VHF
COMMUNICATIONS

High stability oscillator with PLL multiplier for use with DDS components

Wolfgang Schneider, DJ8ES

Contents

The new method of mailing has been slower that I had hoped but I will continue with it because it is helping to keep the postal price for 2013 lower than if I used Royal Mail. There is a subscription form on page 254, I hope that you will re-subscribe.

There is an exciting competition for 2013 introduced on page 254. I hope that it will result in some interesting new articles and authors.

Merry Christmas and a Happy New Year 73s - Andy

Wolfgang Schneider, DJ8ES

High stability oscillator with PLL multiplier for use with DDS components

Often simple TTL oscillators are used as clock oscillator for DDS components. This has the disadvantages that standard oscillators have modest frequency accuracy and considerable variation depending on the ambient temperature. The absolute frequency deviation can be corrected with software but the temperature variation cannot be corrected. This results in the output frequency "walking" to a greater or lesser extent.

1.

General

A possible solution to the issue of frequency stability of clock oscillator for a DDS is the use of a temperature stabi-

Fig 1: Example of a 10MHz OCXO,

lised oscillator, such as an OCXO (Oven Controlled Xtal Oscillator).

Since these are available only in the frequency range 1 to 10MHz at reasonable prices, a corresponding frequency multiplier is required to provide the frequencies needed for the operation of a DDS.

2.

Advantages of an OCXO as a clock oscillator

A wide range of OCXOs are available on the surplus market, the example shown in Fig 1 was used for the tests to generate the clock frequencies. This OCXO can be set exactly to the required frequency of 10.000MHz using an external trimming potentiometer. The frequency variation was ±50Hz from the required frequency as shown in Fig 2.

The problem of temperature behaviour is solved in an OCXO by an integrated heater for the crystal and oscillator circuit. Details regarding the quality of this can be found in the data sheet.

The resulting higher power consumption of the circuit is detrimental to this solution. The OXCO used has a good com-

Fig 2: Frequency variation of the OCXO with the adjustment voltage.

promise in my view. At switch on, the power consumption is just under 200mA at a supply voltage of +12V and falls after a few minutes (warm up) to a value of less than 100mA. The oscillator was as shown in Fig 1 for this measurement i.e. without additional insulation (polystyrene enclosure, or similar).

Alternatively frequency synchronisation with such sources as GPS or DCF77 can be used to improve the long term stability in place of the trimming potentiometer. There are various articles in amateur literature for this type of synchronisation.

Table 1: Jumper settings for the ICS501.

S1	S0	Clk
GND	GND	Input x 4
GND	Open	Input x 5.3125
GND	Vdd	Input x 5
Open	GND	Input x 6.25
Open	Open	Input x 2
Open	Vdd	Input x 3.125
Vdd	GND	Input x 6
Vdd	Open	Input x 3
Vdd	Vdd	Input x 8

3.

A high stability clock oscillator circuit design with PLL multiplier

The circuit shown in Fig 3 is designed for basic measurement of the OCXO, in conjunction with a frequency multiplier. Here the oscillator can be switched between the two PLL multipliers with jumper K7.

The two clock multipliers (IC2 and IC3) each require a +5V supply voltage. This is generated by IC1 (78L05) from the general supply voltage of the module (Fig 1).

The total current consumption is essentially determined by the OCXO.

3.1. ICS501 - LOCO PLL clock multiplier

The use of a ICS501 LOCO PLL clock multiplier is a simple and effective way to create a high quality high frequency clock signal from a low frequency crystal

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Fig 3: Circuit diagram of the test circuit for two different PLL multipliers.

oscillator. The name LOCO stands for LOw Cost Oscillator.

This integrated circuit was designed to replace standard crystal oscillators in electronic circuits. Through the use of PLL technology this component is capable of producing multiples of the input frequency up to 160MHz. The internal multiplier factor is set by external circuitry on S0 and S1. The various options are listed in Table 1.

The ICS501 (shown in Fig 4) is specially designed for the production of clock signals for DDS oscillators or other controllers. The output signal has a low phase jitter and is TTL level. The output current is approximately 25mA.

For the measurements, a multiplier factor of 5 was set using jumpers on K9, according to Table 1. The output signal spectrum is shown in Fig 5, all undesirable secondary harmonics are suppressed by more than 40dB.

3.2. Serial programmable clock source: ICS307

The ICS301 is a serially programmable clock source available in either an SMD or a 16 pin SOIC package. It can generate frequencies between 6 and 200MHz. The input frequency can be a crystal oscillator between 5 and 27MHz.

Similarly to the device described in the previous section, it has been designed for
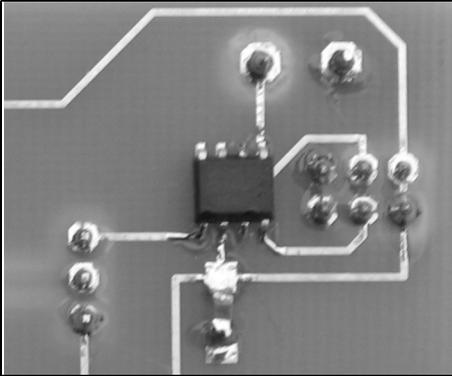
Fig 4: The ICS501 PLL frequency multiplier IC on the test PCB.

Fig 5: The harmonics of the input frequency are more than 40db down in the output spectrum of ICS501 PLL ICS510 multiplier.

use as a clock oscillator. The output signal is a TTL level and low phase jitter.

The programming is carried out via a 3 wire bus connected to K8. For special applications, a second output (K6) is configurable.

Since programming is quite complex as it is for DDS devices is not described in detail but the data sheet [2] gives all the information necessary for that purpose.

4.

Integration of ICS501 in a DDS module with an AD9835

A DDS oscillator module using an AD9835 was used in [3]. *{That article was not published in VHF Communications Magazine because it was an article on Shortwave radio, the DDS section of the article is shown in the appendix – Andy}.* It can be operated with a clock speed of up to 50MHz. In the example above, this clock was created with a standard 48MHz TTL oscillator. Note that the DDS module requires TTL level!

The updated circuit is shown in Fig 6. Here, the function of the DDS downstream filter should be considered fur-

ther. In this regard, please refer to the appendix.

The PLL multiplier with an ICS501 is used instead of the 48MHz TTL oscillator. The frequency multiplication factor is set to 5. Since the OCXO feeding K5 delivers a 10Mhz signal to IC5 the output frequency is exactly 50MHz.

Of course, the software for the ATTiny45 microcontroller (IC1) must be modified for the clock speed of the DDS module AD9835 (IC2).

The operating parameters for the DDS oscillator with PLL multipliers (DJ8ES 085) have only slightly changed compared to the original version from [3]. The power consumption is only 50mA at a supply voltage of +12V.

The output of the module is approximately 0dBm (1mW) depending on of the output frequency. Also note that contrary to the earlier version of the DDS module [3] there is no oscillator module integrated on the board. The clock must be supplied externally. The 10MHz OCXO shown in Fig 1 can be used. The clock oscillator must supply a TTL level (≥3.3V) and is fed on K5.

The PCB layouts are shown in Figs 7 and 8 with the component layout in Fig 9. A
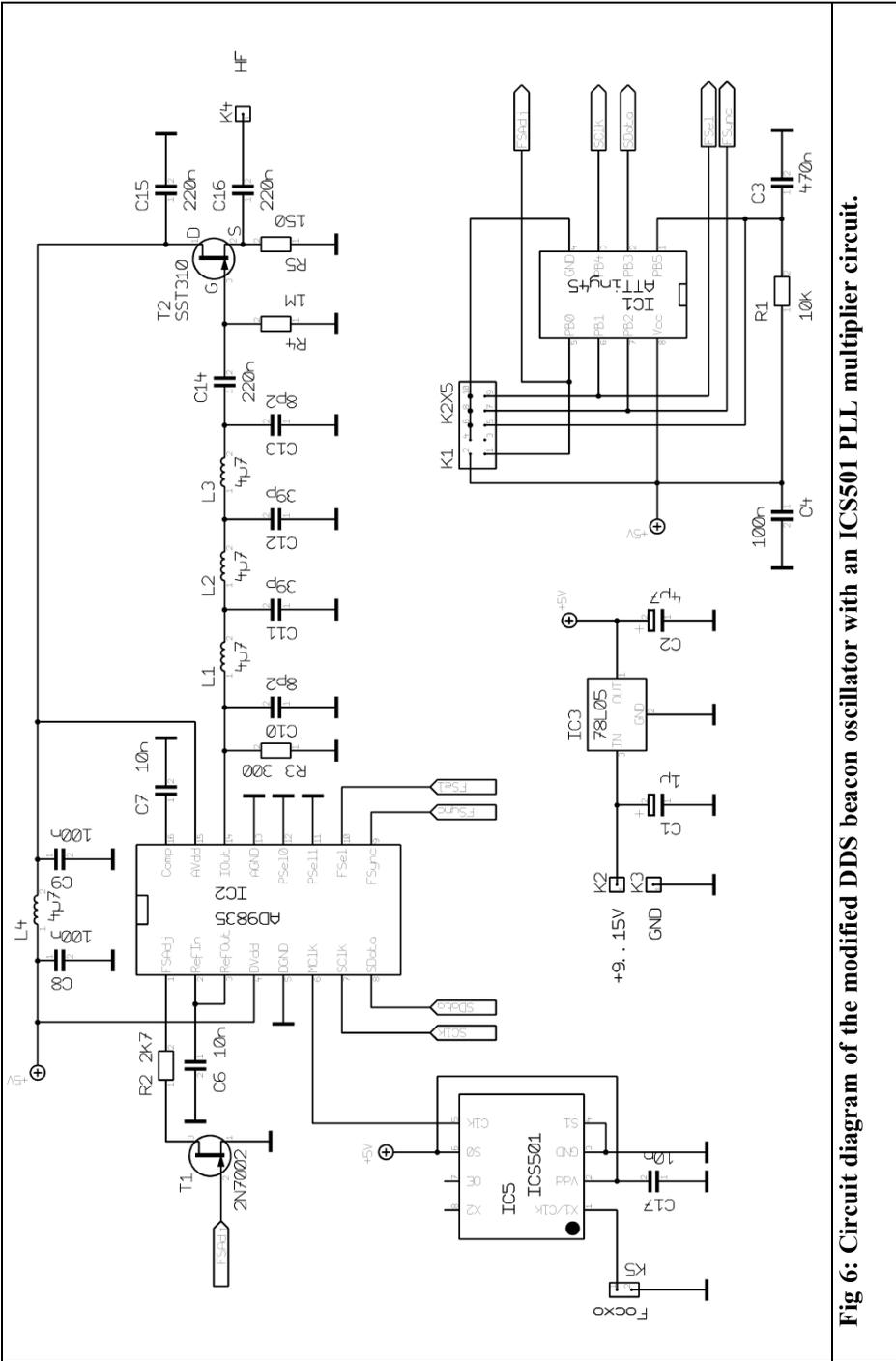
Fig 6: Circuit diagram of the modified DDS beacon oscillator with an ICS501 PLL multiplier circuit.
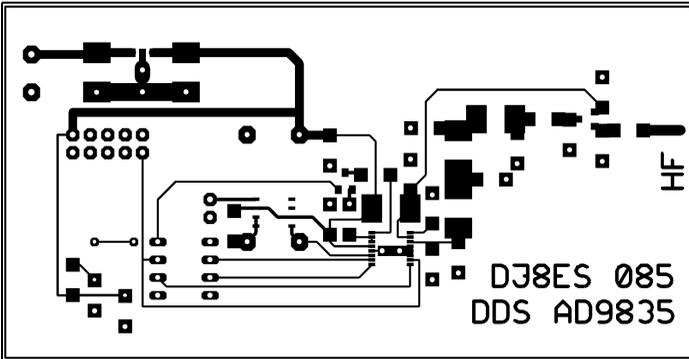
Fig 7: PCB layout for the component side of the modified DDS beacon oscillator.
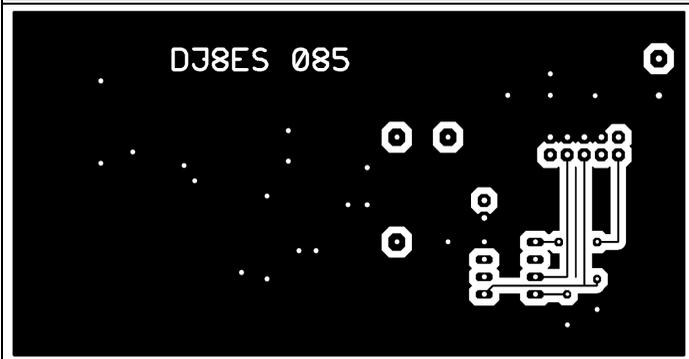
Fig 8: PCB layout for the ground side of the modified DDS beacon oscillator.
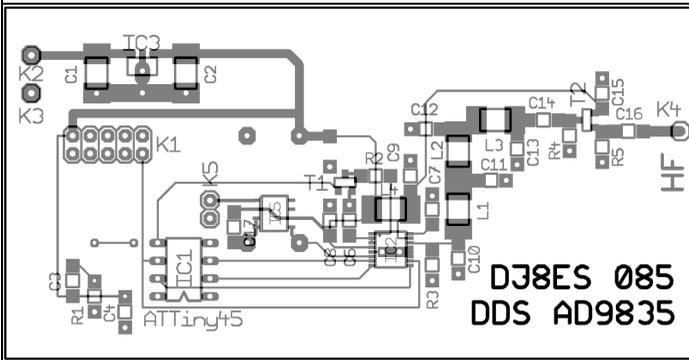
Fig 9: Component layout for the modified DDS beacon oscillator.
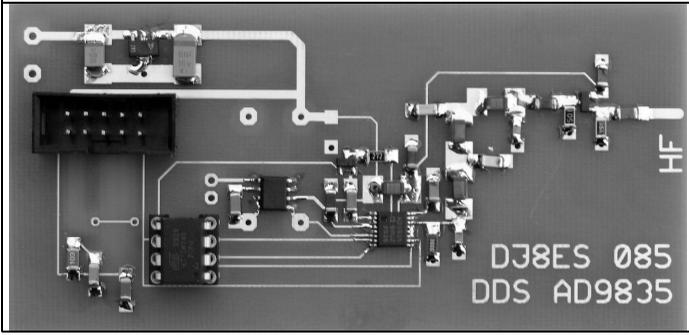
Fig 10: Photograph of the completed modified DDS beacon oscillator.

photograph of the completed PCB is shown in Fig 10. The target files and the software in Bascom are available for those who are interested in copying this design. Please send a requests to VHF Communication Magazine [4].

4.1. Parts list

Here, the parts list on the modified DDS module:

IC1 ATTiny 45, Atmel microcontroller
IC2 DDS chip AD9835, 8 pin DIP, Analog Devices
IC3 Voltage regulator 78L05F, SMD, SOT-89
IC5 LOCO PLL clock multiplier ICS501, IDT
T1 FET 2N7002, SMD
T2 FET SST310, SMD
K1 Pin header 2 rows, 2 x 5 pins, RM 2.54 mm
K2, K3, K4 Solder pin 1 mm
1 x Board 085, DJ8ES 50mm x 100mm mm, double sided with through plated holes

SMD inductance, 1812:

L1, L2, L3, L4 4.7μH

SMD Electrolytic capacitors 25V, 1812:

C1 1μF
C2 4.7μF

SMD ceramic capacitors, 1206:

C3 470nF
C4, C8, C9 100nF
C6, C7, C17 10nF
C10, C13 8.2pF
C11, C12 39pF
C14, C15, C16 220nF

SMD resistors ¼ W, 1206:

R1 10kΩ
R2 2.7kΩ
R3 300Ω
R4 1MΩ
R5 150Ω

5.

Appendix, DDS module from the article on low power shortwave experiments [3]

The modern DDS ICs require very little external circuitry to produce a high frequency sinusoidal output signal. A very interesting application for this is as beacon generator. The control module does everything even the telegraphy modulation.

A high quality low pass filter must be designed to filter the output signal. In addition to the desired frequency (fundamental), a DDS module also generates alias products as: clock + base frequency and clock − base frequency. The same applies to twice and three times the clock frequency. This results in an infinite number of mix products. This can only be removed by a steep sided low pass filter at 30 to 40% of the clock frequency.

5.1. Circuit description

At the heart of the circuit in Fig a1 is the AD9835 DDS IC (IC2) from Analog Devices. It will provide a sinusoidal output voltage into 50Ω between 0Hz and about 33% of it's clock speed. Thus with a clock oscillator of 48MHz (IC4, TTL oscillator) the maximum usable output frequency is approximately 16MHz.

The required spectral purity of the output signal and harmonic suppression can be achieved in the given circuit with a 7 pole Butterworth low pass filter. It is designed for 50Ω and all values are set to standard values.

The broadband amplifier that follows uses a field effect transistor T2 (SST310) doubling the impedance to 50Ω. The output power is approximately 0dBm (1mW).

The DDS module is controlled by a
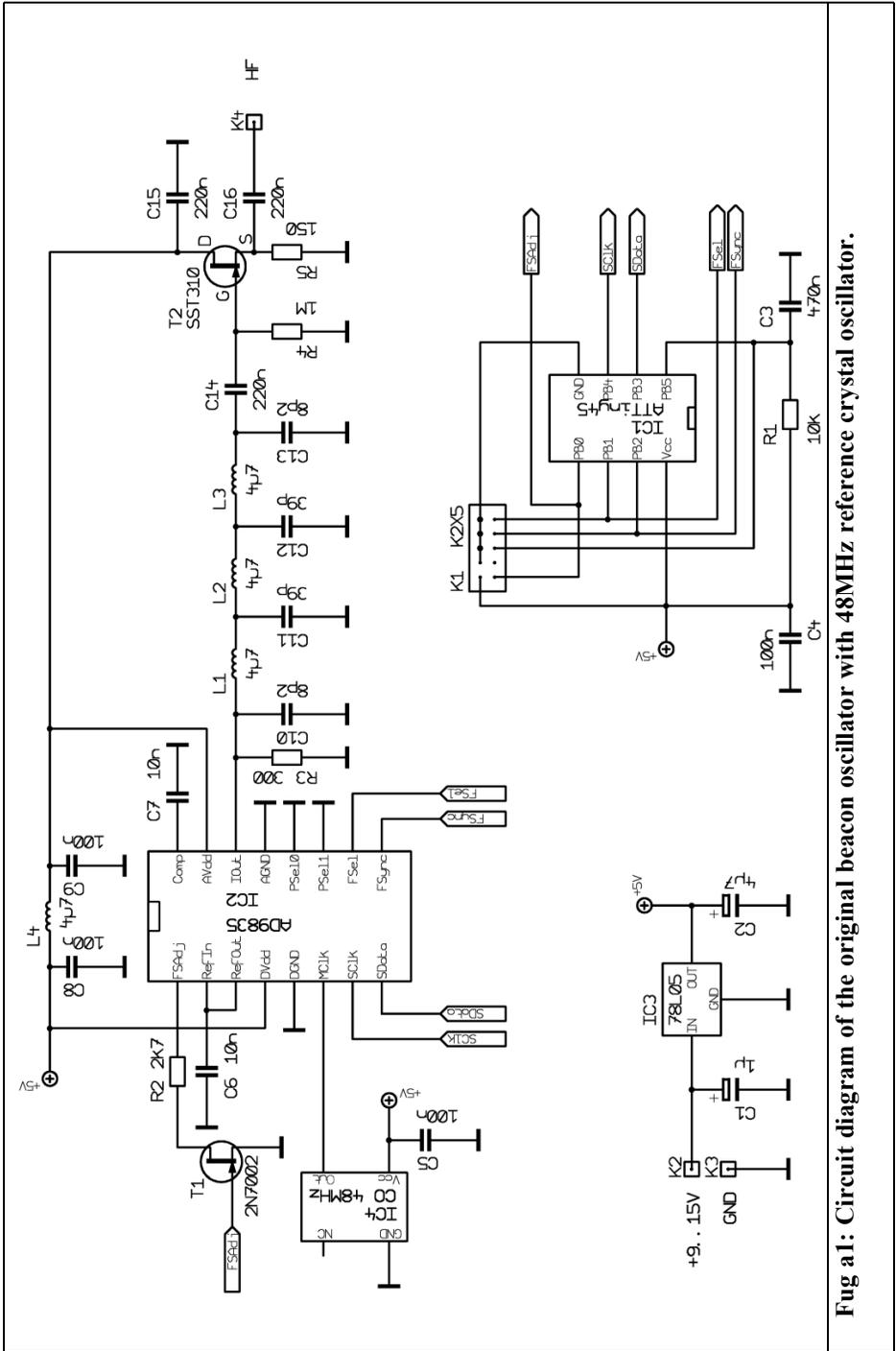
Fug a1: Circuit diagram of the original beacon oscillator with 48MHz reference crystal oscillator.
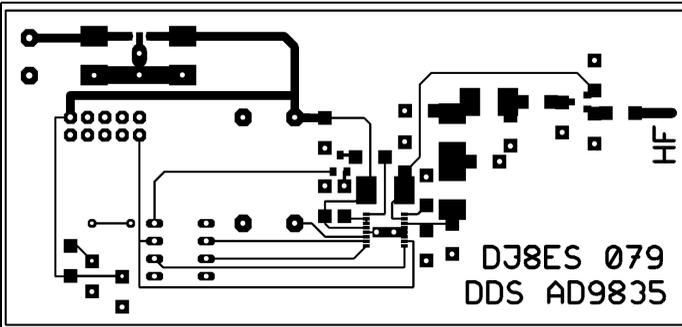
Fig a2: PCB layout for the component side of the original DDS beacon oscillator.
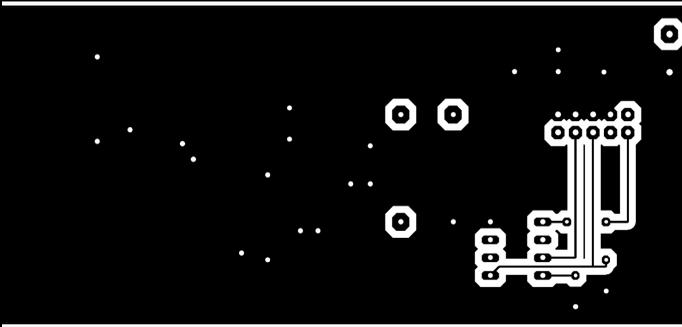
Fig a3: PCB layout for the ground side of the original DDS beacon oscillator.

