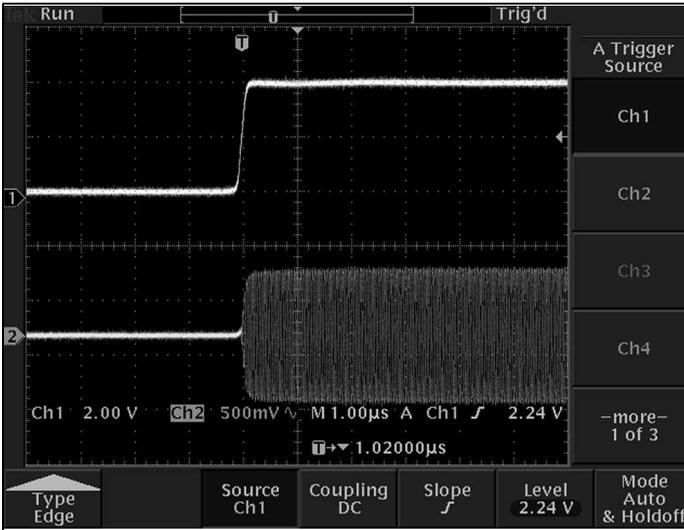
Fig 28: Switch on behaviour of the RF switch with a 9MHz, 0 dBm signal.

rises at higher frequencies.

Fig 28 shows the switched RF signal (9 MHz, 0dBm) in the through position. The upper curve is the switching signal, the lower curve the switched RF signal.

There is not recognisable over-swing at the switching point. The switching time is probably due to the CMOS IC´s and the discrete component driver circuit.

Fig 29 shows the switching off of the RF

signal; the upper curve is the switching signal, the lower curve the switched RF signal.

The relatively long fall time is probably due to the CMOS IC´s.
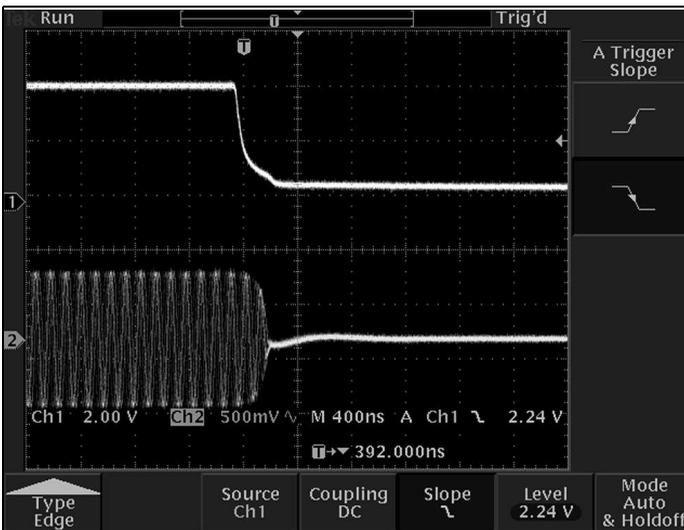
Fig 29: Switch off behaviour of the RF switch with a 9MHz, 0 dBm signal.

Table 1: Parts list for RF switch PCB.	
C1, C2, C3, C6 - 9, C11, C13 -17, C20 - 22	100nF
C4, C5, C10, C12, C18, C18,	10pF
R1, R3, R36, R38	220Ω
R10	68Ω
R4, R5, R8, R11, R14, R15, R17, R21, R22, R25, R25, R29, R33, R34, R39	1kΩ
R12, R26, R27	560Ω
R13, R6	100Ω
R16, R23, R28, R35	62Ω
R7, R9, R18, R19, R30, R31, R2, R37	10kΩ
R20, R32	22Ω
	240Ω
L1, L3, L5, L7, L9 L11	1μH
L2, L4, L6, L8, L10, L12	10μH
D1 - 30	BAT 18
TR1 - 6	ADT1-1WR
Bu1 - 2	SMA
ST1 - 9	1.3mm pin

Table 2: Parts list for control PCB.	
C1, C2	100nF
C3	22μF
R1, R5, R6, R9, R11	100kΩ
R2, R3, R4, R8, R10, R12, R13 - 17, R21, R22, R26	10kΩ
R18, R19,	1kΩ
R22 - 25	10Ω
R27, R28	2k2
R7	47kΩ
D1 - 4	IN4148
IC1	NE555
IC2	CD4023
IC3	CD4011
T1 - 3, T5 - 7	BC548c
T3, T8	BC558c
ST1 - 9	1.3mm pin
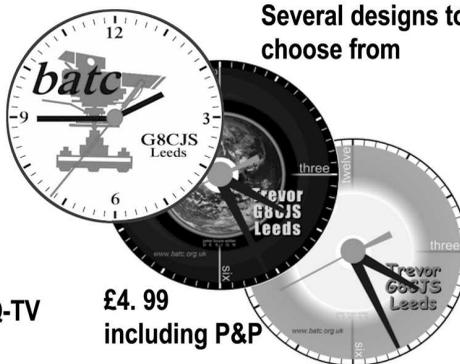
Gunthard Kraus, DG8GB

VLF receiving with an Active Magnetic Antenna and PC sound card interface, part 2

Continued from issue 4/2012

7.

Third project: Antenna using the principle of the field strength meter

This principle is investigated in the following paragraph. The basic circuit is a shorted antenna with the short circuit current measured as shown in Fig 22.

Preparation: only half of the coils were used producing only half the antenna voltage induced by the far field compared to the previous ferrite antenna. But the inductance decreases to approximately one quarter (exactly 1.09mH) and thus the short circuit current and output voltage should rise to double the value from the 4mH coil (see Chapter 4).

The input of an inverting operational amplifier acts as a short circuit. Its voltage gain at 77.5kHz can be determined as follows.

$$Amplification = -\frac{R_{GK}}{X_L} = -\frac{22k\Omega}{531\Omega} = -41.5 times$$

The 22kΩ is the resistor in the feedback path of the amplifier and the 531Ω is just the reactance of the 1.09mH antenna at 77.5kHz because losses no longer play a part. The amplifier has a 50Ω output and the overall gain for the connection to the spectrum analyser is reduced by half (i.e. 20.75).

A comment on the loss:

There is no need to worry about quality of the coils and its influence on the related losses because very quickly the inductive reactance of the antenna, that determines the short circuit current, is much greater than the loss resistance. There can be problems at low frequencies: the reactance continuously decreases
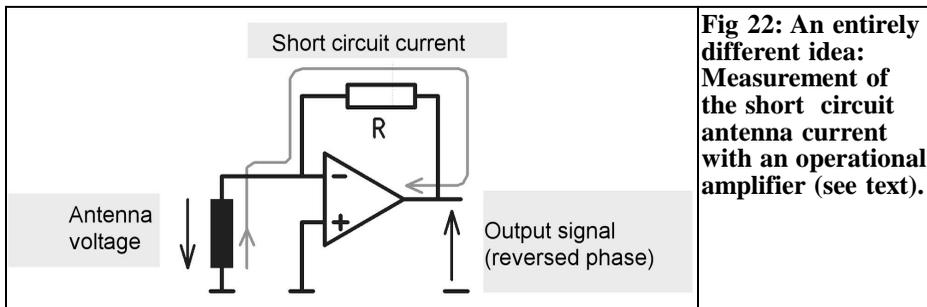
Fig 22: An entirely different idea: Measurement of the short circuit antenna current with an operational amplifier (see text).
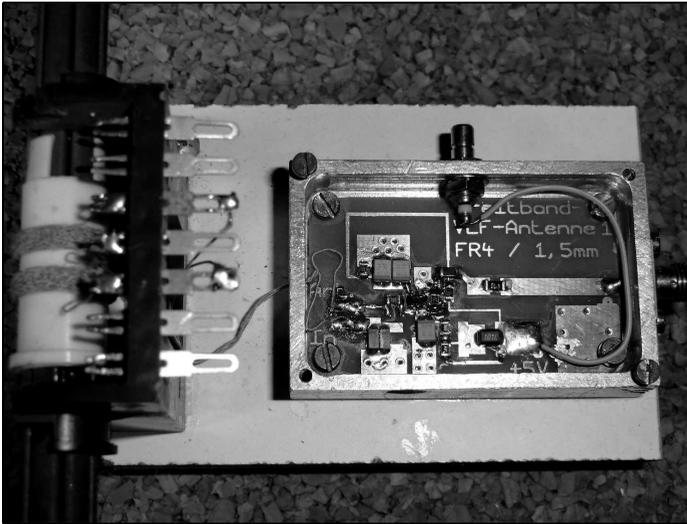
Fig 23: There were no great efforts for the practical application of this broadband system.

and the gain continues to rise according to the above formula. Therefore an additional 10Ω resistor has been inserted in the connection between the antenna and amplifier input. This limits this increase to high final value, with the 6Ω loss resistance of the coil this can increase to: $V_{final} = (0.5) \times (22000/16) = 687.5$. Let's see if that is acceptable with the strong 50Hz fields around.

A PCB was designed with dimensions of 30 x 50mm and fitted into a machined aluminium housing with the necessary connectors (SMA for the RF output / SMB for the +5V supply voltage). Fig 23 shows the overall arrangement with lid removed. The circuit diagram is shown in Fig 24. The PCB layout is shown in Fig 25.
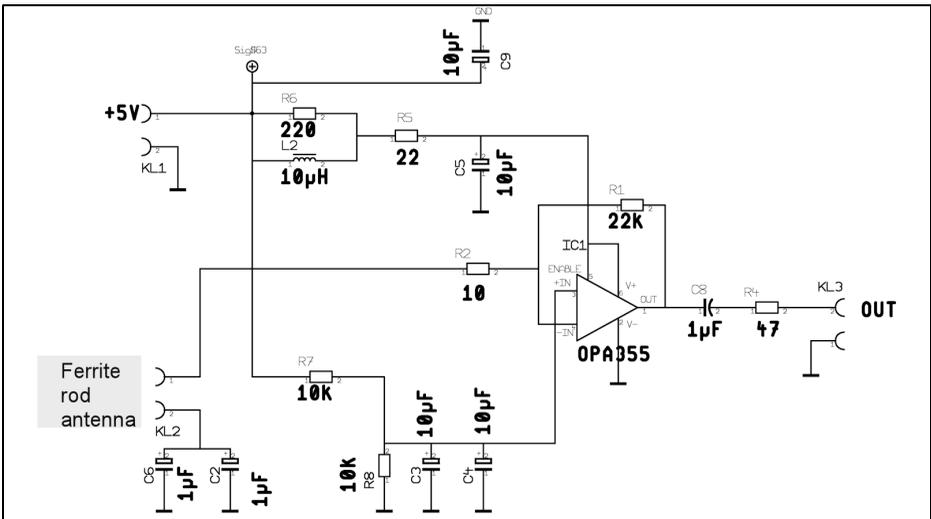
Fig 24: The circuit is quite simple.
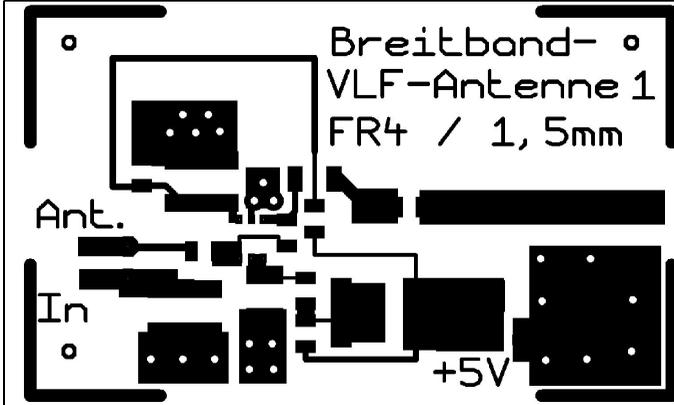
Fig 25: The PCB for the broadband amplifier.

