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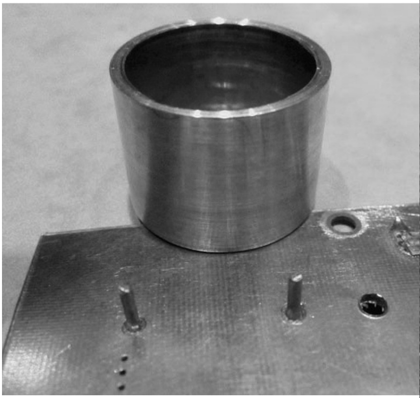

Microwave oscillators using cavity resonators

Carsten Vieland, DJ4GC

Contents

The results of the postal price survey show that a printed magazine is still wanted by the majority, see page 130 for the full details.

One of the overseas agents since the magazine started in 1969, Julio Prieto the Spanish and Portuguese agent, has been unwell and has decided to stand down. Thank you to Julio for his work for the magazine over the years.

There is a correction to the the SMD resistor article in issue 2/2012 on page 154.

73s - Andy

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Results of Postal Price Survey

I put a flyer with each copy of VHF Communications issue 2/2012 and I have received replies from 114 subscribers at the time of writing this article. That is about a 25% response rate so it may not be representative. The results were:

- Continue with a printed magazine with the increased postal prices - 70%
- Change to an electronic delivery only magazine - 28%
- Closed down the magazine - 2%

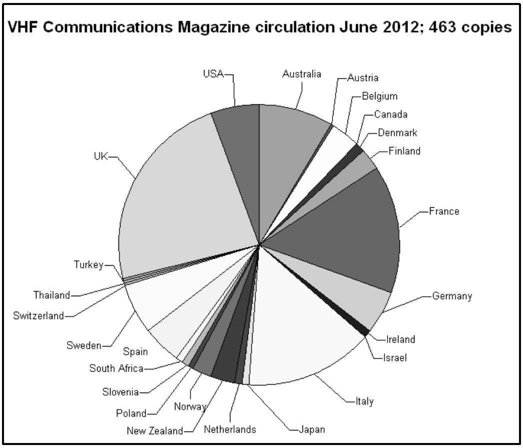
So the majority want the magazine to continue in printed form and say they will pay the increase postal prices.

During the survey several people suggested alternative postal services. If you are reading this article your magazine was delivered via one of those services and the good news is that the postal costs are somewhat cheaper than Royal Mail. If this mailing is a success it means that the prices for next year can be:

- 2013 subscription - £22.80
- Additional price for surface mail - £3.00
- Additional price for airmail to Europe - £5.00
- Additional price for airmail outside Europe - £8.00

That is significantly better than using Royal Mail and I hope it will mean that even more subscribers will be convinced to re-subscribe for 2013.

I have decided to publish details of the magazine circulation so that you can see how difficult it is to maintain this magazine with such a small number of subscribers. The circulation has fallen slowly over the past 10 years from about 1200 copies to its low level of 463 copies by mid 2012. There may be a few more subscribers in the second half of the year but it will not reach 500 copies. This low number means that there is very little money to spend on encouraging new authors to write articles. In turn the lack of articles means the magazine is less attractive to new subscribers. I hope that I will be able to keep the magazine going in printed form with the revised postal prices. If by the time that the first issue for 2013 is ready to be posted I will review the number of subscriptions and if it is critically low I will have to think again.



If you can encourage any of your friends or colleagues to subscribe that will help the situation. If you can write an article for the magazine that will be very much appreciated and there is a small payment for such articles but do not expect to retire on the proceeds.

73s - Andy

Gunthard Kraus, DG8GB

An Interesting Project: A low-loss 10.7MHz bandpass filter with attenuation up to 500MHz

The complete development of this filter is described including the design strategy and simulation. The problems associated with component selection and interpreting the measurement results of the finished product are addressed.

1.

Introduction and specifications

The 10.7MHz frequency is mostly used in FM receivers as an intermediate frequency. There are many finished components to buy (ceramic filters, Crystal filters, etc.) and the use of modern techniques such as digital signal processing is very great. Development and experiments in this field require the correct sources and matching filters. It is often necessary to heavily attenuate unwanted signals far outside the 10.7MHz band.