ATTiny45 (IC1) microcontroller from Atmel. This also produces of telegraph signals. The output level of DDS is adjustable on pin 1 (FSAdj) using the current through R2. A switching FET T1 (2N7002) is used for the CW keying. The modulation depth is approximately 40dB. This is sufficient in the low power short-wave use.

All necessary data for the configuration of the DDS module, the output frequency and the code generator are hard coded in the microcontroller.

Changes can be programmed at any time via the 10 pin ISP port (K1) via software update. Input controls e.g. a knob for adjusting the frequency are not required for a beacon station.

5.2. Building hints DDS module

The full circuit of the DDS module with microcontroller is assembled on a 50mm x 100mm double sided PCB (DJ8ES 079) (Figs a2 and a3). It is constructed with SMD components, only the microcontroller is a plug in DIP package.
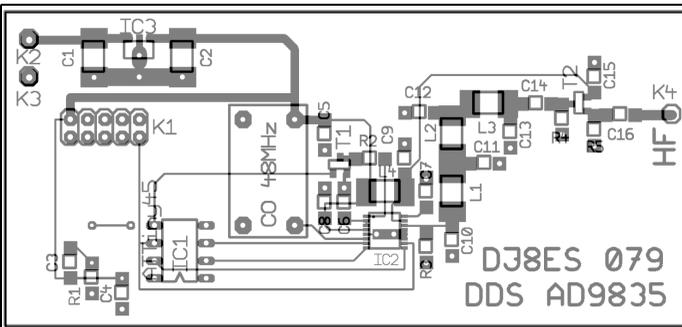
Fig a4: Component layout for the the original DDS beacon oscillator.

A professionally made PCB is recommended due to the very fine tracks in the area of the DDS module and the many vias that are required. The TARGET files are available on request.

The PCB DJ8ES 079 can be assembled in a casual order in accordance with Fig a4, only the DDS module AD9835 (IC2) and the microcontroller ATTiny45 (IC1) should be left until last. Before these two devices are plugged in or soldered, the function of 5V voltage regulator (IC3, 78L05F) should be checked.

The power consumption for the assembly amounts to approximately 110mA. The beacon transmitter should deliver the desired frequency e.g. 7.000.845Hz with an output level of approximately 0dBm (1mW).

Possible deviations of the output frequency in the range of some hertz result from the usual inaccuracy of 48MHz clock oscillator for the DDS module.

5.3. Parts list

IC1	Microcontroller ATTiny 45, ATMEL
IC2	AD9835, 8 pin DDS chip DIP, Analog Devices
IC3	78L05F, SMD, SOT-89
IC4	48.0MHz, TTL, crystal oscillator
T1	2N7002, SMD FET
T2	SST310, SMD FET
K1	Pin header 2 rows, 2 x 5 pins, RM 2.54mm
K2, K3, K4	Solder pin 1mm
1 x	PCB DJ8ES 079, 50mm x 100mm, double sided, through plated

SMD inductance, 1812

L1, L2, L3, L4 4.7μH

SMD Electrolytic 25V, 1812

C1 1μH
C2 4.7μH

SMD Capacitors, ceramic, 1206

C3 470nF

C4, C5, C8, C9 100nF
C6, C7 10nF
C10, C13 8.2pF
C11, C12 39pF
C14, C15, C16 220nF

SMD resistors ¼ W, 1206

R1 10kΩ
R2 2.7kΩ
R3 300Ω
R4 1MΩ
R5 150Ω

6.

References

[1] IDT / ICS, data sheet ICS501 - LOCO PLL clock multiplier: www.IDT.com

[2] IDT / ICS, data sheet ICS307 - serially programmable clock source: www.IDT.com

[3] Low power experiments on short-wave, Wolfgang Schneider, DJ8ES, UKW Berichte, 2/2011, pp. 83-93. The DDS section of the article is reproduced in the appendix.

[4] VHF Communications Magazine email: andy@vhfcomm.co.uk

Correction to the article "The use of electrolytic capacitors in DC power supplies" published in issue 3/2012.

Fig 6 on page 151 has the legends "No" and "Yes" transposed

Konrad Hupfer, DJ1EE

144MHz power amplifier with feedback - A summary report

1.

General

One of the most famous techniques for linearisation of power amplifiers is phase-oriented feedback, as it is often applied in audio amplifier stages. Here it is very broadband, because the phase response can be kept small. For the record, Fig 1 shows the process of linearisation using feedback to the control a very "crooked" field effect transistor.

A gain loss always accompanies feedback. However, the unwanted inter modulation products are reduced by the amount of feedback.

In the present application for a 2 stage

Fig 1: Compensation for distortion of drain current by feedback the voltage U_{GK} derived from I_D and addition to the original input voltage U_{GS}.

Fig 2: Block diagram of feedback applied to the 144MHz PA.
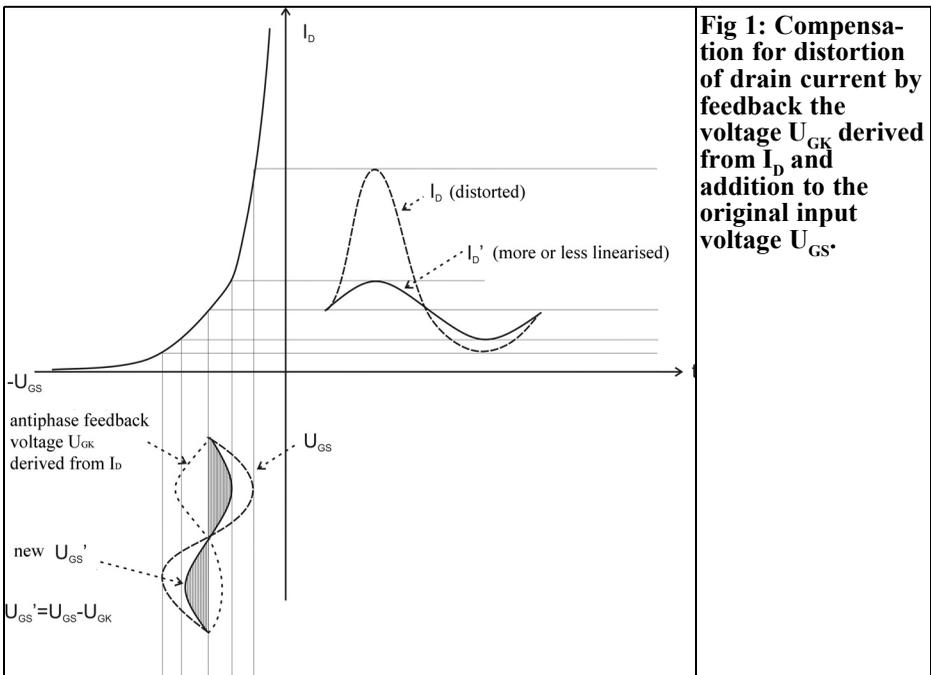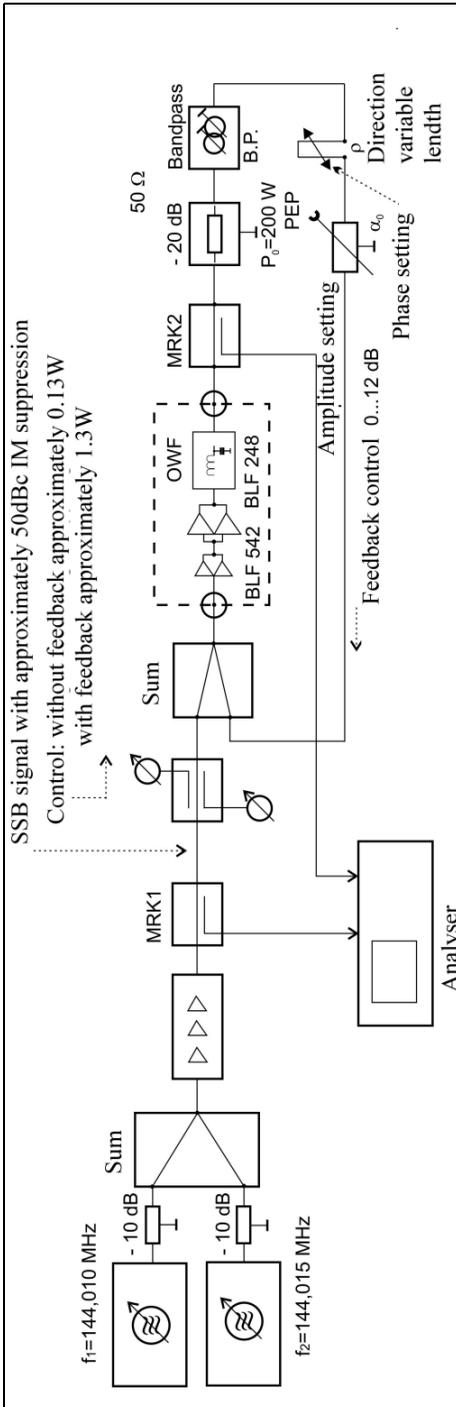
Fig 3: Laboratory style test setup of feedback on a 144MHz PA.

144MHz power amplifier (Po = 200W PEP) a so-called mixed feedback (voltage + current return = performance feedback) is applied. With a feedback factor of 10dB (gain reduction to 10dB!) all IM products are reduced by 10dB.

In contrast to an audio power amplifier phase changes occur in the amplifier matching circuit and lines where the feedback is derived. A band pass filter in the return branch ensures that the feedback is restricted to a narrow frequency range (about 1MHz).

The tried and tested experimental setup of a 200W PEP power amplifier is shown as block diagrams in Figs 2 and 3.

2.

Data and measurements

The following data shows the effect of feedback on a two stage 200W PEP power amplifier for 144MHz;

Gain without feedback: 29dB

Quiescent PA current : 2.0A

Quiescent driver current: 0.82A (class A)

Supply voltage: 28.0V DC
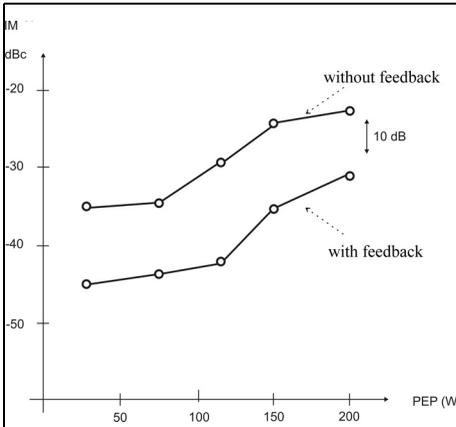
Fig 4: Comparison of IM products with and without feedback.

IM suppression for 200W PEP: approximately 21dBc (!)

Gain with feedback: 19dB

IM suppression for 200W PEP: approximately 31dBc

The IM product suppression improvement is shown graphically in Fig 4.

The "laboratory prototype" shown in Fig 3 is the amplifier circuit with feedback introduced to the overall circuit of the amplifier with commercially available components. I will report on the results of further tests after completion.

3.

Literature

[1] Heinrich Schröder "Electrical engineering" vol. 2, Publication for Radio-Photo-Cinema; GMBH, 1 Berlin 52 (Borsigwalde) 1963/1968

Henning C. Weddig, DK5LV

The filter bank module

**The filters described below are com-
plementary to the IF amplifier [1], the
AGC module [2] and the demodulator
[3] that can be used to build a receiver
for the SW and VHF bands. Other
modules are in the development phase.**

1.

Introduction

This article describes a filter bank suit-
able for the IF amplifier described in [1],
the AGC module developed in [2] and
the demodulator described in [3].

Frederick Krug, DJ3RV presented a ver-
satile IF module for 2m receivers and for
experimenters on the higher bands [4],
whose filter bank was described in [5].
After 30 years since it was published the
semiconductors used are no longer avail-
able and it makes sense to develop the
filter bank.

2.

Literature study

The versatile IF module for 2m receivers
described in [4] was considered. Slightly

modified assemblies are used for the
input and VFO as described by DJ7VY
[18].

Extensive measurements of broadband
amplifiers in grounded gate configuration
are discussed in [5] and the results
tabulated. Also the intercept points of
various KVG quartz crystal filter types
(discrete and monolithic) with two inter-
fering signals are tabulated. These are in
the stopband of crystal filters, but create
an intermodulation signal in the passband
of the filter. From these results we can
see that the output intercept point of the
amplifier stages (approximately
+28dBm) is lower than the input inter-
cept point of the crystal filters
(approximately + 35dBm). However, the
IP is only valid when an input power is
defined.

My own measurements [6] showed that
crystal filters are "non-linear" compo-
nents, which is why it is not good enough
to measure the intermodulation behaviour
at a single input power. Rather, the
intermodulation products for different in-
put values must be measured. If the
results are plotted as a graph, a line that
intersects with a line representing the out
of band interference gives an intercept
point that provides better information
about the behaviour of a crystal filter.

Michael Martin, DJ7VY, has specified
the measuring system and results of

various crystal filters in [7] and [8]. He finds even higher intercept points for the XF9B crystal filters indicating an IP3 of +48dBm. Therefore the mixer would be the limiting factor on the achievable intercept point.

Intercept points of greater than +40dBm can be achieved by developing the so-called H-mode mixer in the medium wave range, as shown in M. Bakker, PA3AKE in [9].

Watkin Johnson announced the HMJ-5 FET mixer with an IP of +35dBm in the VHF and UHF range. Unfortunately when Triquint took over Watkin Johnson this mixer is not produced any more Whether similar mixers from Mini Circuits such as the LAVI-VH+ with data specified as IP about +32dBm achieve this in the 2m or 70cm band would have to be examined by measurement.

Using MMICs, output intercept points of > +38dBm can be achieved. It is apparent that the IP of crystal filters have similar values as the IPs of the mixer or MMICs as amplifiers.

A JFET amplifier in grounded gate using the legendary P8002 is used in front of each crystal filter in the filter bank [5]. The crystal filters are switched with RH12 reed switches. The capacitance of these relay switches is so high that they cannot be used directly in the high impedance branch of the crystal filters. Therefore the relays are placed in the low impedance (50Ω) part of each amplifier i.e. in the source (grounded gate configuration!). According to his statements, an additional impedance transformation of the FET amplifier to lower value, or the compensation of the relay capacitance using links deteriorated the IP too much.

The drain circuit of the JFETs is completed with a resistor chosen to suit the impedance of the crystal filter that is connected via a coupling capacitor. The output feeds a tuned circuit with a low impedance tapping point connected to the centre pin of a second relay .

3.

Concept of the new filter bank

Up to four standard crystal filters from KVG (or Telequarz) were used for the filter bank. These are:
- To receive SSB modulated signal (XF9B), bandwidth 2.4kHz, impedance 500 / / 30pF
- CW signals (XF9M, bandwidth 500Hz, impedance 500 / / 30pF
- FM signals (XF9E, bandwidth 12kHz, impedance 1200 / / 30pF
- Reserve filter with a impedance 500 / / 30pF
- To be switchable.

The switching of crystal filters can be done with RF relays such as Panasonic RK2 or Omron G6Y-12V.

The system impedance at the input and output of the filter Bank shall be 50Ω

The filter are matched with so-called L-transformers in low pass configuration.

4.

Circuit description

Fig 1 shows the circuit diagram of the filter bank. It does not have any special features. The input and output signals are routed from angled SMA connectors to the tracks on the board.

Chapter 3 defined how the relay switching would be made. The contacts of the four relays in the input and output of the filter Bank are connected with 1.27mm wide track. The characteristic impedance of this track for the circuit board material (1.6mm thick FR4) is 76Ω according to AppCad [10]. A 50Ω microstrip line would be 2.9mm wide according to App-Cad. Widening the track to 50Ω microstrip lines near the connectors has been
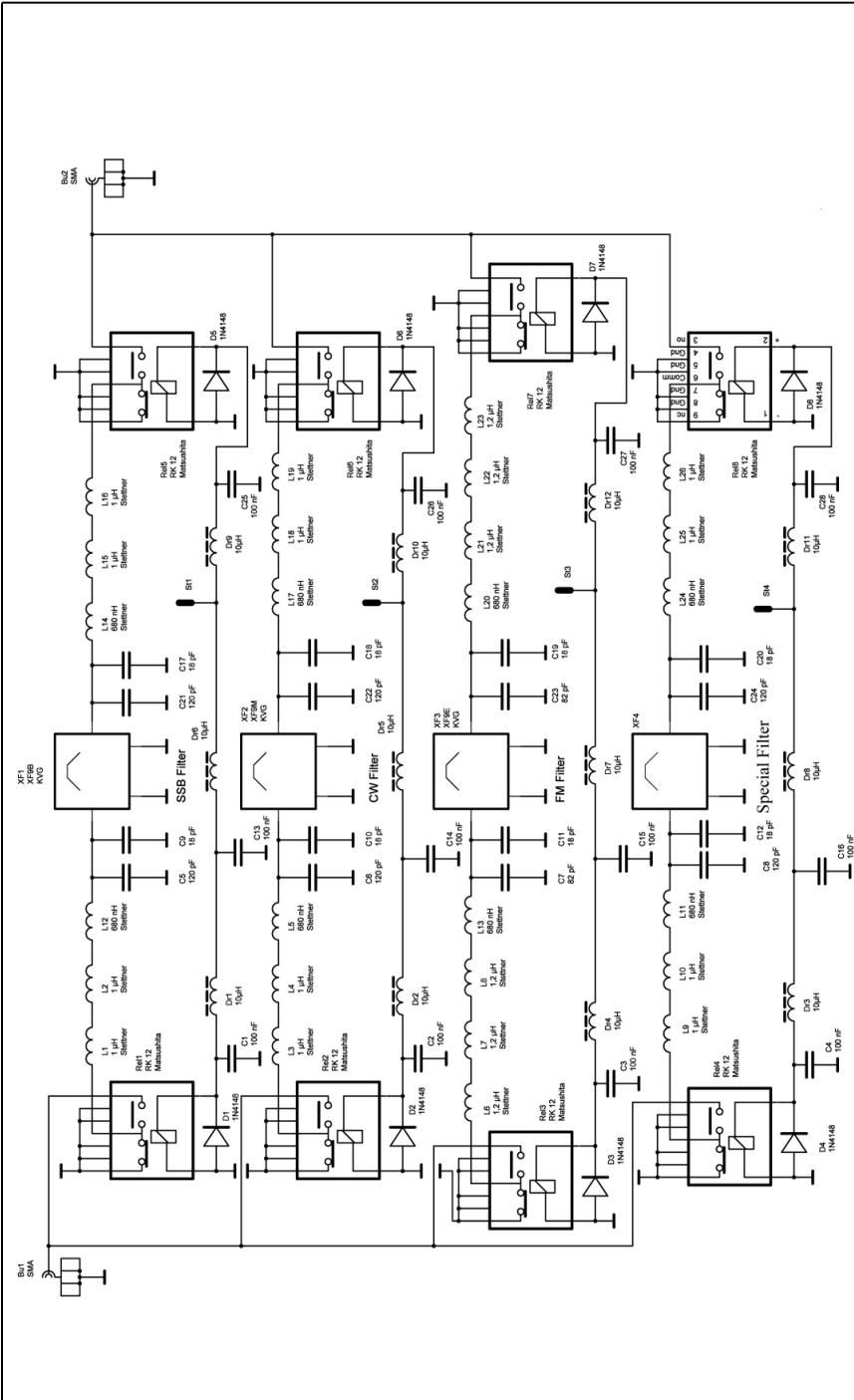
Fig 1: Circuit diagram of the 4 way filter bank.

omitted because at the frequency of 9MHz (0.08λ) and the very short line length for the wavelength causes no significant mismatch. The ground planes on each side of the microstrip line should comply with the minimum distance to the microstrip that is larger than 0.5mm.