Now, we are naturally interested in the reception result. DCF77 produced a level of -71dBm = 63µV that is at least 40dB less than with the two selective antennas - also due to the resonance peaks that are no longer present! From the simulation shown in Fig 26 this gives an induced antenna voltage of 4µV. Theoretically this should be half of 6.6µV i.e. 3.3µV, but this difference of 20% was not traced.

Nevertheless, the receiver works and the stability of the output voltage from 10kHz to 200kHz to is shown in Fig 26.

8.

Fourth project: Repeat project 1, but with less antenna coil inductance

After project 3 there was a ferrite antenna with half the number of turns and one quarter of the inductance available that made me curios: what results would come from the "simple" active antenna with the elaborate isolation amplifier from project 1 – that should be quickly
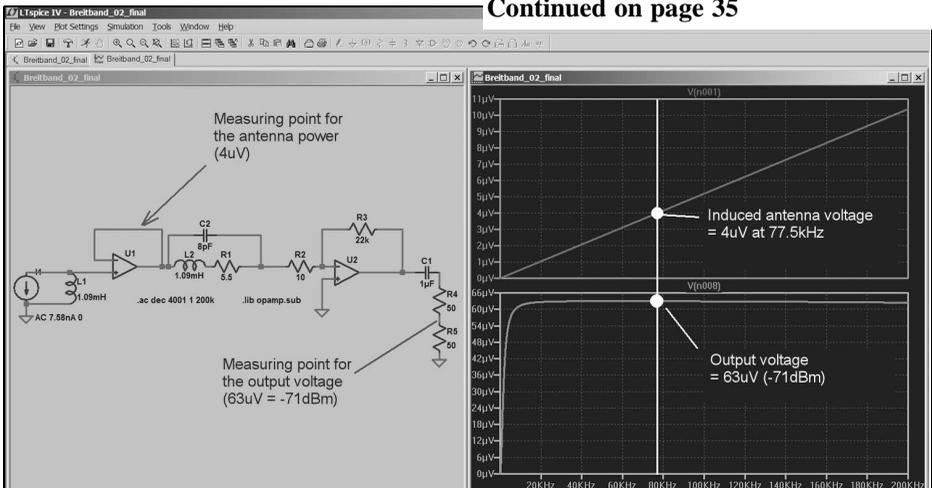
Fig 26: The simulation shows the promised constant output voltage over a wide band for a constant field strength.

Index of Volume 44 (2012)

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2005	Y	Y	Y	Y
2004	Y	Y	Y	Y
2003	Y	Y	Y	Y
2002	Y	Y	Y	Y
2001	Y	Y	Y	Y
2000	Y	Y	Y	Y
1999	Y	Y	Y	P
1998	Y	Y	Y	Y
1997	Y	Y	P	Y
1996	Y	Y	Y	Y
1995	P	Y	Y	Y
1994	Y	Y	Y	Y
1993	Y	P	P	P
1992	Y	Y	Y	Y
1991	P	P	P	P
1990	P	P	P	P
1989	P	P	P	P
1988	P	P	P	P
1987	P	P	P	P
1986	P	Y	Y	Y
1985	Y	Y	Y	Y
1984	Y	Y	Y	Y
1983	P	P	Y	Y
1982	P	P	P	P
1981	P	L	P	P
1980	P	P	P	P
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1974	Y	P	P	P
1973	P	P	P	P
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1970	P	P	P	P
1969	P	P	P	P

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determined?

So, the capacitor was increased to regain resonance at 77.5kHz. With the 1mH (exactly: 1.09mH) reduced inductance a value of 3.86nF was necessary that was easily achieved using a 3.3nF in parallel with a 560nF capacitor (Styroflex versions).

The subsequent reception attempt provided few surprises but rather a confirmation of the considerations:

- The output level after the first stage amplifier with minimum gain fell from -31dBm to -38dBm.
- It should theoretically be a 6dB decrease due to the halved number of turns but a decrease by 7dB can also accepted.

9.

Additional project: Electric shielding

What does additional electrical shielding give to an antenna? Theoretical and practically this should give fewer problems from nearby electrical interference fields and hence a better signal to noise ratio. We must not forget to ground this screen carefully otherwise it is nothing at all. It should not be a closed tube around the antenna; it must be cut lengthwise otherwise the resulting short circuit winding takes away all reception (I tried)!

There is something else to note that was confirmed by an experiment:

In all three projects the current that flows generates a magnetic field in the ferrite antenna. Interestingly, the level at the amplifier output for the same level decreases with decreasing diameter of the shield. Surely, this is caused by the ever increasing eddy currents produced by the antenna current in the shield that causes a decrease in the quality. Increasing the

diameter so that the original level is reached requires that the diameter of the shield be between half and the entire length of the rod. Referring to the beginning of the article Fig 1 shows the answer. It is at that point where the first lines apparently turn in the direction of antenna and produce a portion of the voltage that in turn leads to a current in the winding of the antenna.

10.

Fifth project: Three coil antenna for 136 kHz

The instrument amplifier developed for project 2 with the OPA355 is very "good natured" and broadband (it has been tested up to 10MHz and shows a flat frequency response). So replacing the antenna with a 135.7 – 137.8kHz amateur radio antenna is easy.

10.1 The circuit

The number of turns on the 4mH ferrite antenna must be reduced by a half; even at 77.5 kHz the 2kΩ reactance was very high. The inductance decreases by a quarter and the accurately measured value is 1.09mH. Two of the four part windings were removed and the centre tap relocated to between the two remaining windings.

Next, the quality was measures as described in chapter 5.3, the result was no surprise:

Q = 177 was the value measured at 136.7kHz including a series loss resistance of approximately 5.3Ω . This quality value was also measured for the 77.5 kHz project.

Two pot core inductors were wound for L = 1.09mH. These were the same design as in the first two projects (core material = N28 / AL value = 250). They were wound with 66 turns; don't forget the
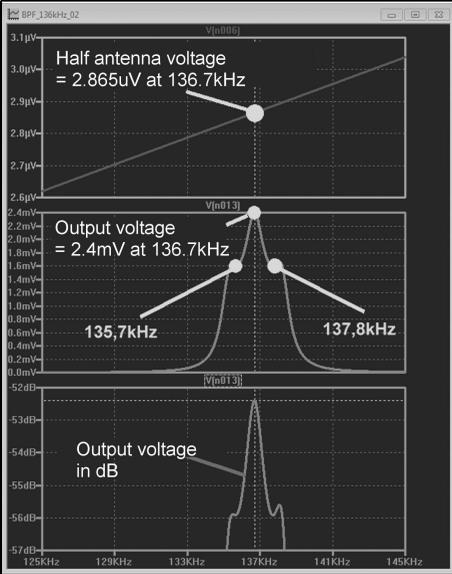
Fig 27: The 136KHz band requires a wider bandwidth so a lot of work was required but a spike still remains.

centre tap. The subsequent measurement of the quality found Q = 361 (with R_{series} = 2.6Ω), that fits well with the value of Q = 367 at 77.5 kHz in Chapter 6.2.

Now came the hardest part; the simulation diagram component values for the filter had to be changed until a fairly useful "filter curve" for the frequency range from 135.7kHz to 137.8kHz was found. Damping resistors across the coils were necessary to give the required range and to reduce the extreme slope in the middle of the band. The compromise curve is shown in Fig 27 with the remaining peak of 3.5dB at 136.7kHz (bottom diagram) that probably has to be accepted. Compare this with the curve for the 77.5kHz version in Fig 20: the ripple is larger but it is only receiving a single narrowband signal for DCF77, therefore plays no major role.

In the simulation diagram Fig 28 only the 3 damping resistors are added to the
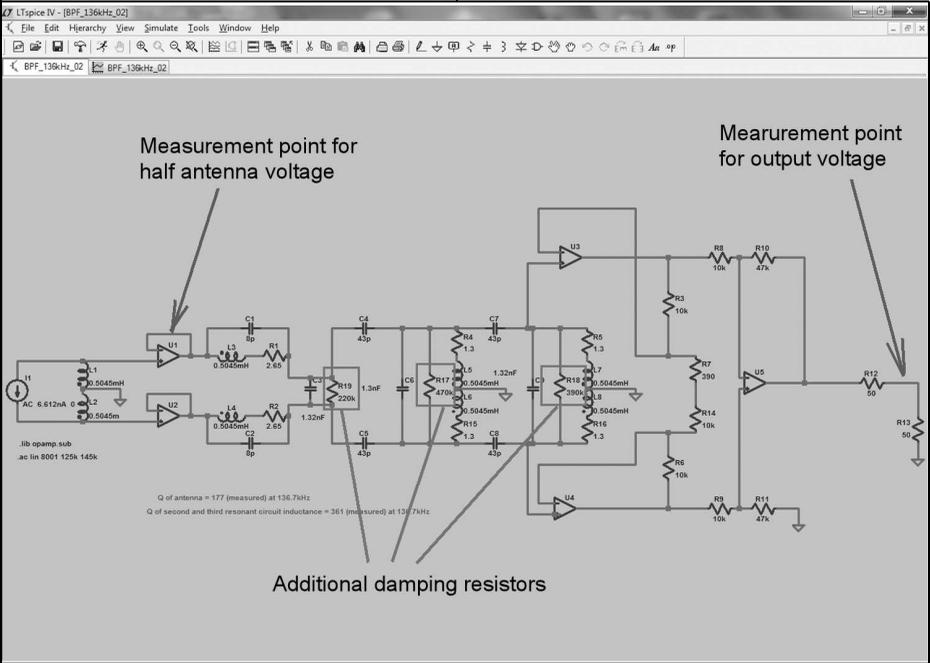
Fig 28: The new simulation diagram for 136kHz.

77.5kHz diagram. Of course the values of the inductors, loss resistance, circuit capacitors and the coupling capacitors in the filter are changed for 136.7kHz.

Thinking of the induced voltage of the antenna compared to 77.5kHz. The number of turns in the winding was halved resulting in only the half voltage. The frequency was increased by a factor of:

136.7kHz / 77.5kHz = 1.764

increasing the induced voltage accordingly by this factor. At the same field strength compared to 77.5kHz this gives

0.5 x 1.764 x 6.5µV = 5.73µV.