At the beginning of a new development, there are always a number of considerations. This leads to a list of the necessary requirements for the concept and finally the specifications. And so the specifications for this project are:

System impedance:	Z = 50Ω
Filter type:	Chebyshev with a maximum ripple of 0.3dB
Centre frequency:	10.7MHz
Ripple bandwidth:	500kHz
Pass band loss:	not more than 6dB
Slope:	Stopband attenuation, at 9 and 12MHz at least 70dB
Wideband attenuation:	Trying for 70dB up to 500MHz

The filter should be housed in a machined aluminium case with screwed on shielding cover. The associated board should be 30 x 50mm or alternatively 30mm x 130 mm. Four screws (M2.5), 3mm from the edge of the board fix it into the housing.

SMA connectors are used for the input and output with their inner conductor directly soldered to the appropriate microstrip lines on the board. The familiar Rogers "RO4003" material is used for the board material with a thickness of 32mil = 0.813mm and double-sided copper laminate (thickness = 35µm). The copper base forms the continuous ground plane, necessary vias from the top to the ground plane are made using silver plated hollow rivets with a diameter of 0.8mm.

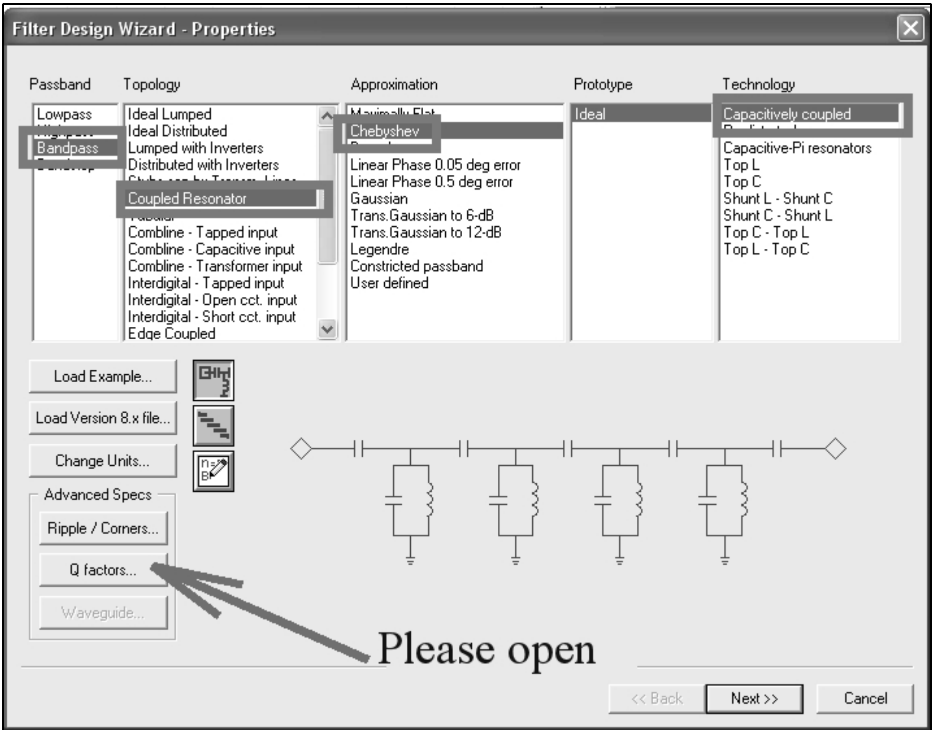


Fig 1: You will be welcomed by the filter designer and then select the filter required.

2.

Development strategy

The following questions need to be clarified first:

- What type of filters meets these requirements?
- What filter level is required?
- What coil quality is at least necessary for the required maximum pass band loss?
- What coil design meets this requirement and fits in the housing at the same time?
- What efforts are required to achieve the high and broadband stop band attenuation?

From previous projects produced it was known that the "coupled resonator bandpass filter type" (narrow bandpass) suits this task well. In this filter all inductors have same value and matching to the source and load impedance is done using capacitive transformation. Therefore, a first simulation assesses what filter level must be selected to achieve the required bandwidth and stop band attenuation. At the same time the Q of the coils needed to give the required maximum ripple in the passband was checked. The realisation of coils of this quality is covered in the next chapter. Finally a prototype was made, first a low filtering level was chosen (for example, filter with n = 3 only requires three coils - less work in design and construction) because it will give the best reproducibility and reliability. Only when all these problems are