This requirement is not fulfilled for the filter Bank because the ground planes only have a distance of 0.5mm from the strip line. This results in an increased amount of capacitance. The configuration used corresponds to a coplanar line. According to AppCAD a coplanar line with the above characteristic has an impedance of 59Ω (1.27mm track width, double sided, distance from the track to the ground plane 0.5mm), with an electrical length of 0.05λ. This impedance is quite close to the desired impedance of 50Ω. According to AppCAD, the impedance of 50Ω in a conductor track width of 1.92mm is achieved. Because the electrical length at the frequency used is very short compared to the wavelength no unwanted impedance transformations can be expected.

I had a "nasty" surprise when making a PCB for a 1.3GHz frequency experiment. The distance from the ground plane to the side of the microstrip was only 0.5mm resulting in a intolerable mismatch. This could only be rectified by milling the close ground planes to a distance of >2mm from the side of the microstrip.

The centre contact of each relay is connected to an L-transformer with the other end connected to the crystal filter. The NC relay contacts are connected to ground so that the input and output of the unused filters connected to ground to minimise crosstalk with the filters not in the signal path.

The matching of the crystal filter impedance (R_F) to the generator impedance (R_G) is calculated for the so-called L-transformer with the following formulas:

$$X_1 = \sqrt{R_G(R_F - R_G)} \qquad (1)$$

$$X_2 = \frac{R_F}{\sqrt{\frac{R_F}{R_G} - 1}} \qquad (2)$$

Where X_1 is the series impedance and X_2 the parallel impedance. Both impedances have opposite sign. If an inductor is chosen for X_1, X_2 must be a capacitor and vice versa. The series impedance has the higher impedance (here R_F) connected in parallel. The parallel impedance has the imaginary part of the crystal filter impedance (30pF) added, i.e. X_2 has the capacitance of the crystal filter added.

The inductors and capacitors for the L-transformer are determined with the formulae (3) and (4):

$$L_1 = \frac{X_1}{\omega_0} \qquad (3)$$

$$C_1 = \frac{1}{\omega_0 X_2} \qquad (4)$$

To match the generator impedance of R_G = 50Ω to the filter impedance R_F 500Ω at the frequency of 9MHz requires the following values:

$$X_1 = \sqrt{R_G(R_F - R_G)} = \sqrt{50(500 - 50)} = \sqrt{22500} = 150\Omega \qquad (5)$$

$$X_2 = \frac{R_F}{\sqrt{\frac{R_F}{R_G} - 1}} = \frac{500}{\sqrt{\frac{500}{50} - 1}} = \frac{500}{\sqrt{9}} = \frac{500}{3} = 166.667\Omega \qquad (6)$$

$$L_1 = \frac{X_1}{\omega_0} = \frac{150}{6.28 \cdot 9 \cdot 10^6} = \frac{150}{56.52} \cdot 10^{-6} = 2.65 \cdot 10^{-6} = 2.65\mu H \qquad (7)$$

The inductor L_1 is formed by connecting two 1206 size SMD 1µH and a 680 nH SMD inductor made by Stelco (coil on a ceramic core).

$$C_1 = \frac{1}{\omega_0 X_2} = \frac{1}{6.28 \cdot 9 \cdot 10^6 \cdot 166.667} = \frac{1}{56.52 \cdot 166.667} \cdot 10^{-6} =$$

$$\frac{1}{9420} \cdot 10^{-6} = 106 \, pF \qquad (8)$$

As mentioned above the imaginary part of the filter impedance (30pF) must be added to the value of C_1 so that results in a total of 136pF for C_1. However, there is circuit capacitance to take into account. In the prototype the capacitance of C_1 was composed of a 120pF and an 18pF SMD capacitor (0805). A adjustment for the lowest ripple in the passband could be made using an SMD trimmer capacitor instead of the 18pF fixed capacitor.

The L-transformer at the output of the crystal filter is constructed like the transformer for the input circuit. The series inductor feeds the centre pin of the relay in the output circuit of the filter bank. Because the SSB and the CW crystal filters have the same impedance, the component values of the L-transformers are the same.

The FM Crystal filter has an impedance of 1200Ω // 30pF requiring the following values for the L-transformer:

$L_1 = 4.24\mu H$,.

$C_1 = 70.674pF + 30pF = 100.674pF$

The inductor is made by connecting three 1.2µH and a 680nH coil.

Note: the calculation of the component values of the L-transformers can be made from the "Tools" tab of the the freeware program RFSim99. The program can be downloaded from [12]. The above formulas and calculations using RFSim99 apply strictly to only two real impedances where the higher value is the the parallel branch of the L-transformer. Maynard Wright, W6PAP in [16] and [17] shows that there are cases in which this condition can be reversed if the higher load impedance has a null value.

To minimise any coupling of the RF signals the control voltages for the relays are decoupled with a choke and a 0.1µF SMD capacitor (1206 size). The third choke is used so that a wide uninterrupted ground plane can be used under the filters, the choke spans this on the PCB. This choke was replaced by a wire link on the prototype.

Diodes in parallel with the relay coils reduce the spikes when the relays are turned off.
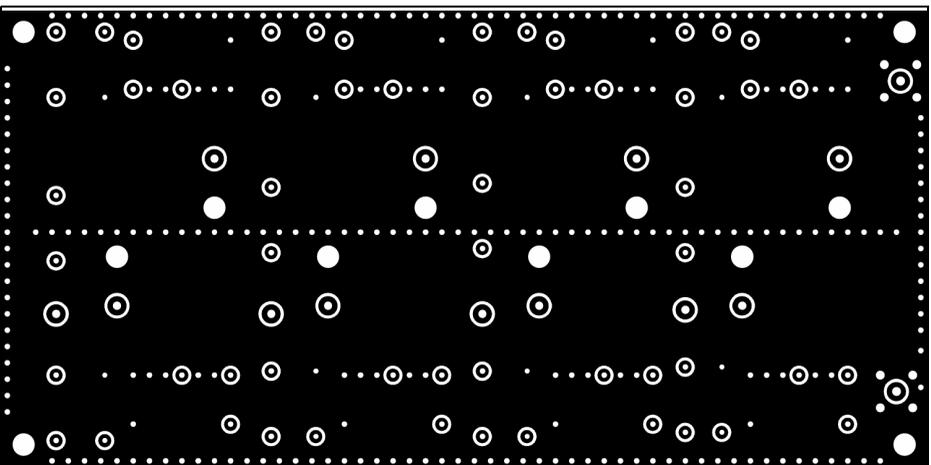

Fig 2: Ground plane side PCB layout (component side) of the 4 way filter bank.
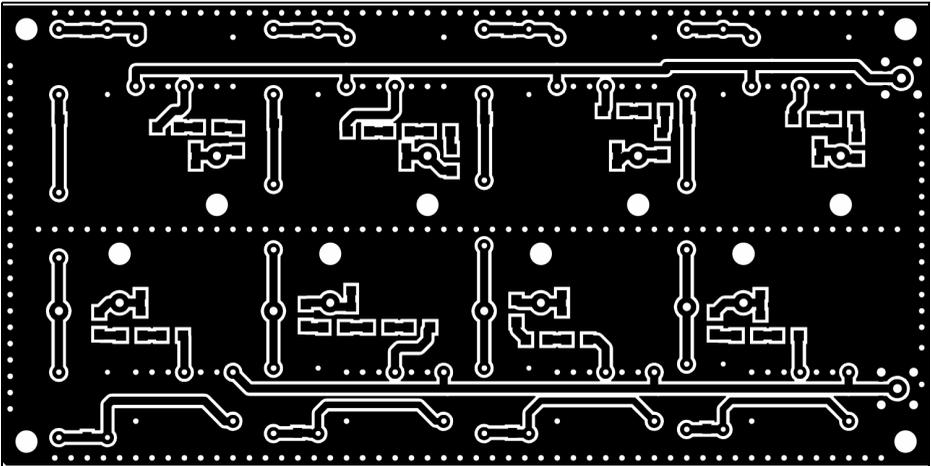
Fig 3: Track side PCB layout (SMD component side) of the 4 way filter bank.

5.

Construction

The filter bank is made on a 145mm x 72mm double-sided printed circuit board. Many holes spaced at 2.54mm intervals are arranged along the edges of the circuit board and in the middle of the

crystal filters. They are used a vias to connect the ground planes (Fig 2) top and bottom (Fig 3). The vias were made for the prototype by "Sewing" a thin single wire strand through the holes then soldering on both sides. This was only done for one edge and the centre of the crystal filters for the prototype. The ground connections of the relays were connected by means of pieces of wire on both sides to the ground of the circuit board. The
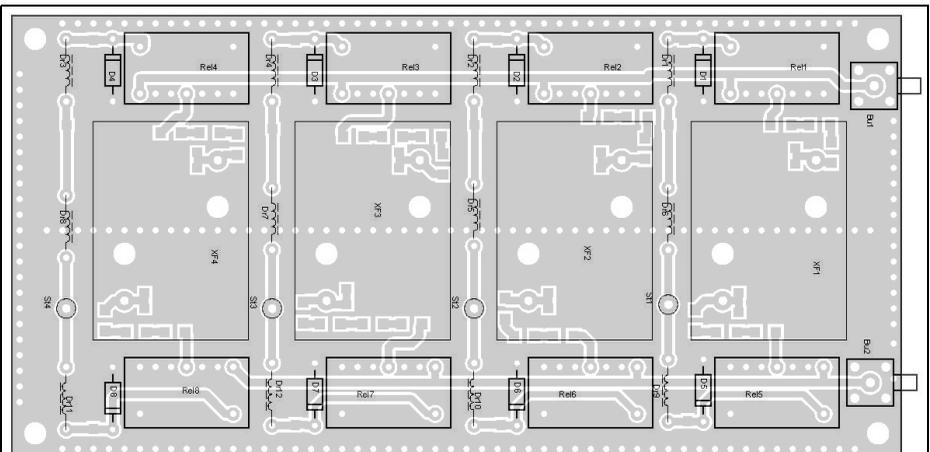
Fig 4: Component side layout of the 4 way filter bank PCB.
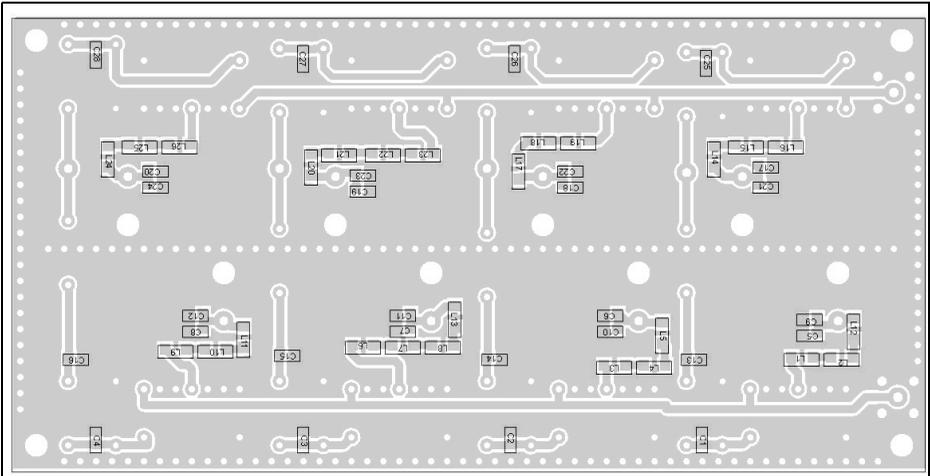
Fig 5: SMD side layout of the 4 way filter bank PCB.

ground pins of the SMA Sockets should be soldered on both sides.

The leaded components and Crystal filters are fitted on the component side of the board (Fig 4), the SMD components are fitted on the track side (Fig 5).

To make a better ground contact between the bottom of the crystal filters and the ground surface of the circuit board you could tin plate the surface on the top of
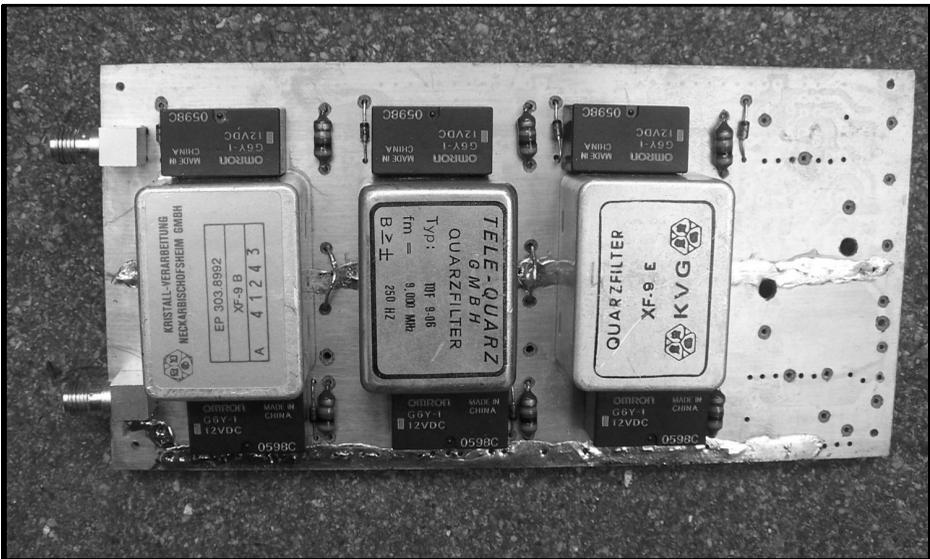

Fig 6: Photograph of the completed prototype showing the XF9B SSB filter, TQF9-06 CW filter and the XF9E FM filter. The vias along the edge and in the middle under the filters can be seen.
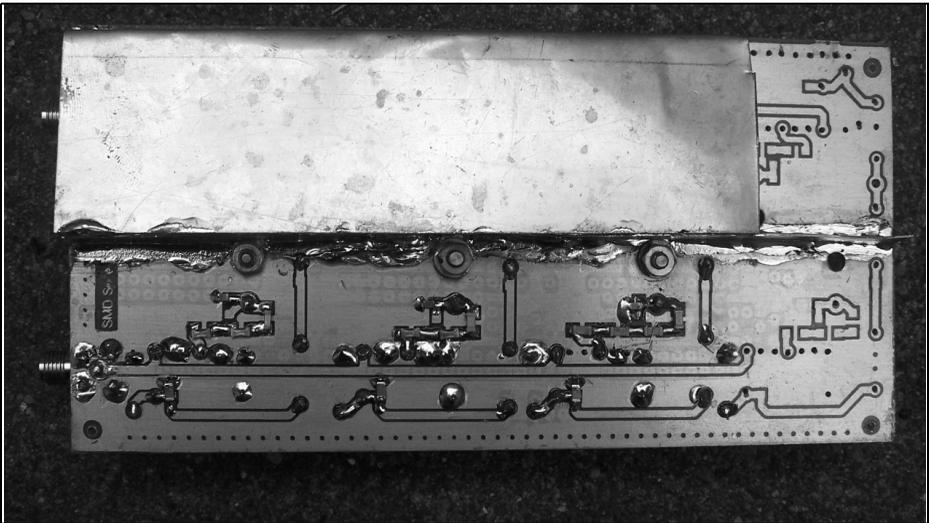
Fig 7: The bottom of the filter bank has an L shaped shielding plate covering the upper half. It is soldered to a metal bar that runs along the middle of the circuit board length.

the circuit board.

A photograph of the top of the completed board is shown in Fig 6 and the bottom is shown in Fig 7.

6.

Measurements

The filter curves of three Crystal filters used in the filter Bank were measured using an FaNWT network analyser from Funkamateur [13] using the WinNWT software version 4.10 by Andreas Linde-nau, DL4JAL [14].

When measuring narrow band filters the sweep speed (frequency change per unit of time of the sweeper source) should not be too high to give enough time the overshoot to steady otherwise you will get distorted curves. A slow sweep time can be adjusted by choosing an interval of 1ms.

The filter curves measured looked prom-ising, only the CW Crystal filters showed

an unusual shape.

The dynamic range of the FaNWT is not up to that of expensive devices therefore the filter curves were also measured with a Rohde & Schwarz vector network analyser. To exploit the full dynamic range and resolution of this equipment to 1KHz a sweep time of 30 seconds was used.

6.1. Surprise in measurements

The surprise came when determining the out of band attenuation of the filter. This was 80dB but could be increased to over 90dB (and thus in the "noise" floor of the analyser) if a shielding plate (30mm high, 145mm long) was held along the centre of the filter bank. It was not necessary to solder the plate. A reduction of the height to 10mm did not result in better out of band attenuation!

Since it was not useful to have this 30mm shield below the PCB giving an overall height of the module >50mm (the filters are about 200mm high) another solution was sought and found. This took the form of a 10mm high sheet metal strip sol-dered on the bottom of the filter bank in
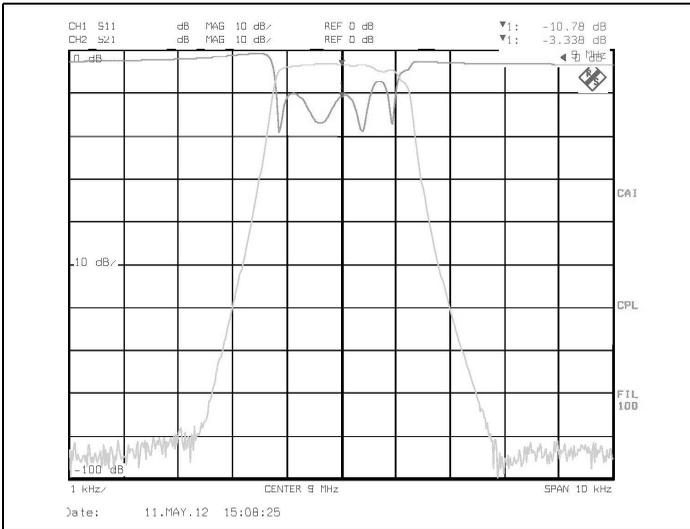
Fig 8: The insertion loss (S21) and the input return loss (S11) for the XF9B SSB crystal filter with a frequency sweep of ±5kHz. The vertical scale is 10dB per square.

the middle of the long side. A second L-shaped curved metal sheet covers the crystal filter area. It was not necessary to make this a "watertight" compartment on the prototype i.e. soldering the plates together. The crystal filters were not separated with metal bridges.

6.2. Test results

6.2.1. SSB crystal filter

Figs 8, 9 and 10 show the response curve for the XF9B SSB filter they show the insertion loss (S21) and the input return loss (S11) with a sweep width of 10kHz, 5kHz (the smaller range improved the vertical resolution of 2dB/grid) and 50kHz. Fig 9 shows the 3dB bandwidth of 2.4kHz given in the manufacturer's specifications. Fig 10 shows the out of band attenuation achieved of 90dB.
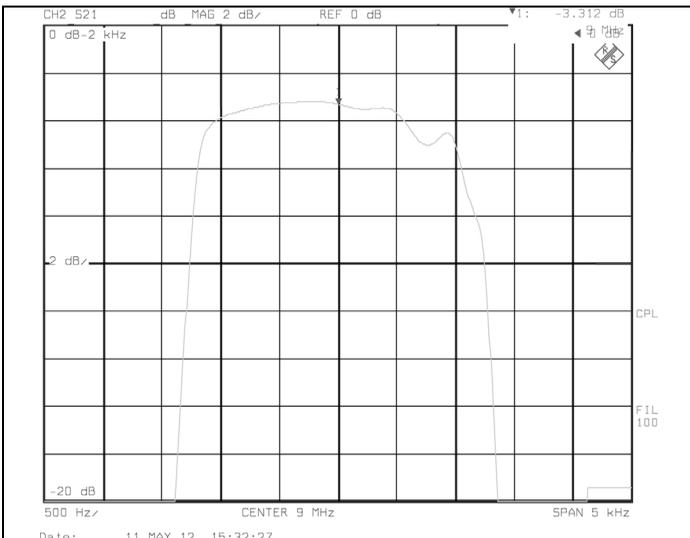
Fig 9: The insertion loss (S21) and the input return loss (S11) for the XF9B SSB crystal filter with a frequency sweep of ±2.5kHz. The vertical scale is 10dB per square. The 3dB bandwidth of the filter can be seen at 2.4kHz.
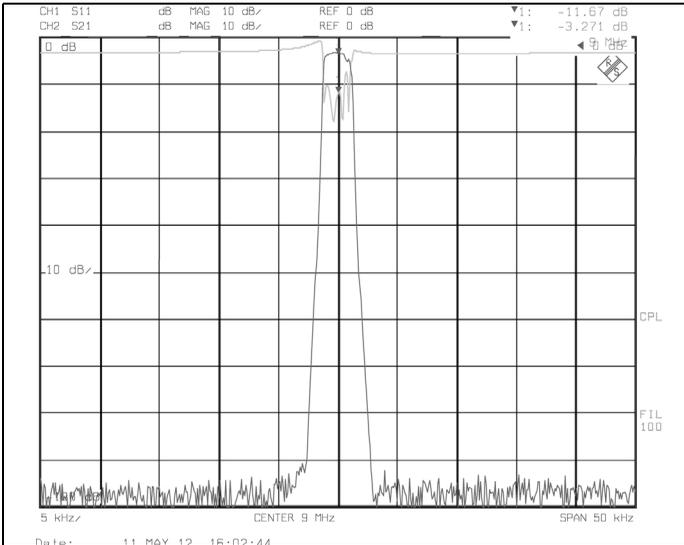
Fig 10: The insertion loss (S21) and the input return loss (S11) for the XF9B SSB crystal filter with a frequency sweep of ±25kHz. The vertical scale is 10dB per square. The out of band attenuation of 90dB can be seen.