The 4mH inductance has been reduced to 1mH (due to the half number of turns). The constant current source at the input of the simulation diagram must be adjusted to give exactly the newly calculated 5.73µV at 136.7kHz (or 2.865µV as half voltage to ground). This is a current amplitude I of 6.612nA. Of course the output voltage of 2.4mV is decreased compared with the 77.5kHz version. But this is really not a problem: all measurements and simulations have been made so far with the minimum gain setting and there is enough gain in reserve to compensate.

Fig 29 shows the circuit diagram with the changed component values and recognisable circuit changes to be used in the next chapter.

10.2. A tour of the 136kHz band

After the circuit was built and the coils matched to give the required filter curve it was exciting: what will be found on this band with a spectrum analyser? You should know exactly what antenna power corresponds to what level on the analyser screen as follows:

For the minimum gain setting an antenna signal of 1µV and f = 136.7kHz produces a voltage at the analyser input of 2.4mV / 5.73 = 419µV. This corresponds to a level of -54.5dBm.

That seems small but it is no problem. Turn up the gain using the potentiometer so that an induced antenna voltage of 1µV gives an output signal of -40 dBm = 2.24mV. Such voltage values still don't stress the spectrum analyser.

So the arrangement is quite optimistic for scanning the 136kHz band but except for an increased noise level nothing was found! The measurement bandwidth was reduced as far as possible and the sweep time reduced accordingly then some tiny and barely visible "spikes" in the noise were found in the now extremely slow display. An amplitude calculation using the measured level was still under 0.1µV and thus it became clear:

A sweep of the 136kHz band with a spectrum analyser is not the correct technique because the extremely slow sweep rate of signals such as QRSS telegraphy cannot be seen because the amplitude history in time of the signals is lost.

Only one evaluation technique remains, DSP with a modern waterfall display program using a PC sound card!

10.3. Additional developments required for VLF reception using a sound card

The 136kHz amateur band is too high even for external USB sound cards with a 192kHz sampling frequency. There are sound cards (better: professional sound machines...) with a 384kHz sampling frequency but at an horrendous price. So the next idea was frequency conversion. This idea was just right after some pondering:

The operational amplifier in the output stage has a CMOS compatible "enable" connection. Using a 125kHz square wave signal suddenly makes it a single ended mixer. The large conversion loss associated with this method can be compensated easily using the gain potentiometer and the difference frequency at the output of the "mixer" is now at 136.7kHz - 125kHz = 11.7kHz. This value can be processed

Fig 29: The complete circuit diagram for 136kHz that was used to develop the simulation diagram.
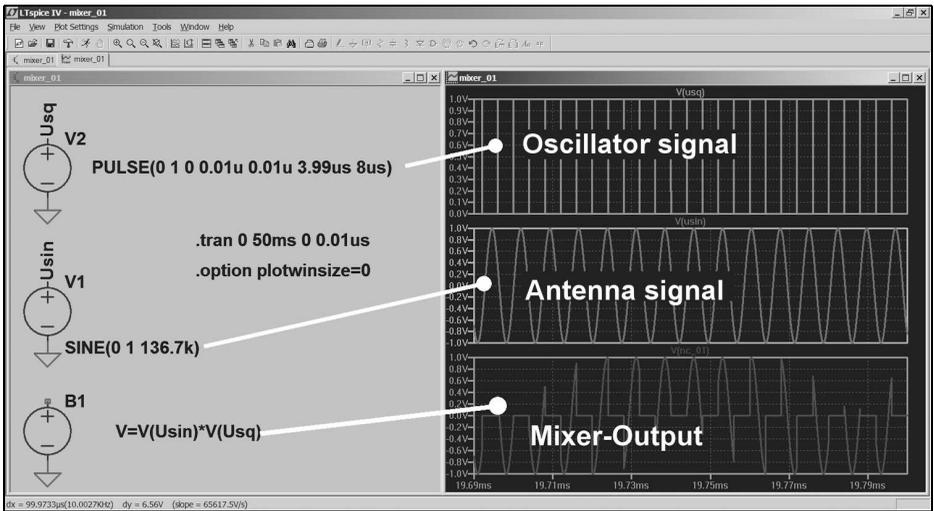
Fig 30: How to add a mixer to the amplifier simulation.

easily by all modern sound cards because the sampling frequency is usually adjustable to either 44.1kHz or 48kHz.

A spice simulation was immediately launched to verify the idea. A "bv" voltage source is used as a single ended mixing stage multiplying a sinusoidal signal at 136.7kHz (peak = 1V) with a square wave signal at 125kHz (minimum value = zero V / maximum value = + 1V). Fig 30 shows this circuit with the result in the time domain. The frequency spectrum generated can be seen in Fig 31 and now it is not difficult to identify the desired difference frequency and the other signals produced.
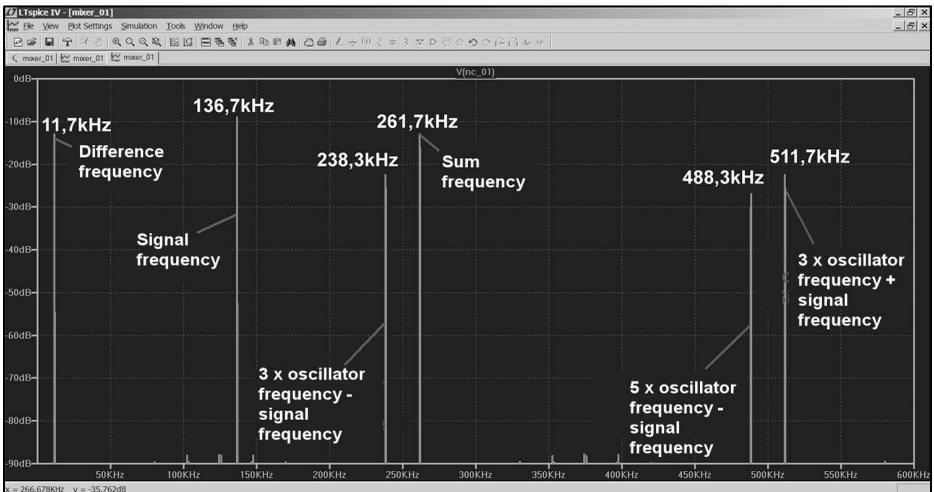
Fig 31: The frequency spectrum produced by the operational amplifier working as a mixer using an oscillator frequency of 125kHz.
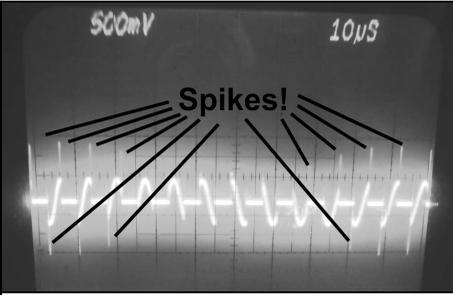
Fig 32: This clearly shows the culprit of the problems with the PC sound card.

A tip:

If you are interested in the exact value of the conversion loss in the conversion, show the results using an FFT chart with "linear representation of the amplitude". First convert the sine wave voltage at 136.7kHz, the FFT produces a single line with an amplitude of 0.707V - the RMS value for the selected peak value of 1V. Then repeat the FFT on the output signal and the sum and difference frequency, each gives an RMS amplitude of 225mV. This gives a conversion attenuation of 10dB that was confirmed by a

controlled measurement of the circuit.

The practical application is as follows: the enable pin is separated from the supply voltage on the PCB with a scalpel and fed directly to a separate port on an SMA connector. So that the circuit reverts to amplifier operation if the external oscillator signal is removed the enable pin is connected to +V via a pull-up resistor of 100kΩ.

A TTL function generator at 125kHz was connected after this change and the output signal of the mixer stage looks exactly as required by the theory (according to Fig 30 and 31).

But the hope of a quick solution proved to be naive because the sound card or the program "Spectrum Lab" produced hysterical reactions when the mixer output was connected to the microphone input, not just the desired difference frequency of 11.7kHz. It took some effort to find out the cause that can be seen on a screen shot taken on a 100MHz oscilloscope in Fig 32:

Incredibly narrow, barely recognisable needle pulses (spikes) were produced at

Fig 33: Once again: Design of a Tschebyschef lowpass filter but this time with an online calculator on the Internet.
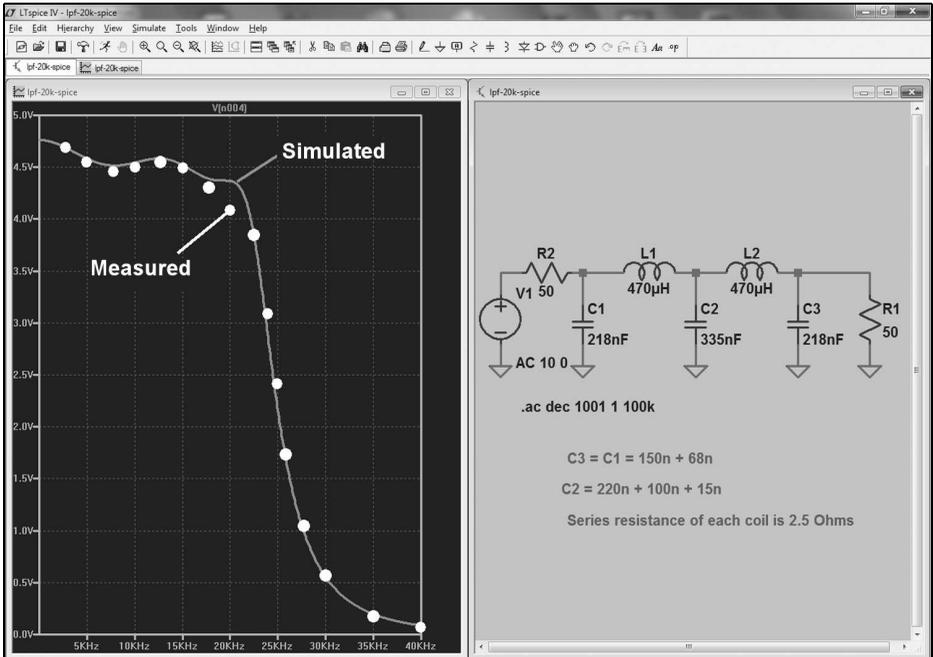
Fig 34: Theory and practice of the 22kHz lowpass filter.

every on and off switch of the mixer, sometimes with amplitudes above 1V! That was probably just too much for the sound card input.

10.3.1 A lowpass filter for 20kHz required

An extra 20kHz lowpass filter with Z = 50Ω was required as quickly as possible to limit these needle pulses and to ensure that only the desired difference frequency of 11.7kHz is fed to the sound card. The following filter data was used:

- Impedance Z = 50Ω
- Cutoff frequency approximately 20 - 22kHz
- Filter degree N = 5
- Tschebyschef type with ripple 0.3dB

But in the meantime a new PC with Windows 7 had arrived in my workshop. Traditional DOS filter programs such as "fds" would suddenly no longer run and

it was necessary to design the filter with "Chebyshev PI LC Low Pass Filter Calculator" from the Internet (Fig 33). It was also necessary to overcome the problems:

- The first simulation showed that the value of the inductance required was 500µH. The cutoff frequency is not critical so 470µH was used because this is a standard inductor value in the E6 series. It looked like a hole mounting 1W resistor, not expensive but had a quality factor of approximately Q = 40 - 50 (supplier: Reichelt Elektronik, axial design from the L-HBCC Ferrites range, length approximately 15mm).