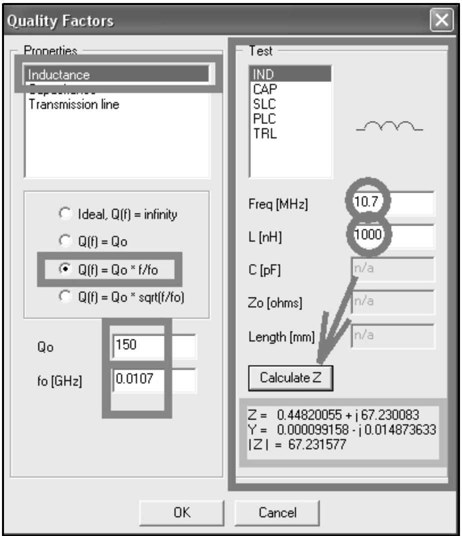


Fig 2: Do not forget to input the coil quality. In addition the other components values can be specified.

resolved satisfactorily can the full version and the optimisation of the stop band attenuation be addressed.

3. Filter circuit and coil quality

The indispensable development tool is the free RF CAD software "Ansoft Designer SV" that contains an excellent filter calculator. As mentioned in earlier publications, Ansoft has unfortunately taken this Designer SV off of the Internet. But the author has been permitted to keep a download [2].

When the program is started there is an option under "Project" for "insert filter design". The corresponding menu entries (Bandpass / Coupled resonator / Chebyshev / Ideal / Capacitively coupled) as shown in Fig 1 control the performance. The diagram produced shows that you have done done everything correctly and you can click on "Q factors" to set the quality of the coils. The left half of the menu shown in Fig 2 ensures the correct entries (Inductance/Q(f) = Q_o * f/f_o / Q_o = 150/f_o [GHz] = 0.01) and Q_o uses a quality = 150 at 10MHz that varies linearly with frequency. In the right half of the menu you will find a useful option that can be used to test the reactance and

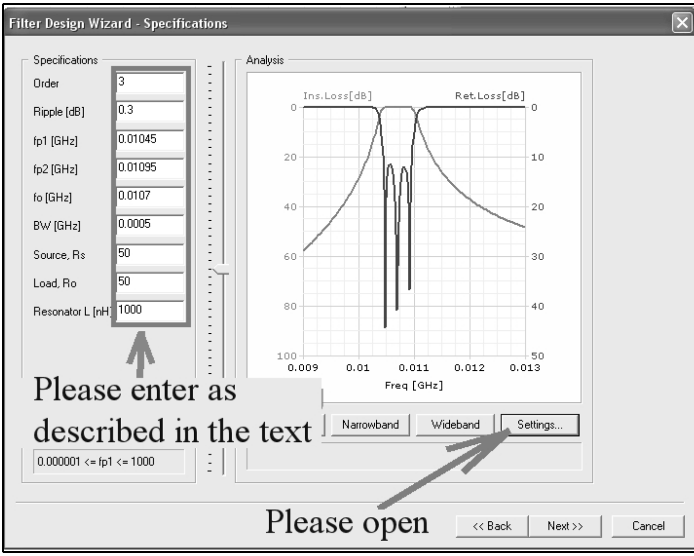


Fig 3: Enter the filter details very carefully (see text for the order) and open "settings".

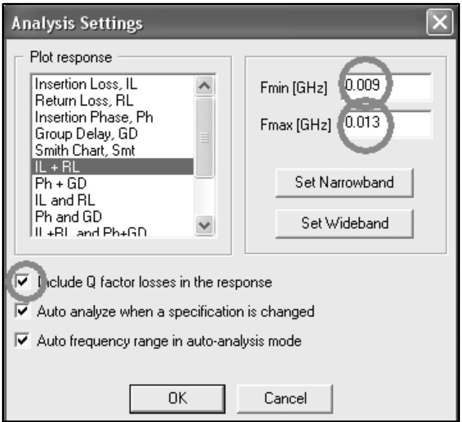


Fig 4: The frequency range is specified and the coil quality included.

loss resistance as a series connection or parallel connection! A 1000nH inductor should have a Q = 150 at 10MHz.

After clicking on "Next" to get the actual settings menu of the filter (Fig 3), note the order of input. The two first entries (Filter order = 3 and maximum ripple = 0.3dB) are harmless. Then the frequency must be specified first with f_o = 10.7MHz = 0.0107GHz followed by the desired bandwidth BW = 500kHz = 0.0005GHz. Then click the value box for the lower or the upper cutoff frequency f_{p1} or f_{p2} ranges so that they are the correct values.