6.2.2. CW crystal filter

This TQF 9 crystal filter was manufactured by Telequarz purchased via the Internet. Unfortunately I did not measured this crystal filter when it was received and made a claim...

Figs 11, 12 and 13 show the response curve for the CW filter they show the insertion loss (S21) with a sweep width of 2.5kHz, 5kHz (the smaller range im-

proved vertical resolution of 2 dB/grid) and 50kHz. Fig 11 shows that the crystal filter has a defect. An extension of the sweep time gave no improvement so that measurement errors due to a high sweep rate can be excluded. It would be nice to have access to a matching wand and "just" pull the filter curve to correct it because a resonator in the filter has a high resonant frequency. Unfortunately
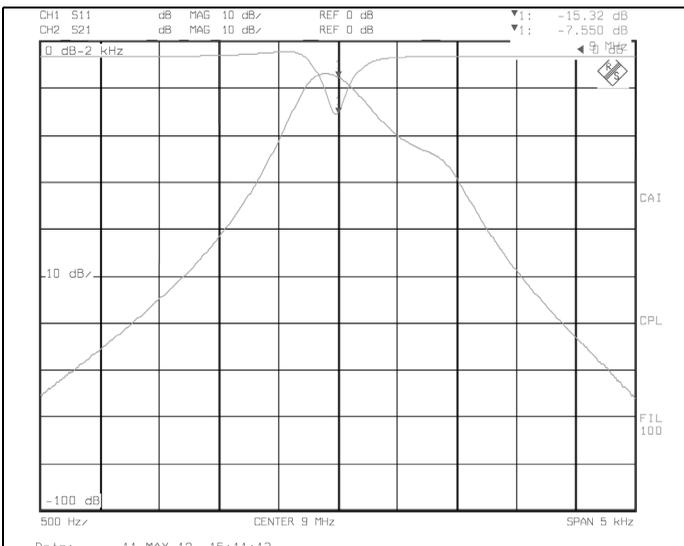
Fig 11: The insertion loss (S21) for the TQF9-06 CW crystal filter with a frequency sweep of ±2.5kHz. The vertical scale is 10dB per square. It shows that one resonator has a resonant frequency that is too high.
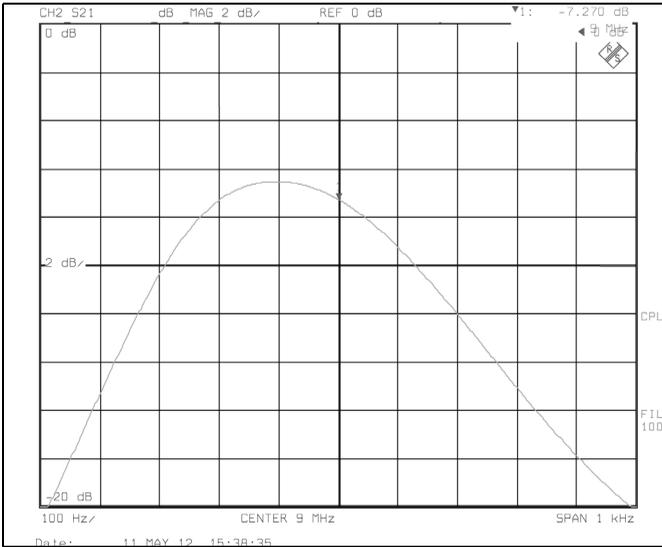
Fig 12: The insertion loss (S21) for the TQF9-06 CW crystal filter with a frequency sweep of ±500Hz. The vertical scale is 2dB per square. It shows that one resonator has a resonant frequency that is too high. The insertion loss is > 6dB and the centre frequency is shifted down.

that is not possible with a crystal filter. The filter case would have to be opened and the defective crystal found and replaced by a matching crystal. The "bent" filter curve can be seen even better in Fig 12. Fig 13 shows the out of band attenuation achieved of 90dB.

6.2.3. FM crystal filter

Fig 14, 15 and 16 show the response curve for the XF9E FM filter they show the insertion loss (S21) with a sweep width of 20kHz, 10kHz (the smaller range improved vertical resolution of 2 dB/grid) and 50kHz. Fig 15 shows that the 3dB bandwidth is 12kHz as given in the manufacturer's specifications. Fig 16 shows the out of band attenuation achieved of 85dB, is slightly less than that of SSB and CW Crystal filters.
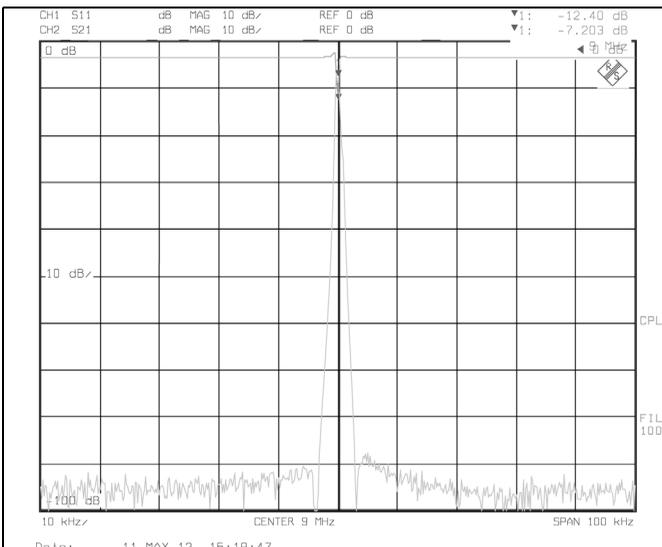
Fig 13: The insertion loss (S21) for the TQF9-06 CW crystal filter with a frequency sweep of ±50kHz. The vertical scale is 10dB per square. It shows that one resonator has a resonant frequency that is too high. The out of band attenuation is approximately 90dB.

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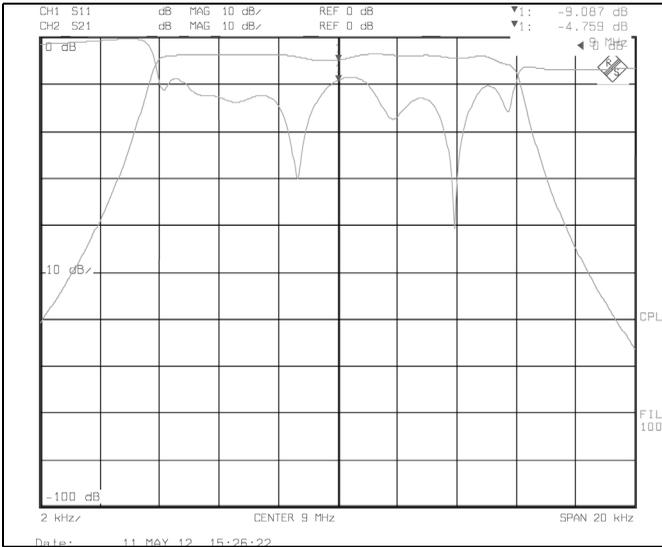
Fig 14: The insertion loss (S21) for the XF9E FM crystal filter with a frequency sweep of ±10kHz. The vertical scale is 10dB per square. The insertion loss is > 6dB and the centre frequency is shifted down.

6.3. Intermodulation measurements

First measurements of the intemrodulation performance of the crystal filters XF9B and XF9M using out of band interfering signals producing inband IM products showed an IP of about +30dBm. To ensure that these results are not introduced by an inappropriate test setup (e.g. generating IM products due to an insufficient decoupling of

the two signal generators) different signal genrators/spectrum analyser (in the QRL of the author) or the addition of decoupling amplifiers or the use of lowpass filters to suppress the harmonic content of the signal generators have to be considered and tested.

A separate article about these measurements (upgraded version of [6] is in preparation.
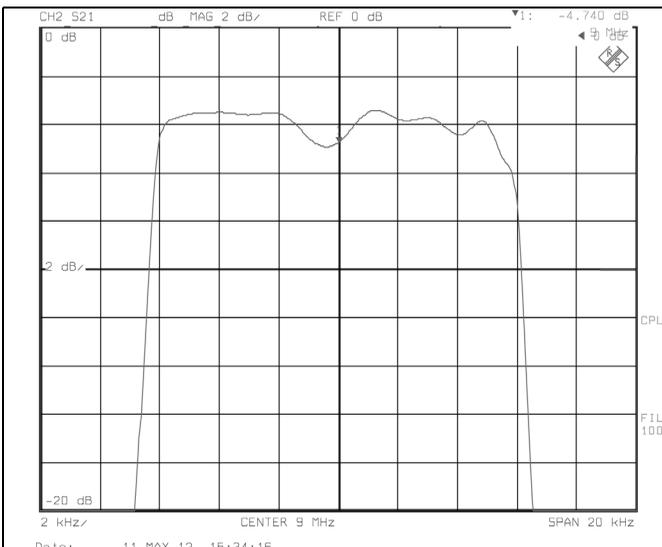
Fig 15: The insertion loss (S21) for the XF9E FM crystal filter with a frequency sweep of ±10kHz. The vertical scale is 2dB per square. The insertion loss is approximately 4dB.
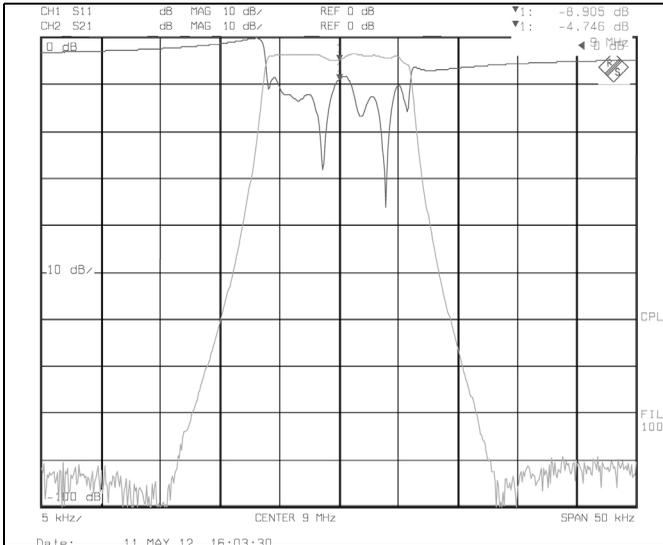
Fig 16: The insertion loss (S21) for the XF9E FM crystal filter with a frequency sweep of ±50kHz. The vertical scale is 10dB per square. The out of band attenuation is approximately 85dB.

Jörn Bartels DK7JB [19], his site is "unfortunately" in the German language has a nice long article about his effort to build a high performance transceiver. He also describes his test setup for measuring the IP of mixers and other active/passive components like bandpass filters, shows first measurement results on the IM products of homemade ladder type crystal filters.

This article ("Basteltagebuch") is updated from time to time.

Martein Bakker PA3AKE has presented some interesting test results on crystal filters from International Radio (Inrad) [15]. He also confirmed the non-linear behaviour of crystal filters in the pass band as well as interference in the stop band where IM product fall in the pass band. His results show that the intercept point must be determined by measuring the modulation products of interest for different input levels. It is still striking that the intercept point for interference and inter modulation signals in the pass band of crystal filters result an intercept point about 10dB lower than for interference in the stopband of the filter.

7

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[2] The AGC module, Henning C. Weddig, DK5LV; VHF Communications Magazine, 1/2008 pp 26 – 30 and 35 - 42 , 4/2009 pp 242 – 250, 1/2010 pp 42 – 53 and 1/2011 pp 36 - 46

[3] The demodulator module; Henning C. Weddig, DK5LV, VHF Communications Magazine, 4/2011 pp 215 - 231

[4] A Versatile IF Module Suitable for 2m Receivers, or as an IF Module for the SHF Bands; Frederick Krug, DJ3RV, VHF Communications Magzine, 4/1981 pp 244 - 250

[5] A Versatile IF Module Suitable for 2m Receivers, or as an IF Module for the SHF Bands; Part 2: Matching stage for the crystal filter; Frederick Krug, DJ3RV, VHF Communications Magzine, 2/1982 pp 112 - 124

[6] Measurements of inter modulation performance of crystak filter nonlinear behavior of passive component! Henning C. Weddig, DK5LV, proceedings 48. UKW Tagung Bensheim 2003

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[8] Improvement of an intermodulation free dynamic range of receiver input parts in the medium wave range; Michael Martin, DJ7VY, CQ-DL 2/1982; Page 66 to 69

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http://martein.home.xs4all.nl/pa3ake

[10] Freeware package AppCAD,
http://www.hp.woodshot.com/

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[12] Freeware program RFSim99; Downloadable from the website,
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[13] Funkamateur network analyser;
http://www.box73.de/index.php?cPath=112_113_132

[14] Software WinNWT4; Andreas Lindenau DL4JAL, www.dl4jal.eu

[15] IF Selectivity Board - filter; Quality Quartz filter; Martein Bakker PA3AKE,
http://martein.home.xs4all.nl/pa3ake/hmode/IF_filters.html

[16] Octave for L-networks; Maynard Wright W6PAP, QEX March/April 2012, page 44-46

[17] More octave for L-networks; Maynard Wright W6PAP, QEX May / June 2012, page 20-23

[18] A Modern Receive Converter for 2m Receivers Having a Large Dynamic Range and Low Intermodulation Distortions, Michael Martin, DJ7VY, VHF Communications Magazine, 4/1978 pp 218 - 229

[19] Basteltagebuch Eigenbau TRX 21012, Jörn Bartels DK7JB
http://www.bartelsos.de/dk7jb.php/selbstbau-trx-2012

Download of his file:
http://www.bartelsos.de/index.php?dl_file=CKULUVNTA0

An apology to Henning, I should have published part 2 of his Test Switch article in issue 2/2012 but I forgot. I will publish it in issue 1/2013. It was a prize winner at "UKW Tagung 2012" in Bensheim - Andy

Andy Barter, G8ATD

Electronic compass Windows display

1

Introduction

In issue 3/2012 I described an electronic compass that I had designed and built. From my field test I found that the LED display was not bright enough and said that I would try to write a Windows program to display the compass bearing in a window on a PC.

As I stated I am not a programmer but I can write programs that do the job even if they are not as good as they should be. I needed to find a suitable programming language to use for the project, it needed to be:

- Easy to use
- Able to produce a standalone programs that will run on a Window XP or Windows 7 PC
- Most important it must be free

Because I have used the Basic programming language and the PICAXE is programmed in a Basic like language I searched to find Basic languages for Windows. I found Just Basic [1] that seemed to fit the requirements so I started to work my way through the tutorial to make sure that I could produce the program that I wanted.

The program requirements were:

- Read the compass data from a com port
- Open a window on the PC and display the bearing read from the com port
- Display the bearing as an arrow pointing in the correct direction

2

Modification to the compass head unit

The first problem that I found was that the baud rate that I chose for the RS232 link from the compass head unit to the display unit is not one of the standard speeds supported by Windows. The com port also needs a terminator for a string of data. This requires a small change to the program in the PICAXE microcontroller program to change the baud rate and add a carriage return to the end of the data being sent. Table 1 shows the changes to the program. Of course the baud rate of the original display unit must be changed as well if it is to be used in the future. The program change required is shown in Table 2.

I modified the head unit and connected it to the serial port on my PC with the data from the Head Unit connected to pin 2 of

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Table 1: Changes required to the Head Unit PICAXE program.

Original line:
serout C.7, T600_4, ("BRG",BTTH, BTH, BHND, BTEN)   ; send data at 600 baud to pin C.7

Change to:
serout C.7, T1200_4, ("BRG",BTTH, BTH, BHND, BTEN,&D)   ; send data at 600 baud to pin C.7

the 9 way D type (it has a real serial port, more about PCs without real serial ports later) and ran the dumb terminal program to check that I could receive the bearing data. Sure enough the terminal display showed the message "BRGnnn".   All I had to do now was to write the required program.

3

The program

Fortunately one of the example programs that comes with Just Basic when it is downloaded is a clock program [2]. This contained most of the code that I needed to produce a display showing an arrow pointing in the direction of the bearing being read by the compass. After several attempts I had a program that would do just that. The main trouble I had was understanding the method of deleting the data stored for the arrow when it is removed from the display and re-written in the new position. Each time a graphic is drawn with Just Basic the data required is stored in a "segment" and the old segment must be deleted otherwise the memory used by it is not released and the memory that the program is using keeps on expanding. I had several programs

that grew like Pinocchio's nose until I worked out how the clock program achieved the task.

The next task was to read the data from the serial port. That was extremely easy with Just Basic so that part of the program worked very quickly. The only operational problem was that the com port number was hard coded in the program. The solution was to pick up the com port number from a file called "compass.ini" (this file must be in the same directory [folder] as the program). The file should contain just the serial port number, the program opens the file when it starts and uses that number to open the com port.

Now I had a working program, the next task was to see if it could be saved in a form that I could load on another PC. I also wanted to see if the program would work on a PC running Windows 7 and I could use a USB serial port because my Windows 7 laptop does not have a real serial port.

Making a transportable program is easy and fully described in the Just Basic help file. Just save the program as a token file (.tkn) and save a copy of the Just Basic runtime engine (JBRUN101.EXE) with the same name as the tokenised program along with several support files and copy them all to the target PC then run the

Fig 2: Changes required to the Display Unit PICAXE program.