- The circuit capacitors used were wonderful metallised polyester SMD capacitors for this frequency range that have high quality values (supplier: Reichelt electronics). Unfortunately the values available in the E6 series mean that up to three capacit-

Fig 35: The two versions of the lowpass filter that were built.

ors must be connected in parallel for the values from the filter calculator (desired maximum deviation: 1 – 1.5%).

The final combinations were:

L1 = L2 = 470µH

C1 = C3 = 218nF (150 + 68nF)

C2 = 335nF (220nF +100nF + 15nF)

Simulation and measurement results are shown in in Fig 34. Because an inductance value of 470µH was used the desired ripple corner frequency of 20kHz has increased slightly but it is enough (along with the extremely good selection of a three stage filter at the antenna input) at a sampling frequency of 48kHz to meet the Shannon Nyquist condition (no ADC signals allowed above half the sampling frequency). It is good to see that at frequencies up to about 15kHz measurement results exactly match the simulation with a coil with 2.5Ω series resistance as specified in the data sheet. Then losses increase in the wire (skin effect) and in the ferrite material. The measurements were made with a function generator (RI = 50Ω) and an oscilloscope measuring the voltage across the 50Ω terminating resistor. Therefore linear representation is used for the simulation results.

Fig 35 shows the two prototype filters manufactured in two different housings.

The longer version of the housing is intended for a 30mm x 130mm PCB space enough to separate the two inductors because they do not have an enclosed magnetic shell. The axial bobbin design has a strong stray field and therefore mutual coupling should be avoided. But the fears were unfounded as shown by the complete success of the smaller version PCB (30mm x 50mm). Result: completely identical filter curves but now a wonderfully compact module. A wide frequency test showed no effect on the attenuation up to 5MHz.

10.3.2. The conversion oscillator

Happily this problem was very easy to solve because the note "CMOS compatible" for the enable input of the OPA355 pointed the way. A commercial 1MHz quartz oscillator in metal housing was combined with a 74HC93. This contains four flip-flops connected in series and easily allows frequency division by 2, 4, 8 or 16. A quartz oscillator should have sufficiently small noise sidebands to prevent an increase in the module noise. A small jumper is used to select the desired division of the oscillator signal and connect it to the output. Thus the frequencies available are:

1MHz, 500kHz, 25 kHz, 125kHz and 62.5kHz

The last two values are just what you
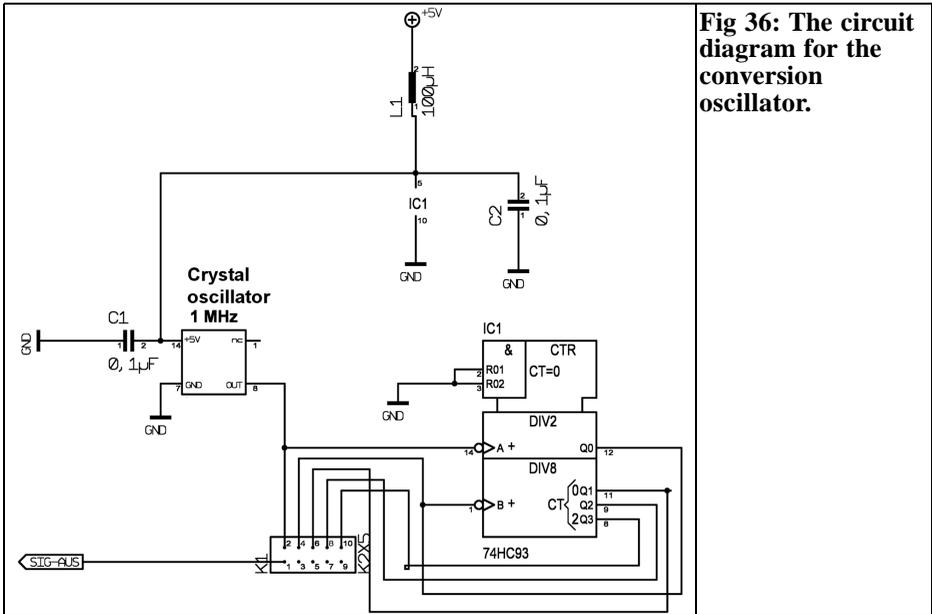
Fig 36: The circuit diagram for the conversion oscillator.

need for the 136kHz band or DCF77.

The circuit diagram is shown in Fig 36, the PCB layout in Fig 37 and the finished unit in aluminum housing in Fig 38. A compromise was required because SMD components were not used (quartz oscillator, connector) so some components were placed on the top side of an SMD circuit board.

10.4. The test run

The complete system for testing reception is shown in Fig 39, it consists of the following equipment. A signal generator with an internal resistance of 50Ω feeding a short circuit loop with a V_{output} of 4.5mV (V_{output} for an output level of -40dBm) and thus produces a current of approximately $90\mu A$ at 136.7kHz in the loop. The 11.7kHz output signal displayed on the oscilloscope screen has a peak to peak value of about 10mV and so the PC sound card and the software were satisfied.

(If you take a closer look at the picture the quality of the filtering can be seen: the whole system works fine at a distance of 1m from the 19inch CRT monitor of
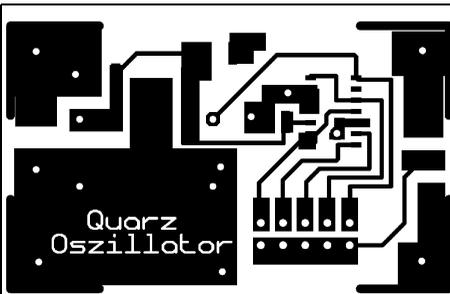
Fig 37: The PCB layout for the conversion oscillator.
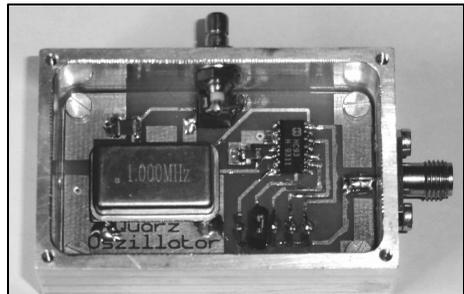
Fig 38: The prototype conversion oscillator.
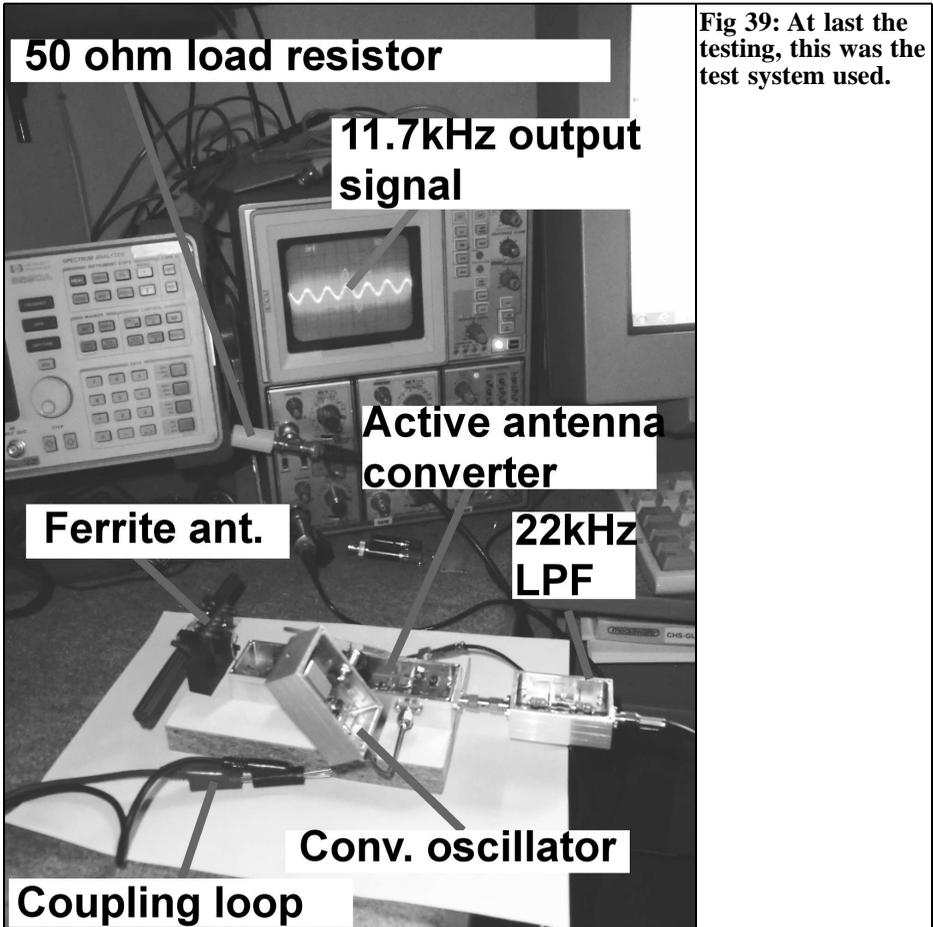
the PC with its stray magnetic fields).

The sensitivity of the new system with filtering at 136.7kHz should be calculated. The specification in Chapter 10.1 can be adjusted as follows with reference to the dB filter curve in Fig 27 to correct for other frequencies in the band:

- Set the minimum gain for an antenna signal of 1µV and a frequency f = 136.7kHz giving a voltage at the unfiltered output of 2.4mV / 5.73 = 419µV. This corresponds to a level of -54.5dBm.

- This signal on 11.7kHz produced by the gives an additional attenuation of 10dB. Thus the level at 11.7kHz is -64.5dBm at 50Ω resulting in an output voltage of 133µV for an antenna signal of 1µV at 136.7kHz.