Both the source and the load resistance are 50Ω and inductors used are 1µH = 1000nH. You can already see the filter curve but the frequency axis needs to be improved. Click on "Settings" and enter a sensible frequency range of 9 to 13MHz in the popup menu (Fig 4), this gives a better fit for the axis. To take the quality of the coils into account tick the check box as shown and make the calculations with "OK", the result can be seen in Fig 5. This shows not only all the component values but also the values of S11 and S22

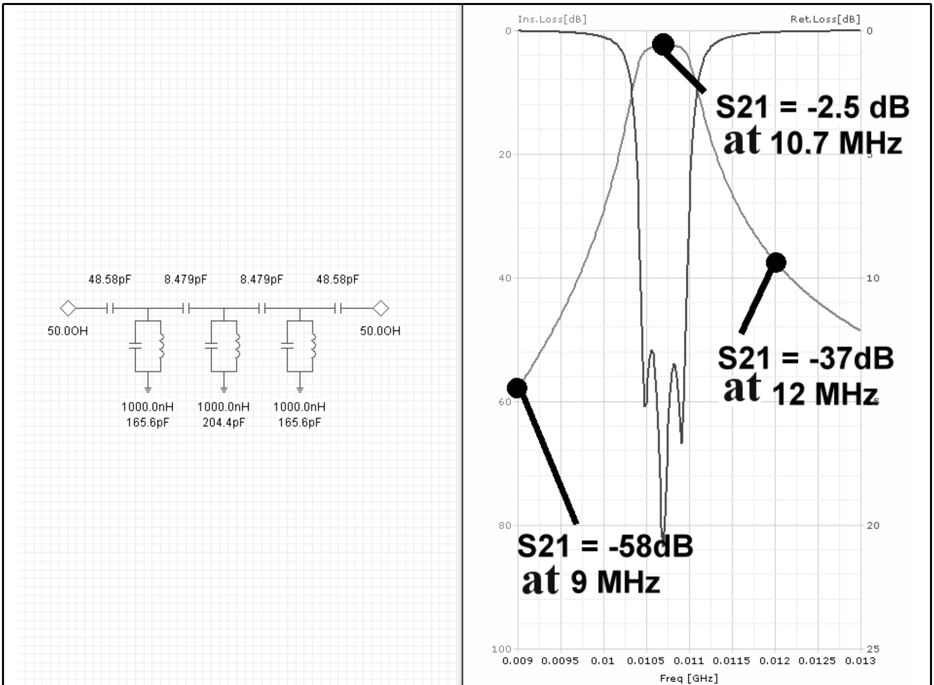


Fig 5: This display provides all important filter properties and shows that the filter level must be increased.

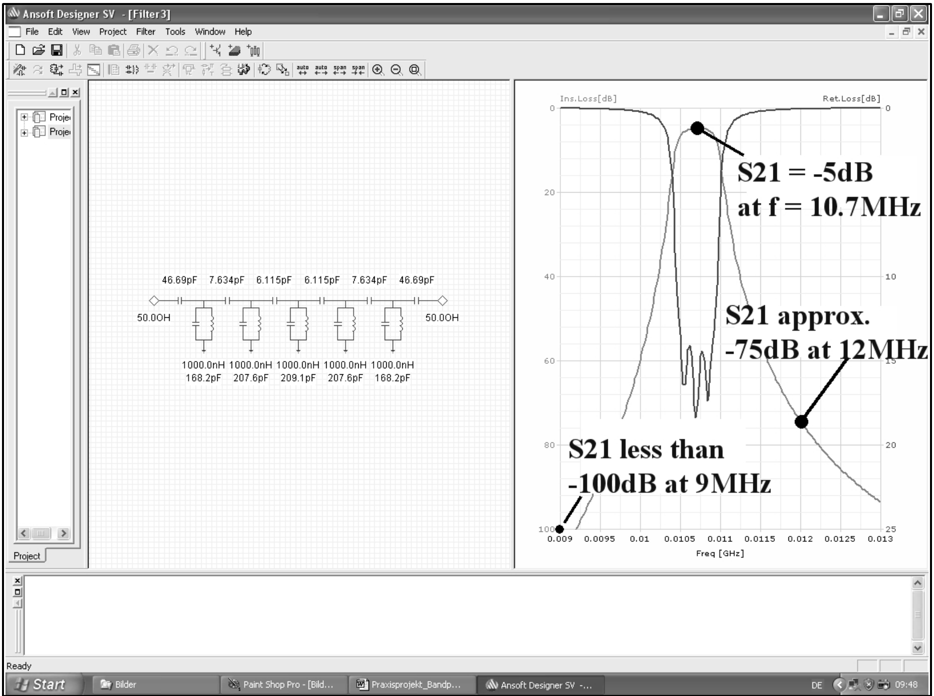


Fig 6: The results for the filter with n = 5 and it fits the requirements.