Original line:
serin C.0, T600_4, ("BRG"),b2, BEARING, BEARING_1, BEARING_2 ; read the serial port

Change to:
serin C.0, T1200_4, ("BRG"),b2, BEARING, BEARING_1, BEARING_2 ; read the serial port

Table 3: Just Basic program to display the compass in a Window on a PC.

```
'Electronic Compass display - September 2012
'Set the size of the window
        WindowWidth = 180
        WindowHeight = 210
'Do not show the Just Basic main window so that only the compass display is shown
        nomainwin
'Open compass.ini and get the com port number
        open "compass.ini" for input as #ini
         port$ = input$(#ini, lof(#ini))
        close #ini
        let N = val(port$)
'Open the comm port defined by the number in the file compass.ini
        open "com";N;":1200,n,8,1,ds0" for random as #com2
'enable the com error handler
        oncomerror [commerror]
'Open the window to draw the compass display
        open "Compass" for graphics_nsb as #draw
'Detect closing the window so that things can be tidied up
        print #draw, "trapclose [exit]"

'Draw the compass outline with cardinal point markers
        print #draw, "home ; size 2; down ; circle 55"
        print #draw, "place 80 25"
        print #draw, "font courier_new 12 bold ; color black"
        print #draw, "\N"
        print #draw, "place 80 160"
        print #draw, "font courier_new 12 bold ; color black"
        print #draw, "\S"
        print #draw, "place 10 90"
        print #draw, "font courier_new 12 bold ; color black"
        print #draw, "\W"
        print #draw, "place 150 90"
        print #draw, "font courier_new 12 bold ; color black"
        print #draw, "\E"
        for angle = 0 to 270 step 90
        print #draw, "up ; home ; north ; turn "; angle
        print #draw, "go 55; down ; go 5"
        next angle
        print #draw, "flush"

'Set some initial values and start the timer to trigger reading the com posrt and displaying the bearing
        let b = 0
        let flag = 0
        timer 100, [main]
        wait

'The main program loop, checks the flag for first time through, removes last display if not
'Check the com port, display the result
        [main]
        if flag > 0 then gosub [remove]
        let flag = 1
        gosub [checkdata]
        gosub [display]
        wait
```

Fig 3: Just Basic program to display the compass in a Window on a PC.

```
'Display the bearing as text and graphic
        [display]
        print #draw, "place 5 10"
        print #draw, "font courier_new 12 bold ; color black"
        print #draw, "\Bearing\ "; b
        print #draw, "up ; home ; north; turn "; b
        print #draw, "size 3 ; color black ; home ; down ; go 50"
        print #draw, "size 3 ; color black ; turn 150 ; down ; go 10 ; turn 180 ; up ; go 10 ; turn 240 ;
                        down ; go 10 "
        print #draw, "flush"
        print #draw, "segment segid"
        return
        'Delete the last data from memory to stop memory loss and remove the displayed data
        [remove]
        print #draw, "delsegment "; segid - 1
        print #draw, "up ; home ; north; turn "; b
        print #draw, "size 3 ; color white ; home ; down ; go 50"
        print #draw, "size 3 ; color white ; turn 150 ; down ; go 10 ; turn 180 ; up ; go 10 ; turn 240 ;
                        down ; go 10 "
        print #draw, "place 5 10"
        print #draw, "font courier_new 12 bold ; color white"
        print #draw, "\Bearing\ "; b
        return
'Check to see if there is any data at the com port
        [checkdata]
        numBytes = lof(#com2)
        if numBytes = 8 then gosub [getdata]
        return
'Get the data from the com port
        [getdata]
        a$ = input$(#com2, numBytes)
        b$ = mid$(a$,5,3)
        b = val (b$)
        return
        [commerror]
'disable the com error handler
        oncomerror
'display message that there has been an error
        confirm "Error when trying to open com port, press any key to close "; answer$
        goto [exit_1]
        [exit]
'ask the user if they want quit
        confirm "Do you really want to close the electronic compass?"; answer$
        if answer$ <> "yes" then wait
        [exit_1]
'If the answer was yes, close the window and com port and end the program
        close #draw
        close #com2
        end
```
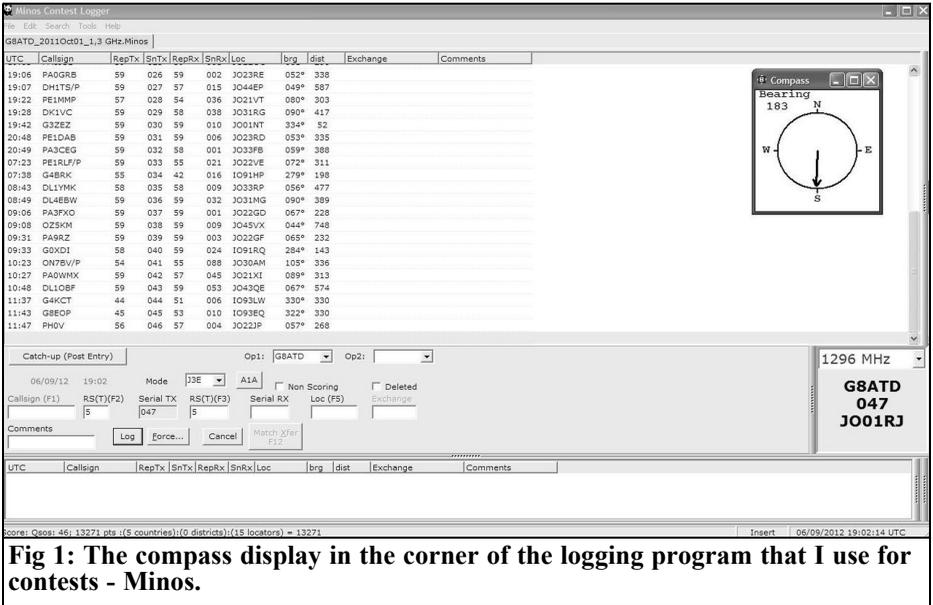
Fig 1: The compass display in the corner of the logging program that I use for contests - Minos.

renamed Just Basic runtime engine. I was pleased to see that it ran on my Windows 7 laptop but the display was very erratic. I suspected that that was due to the USB serial port, to confirm that I ran the program back on the PC that I had used to develop the program but used the USB serial port. Sure enough that was erratic. After some experimentation I added a test in the subroutine "checkdata" that checks the com port to see if it has data. The test only allows the data to be processed when there are 8 data bytes available. This made the Windows 7 display stable but just a bit less responsive that when using a real com port. There was no effect when using a real com port.

The final program is show in Table 3 with an example of the display in the corner of my logging program is shown in Fig 1.

The only problems that I have not mastered are proper error handling if the ini file is missing or the wrong com port is specified in the ini file but it is such simple program that these are fairly obvious. Maybe when (or if) I ever

become a proficient programmer I will be able to sort these out.

4

Conclusion

The program is a useful addition to the electronic compass. I am sure that readers with more experience of Windows programming can produce something much better. If anyone does then I am happy to publish a follow up article.

5

References

[1] Just basic can be downloaded from: http://justbasic.com/

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Gunthard Kraus, DG8GB

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

It is often good to see what's new - so for a change why not a trip to very low frequencies? This article shows how modern components and techniques offer attractive opportunities to implement old or new principles. The circuit simulation answers many questions and solves most of the problems.

1.

What's where?

First, the way that the long wave frequencies are allocated to radio amateurs will be described. The upper limit of interest is set by commercial radio station programs at 150kHz. Amateur radio activities in Europe go down to 137kHz. I found the following overview on the Internet:

Amateur radio operation on long wave is allowed in Germany, in a small segment between 135.7 and 1378kHz. The following "unofficial" band plan is used by most amateurs in Europe:

- 135.7 to 135.8:   QRSS ("Visual CW")

- 135.7 to 136.0:   TX tests & beacons

- 136.0 to 137.4:   normal CW ("conventional")

- 137.4 to 137.6:   data modes

- 137.6 to 137.8:   QRSS ("Visual CW")

It is equally lively below 100kHz between 50 and 80kHz. The well known time signal transmitter DCF77 is on 77.5kHz. Under 50kHz all kinds of things are possible such as the OMEGA navigation, U-boat communication and interesting geological effects. There is more information on the standard home page [1]. It has the descriptive title "VLF activities below 22kHz" and provides information about many interesting things (e.g. what are "Schumann resonances"?) and provides links to virtually everyone active this area.

You should obtain the relevant frequencies and a band plan for your area. You can find this on the Internet or in a specialised book for "world receiver owner" that shows the transmitter frequencies. Without such an overview listening is pointless and you will loose interest quickly because you don't know where to find a specific station or an interesting announcement.

2.

The issue of antenna - a small resume

Vertical polarisation is usually used for transmission so that the electric field lines are perpendicular to the ground.

The magnetic field is at right angles to the electric field component and therefore parallel to the Earth's surface. This "field polarisation" must be used because propagation is mainly by ground wave. The waves penetrate deeper and deeper into the Earth with decreasing frequency. Horizontal electric fields would cause corresponding currents in the ground and heating effects. This absorbs the power and costs the far field range.

So wire antennas that take advantage of the electric field component for reception must be parallel to the earth surface. You should remember that such antennas are usually much smaller than the wavelength (the wavelength is 2000m at a frequency of 150kHz) and not resonant. So they are basically just a capacitor with two plates (Earth and antenna wire) where the incoming electric field generates a voltage. This voltage is increased with a greater field line length that can be captured between the "plates". Therefore it is important to suspend the wire as high as possible with a good ground plane. It is precisely the relationship:

Antenna power = Incoming electric field strength x effective height of the antenna

The antenna capacity then determines the internal resistance of the resulting voltage source.

The low frequencies giving increasing wavelength unfortunately completely exclude such antennas as dipoles and Yagis with their good resonance effects. This leaves only the wire antennas with capacitive internal resistance that is unpleasant. An electromagnetic interference fog is always present at this frequency range that not only consists of atmospheric disturbances but also from all types of consumer products: switch mode power supplies in computers, TV sets, vacuum cleaner motors, fluorescent lamps, energy saving bulbs are some possibilities.

The far filed of the electric and the magnetic field obey the following formula:

$$\frac{E}{H} = 120 * \pi = 377\Omega$$

where E is the electric field in V/m and H as magnetic field in A/m.

The electric field is often predominant in the near field therefore it makes sense to try to use the magnetic field component for receiving. The "magnetic field" immediately brings to mind "induction". The incoming magnetic field component can induce a voltage in a coil, this is the principle of the loop antenna.

When circuits are used where the electrical field component cannot affect the performance, improvements are expected. However an "electric antenna" is far superior to a"magnetic antenna" in the amplitude of the antenna power generated from the incoming field. It is no problem at all to produce up to a factor of 50 or 100 times greater voltage from a simple wire antenna mounted high up compared to the loop antenna. Luckily it not only depends on this voltage amplitude but crucial, and especially for weak signals, on the noise signal and because of that fact the loop often looks better in real use.

Now for the active antenna. As in a "normal" antenna the receiver input is used as a terminator. This results in the "antenna aperture" that together with the field strength produces the active power fed to the receiver input. Some calculations are of course required because of the many resistances involved ranging from the complex internal resistance of the antenna and all possible losses.

The active antenna is basically an amplifier (regardless of whether a wire antenna or a magnetic antenna is used) that feeds the receiver with the amplified open loop voltage generated in the antenna. Therefore the signal can be transmitted over greater distances.

3.

Measuring instruments required for your own LW experiments

You can only test and implement these ideas and simulation circuits if the appropriate test instruments are available. Each developer probably has an oscilloscope but a method to determine inductance is also very important (LCR Meter, professional L bridge etc). You also need a sine wave source with adjustable amplitude and exact frequency display but a separate frequency counter can be used. A more difficult requirement is some kind of receiver with a precise level indication. There are fairly cheap solutions: signal generators can be found at amateur radio flea markets virtually at scrap price, these often have a matching selective level meter. It does not matter if these are valve instruments any "antique box" will do the job. The receiver sensitivity needs to be as high as possible to find weak signals. The upper frequency limit may not be too high, older devices suffer from scale resolution and the lowest usable input frequency. The combination of the W&G PS3 and SPM3 devices proved themselves. Both are completely transistorised, nice and small, and their upper cutoff frequency of 512kHz is enough. A built in battery gives the opportunity to work in a low noise area - in the middle of the prairie! - to carry out measurements. So "SPM3" is an important insider tip in the scene ....

But the ultimate device "HP8556A" was found last year in flea market section of HAM RADIO. It is a plug in unit for the HP spectrum analyser (HP141T / 8555 / 8552B).) that I have used for many years in my own workshop. This very rare and probably little sold plug in type "HP 8556A" has two frequency ranges from 0 to 30kHz and 0 to 300kHz. The minimum measurement bandwidth is 10Hz

and the smallest sweep is 20Hz per division. With a sensitivity of better than -120dBm it is the "perfect measuring machine" and you can "hear the grass grow" for this purpose. Ultimately this module was the final impetus to tackle this project.

4.

Back to basics

As mentioned earlier, an active magnetic antenna is to be considered as a simple coil in a changing magnetic field that induces a voltage. This complies with the following relationship that can be found in any physics text book and comes from James Maxwell:

$$U_L = N \cdot \left(\frac{d\phi}{dt} \right) = N \cdot \left(\frac{\Delta\phi}{\Delta t} \right) \qquad (1)$$

Where "N" is the number of turns and in the brackets is the rate of change of magnetic flux ϕ in the coil.

For a frame antenna in air (and shown later with a mental trick for a ferrite antenna) the magnetic field strength "H" can be substituted:

$$U_L = N \cdot \left(\frac{A \cdot dB}{dt} \right) = N \cdot A \cdot \left(\frac{\mu_0 \cdot dH}{dt} \right) = N \cdot A \cdot \mu_0 \cdot \left(\frac{dH}{dt} \right)$$

$$(2)$$

Where:

- B corresponds to the magnetic induction (or field line density per unit area)

- A is the cross sectional area of the frame that the field lines pass through

- μ_0 is the absolute permeability constant from physics

Now take a sinusoidal magnetic field of
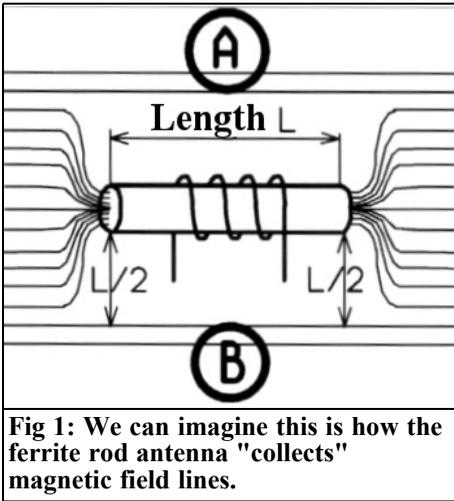
Fig 1: We can imagine this is how the ferrite rod antenna "collects" magnetic field lines.

strength:

$$H = H \cdot \sin(\omega \cdot t)$$

and inserts into the above equation to get the following relationship:

$$U_L = N \cdot A \cdot \mu_0 \cdot \left(\frac{dH}{dt}\right) = \omega \cdot N \cdot A \cdot \mu_0 \cdot [H \cdot \cos(\omega \cdot t)] =$$

$$2\pi \cdot f \cdot N \cdot A \cdot \mu_0 \cdot [H \cdot \cos(\omega \cdot t)]$$

From this you get the peak of the induced voltage:

$$\hat{U}_L = 2\pi \cdot f \cdot N \cdot A \cdot \mu_0 \cdot \hat{H} \qquad (3)$$

After transformation the electric field strength is:

$$\hat{U}_L = 2\pi \cdot f \cdot N \cdot A \cdot \mu_0 \cdot \frac{\hat{E}}{120\pi} = \frac{f \cdot N \cdot A \cdot \mu_0 \cdot \hat{E}}{60}$$

$$(4)$$

The induced voltage increases (for constant field strength) linearly with the frequency, the number of turns and the effective area.

Of course we are trying to make a large antenna to receive very weak signals at a certain frequency. In principle it does not seem to matter whether more turns or a wider frame are used. But the following limits are quickly found:

- More turns unfortunately increase the winding capacity and eventually lead to the natural resonance of the antenna that restricts the use for increased frequency. The following techniques help: e.g. a ferrite antenna with a coil using a cross wrapping technique or a frame antenna with less turns and a significant distance between the individual turns of the antenna.

- Not everyone can enlarge the area of the framework (Images of frame antennas with more than 5m x 5m behind a house can be found on the Internet).

However you can not always do what you want. Those who share her love for the "Schumann resonances of the globe" must deal with the problems of inducing small voltages with frequencies down to to 10Hz or less. Then you must say "up with the number of turns and the frame size" so that you will be able to detect anything. There are links to such experiments on the Internet for the number of turns from 60,000 to 100,000.

A word on ferrite antennas: they save on size but performance depends less to the diameter of the rod than on its length. This can be seen clearly in Fig 1. Suppose the ferrite material has a very large permeability: then it will have almost zero magnetic resistance for the magnetic field lines and as many field lines will try to take this "convenient" path as possible. To get a mental picture of this, if the "detour" to the ferrite rod and back is the same or greater than the direct route through the air then the field lines will not take the detour. These limits are the areas "A" and "B" in Fig 1. Thus there is a circular effective with a diameter that roughly corresponds to the length of the ferrite rod (assuming that the magnetic resistance of the rod to zero).

Logically doubling of the length should double diameter of this effect and hence quadruple the surface area giving a four fold voltage increase. Anyone who glues two identical antenna rods together will be disappointed: the glue line makes a small air gap that takes away some of this effect due to its magnetic resistance. From my own experience (because the practical ferrite material does not have infinitely high relative permeability) doubling the length gives triple the voltage compared to the simple rod with an unchanged number of turns in the coil.

Finally the answer to the question: "what part does the diameter of the ferrite rod itself play?" The limits are as follows:

- When using long ferrite rods the length of the rod determines the active area of the coil (as just described). However ferrite is basically a "non-linear material" therefore when stronger signals are being received or different greater voltages are induced, unpleasant things like harmonic distortion, intermodulation and cross modulation occur. A larger cross section of the rod helps to reduce the magnetic induction in the antenna rod and so the unpleasant effects.

- For very short antenna rods (like those often seen in radio controlled clock modules) the "effective diameter of the capture area" is more due to the diameter of the coils than the length of the rod and the induced voltage decreases accordingly. Distortion is a problem to these antennas as well but it is reduced because less ferrite is involved giving less distortion for strong signals.

The directional characteristics of frame antennas in their simple form needs no further explanation because the rotation method when used as DF antenna are well known. But one thing that is no obvious:

The signals from a frame antenna are always considerably less than that from a wire antenna but the signal to noise ratio can be much better as described previously. Therefore a simple and cheap way to increase the signal from a frame antenna is to make it a tuned circuit with a high quality capacitor so that the circuit voltage is increased.

Another principle that should be mentioned:

If you short circuit a ferrite antenna you get an interesting effect with increasing frequency. In accordance with the law of induction the induced voltage increases linearly with frequency but its inductive reactance also increases. Neglecting the loss resistance, if the the short circuit current in the antenna is measured it remains constant according to the ohms law over a very wide frequency range (as long as the magnetic field strength does not change). Most commercially available field strength meters work on this principle by showing the short circuit current on a suitably calibrated display. The secret that must be kept in mind is:

An increased number of turns "N" in the antenna causes the induced voltage to increase linearly with "N". However the inductance and inductive resistance increase proportional to N^2. Thus the short circuit current will decrease at the same rate, when increasing the number of turns. Therefore most field strength meters use only a single turn (and enlarge the capture area instead if possible).

5.

First project: Resonant antenna with an amplifier for DCF77

5.1. Concept

Usual DCF77 receivers such as radio controlled clocks use only a small ferrite rod antenna for the receiver and then
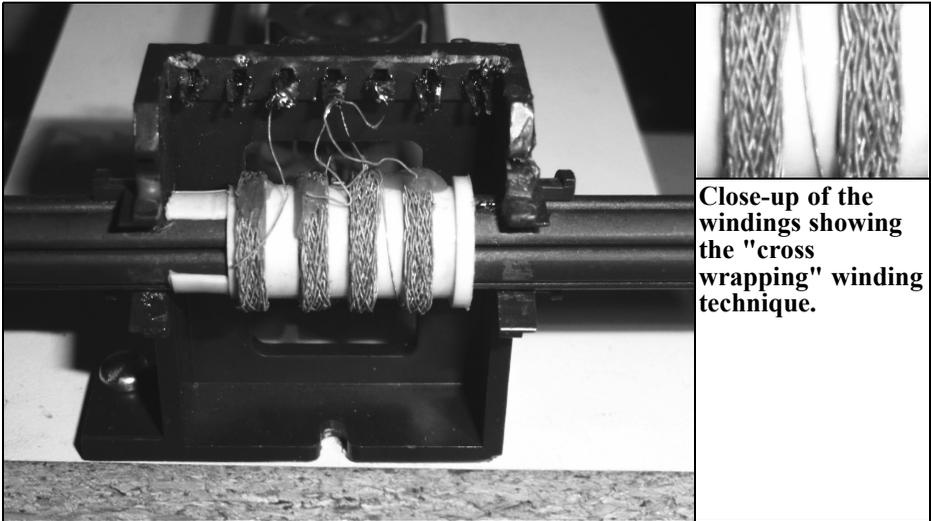
Close-up of the windings showing the "cross wrapping" winding technique.

Fig 2: A photograph of the ferrite rod antenna showing the individual windings of Litz wire. These windings reduce the self capacitance and hence increase the self resonant frequency.

filter out the DCF77 signal with a simple but extremely narrow band crystal filter. Combined with a special receiver IC this makes as a very compact low cost and high quality solution that works well in various environments.

The starting point for this project should be a ferrite rod antenna with a parallel capacitor interfaced to an adjustable gain high isolation amplifier to raise the signal level and give a 50Ω output.

5.2. Preparation

- First, two disused but identical ferrite rod antennas from old radios were found and all windings except the long wave winding were completely stripped (because if there is any other windings present you will fight with quite peculiar resonance effects - this as a warning). The long wave winding, that was divided into four individual windings, was deliberately positioned exactly in the middle of the rod (Fig 2). This was measured with a classic inductance meter, "LARU" from Rohde & Schwarz as an in-

ductance of about 4mH (4.05 mH for the first and 3.98mH for the second antenna found).

- An "Ideal operational amplifier" was used in the simulation. However the following are required to make a practical circuit: the property "ideal" used for the operational amplifiers in practice means up to approximately 200kHz. The starting point is a maximum gain of approximately 40dB of gain with an addition 20dB reserve. 60dB is a thousandfold voltage gain so the gain bandwidth product required is

    200kHz x 1000 = 200MHz

  (Whoops - who would have thought?). The SMD version of the OPA355 from Texas Instruments was chosen. This requires a supply of only 5V between the pins "V+" and "V" (caution: only 7.5V is the maximum allowed). The Spice model was found on the Internet.