10.4.1 Which program can be used for the PC?

The Internet is teaming with DSP programs and each has its merits. But after a thorough examination, some requests from "clever people" and some tests, the decision is no longer difficult:

- "Spectrum Lab" is the perfect machine. It is really a "Jack of all Trades" with additional packages and options for every conceivable problem, any mathematical possibility or any
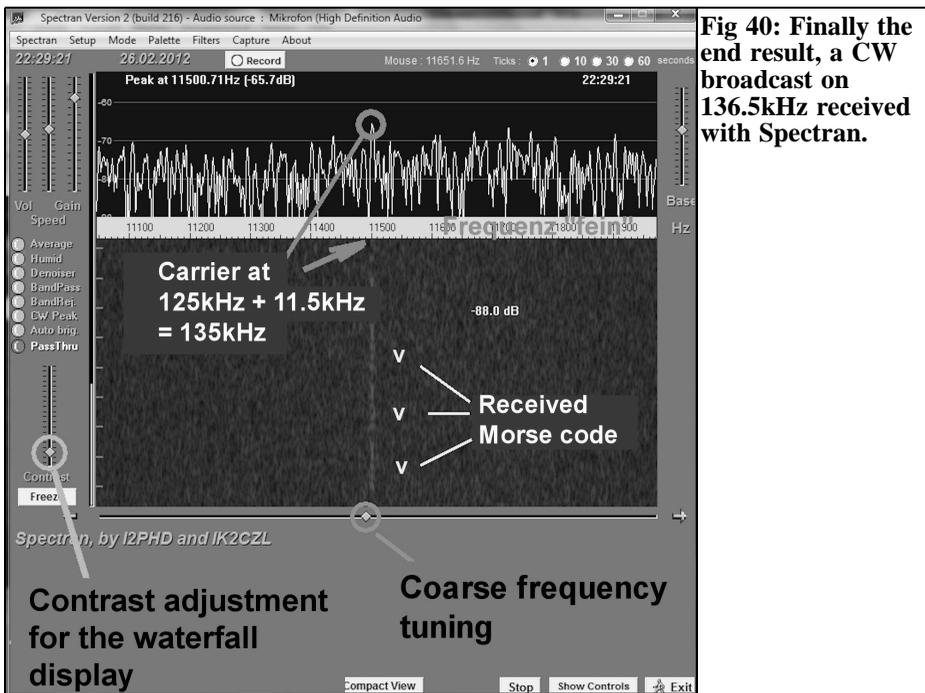
Fig 40: Finally the end result, a CW broadcast on 136.5kHz received with Spectran.

mode. But it is not for the casual user or novice because it has many facilities only suitable for an expert.

- To cut a long story short: to start you need a beginners program with only the most important functions but simple and self-explanatory operation. "Spectran" is such a program that I used to produce the screen shot shown in Fig 40 that was made after a training period of one hour!

These are some tips:

- In addition to rough selection of the frequency (using the slider control to quickly go to a particular area of interested) the frequency can be finely adjusted. Place the cursor in the spectral display area (marked on Fig 40 with "Fine frequency") hold down the left mouse button and move the mouse back and forth. The scale then follows the movement of the mouse.

- In addition the signal is shown on the waterfall display. The correct setting can be found very quickly.

10.4.2. Reception with Spectran

The screen shot shown in Fig 40 did not appear immediately by itself without any intervention on the PC screen. The following explanation describes the correct steps:

- The first task is on the Windows taskbar, use "Setup" and select "sound card" then select "Input". That must be set correctly otherwise nothing works. The next two pictures show how it go further.

- Fig 41 shows how to set the appearance of the calculated spectrum with "Base" and "Gain".

- Figure 42: this button to hides all the essential settings. They begin with the choice of the sampling rate of 48kHz. This gives a base band range of less than 24kHz.

It continues with the measurement
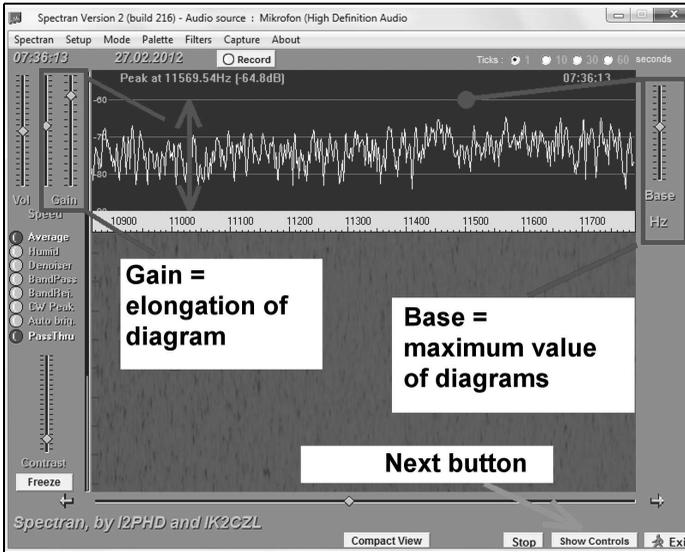
Fig 41: Adjusting the Spectran display to meet your own requirements.

bandwidth that can be changed, the current setting is 1.5Hz.

Followed by "Frequency offset" = frequency shift. The frequency display can be corrected for the actual frequency of 125kHz oscillator.

With "Speed", you can change the flow of the waterfalls. Second markers are inserted in the waterfall.

Mean value calculation is carried out with the "average" function. This fishes out very weak signals from the noise. The drawback of this is that transitions of the CW carrier from on to off and vice versa are extended.

• The other buttons (Humid / De-noiser / Band Pass / Band Rejection / CW Peak / Auto) can be tested occasiona-
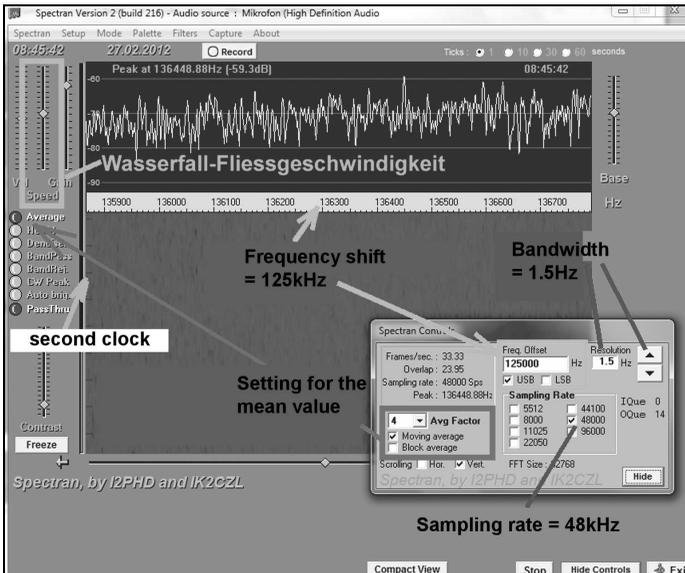
Fig 42: But everything works properly only with the correct setting.
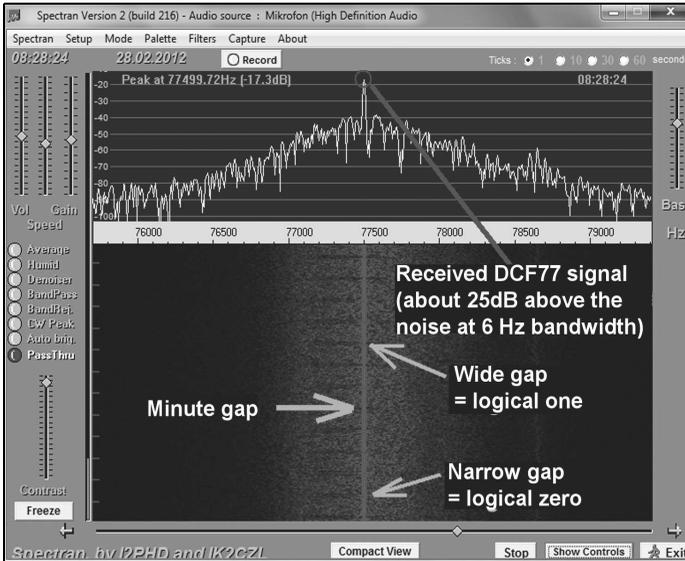
Fig 43: What worked at 136kHz still works at 77.5kHz.

lly, as well as all the other buttons. This also applies to the menu "Setup / Mode / Palette / Filters / Capture" on the Windows taskbar. There are many very interesting additional features but for now just enjoy the results of the setting shown in Fig 41 and 42 to produce a clean display.

11.

Conversion to a DCF77 receiver

After all the problems of receiving signals on the 136kHz band were solved I could not resist the psychological experimenter urge to convert the triple antenna DCF77 receiver. The work required was
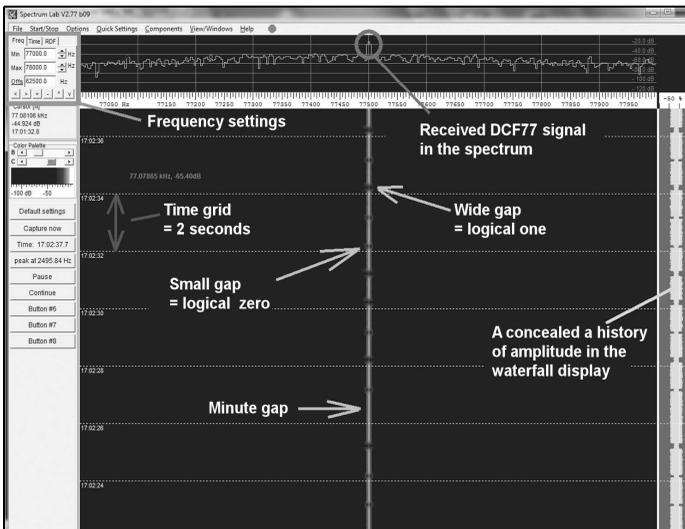
Fig 44: The next challenge: the same thing again, but now with Spectrum Lab.
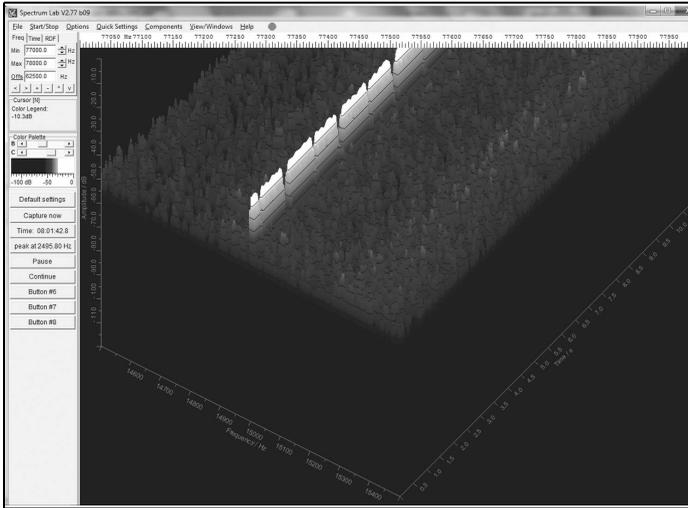
Fig 45: This 3D waterfall alone is enough incentive to persevere with this programme and to explore its possibilities.

acceptable; the enable connection from the final operational amplifier was disconnected from the power supply and fed to its own port with a 100kΩ pull-up resistor. A 22kHz lowpass filter (long version) was already built. The conversion oscillator module had to be setup by moving the jumper to "62.5kHz" and an offset of 62.5kHz set in Spectran. The result is shown in Fig 43 and this is a good reward for the effort.

Another possibility only arose when looking at this screen: the spectrum shown in Fig 43 shows the transmission curve of the triple bandpass filter in the noise. Even the smallest rotation of one of the matching cores can be seen immediately. Thus the fine adjustment and elimination of residual unbalance is a pleasure.

12.

What remains to be done?