in dB are shown graphically. It is good to know that with a filter level n = 3 the attenuation at 9 and 12MHz is 58dB and 38dB respectively and not the required attenuation of 70dB. The passband loss at 10.7MHz is, however, a pleasing 2.5dB.

A new design process will start immediately, this time with the filter level n = 5.

Because the input procedure is the same for the new filter level the new design offers no problems and it produces the simulation result shown in Fig 6. Now the attenuation at 12MHz is the required 70dB (at 9MHz there is more than 100dB) and the attenuation in the passband reaches just 5dB (6dB would be allowed).

Thus the goal is clearly defined. The only real problem is the required minimum coil quality of Q = 150 because firstly that level is quite high and secondly the coils need to be small so that they can be fitted without mutual interference on the board and in the chosen shielding housing. Unfortunately both requirements contradict each other.

4.

The fight with coils

The answer is in the technique: "no trade shows no knowledge". Identification of high quality coils at high frequencies (here: at 10MHz and 100MHz) would be interesting and requires an expensive piece of measuring equipment rarely available as on the used market. It is good if you have had many years at the electronic flea markets and could not resist some purchase that has paid off

Fig 7: A return to my professional past: The Boonton RX Meter waiting for the first measurements. Never forget what the yellow sticker says (Sticker on the top left of the instrument says "Attention 110V").

now. In the hidden store I have a "Boonton RX meter" from 1960. It is an impedance measuring bridge for the frequency range 10MHz to 300MHz (principle: modified Schering bridge test oscillator with a simple superheterodyne receiver to measure the bridge voltage - everything in valve technology). The historic equipment works after refurbishment as shown in Fig 7.

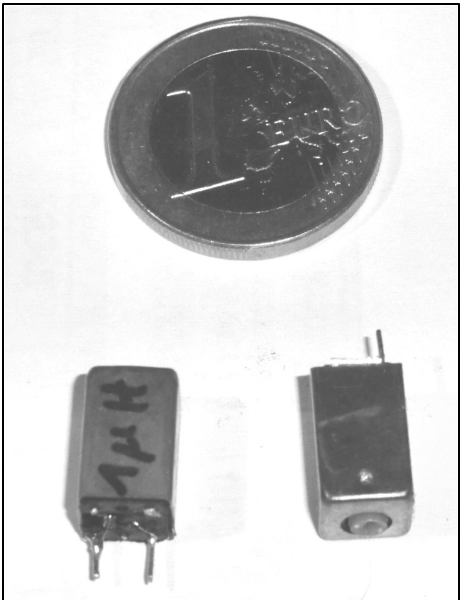


Fig 8: Widely used and proven: the shielded coils from Neosid with a matching core and a size of 7.5 x 7.5mm. Unfortunately a Q no greater than 100 was achieved.

A bridge is a fine and accurate device in spite of some strange operation and the necessary brain power to make and evaluate the results with the calculator. The bridge determines inductance as "negative capacitance" and the adjustable range is limited to 100pF. Therefore to measure L = 1µH the measurement frequency was increased to 16MHz to keep the "negative equivalent capacitance" below 100pF. The yellow sticker on the top of the unit (attention: 110V voltage) requires attention as well as an additional external variable transformer otherwise it is goodbye valves...

First of all the familiar and often used Neosid coils with adjustable core (base size: 7.5 x 7.5mm) were examined (Fig. 8). A series of tests with different core materials, designs and wire gauges always gave the same result: it was not possible to create a coil with a Q of more than 100 for a 10MHz filter.