- In the Spice simulation the ferrite rod antenna is considered as a

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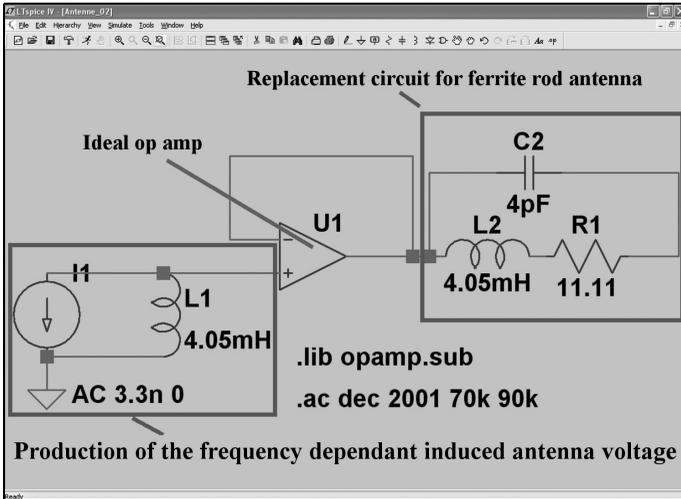
Fig 3: A Spice simulation of a ferrite rod antenna can be carried out using these "tricks".

voltage source whose voltage increases linearly with frequency (at constant reception field strength). Because this is not something that is available for Spice a simple replacement circuit was constructed from a constant current source, an inductance of 4.05mH and an ideal op amp used as impedance converter and used to generate the frequency dependent coil voltage. The output resistance is almost zero ohms so this can be connected to the actual replacement circuit of the ferrite rod antenna and capacitor. This antenna replacement circuit consists of the antenna inductance (4.05mH), a series resistance as a representation of the coil losses and the coils own capacity giving the natural resonance of the antenna.

Fig 3 shows the simulation circuit with explanations - the details and justifications for the component values follow later.

5.3. Sub project: Determining the Q of a ferrite rod antenna

Quality values are required for the filter, precise knowledge is important for a correct simulation. But what to do if

there is no suitable Q factor meter for this frequency range?

Therefore, the following measuring technique was used:

The output of a professional signal generator with the internal resistance of 50Ω feeds a signal meter with an input impedance of 50Ω over a 50Ω cable (RG58). The connecting cable was broken in the middle and both ends connected to the existing solder eyelets on the ferrite rod antenna. This was actually a series connection comprising of the ferrite rod antenna (4.05mH) and a high quality 1.039nf Styroflex capacitor (the capacitor consisted of a 1nF and a 39pF capacitor in parallel). This resulting series circuit had a resonant frequency of approximately 77.5kHz and therefore only the pure series loss resistance of the coil was left to absorb power at this frequency. The losses in the Styroflex capacitor were neglected so only the losses in the coil remained.

A short circuit was place across the break in the connecting cable and the output level of the signal generator increased until the signal meter showed a full scale of 100mV. Thus, the voltage of the signal generator was 200mV. The shorting link was removed and the voltage at the
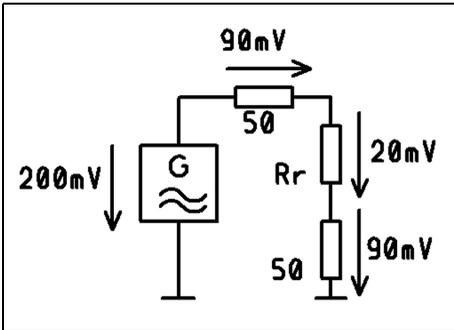
Fig 4: The equivalent circuit for determining the coil losses and the Q. (see text.

receiver input fell from 100mV to 90mV, this is explained in Fig 4 and as follows:

- The coil current could be calculated easily from the voltage at the receiver input (90mV) and it's input impedance (50Ω):

$$I_{coil} = \frac{90mV}{50\Omega} = 1.8mA$$

- Because there must also have been 90mV developed across the 50Ω internal resistance of the signal generator then 20mV must have

been developed across the loss of resistance of the coil. Thus the resistance was:

$$R_{series} = \frac{20mV}{1.8mA} = 11.11\Omega$$

- The Q of the ferrite rod antenna is thus:

$$Q = \frac{X_L}{R_{series}} = \frac{2 \cdot \pi \cdot f \cdot 4.05mH}{11.11\Omega} = \frac{2 \cdot \pi \cdot 77.5kHz \cdot 4.05mH}{11.11\Omega} = 177.5$$

Fig 5 shows the equipment used to make these measurements, more detail can be seen in the picture. The frequency counter was connected to the VCO output of the HP3410A AC microvoltmeter. A terminating resistor of 50Ω was connected to the input of the microvoltmeters because the voltmeter itself has a very high input impedance. It is interesting to see how the two 50Ω cables were mounted on a wooden board and connected with bare wires to the ferrite antenna and the series resonant circuit. This meant that it was easy to bypass the tuned circuit with a screwdriver, adjust the output level so that the voltmeter displayed the desired full scale (100mV) and remove the screwdriver. The fre-

Fig 5: The equipment used to measure the coil losses and Q.
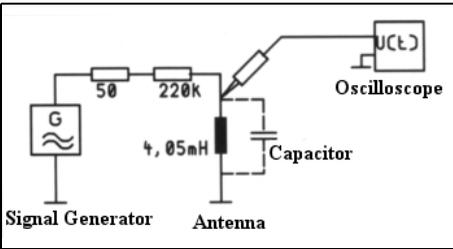
Fig 6: The natural resonant frequency of the antenna was determined with this circuit.

quency can be changed until resonance is reached giving a maximum on the display of the voltmeter.

5.4. Another sub project: Determining the natural resonance

The circuit used to determine the natural resonant frequency of the ferrite rod antenna is shown in Fig 6. A 220kΩ resistor is connected between the signal generator and the ferrite rod antenna that is the parallel circuit being measured. This should avoid excessive damping by the 50Ω internal resistance of the signal generator. No additional capacitors are connected to the antenna. The voltage resulting across just the antenna is measured with an oscilloscope using a 10: 1 probe, it showed a maximum at 661kHz. For this resonant frequency the circuit capacitance has a value of:

$$C_{coil} = \frac{1}{(2 \cdot \pi \cdot f)^2 \cdot L} = \frac{1}{(2 \cdot \pi \cdot 661kHz)^2 \cdot 4.05mH} = 14.3\,pF$$

Looking at the probe gives the following information:

Tektronix 10x 10.5pF 10M 1m

Only the input capacitance of 10.5pF is of interest that must be subtracted from the value measured. This leaves about 4pF (3.8pF to be precise) as the self capacitance of the coil that is a very small value that has been achieved by the cross wrapping technique of the winding and the division of the windings into four parts. The divided windings reduce a capacitance of 16pf for each individual winding as show in Fig 7. This effect was exploited at the beginning of wireless technology more than 100 years ago.

5.5. The amplifier

This is designed as a two stage amplifier. The first stage is a non inverting stage therefore has a very high input impedance to prevent damping of the ferrite rod antenna and capacitor (Fig 8). Its voltage gain can be varied using the potentiometer R9 between 11k and 111k. The second stage (inverting amplifier) provides an additional 10 times fixed gain with an output impedance of 50Ω. So it is ready for the application with gain settings between a minimum of 5.5 and a maximum of 555 when connected to the spectrum analyser with its input impedance of 50Ω

Can you guess: Unfortunately these requirements are associated with a high level of effort for the practical implementation and the final circuit diagram was quite complex. There is for example the problem of only working with an operating voltage of +5V. The inputs of the op amps must have a bias of +2.5V that is generated by the resistors R7 and R8 and
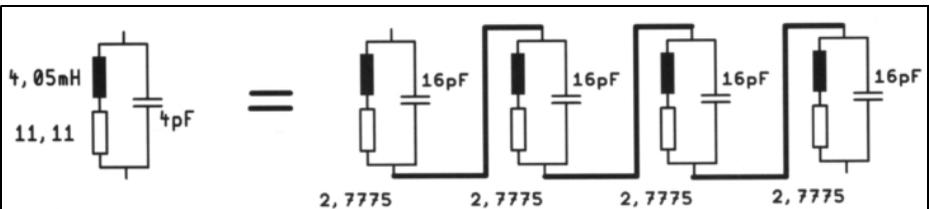
Fig 7: This shows the method used to reduce the self capacitance referred to in Fig 2. (see text).

Fig 8: This module was designed to amplify the weak signals from the ferrite rod antenna. (see text).
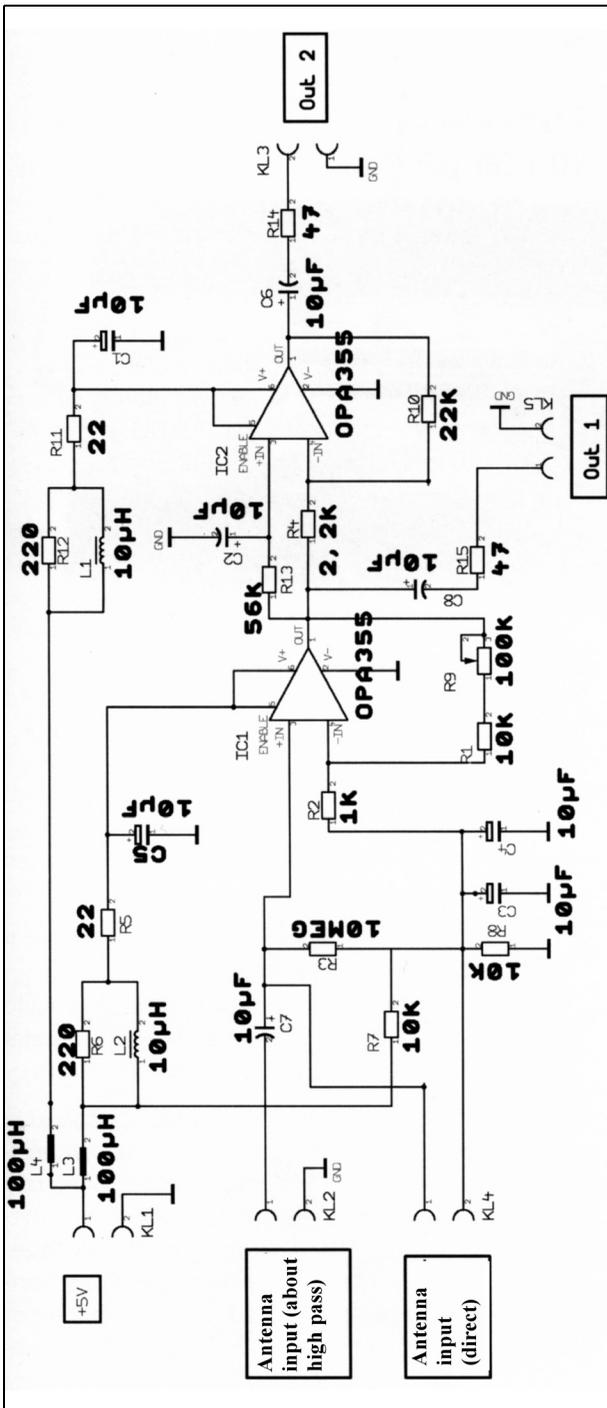
then isolated. DC coupling between the first and second stages reduces any gain loss at low frequencies. It is inevitable that the power supply has to be isolated with elaborate decoupling.

Fig 9 shows the first prototype with an antenna. It can be seen that some details have been changed after measurements and investigations. For those who would like to build the amplifier, Fig 10 shows the final state of the circuit board. It is of course in proven GHz technology: lower continuous ground plane, carefully separated ground islands on the top that are connected to the ground level with vias, clear separation of both stages and RF outputs using SMA connectors. The +5V power for the amplifier is fed in on a female SMB connector and it is all mounted in a machined aluminium housing with lid. These are details that are learned with flesh and blood by UHF and GHz developers. Therefore they should work even at 77kHz without any problems.

Before the first attempts were made to receive signals the tuned circuit was set up with a small trimmer capacitor (10-50pF) in place of the 39pF capacitor and carefully adjusted to

Fig 9: The first prototype active antenna for DCF77 with a single coil works well.

77.5kHz. To make this adjustment a length of RG58 coax cable with a BNC connector on one end and two banana clips on the other end was connected to the output of the a signal generator with a digital frequency display. The two banana plugs were connected thus forming a short circuit loop that was placed near the ferrite rod antenna, the short circuit current induces power into the antenna. The output level was increased until an output voltage of 1V at 77.5kHz was measured with an oscilloscope connected to the output of the amplifier. The trimmer capacitor was adjusted to give the highest voltage at this frequency.

5.6. Measurement results and evaluation using a Spice Simulation

Now for the exciting part; the trial to

Fig 10: The PCB layout shows the additions needed for a high gain amplifier in a small space. (see text).
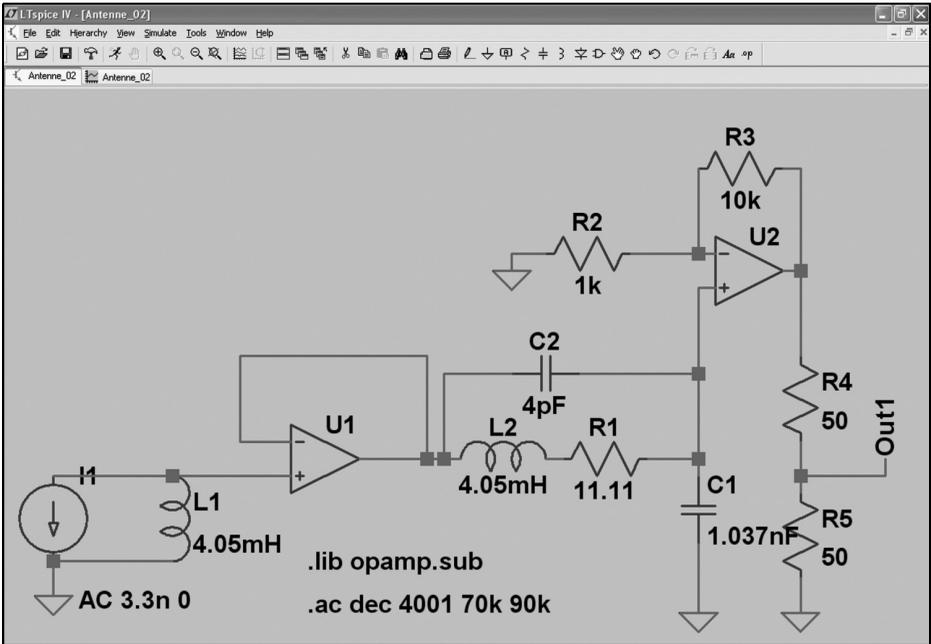
Fig 11: The circuit for simulation of the antenna plus amplifier.

receive the real DCF77 signal in the basement workshop. Result: a display of 31dBm 6.3mV measured at the output "OUT1" with a minimum gain (G = 5.5) using the spectrum analyser. A Spice circuit diagram as shown in Fig 11 was produced to check the results. The first trial simulation showed that the LTSpice "Ideal standard OP opamp " gave virtually the same results as a simulation with the OPA355 with all the power supply components. All further simulations were
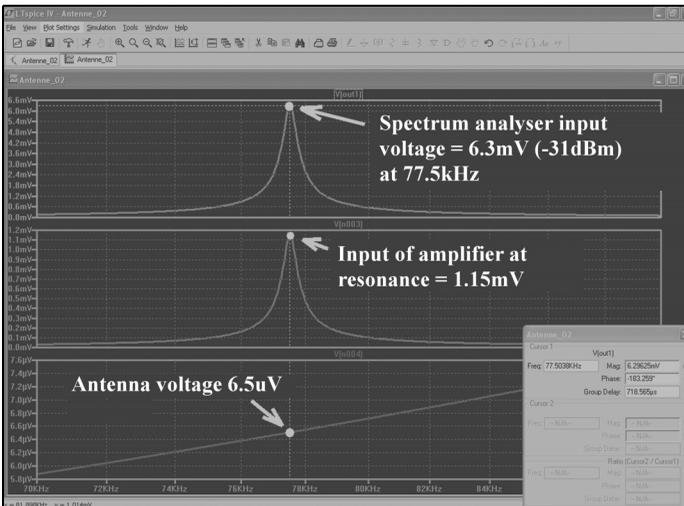
Fig 12: Simply fascinating: more accurate knowledge of the component data means that all the signals in the simulation circuit can be measured correctly.
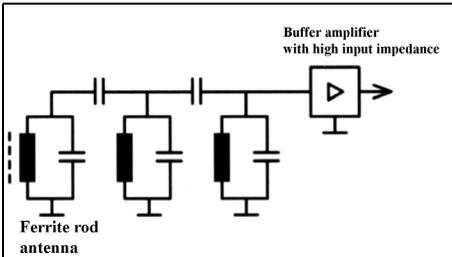
Fig 13: Now for something different: developed from the simple circuit, a circuit with three bandpass filters and a buffer amplifier.
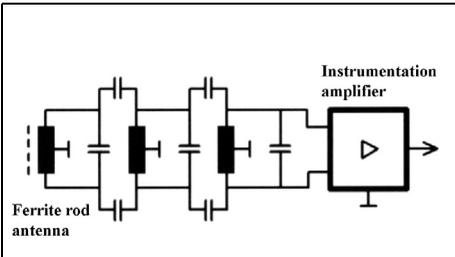
Fig 14: It becomes a fairly elaborate "beast" when it is made symetrical.

made with the simplified circuit. But do not forget to include the Spice model for the OPs with the directive

.lib opamp.sub

to make the simulation.

The simulation results for the simplified circuit are shown in Fig 12. The signal measured at the input of the analyser of 6.3mV requires a constant current of 3.3nA from the "antenna voltage source". This induces a signal of 6.5µV that is increased by the resonance effects at 77.5kHz to 1.15mV at the input of the amplifier.

Note:

The level measured by the analyser in my basement is not constant as claimed because of the fading at this frequency. This can be observed at up to 3dB in the course of one day or from one day to the next. This is because the propagation is affected by the weather with high humidity being most noticeable. Shortly before dawn fading of 2dB with a frequency of about 1Hz can happen suddenly, a new experience for me. Perhaps I will use this "machine" later, together with a notebook as a recording station, to monitor the field strength for 24 hours and let you know more.

6.

Second project: An antenna with triple bandpass input filter

6.1. Concept

During measurements discussed above quite a high noise level was noted. Therefore it seemed appealing to produce a highly selective version of the active antenna. There was an ulterior motive for this selectivity: I planned to feed this 77.5kHz into a USB sound card with a D/A sampling frequency of 192kHz. A direct conversion receiver is theoretically possible up to half the sampling frequency (i.e. 96kHz). The DCF77 signal is well within this limit. With the high attenuation of the filter and the selection achieved it should be possible. The Nyquist / Shannon condition (no useful signal can be recovered from more than half the sampling frequency otherwise there are undesired effects of "Aliasing") is almost perfectly meet. Using a sound card gives an easy entry into this interesting area.

Figure 13 shows the proposed principle: The ferrite rod antenna is used as the first coil in a triple bandpass filter with capacitive coupling. The output of the filter is fed to an amplifier and increased sufficiently. The resulting output voltage with an internal resistance of 50Ω is available on an SMA connector.

The technique to make this circuit less
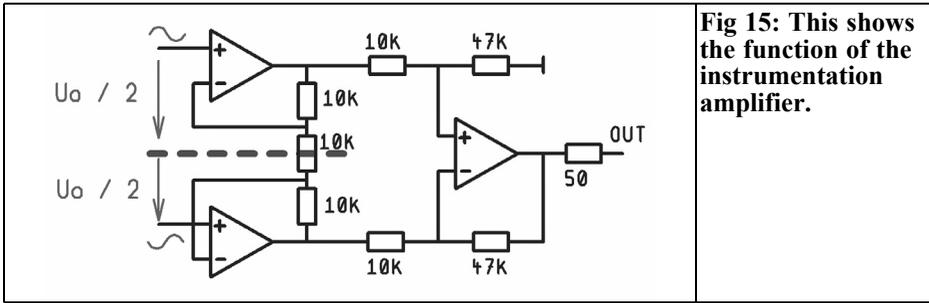
Fig 15: This shows the function of the instrumentation amplifier.

susceptible to electrical interference fields is to make it symmetrical. The ferrite rod antenna coil and the two other coils should have a centre tap and amplification of the filter output signal should use an instrumentation amplifier as shown in Fig 14. The amplifier consists of three operational amplifiers, it has a symmetrical input and the amplified signal is fed to a single ended 50Ω output. To finish off a further experiment was carried out with a grounded conductive shielding round the antenna that further reduces the impact of unwanted electrical fields. So there was plenty of work to do.

6.2. Preparations

- The ferrite rod antenna from the first project was used with tha data: L = 4.05mH, Q = 177, R_{series} = 11.11Ω. It only requires a centre tap.

- Two ferrite cores with an AL value of 250nH were found in the junk box to use for coil 2 and 3, they were made from N28 material that is well suited to this frequency range. They also had a centre tap. It is easy to work out the number of turns required, for a 4mH coil and an AL value of 250nH you need the following number of turns:

$$N = \sqrt{\frac{L}{A_L}} = \sqrt{\frac{4000000nH}{250nH}} = 126$$

- The question is: what is the Q of these coils? Using the test procedure described in section 5.3 these

coils (with 126 turns of 10 x 0.05mm litz wire) gave a voltage of 95mV instead of 100mV without series. As a result only 10mV are developed across the loss resistance and it's value is:

$$R_{series} = \frac{10mV \cdot 50\Omega}{95mV} = 5.3\Omega$$

This gives a coil Q of

$$Q = \frac{2 \cdot \pi \cdot 77.5kHz \cdot 4mH}{5.3\Omega} = 367$$

Which is a very good value.