There are many things to do, some of my ideas are:

- One idea that has been forming in my head for some time is the measuring of the local magnetic field strength using the received signal. There are

details about this, mostly on the Internet. You build a round frame with a one metre diameter coil that is fed with a precisely known and constant power. The magnetic field strength within such a framework is almost homogeneous and exactly predictable. The ferrite rod, with its circuit, is placed in the middle of the coil, calibrated and compared with the prediction (remember the beginning of this article? The theoretical diameter of circular "frame" gave rise to the parameters of the ferrite antenna used, a rod length of 14cm representing an area of 154cm^2. That would be an interesting comparison and thus a good control for these experiments). Major errors caused by the circuits in the milled aluminium housing inside the frame are not expected because aluminium is non-magnetic (correctly: paramagnetic) and does not affect the field distribution. But the eddy currents generated in this aluminium housing could weaken the field produced by exciting coil. This is a really interesting project.

- The broadband field strength meter from chapter 7 could be used with a modern "image reject mixer" and a

DDS oscillator to investigate the whole range of 10kHz to 200kHz. A new two-stage post amplifier with an external knob for manual gain setting could be used to compensate for clipping by strong electromagnetic fields.

- It is high time to use the "giant machine" called "Spectrum Lab". Fig 44 shows an interesting display with a completely new 3D waterfall display in Fig 45.

- Change the electronics for the next RF and microwave projects to deal with.

- Or ….
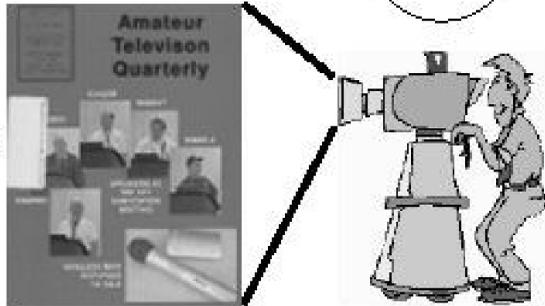
André Jamet F9HX

Beware of counterfeits!

Today, the temptation is great to acquire components and devices at costs well below the normal price. Indeed the Web gives us easy and efficient access to search and find even the rare birds.

Alas, what disillusionment! We do not always admit these setbacks being too upset by the mistake!

Several sources feed these sites with low priced items. Some are serious, such as obsolete stock, bankruptcy of a company, change of activity, etc. Even second hand components can be very valuable if their removal was done correctly. If this is a purchase on eBay the references must be carefully examined as well as the guarantees and information available on the seriousness of the seller. Check that defective products can be returned.

Unfortunately, other sources offer products at low prices. There is a "Hot bed" of simple mailbox advertisements that respond to the calls for quotations or provide a catalogue that claims to have rare or very expensive components at very attractive prices. The product delivered are often counterfeit that do not meet the characteristics corresponding to its designation. Any claim is impossible and the supplier has often disappeared!

I have discovered beautiful packs of SMA connectors with a white insert that is not Teflon®, or more subtle, with incorrect dimensions. I bought some TIM1011-15 L transistors supposed to give 15 watts at 10GHz. They didn't work and the brand name on the case was removable by a drop of saliva!

Fashion is an important area for counterfeiting as well as luxury objects. This causes huge losses from sale of branded items. An example is counterfeit champagnes and wines!

Counterfeiting abounds in tooling, domestic appliances and unfortunately the spare parts for automobiles. And this is the danger, because a bonnet that is too thin or made from poorly ribbed sheet metal will not play its role in a head on collision.

It is even more serious when manufacturers and distributors are the victim of counterfeit goods without their knowledge. A cascade of firms is involved in the path of a sale making it difficult to trace and possible cheat.

Professional electronics journals publish warnings for procurement of services to raise awareness of the problem. Even the critical sectors of aeronautical, space, military and medical are affected.

Without wishing to refer to scapegoats it must be noted that counterfeiting is almost officially developed in some countries.

For us, individuals and amateurs, the risk is mainly financial. But we do like to be safe. Therefore vigilance and dissemination of the information received on detected defects is important.

Marty Singer, K7AYP

A useful coax latching relay control circuit

1.

Introduction

Many of our VHF/UHF and microwave designs use a coaxial latching relay. The typical dual-coil part comes in many configurations. Characteristic specs include various latching coil voltages, SMA or N type connectors, SPST, SPDT switch configurations, and even multiple contacts requiring logical control. Manufacturing techniques have made these components excellent signal switching devices for VHF through microwave frequencies. Isolation between contacts reach -70 to -80dB. Appropriate connectors allow switching 1kW.

Latching relays in my parts stock require a voltage pulse for a specified time duration. Available data sheets show the following typical information:

- Pulse voltage: +12V, ±10%,
- Pulse duration, each latch coil: 20ms minimum, 1s maximum
- Time to switch, typical: 15ms
- Operating pulse current, typical: 0.2A
- Connector, SMA female, typical power rating: 100W maximum

Internally, they have two solenoid-like windings that control metallic wiper-arms for the relay contacts. All are enclosed in a permanent metal package. Manufacturers advise to only operate them with pulse excitation.

This article describes a latching relay pulse circuit used to switch coaxial relays. The circuit combines latching coil drivers with pulse generation. Drive signals only need to change voltage level and polarity to activate the relay. No external pulse-shaped forcing function is necessary.

2.

Circuit description

"Traditional" positive-going pulses implemented for relay activation increase amplitude to a voltage that remains for a given time and then returns to 0V. If the amplitude over time meets relay specification the latching relay toggles to its complementary contact connection.

For this relay driver circuit, each internal latching coil is activated from stored voltage on a capacitor. The capacitor potential is switched across a latching coil where it decays over time. If enough time and energy is present, the relay coil activates the contacts to their alternate position. The energy will be defined a "pulse", with additional design informat-
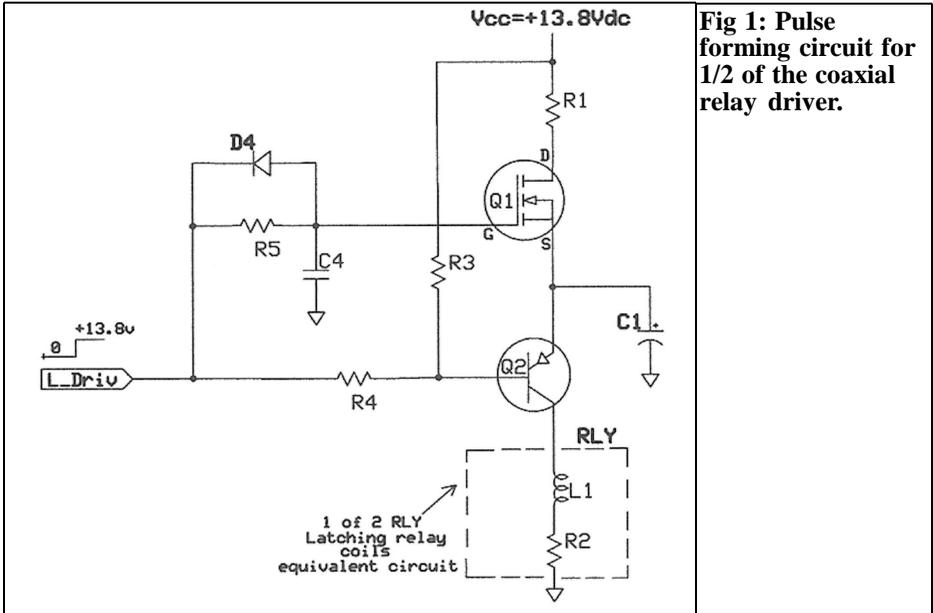
Fig 1: Pulse forming circuit for 1/2 of the coaxial relay driver.

ion in the circuit design descriptions.

The fundamental circuit configuration driving each relay latching coil is shown in Fig 1

The circuit consists of the two transistors Q1, Q2 and several passive components. A moderately large value capacitor, C1, is connected between the junction of Q1-Q2 and ground. One of two latching relay coils is represented in RLY by its inductor (L1) and resistor (R2) equivalent circuit components (surrounded by dotted lines). Other coil equivalent parameters exist, but should not affect circuit operation. The V_{CC} power supply for this circuit is +13.8V DC, commonly supplied by power supplies these days.

Two circuit operations are performed. First, a potential is placed into capacitor C1. Then, C1 is switched across the coil where its charge activates the relay while it decays from its initial value. Each operation will be discussed next.

First, N-Channel D-MOS-FET Q1 is enabled by a +13.8V, **L_Driv** input signal, greater than its $V_{GS(th)}$ specification. Device Q1 conducts and current flows

from V_{CC} through resistor R1, charging capacitor C1. If **L_Driv** remains +13.8V long enough it appears that C1 charges to almost +13.8V DC (99.24% of V_{CC}). These parts with V_{CC} form a conventional R-C charging circuit.

The second circuit operation performs "pulse" generation and RLY coil switching. It begins when **L_Driv** changes polarity to 0V. Device Q1 turns "off" ending the C1 charging cycle.

Transistor Q2 is biased to conduction using resistors R3, R4. Potential stored in capacitor C1 is connected across RLY coil (L1-R2) and decays, characterised by a timed exponential discharge with coil resistance R2. The amplitude of discharging voltage over time forms the relay coil energy pulse.

To insure that charging and discharging is independent of each other, resistor R5 and C4 delay Q1 turn-on until Q2 is disabled when **L_Driv** switches to +13.8V. Diode D4 discharges C4 when Q1 is disabled and C1 is switched to the latching relay coil (**L_Driv** = 0V).
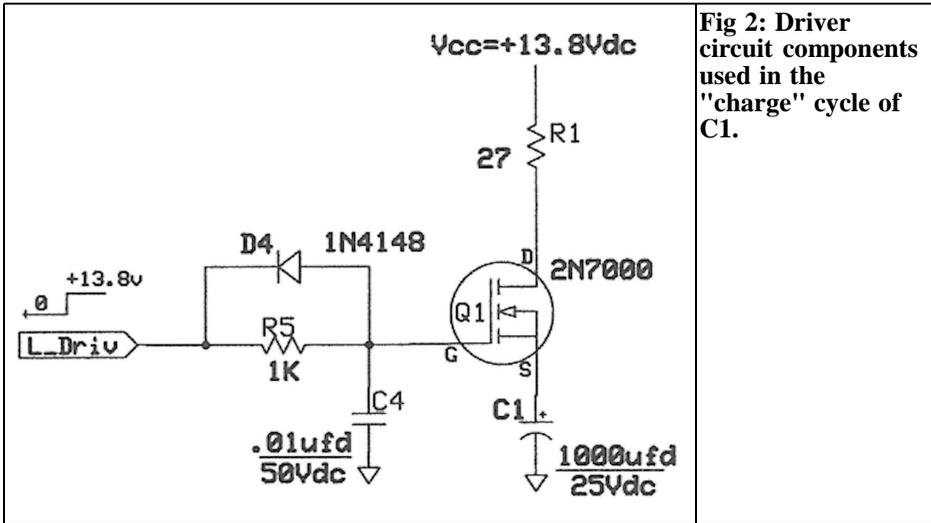
Fig 2: Driver circuit components used in the "charge" cycle of C1.

3.

Circuit design

3.1 Circuit Charging Cycle

For this and the following section the driver circuit in Fig 1 is separated into its charge, and discharge circuits. Circuit design is discussed and key component parameters are determined.