6.3. Simulation and measurement

The somewhat complicated instrument amplifier should be considered first. It is quite complicated due to the symmetrical arrangement, the three active elements are shown in Fig 15.

The signal, symmetrical around ground, from the triple bandpass filter is fed to each input of the high impedance non inverting amplifiers. An imaginary line can be drawn through the circuit to show the symmetry with each input amplifier reinforcing the signal by the amplification factor set by the non inverting circuit:

$$V = 1 + \frac{10k}{5k} = 3$$

So each half of the antenna signal is filtered and reinforced by a factor of 3
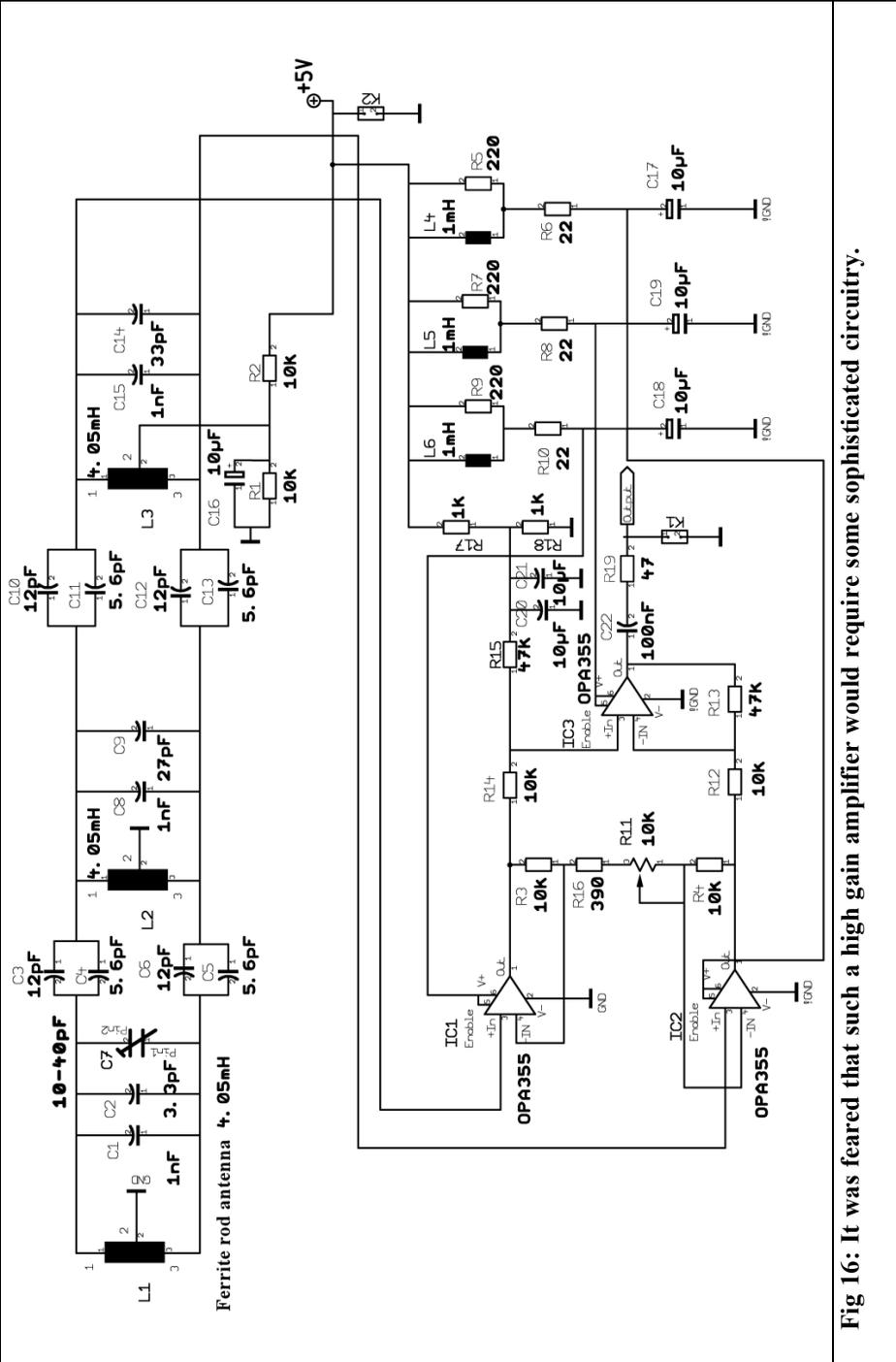
Fig 16: It was feared that such a high gain amplifier would require some sophisticated circuitry.
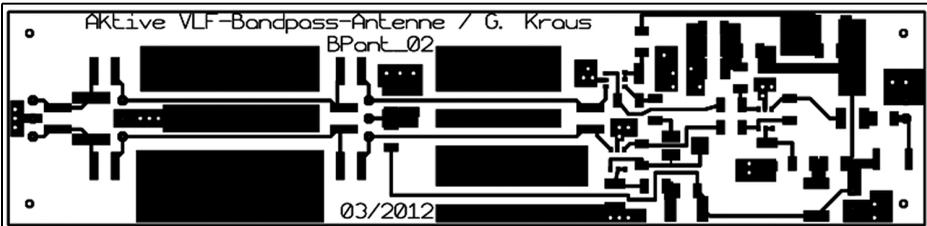
Fig 17: The PCB layout for the three bandpass filters and following amplifier.

but the symmetry isn't lost.

Three things happen in the circuit that follows:

- The two signals have been amplified and their phase is still reversed. They are then summed with respect to ground and the exiting phase shift of 180° is eliminated. So the result is nothing more than the amplified signal induced in the ferrite rod antenna that had been filtered.

- This unbalanced antenna signal is amplified in the final stage by a factor determined by the circuit components:

$$V = \frac{47k}{10k} = 4.7$$

- The final stage has a 50Ω output impedance. This gives a reduction by a factor of 2 when measured with a spectrum analyser. Therefore the overall gain of the arrangement is:

$$V_{total} = 0.5 \cdot 4.7 \cdot 3 = 7.05$$

The practical execution is the circuit diagram shown in Figure 16. There are
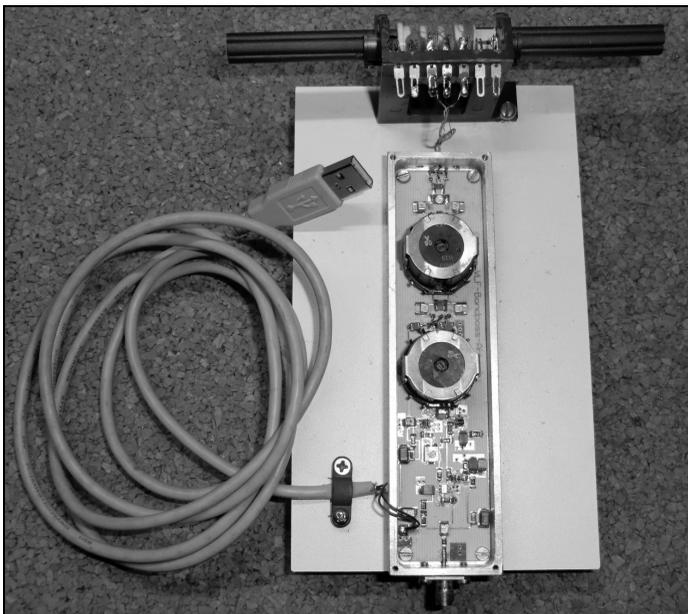
Fig 18: The prototype of the three bandpass filter version with a special power cable.

Fig 19: The effort required is considerable even for the simulation circuit.
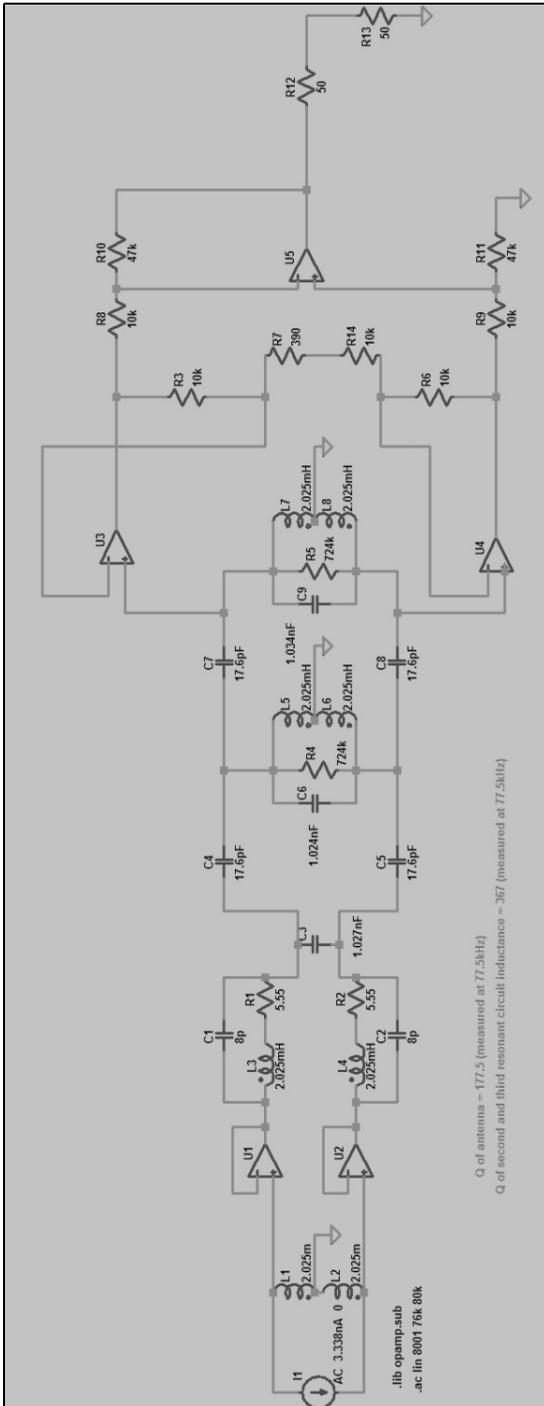
some interesting details:

- The centre tap of the last coil is not directly connected to ground. It feeds the instrumentation amplifier bias of 2.5V generated from the +5V supply.

- A 10kΩ potentiometer is used instead of the 10kΩ resistor shown in the feedback that was the line of symmetry. This can increase the gain of the amplifier and the 390Ω resistor limits the maximum value. Thus the overall gain of the amplifier can be adjusted between 6.87 and 122.8.

- There is a 10 - 40pF trimmer capacitor on the input circuit of ferrite rod antenna for precise adjustment.

For anyone who want to know more details, Fig 17 shows the layout of the printed circuit board and Fig 18 shows the completed prototype. There is a practical tip for this prototype: look closely at the supply cable shown in the picture, it is a standard USB cable that supplies a ground and +5V supply (an SMB connector was off course used for the final version). The reason for this is simple: this is the easiest, cheapest and best way to provide a +5V supply with one cable connected directly to the PC or notebook USB port. An auxiliary power supply for USB peripherals with higher power consumption of more than 500mA" can be purchased for a few Euros; external USB hard drives are their main application. These additional

Fig 20: This shows all interesting voltage gradients up to the induced antenna voltage.

Antenna voltage = 3.3uV at 77.5kHz

135uV in the middle of the tuned circuits

364uV at the input of the amplifier

5mV at the output of the amplifier

power supplies serve their purpose well, they are precisely regulated and still very small. They have a long power cord so the antenna is far enough away from the noise source generated by the switch mode power supply.

Now for the exciting part, the antenna is carefully tuned (see Chapter 5.4) and then the first measurements made with the spectrum analyser:

The analyser showed a DCF77 signal level of -33dBm that corresponds to a voltage of exactly 5mV RMS.

Now the circuit was simulated for AC
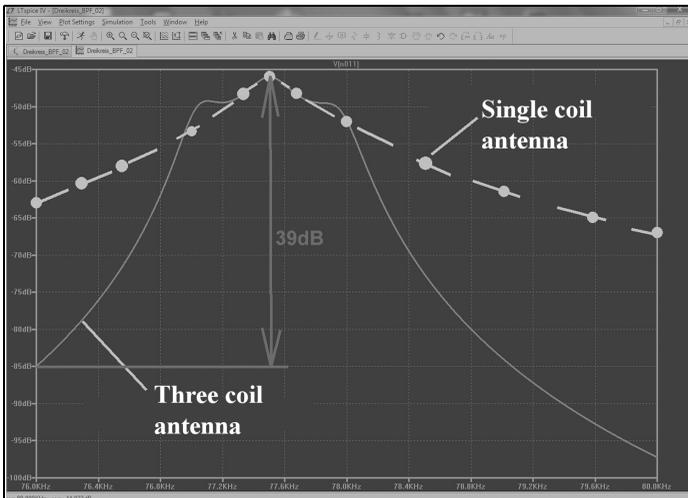
Fig 21: The selectivity properties of the three coil bandpass filter are fascinating compared to the single coil.

Single coil antenna

39dB

Three coil antenna

sweeps from 70 – 90kHz. Unfortunately the elaborate circuitry and the simulated signals should be on separate Figs because the interesting details and component values are difficult to read. The circuit created is shown in Fig 19. The losses of the second and third tuned circuits with a Q of 367 were taken into account as parallel resistor of 724kΩ. Fig 20 shows the simulated voltage gradients. It is interesting to see the effect of the individual bandpass filters on the frequency response. Fortunately the resonant frequencies of the individual circuits as well as the coupling capacitors can be easily changed on the PC screen to search for the best compromise for the frequency response.

This best result from the simulation is the 3.3µV antenna voltage, this is only half of the symmetrical signal because it is measured with respect to ground. If it is doubled we get 6.6µV and almost the same as the 6.5µV from the simple antenna in Chapter 5.5. Such results are a real joy. With such correct simulation results all other interesting signals can be obtained directly from the simulation diagram giving a more accurate knowledge of all circuit details.

One thing that should not be missed that confirms the reason for the triple bandpass filter is the direct comparison of the selectivity curves in dB for the two antenna versions shown in Fig 21. The curve for the triple bandpass filtered antenna shows that at 76kHz the attenuation is almost 40dB and 52dB at 80kHz! compared to the value at 77.5kHz

to be continued

X

References

[1] http://www.vlf.it/ (Radio waves below 22kHz) To be found on the Internet
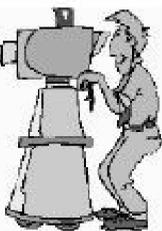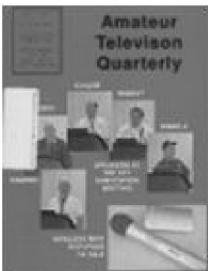
Paolo Petrini, IW1ACL

A modern Test Set for Network Analysers

This article describes a design by Paolo Petrini. It has not been constructed but covers some interesting concepts - Andy

1.

Overview.

The Network Analyser (N.A.) is basically a set of RF/microwave receivers feeding a calibrated receiver able to perform some basic functions such as computing, in modulus and phase or the ratio of two signals. The simplest N.A. has two receiver inputs, one called test and the other used as reference; modern instruments have four or more independent receivers.

In order to fully characterise a two-port device (the most common task) defined by its scattering parameters, we need a piece of equipment (Test Set) that transfers the power signals a1 a2 b1 and b2 to the receiver. Typically, when measuring forward parameters S_{11} and S_{21} a_1 is used

as reference, while a2 is used as a reference for S_{22} and S_{12}. See Fig 1. (Remember that a1 and a2 are the incident power signals, while b1 and b2 are the reflected power signals).

This solution that reverses the driving port from 1 to 2 depending on the parameter investigated, is perfect when dealing with passive circuits such as filters. This solution was chosen because real time measurements (and displayed results) were possible with all analogue instruments available in the 1970s.

But what happens with an amplifier under test? This simple solution works only under the assumption of linearity i.e. very small signals. Whilst in linear mode the output stage can be driven (b2) and the measured impedance is correct.

However in most cases the amplifier is not operating in linear mode (class C) or the investigation has to be performed at a given power level, so a different set up is needed.

Unfortunately this solution requires a digital computer, because real time display is almost impossible.
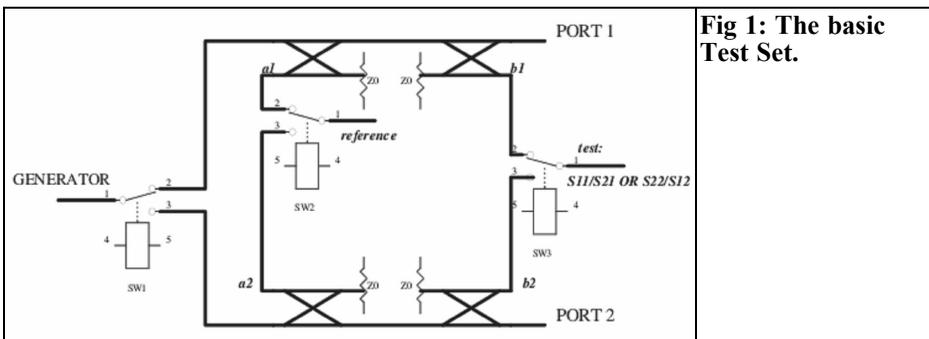
Fig 1: The basic Test Set.
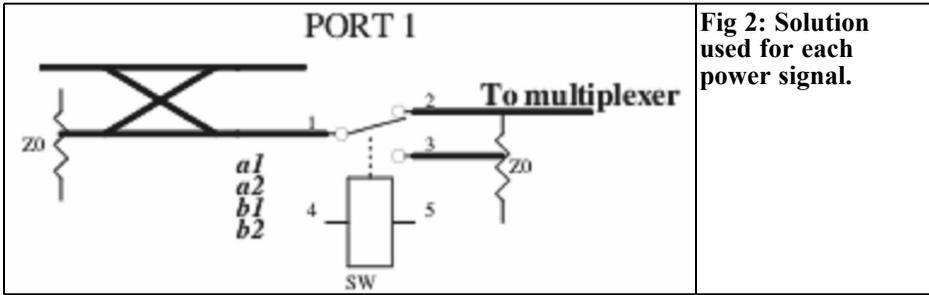
Fig 2: Solution used for each power signal.

First of all a common reference signal is needed, that is a portion of the generator output (using a power splitter or a directional coupler; this signal is fed to the reference input of our N.A).

Then all the power signals are acquired either directly with four independent receivers or multiplexed. In this article we assume a N.A. with one reference input and one test input. Multiplexing can be done with mechanical or solid state switches; in any case provision for at least 80dB isolation is mandatory, and all unused ports must be terminated.

In Fig 2 we see the solution that has to be adopted for each power signal. The wanted signal is acquired by means of a broadband directional coupler and feeds an SPDT switch that connects the coupler's output to a low reflective load or the multiplexer. When connected to a load we assume an attenuation of at least 40dB towards the mux.

The multiplexer as well will have at least 40dB channel to channel isolation, so the >80dB requirement is met. The complete setup is shown in Fig 3. Depending on the type of switches and components available, a different measurement strategy is needed.

2.

Circuit description

The generator is connected to a broadband 6dB power splitter, or if preferred to a broadband 10dB directional coupler; the latter solution provides a 5dB stronger signal for the measuring ports (recommended above 10GHz).

One of the ports, by means of K2, is either terminated or connected to K1 that then selects the attenuated generator or $a1_{ref}$ (for compatibility with the older type of setup) as reference for the network analyser. The other port goes to the circuit made of K3 K4 K6 that connects the generator to port A or port B and terminates the other port. As an alternative the user can buy a terminated transfer switch (such as HP 33312C/15 [8763C] series or similar devices).

Branch A and branch B go to the corresponding measuring ports via two medium to high directivity directional couplers. They are coupled by means of the circuit shown in Fig 2, the ports go to four positions transfer switch which finally is connected to the test port of the network analyser.

So the analyser will measure A1/Ref, A2/Ref, B1/Ref, B2/Ref from which the S-parameters are derived by division (complex math)

3.

Solid state switches.