In Fig 2, components are shown for only the driver's charging circuit. The D-MOS FET is the switch between charging capacitor C1 and resistor R1 using the +13.8V DC power supply. Charging C1 involves two important operating conditions:

- Using N-Channel D-MOS FET technology, an average +2.0V gate-to-source voltage [$V_{GS(th)}$] difference exists to keep Q1 conducting (for my parts stock FET devices). The Q1 gate voltage from **L_Driv** is +13.8V. The capacitor C1 charging cycle will end when: (+13.8V) – (+2V) = +11.8V is reached. This voltage is labeled V_{C1}.
- In Section 3.2 with Fig 3, it is shown that C1 does not fully discharge when

switched across the latching coil by transistor Q2. There will be +2.7V remaining, called a "pre-charge" on capacitor C1. This voltage is labeled $V_{pre-charge}$.

From these conditions, voltage on C1 changes between +11. 8V (charged) to about +2.7V (discharged) during every C1 operating cycle.

Peak current through Q1 must be determined to select a FET. According to pulse theory, peak current at "time 0" for this circuit with R1 = 27Ω is:

$$I_{peak} = \frac{(V_{cc}) - (V_{pre-charge})}{R1}$$
$$= \frac{[(+13.8V) - (+2.7V)]}{27} = 0.411A$$

By using a D-MOS FET device, peak current through Q1 could be varied by the pre-charge voltage and resistor R1 [$R_{DS(on)}$ resistance of Q1 is not considered]. To determine the time taken to reach full charge (+11.8V) the value of pre-charge voltage and the +11.8V limit set by Q1 $V_{GS(th)}$ are determining factors. Calculations and testing show the time for C1 to reach five time constants is
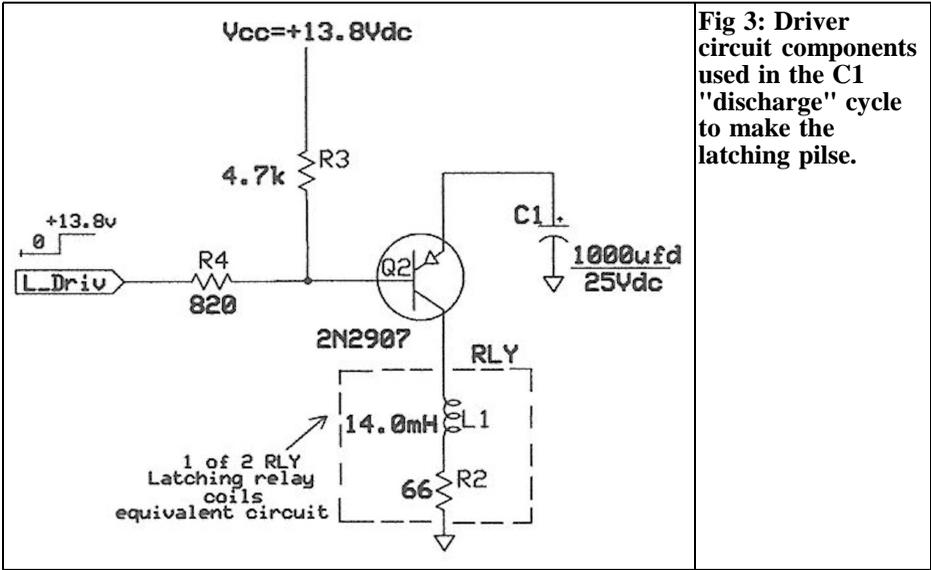
Fig 3: Driver circuit components used in the C1 "discharge" cycle to make the latching pilse.

about 60ms. This time delay is always necessary to ensure enough charged potential on C1 for pulse generation (excluding any mechanical constraints).

My part stock provided a 2N7000, N-Channel Enhancement-Mode, D-MOS FET. A few applicable specifications are: $V_{GS(th)}$ threshold = + 2.0V, typical; $R_{DS(on)}$ = 5Ω, max @ 0.5A; $I_{D\text{-pulse}}$ = 0.6A.

Peak power dissipation for R1:

$$P_{R1} = \left(I_{peak}\right)^2 \times R1$$
$$= (0.411)^2 \times 27 = 4.56W$$

Consider this wattage in your intended application i.e., the short time duration peak current is present and the driver circuit low operating duty cycle (both in active use during an operating session and how often your equipment with this pulse circuit is used). I cycled a driver circuit with a 27Ω, one watt, metal film resistor for 7 days, 24 hours per day at a 100ms circuit toggle rate. I felt no excessive component heating (over what a 1W resistor should be). Measurements

before and after the test period showed no change in resistance value. Choose your own "comfort level".

Remaining components in Fig 2 control Q1 switching. Driver circuit polarity changes toggle C1 between independent charge and discharge events. In that way, potential energy is repeatedly stored without extraneous current or leakage paths and repeatable voltage is applied across the latching coil. The R5-C4 combination causes about one µs time delay before **L_Driv** reaches the +2.0V $V_{GS(th)}$ turn-on threshold, ensuring the capacitor discharge cycle (described in Section 3.2) has completed. Diode D4 speeds Q1 turn-off so a charge cycle begins without any V_{CC} potential supplied to the discharge circuit.

3.2 Circuit Discharge Cycle

This section identifies Q2 operating requirements, its part number and values for R3, R4. Testing is described to determine reliable relay operation with minimum operating voltage.

Fig 3, shows only components involved with using the potential charged on C1 to pulse the latching coil. Components L1,

R2 represent the series equivalent circuit for a latching relay coil in RLY. Their values were determined by measuring seven different sets of latching relay coils (14 total coils) made by two manufacturers to see if this circuit design must consider large variation in their values. Inductance was measured to be ±3mH across the 14 coils. Resistor R2 equivalents measured about ±5Ω. Fig 3 shows values for the relay I selected for calculations and testing. These values are strictly generic, from a random group of relays in my parts stock. A particular relay inductance and series resistance for your circuit should be measured.

The relay circuit analysed is almost a "classic" R-L circuit consisting of L1, R2 and a voltage source. However, voltage source V_{C1} is not constant, rather a decaying voltage switched across the latching coil by transistor Q2 when **L_Driv** changes state from 0V to +13.8V. In the R-L network, current will not change instantaneously when V_{C1} is applied. To determine peak current, time to reach five time constants is computed, then a decaying voltage on C1 at that time is calculated. From these terms, current is found that closely compares to a maximum current transistor Q2 must deal with.

Five time constant calculation for the R2-L1 network:

$$t_{R2-L1} = 5 \times \frac{L1}{R2} = 5 \times \frac{0.014H}{66\Omega} = 1.06ms$$

Voltage on C1 at the five time constant point:

$$E = V_{C1} \times [e^{-t/(R2)(C1)}]$$

Where:

E   = voltage across coil after 5 time constants

V_{C1} = Initial voltage on C1 when it was switched across the relay coil, +11.8V

t   = five time constant duration, 0.0011s (rounded)

R2   = measured latching coil resistance, 66Ω

C1   = capacitor providing coil voltage, 0.001F

$$E = +11.8 \times \left[e^{-0.0011/(66)(0.001)}\right] = +11.61V$$

$$\{1\}$$

Knowing the voltage at five time constants, circuit current is:

$$I_L = \frac{E \times (1 - e^{-t[R2/L1]})}{R2} \quad \{2\}$$

Where:

E   = computed voltage on C1 after five time constants, +11.61V

I_L = high current after five R2-L1 time constants, A

t   = five time constant, 0.0011s (rounded)

R2   = measured latching coil resistance, 66Ω

L1   = measured latching coil inductance, 0.014H

$$I_L = 11.61 \times \frac{[1 - e^{-0.0011(66/0.014)}]}{66} = 0.175A$$

I used a 2N2907 PNP silicon transistor for Q2. A few specifications: $I_C = 0.6A$ max; h_{FE} @ 0.150A = 200, typical.

Values for resistors R3, R4 are computed when **L_Driv** is 0V. A DC beta (h_{FE}) of 200 is used:

(a) Q2, base current sourced into the base bias design:

$$I_{Q2(base)} = \frac{I_L}{h_{FE}} = \frac{0.175A}{200} = 875\mu A$$

This base-supplied current is small compared to current from V_{CC} flowing through base bias resistors R3, R4 so it is neglected from further calculations. I

selected a +2.0V base bias voltage making the Q2 emitter voltage +2.7V.

The +2.7V will be the lowest voltage capacitor C1 reaches when it discharges across the latching coil and is the voltage defined as the pre-charge voltage in Section 3.1.

At that point, Q2 will begin cut-off since the junction is back biased. The +2.7V was known to be below any coil operating voltage (by a good margin) for my +12V coaxial relays (see Section 3.6 for additional comment).

(b) With I_{Q2}(base) neglected, R4 = 820Ω and V_{CC} = supply source, R3 is :

$$V_{bias(2V)} = V_{cc} \times \frac{R4}{R3 + R4}$$

$$= +13.8 \times \frac{820}{R3 + 820}$$

R3 = 4839Ω {used 4.7kΩ}

Measured results closely follow these computed figures. Current flowing through R3, R4 is 2.5mA.

3.3 Circuit Operation

This section details how the capacitor C1 charge-discharge operates the relay. Latching coil activation voltage limits are investigated and calculations show fabricated pulses meet reliable test results.

The key operating criteria for this relay driver circuit is; if a pulse amplitude applied to a latching coil is no greater than manufacturer's maximum specification (+13.2V) and 20ms later the pulse amplitude is no lower than the lowest pulse voltage specification, the relay will reliably toggle when properly controlled. The "classical" pulse shape often recognised is not necessary, only that applied energy must meet/exceed relay energy and time requirements.

I will amend the words… lowest pulse voltage amplitude… with "minimum voltage amplitude measured with my relays on a test fixture demonstrating reliable operation".

Coaxial relays in my parts stock are manufacturer-specified to operate reliably with a minimum +10.8V pulse. I wanted to determine if they could reliably operate below this voltage for 20ms because this would "relax" some driver circuit design limits.

A pulse latching test fixture was constructed. Seven, latching relays were initially subjected to a +13V pulse for 20ms using a pulse generator with variable voltage amplitude control. The test fixture included a circuit to steer the generator signal output to alternate sides of the relay-under-test (RUT) between coil drive signals **L_Driv**, ***L_Driv**. Over time, each RUT received a carefully measured, lower voltage amplitude, 20ms pulse at 0.4s toggle rate until it no longer operated reliably (i.e., erratic toggling, dead-stop operation). Hardly a "true reliability test", but one that gave "first order" minimum pulse amplitude/time operating information useful in "above average" ham radio applications.

Six, relays tested unreliable at +7.7V / 20ms, so a reliable minimum operating voltage for them was defined at +8.0V / 20ms pulses. One, relay demonstrated unreliable operation below +8.2V / 20ms pulses so its low safe operating voltage was defined at +8.5V / 20ms pulses. Note the manufacturer's lower operating limit for both brands tested is: (+12.0V) – 10%(+12.0V) = +10.8V.

One, of the six most reliable relays was used in circuit development shown in Figs 2 and 3. No other multi-variable tests; changing pulse amplitude, pulse width time, or relay toggle rate were done.

We now compute the voltage level on C1 after 20ms:

$$V_{pulse} = V_{C1} \times \left(e^{-t/(R2)C1} \right) \quad \{3\}$$

Where:

V_{pulse} = voltage across the coil (and capacitor C2) after time t

V_{C1} = capacitor C1 stored voltage (i.e., voltage source) at beginning of discharge, +11.8V

t = time after beginning voltage discharge (relay spec), 0.020s

R2 = nominal latching coil resistance, 66Ω

C1 = discharging capacitor used to generate a latching pulse, 0.0010F

$$V_{pulse} = +11.8 \times \left(e^{-0.020/(66)(0.0010)} \right) = +8.72V$$

This voltage was verified on the test fixture. This voltage is below manufacturer specification, but 0.72V above the minimum reliable operating voltage found for six, tested relays.

The circuit design was further exercised by placing six of the relays in the test fixture, cycling 7 days/24 hours a (noisy) day! During test, computed voltages and time events were measured to be within component tolerances and specifications. There were no operating failures during test.

In summary, a repeatable pulse shape has been formed with a +11.8V peak amplitude, and +8.72V minimum. It successfully operates tested coaxial relays found in my parts stock.

3.4 Putting the Pulse Circuit Together with Relay

The complete relay circuit design takes shape in Figure 4:

The single pulse circuit previously described (Figs 1, 2 and 3) is duplicated so one pulse circuit connects to each RLY latching coil. No bypass capacitors are shown to help clarify circuit description. Back EMF diodes ("catch diodes") D1, D2 are added. Input drive signal **L_Driv** has a compliment ***L_Driv** signal driving the added pulse circuit.

A DPDT toggle switch, wired with +13.8V DC and ground, can make these drive signals, or a single input drive signal from an IF radio T/R Switch output, together with a transistor inverter, would make both signals. Both drive signals must switch complement polarity at the same time to properly operate their respective pulse circuits.

3.5 What about the `Power-Up` State?

What contact position will RLY initialise to? If this circuit controls a "transmit"-"receive" signal path, it is likely relay contacts would be initialised to "receive". The circuit to do this is shown inside the dashed outline box in Fig. 4 labelled "Power-On Reset".

The reset circuit will work connected to either pulse circuit (not both). The P-Channel Power FET (Q5) connects across Q1, Q2. The FET gate connects to capacitor C3, and R9 forming a charging circuit referenced to V_{CC}. When V_{CC} initialises, $V_{GS(th)}$ is 0V, turning "on" Q5 so it connects V_{CC} and R1 directly to the preferred latching coil. Voltage across the coil establishes the relay start up condition. Device Q5 remains "on" while $V_{GS(th)}$ rises at a R9-C3 rate. At FET $V_{GS(th)}$ threshold, Q5 "opens" and the power-up circuit has no further circuit influence until the next power-up cycle. Diode D3 allows fast C3 discharge when V_{CC} is removed. The R8-C3 values provide about 500ms reset time for my system power-up needs, and sources about 0.148mA into the relay coil (manufacturer has 0.2A maximum/1s pulse specification. Transistor Q5 on-resistance $R_{DS(on)}$ is neglected. Since most FET $V_{GS(th)}$ voltages are not precisely specified, my R9-C3 values may need changing for your power-up timing cycle.

3.6 Final Design Comments

The relay toggling rate is not dependent on waiting for a pulse-forming capacitor to reach 0V after connection to a latching coil. Referring to Fig 4, testing showed +8V is the lower voltage limit for reliable
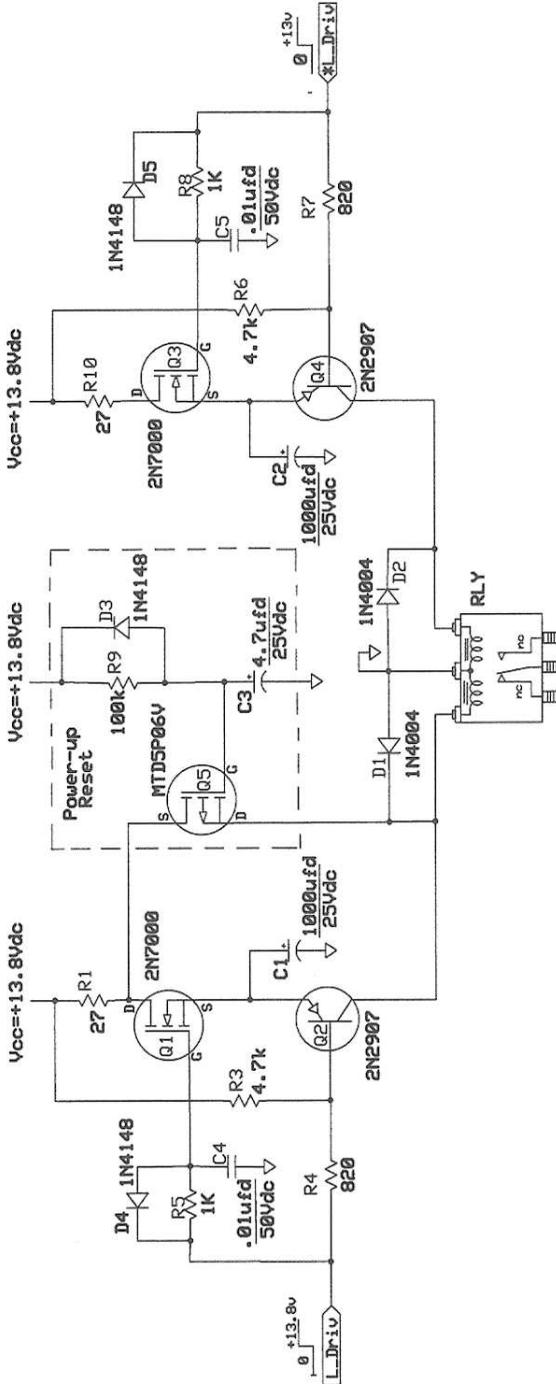
Fig 4: Full driver circuit + power-up.

relay operation (six, part stock relays). Further C1 (or C2) discharge *below* this voltage serves no useful pulse generation purpose. The switching rate of the relay can be increased if the discharging potential across the relay from C1, (or C2) is itself limited after going safely *below* the minimum relay operating voltage. At that point, input drive signals can switch, starting a capacitor recharge cycle. Of course, necessary charging time for C1 (and C2) must be met. An alternative is to change Q2, Q4 base bias resistors so they set Q2 (and Q4) emitter voltages above +2.7V, but safely *below* the relay +8V minimum, operating voltage.

More capacitance can be added to the 1000µfd storage capacitors. The V_{pulse} maxima stays at +11.8V (V_{C1}), but V_{pulse} minimum increases during capacitor discharge (per equation 3, at 20ms, discharging voltage from C1 (or C2) increases from +8.72V to about +9.6V when each 1000µfd capacitor value is increase to 1470µfd).

In all cases, I suggest checking circuit parameters to verify components are operating reliably within specification.

3.7 Part Selection Comments

- R3, R4, R6, R7, R9: 1/8W 2%-5% tolerance, metal film or carbon film resistors.

- Q2, Q4: the 2N4918, a member of the 2N4918 thru 2N4920, medium-power, PNP, silicon family, is a substitute. Switching speed and high f_T are not design factors. Adequate collector current capability w/o high $V_{CE(sat)}$ is important.

- Q5: A P-Channel Power FET with $I_{D\text{-pulse}}$ = 18A; I_D(cont) = 5A. A bit excessive for this application, but it was in my parts stock. The $V_{GS(th)}$ is +2.8V, typical, and is nice for this application. There should be many other substitutes for this part.

- C1, C2, C3: inexpensive, aluminium electrolytic capacitors (other types will work), with typically wide toler-

ance range and a 25V DC specification. Low leakage is ideal, but not necessary. I found repeatable pulse amplitudes and pulse-widths using low cost components. If wide tolerance parts are used, circuit time constants, voltage amplitudes and pulse-widths may be different than I computed, or tested.

- Q1, Q3: low cost P-Channel Enhanced MOSFET. The BS170 or 2N7002 are possible substitutes. Fast switching speed or high f_T is not important. One, key specification is the actual $V_{GS(th)}$ threshold voltage since it defines the maximum voltage (V_{C1}) charged on C1, C2. Another is maximum drain-source pulse current.

- RLY: Ducommum Technologies, p/n 2SE1T113A, SPDT coaxial relay, or equivalent. Any relay should be tested to determine the minimum voltage with specified pulse-width that reliably toggles the device. Measure latching coil inductance and resistance for necessary calculations.

- D1, D2, D3: low cost silicon diodes, part numbers on the schematic. Many other types can be substituted.

4.

Conclusion

An easily implemented coaxial relay circuit which combines switching control with pulse forming circuitry has been presented. Critical operating parameters and circuit precautions were discussed. Another circuit feature is input voltage levels are used to activate the latching relay coils, eliminating need to implement external pulse-forming, or pulse signal sources. After input drive signals change state, they may remain quiescent for any duration before changing to toggle the relay.

Standard value, inexpensive components are used. Special circuit, or relay mounting is not necessary. The relay is not subjected to heat dissipation or electrical pulses outside manufacturer application except for operating the latching coil below manufacturer minimum pulse voltage specification. The below minimum pulse voltage was shown to still operate relays reliably.

A power-on reset circuit is included to set relay contact position after a system start-up.

The circuit is adaptable to other V_{CC} values, relay voltage specifications, etc. I hope the computation processes show how substitute components can be used, and more novel extensions to circuit design and operation can be implemented for your Ham projects.

Gunthard Kraus, DG8GB

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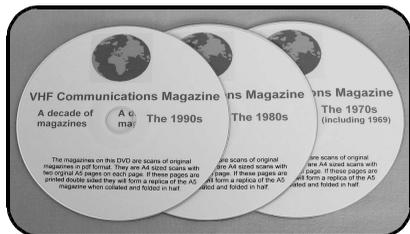
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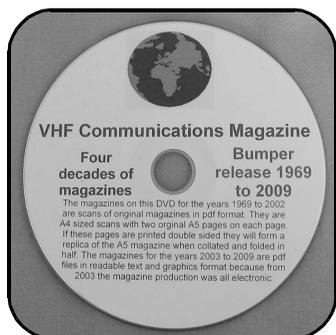
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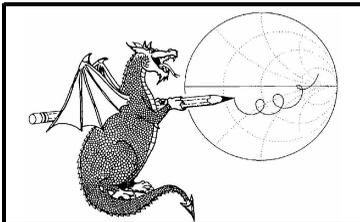
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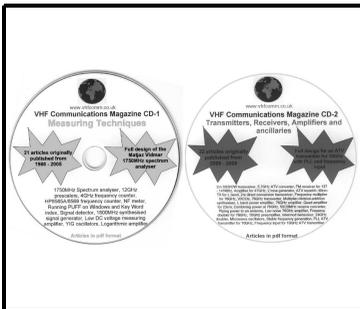
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