This type of switch can be operated infinite number of times so, for each frequency of interest, all the parameters are acquired, plus eventual other values
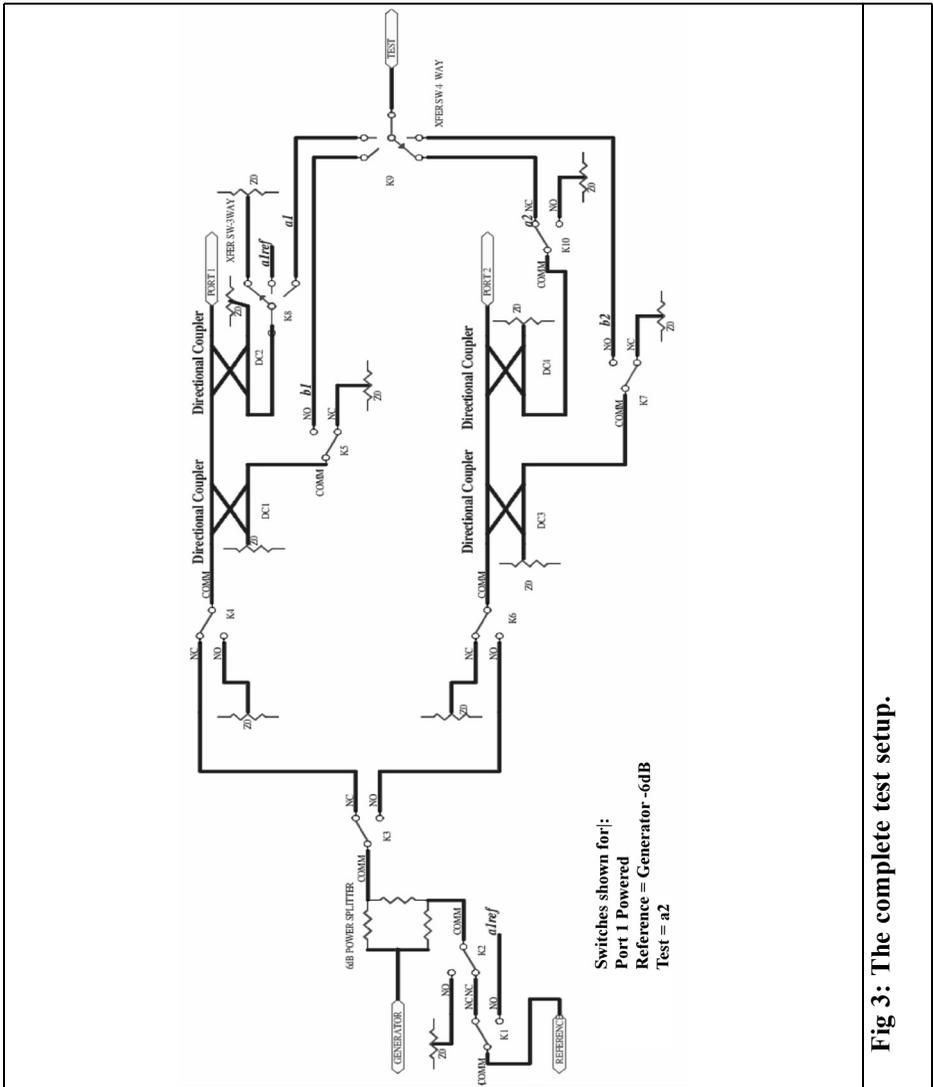
Fig 3: The complete test setup.

(in addition to power signals, we could also measure power delivered to a load, currents, harmonics and so on).

The generator power can be adjusted in order to fulfil the requirements, such as P1dB level or maximum allowed harmonic level.

Thanks to solid state switches, the measurement will be very fast, and a quasi real time display is possible.

Using surface mount GaAs switches a 2 - 3GHz test set is feasible in a small PCB, nice companion for the new and low cost VNAs such as MiniVNA Pro, or others. The only large components are the directional couplers (4 single or 2 double [such as the old HP778D] if a very broadband unit is desired. Smaller couplers are available, but they are narrow band or have a limited directivity.

4.

Mechanical switches

Here we have two limitations: low speed, in the order of tens of milliseconds, and a maximum number of operations with guaranteed performance, and the repeatability is fair. Many RF/microwave switches have only one million cycles rating. For this reason, for each position of the switches, a complete frequency sweep is taken. So switches will not wear out, but of course a real time display will be impossible. These switches may require complex driver circuits, as when dealing with single coil latch relays (positive pulse for closing and negative for opening the contact, or vice-versa)

5.

A clever solution

By adding some complication (relays K9 and K7) to the circuit shown in Fig. 3 we can have quasi real time display for S_{11} and S_{21}.

This is obtained through wiring A1 to the reference input by means of extra switches. Another useful option is to mount a transfer switch with built in terminations in order to maintain the possibility of having port 2 active. Other accessories are bias networks and DC blocks.

6.

High power

The high power test set, typically used to characterise active devices and power amplifiers, has extra requirements, inherent to power levels which otherwise

would destroy the VNA. All the components in PORT 1 and PORT 2 branches must withstand the power levels produced by the amplifier; in addition some attenuation (maybe a programmable attenuator) is needed in the path to TEST and REFERENCE

7.

Software

After one or four frequency sweeps, depending on the switches available, four arrays of complex data (Real/Imaginary or Modulus/Phase components) will be available: A1/ref, A2/ref, B1/ref, B2/ref; let us call them c1, c2, c3, c4 for simplicity. The raw S parameters will be then:

$S_{11}=c3/c1$, $S_{22}=c4/c2$, $S_{21}=c4/c1$, $S_{12}=c3/c2$

At this point a full calibration is needed to compensate for the imperfections in the system: it is unlikely that with many switches, loads, cables, couplers (unless very expensive and broadband) the response will be immune from ripple.

So the classic calibration techniques are needed: characterisation of known loads and through cables in order to compute correction coefficients. These techniques are discussed in many papers, and sometimes the software routines are available for download.

The application software (written in any traditional language or Labview) acquires the raw parameters, applies the correction coefficients, and plots the data.

8.

A single board solution

Just as a suggestion (The part numbers

"suggested" are MiniCircuits because the data sheets were easily available in PDF; however any other component with similar or better characteristics can be used) the following surface mount components could be used to produce a microstrip single board test set usable from a few MHz up to about 1.5GHz:

- Power splitter: a 6dB 0 degree resistive power divider made of 3 x 50Ω (Use 2 x 50Ω resistors in parallel) 0805 SMD resistors.
- Loads: 2x 100Ω 0805 SMD resistors in parallel
- Directional couplers: 4 x ADC-10-04 with 10dB coupling and 40dB directivity from 5 to 200MHz (The coupler is specified 0.005 to 1GHz, but calibration will artificially extend its frequency range).
- SPDT switch: M3SWA-2-20 that has built-in 50Ω resistors automatically connected to the inactive port; the control voltage is TTL.
- SP3T/SP4T switch: GSWA-4-30 SP4T with TTL driver and built-in 50Ω resistors automatically connected to the inactive ports

Fig 4 shows the circuit diagram for the proposed test set. As the reader can see, the circuit diagram is not complete. Most power lines and ground connections are not indicated for clarity. Please feel free to contact the author for any questions email: petrini.p@tiscali.it.

- The test set is powered at plus and minus 5V
- RFIN receives RF power from a synthesiser
- REF is the reference power to the network analyser
- TEST is the measuring port of the network analyser
- PORT 1 and PORT 2 are the test ports for the D.U.T. (device under test)
- Six TTL control lines are also required.

9.

A single board companion for MiniVNA Pro

MiniVNA Pro users can build a subset of the schematic shown in Fig 4, because it measures the signal loss or gain between the port named 'DUT' and 'DET'; DUT combines the function of RFIN and REFERENCE, while DET is the receiver input. The revised circuit diagram is shown in Fig 5. Two switches, one coupler and one power splitter are eliminated, but at the expense of not being able to do the traditional direct measurement of S_{11} and S_{21}. Five TTL control lines are also required. Of course a good part of the MiniVNA Pro control software has to be rewritten, with the added functions and control lines interface (Maybe with a firmware modification 5 TTL control lines can be found). With clever programming quasi real time display will be possible. In addition the classic 12 term calibration must be implemented.

10.

Electrical and mechanical hints.

In order to have repeatable measurements a few rules should be followed:

- Stiff mechanical construction
- High quality connections
- For solid state switches the control voltages and currents shall be as stable as possible.
- Stabilised power supply
- Good ventilation for low temperature variations, that is low phase drift in cables and constant performance in switches
- If the test set is built in microstrip technique with surface mount
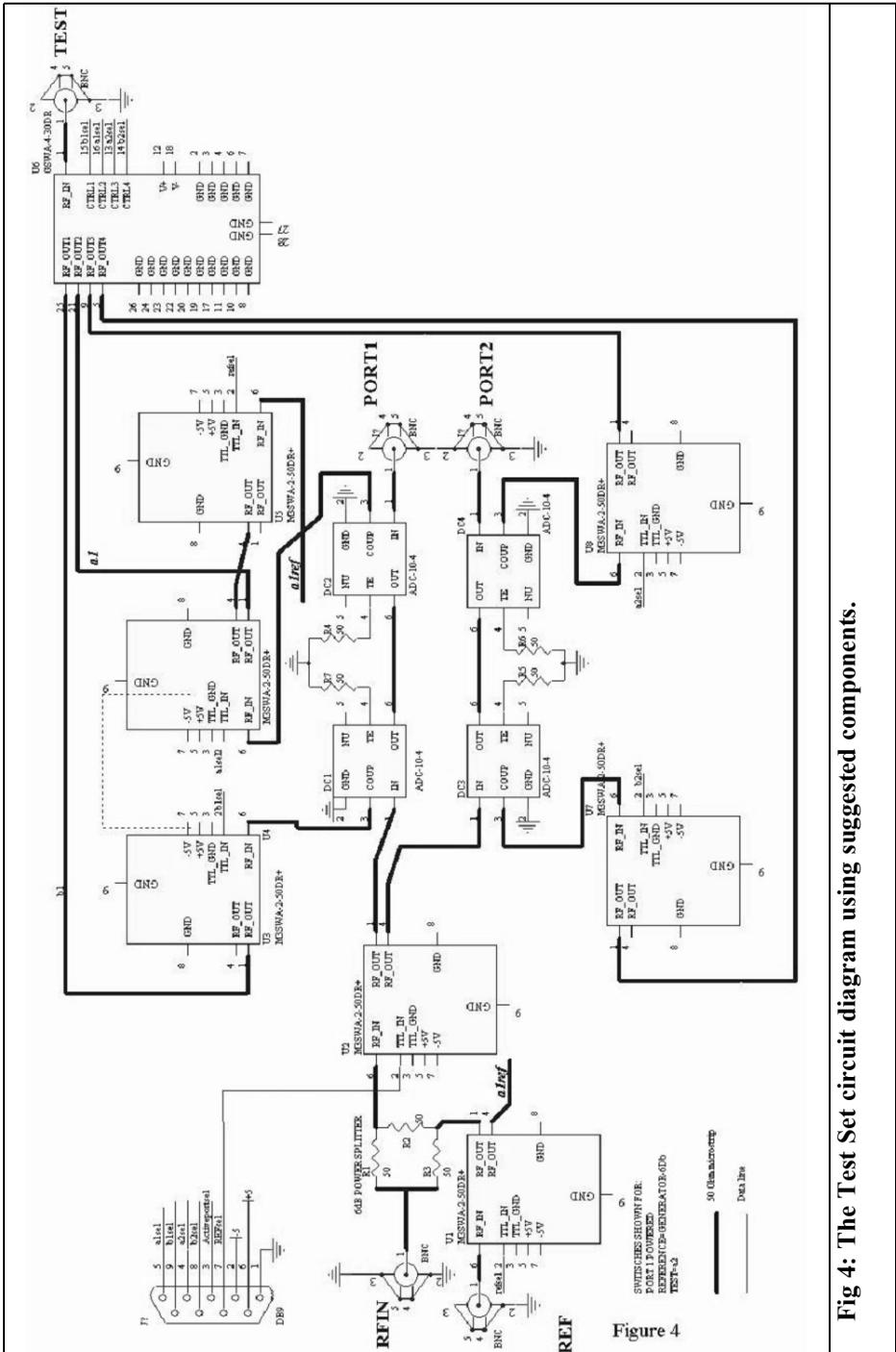
Fig 4: The Test Set circuit diagram using suggested components.

Figure 4
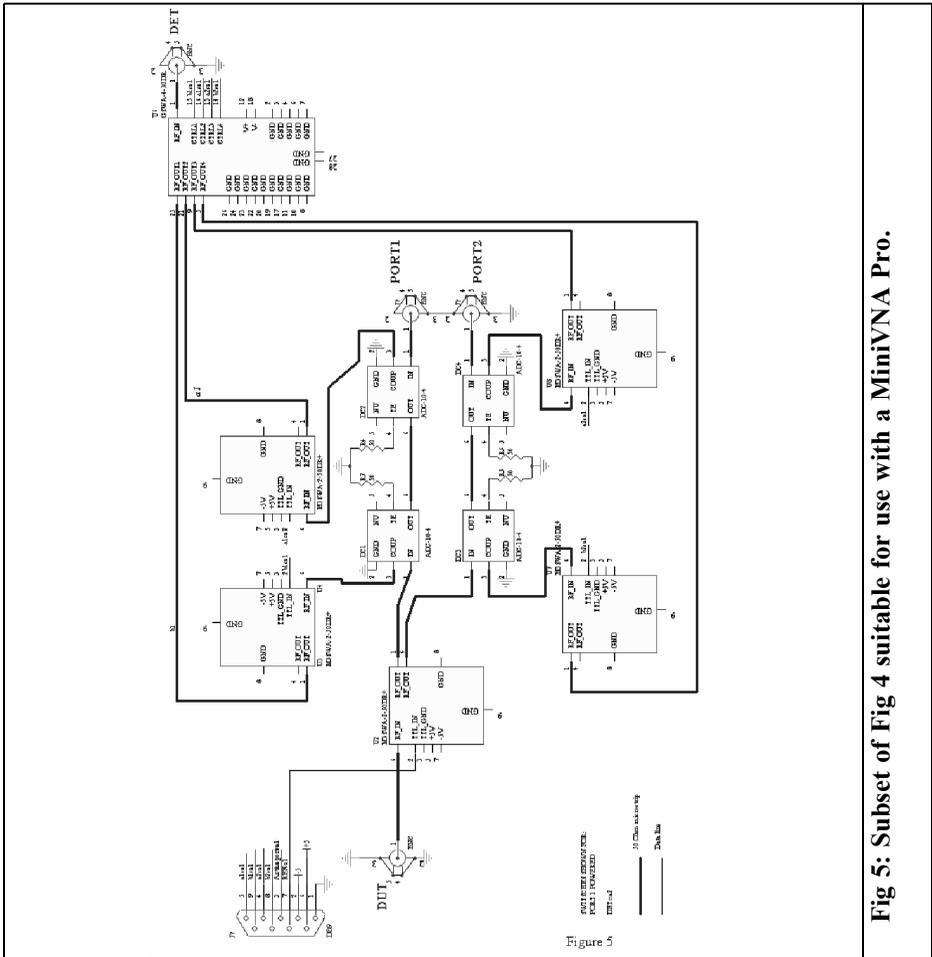
Fig 5: Subset of Fig 4 suitable for use with a MiniVNA Pro.

Figure 5

switches, the mounted PCB should be protected from contamination, and the containing box must be free from resonance at high frequency (cavity modes)

- Use of coplanar 50Ω transmission lines for better screening

11.

Extra possibilities

This setup enables a number of interest-

ing possibilities:

- Evaluation of P1dB and associated gain Vs frequency

- Load pull measurement in association with a computer or manually controlled tuner which changes the impedance seen by the active device under test

- Harmonic tuning in association with a computer controlled tuner and bandpass filter

Gunthard Kraus, DG8GB

Internet Treasure Trove

Mixer Basics Primer - A Tutorial for RF and Microwave Mixers

A very nice introduction to this topic, the rest of the home page and the other documents from this company are very interesting e.g. mixer up to 65 GHz. Another application note that is equally important is: T3 mixer primer - A tutorial on high linearity, high IP3 RF & microwave mixers

Address:
http://www.markimicrowave.com/assets/appnotes/mixer_basics_primer.PDF

Wireless solutions - design resources

A great service from the distributor Farnell. There is a huge collection of application notes on electronics and communication technology indexed by company.

Address: http://UK.Farnell.com/JSP/bespoke/bespoke7.jsp?bespokepage=common/en/technology-first/applications/wireless/design-resources.jsp

Antenna circuit design for RFID applications

This application note is from Microchip, a company that is well known for their micro-controllers. They are expanding their product range and therefore such application notes can be found from them. There are more interesting things as well.

Address: http://WW1.microchip.com/downloads/en/AppNotes/00710c.PDF

Links to the best amateur radio pages

The title speaks for itself. The list is in German but it is fairly self explanatory and includes many useful links.

Address: www.mydarc.de/dd0um/links.htm

Directional BiQuad antenna

This web page is interesting not only because of the many pictures showing how to make these antennas, also be-

cause of the many links on this topic and the interesting details about the properties of a "Dual BiQuad".

Address:
http://www.seattlewireless.net/index.cgi/DirectionalBiQuad

Antenna theory

Who wants to learn about the latest in aerial technology or refresh old knowledge. Here you will encounter a typical modern Internet tutorial with a huge range to inform.

Address: http://www.antenna-theory.com/

An integrated 50GHz SiGe sub harmonic mixer/downconverter

It is always interesting to look at the "forefront of technology" and to see what is feasible now.

This brief article provides the necessary background information with the circuit details and measurements.

Address: http://137.110.119.31/papers/2007_1.pdf

Service Manual download

If you have older instruments in the shack it is always worth looking for the appropriate manuals. In addition to the

famous standard place (BAMA) you should sometimes consult page like this one.

Address: http://elektrotanya.com/hp_8590a.PDF/download.html

SPICE models

You often have to search for a long time to find the SPICE models for specific components required for simulation circuits.

So here are a small set of addresses that you can browse one by one.

Addresses:

http://www.TI.com/ADC/docs/midlevel.TSP?ContentId=31690

http://www.ecircuitcenter.com/circuits.htm

http://www.analog.com/en/amplifier-linear-Tools/topic.html

http://www.diodes.com/zetex/?ztx=3.0/3-10

http://test.tebes.at/DownLoad/Y/tools/5Spice/CD_Spice-models/vendor%20List/spice-models-collection/

http://www.Tech-DIY.com/spice.htm

http://www.webtopicture.com/spice/spice-models-library.html

Win €150 worth of components

Franco Rota of RF Elettronica is sponsoring a competition to encourage new articles for VHF Communications Magazine in 2013.

For some time I have been trying to find ways to encourage new articles for the magazine. Franco and I have discussed this and he has come up with the idea of a competition for the best article submitted to me for publication in VHF Communications Magazine in 2013. Franco will sponsor the competition in return for advertising in the magazine. There will be a voucher worth €150 to spend with RF Elettronica for first prize and a €75 prize for second prize.

For those who do not know about RF Elettronica there will be an advertisement on the back page of the magazine in 2013 or visit the web site - http://www.rfmicrowave.it/. Franco has a vast range of electronic components with many that microwave constructors find difficult to get anywhere else.

To take part in the competition submit your article on a subject that you think will be of interest to the readers of VHF Communications Magazine to me by email or post. I prefer the text in Word format and any pictures or diagrams in jpeg format but I can work with most formats if you cannot manage these. If you have never written an article for a magazine before do not worry I will help with the format of an article and work with you to get it right before it is published. All you need to have is the basis of a good interesting article, probably based on something you have made or developed, with a description and some diagrams or pictures.

The readers of the magazine will judge the articles. There will be a voting form on the web site or postal votes to decide the winners in autumn 2013 with the winners announced at the end of 2013 and in the following magazine.

Good luck, 73s Andy

Send articles by email to: andy@vhfcomm.co.uk

Send articles by post to: Andy Barter, K M Publications, 503 Northdown Road, Margate, Kent, CT9 3HD, UK

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Volume No.44   Winter   Edition 2012-Q4

Web: http://www.vhfcomm.co.uk

ISSN 0177-7505

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VHF Communications Magazine has been published since 1969. Up to 2002 it was produced by traditional printing methods. All these back issue have been scanned and converted to pdf files containing images of the A4 sheets that formed the A5 magazine when folded in half. These have been put together on DVD in decade sets.

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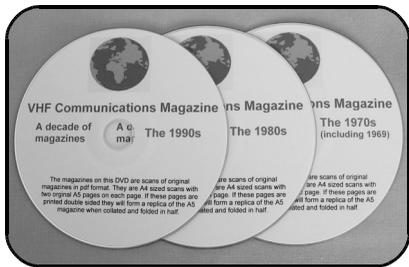
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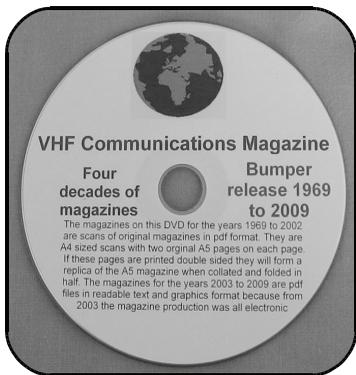
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