Fixed core kits as an alternative are significantly larger and relatively expensive. Air core coils are a non starter for a 1µH coil with a high quality of 150 at 10MHz and therefore only the familiar Amidon torroidal cores remained as an alternative. They are readily available and inexpensive and quality information is available from the Internet that gives hope. Also they promise very low cou-

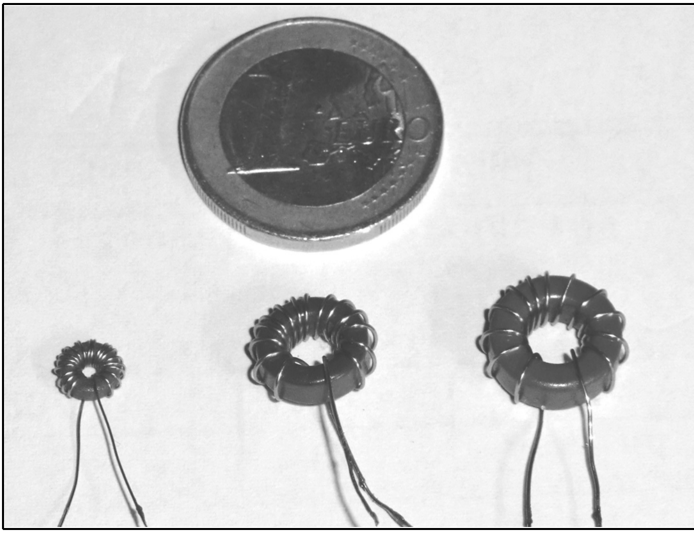


Fig 9: The three Amidon torroids tested.

pling due to the closed field lines that is why additional shielding of individual circuits is not required. It only remains to choose between ferrite and fine iron powder as a magnetic material in the core. Very high permeability ferrite (this means high AL value and low windings!) give very pleasing quality values. There is another problem, a closed core causes early saturation and non-linearity with increasing current leading to signal distortions. Pure iron powder with its microscopically fine iron particles contains enough plastic binder between the single grains to give an air gap with the effect of a much higher saturation. So the eddy currents in the the tiny grains are minimal. Therefore the priority was given to iron powder cores and a small range were ordered for experiments. The types chosen were T20-2, T37-2 and T44-2. The designation system is easy, the first number indicates the diameter in 0.01 inch while the number "2" after the dash says: "Optimum frequency range = 1 to 30MHz". The torroids also have a colour range, red and brown for this frequency range.

These three cores were wound with enamelled copper wire to give an inductance in the order of 1µH with turns distributed around the entire circumference (Fig 9 shows a size comparison with a Euro coin). The inductance values were measured with an old "LARU" from Rohde & Schwarz. The measurements with the Boonton RX-meter went further. As mentioned earlier, 16MHz had to be used as a measurement frequency to stay in the range of the maximum "-100pF". This resulted in the interesting Table 1.

Table 1: Comparison of the different cores with their inductance an Q values.

Core type	Core outside diameter	Turns	Wire diameter	Inductance	Loss resistance	Q at 16MHz
T20-2	5.08	19	CuL 0.2	0.97µH	12.9kΩ	132
T37-2	9.53	15	CuL 0.3	1.02µH	18.8kΩ	183
T20-2	5.08	19	CuL 0.2	0.99µH	15.9kΩ	159

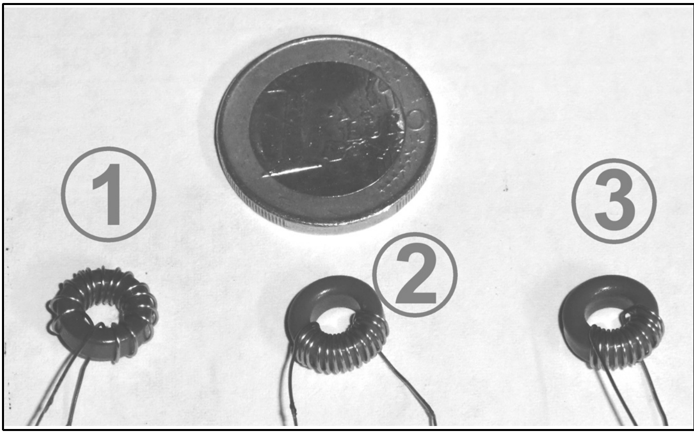


Fig 10: Three test coils with different windings and winding patterns.

The results show that the T37-2 core gives the highest quality of Q = 183 in this design, therefore it was chosen for the filter. To round things off another coil was wound on this type of core (number 2 in Fig 10) with 15 turns but this time the coils close together. This resulted in increased inductance to approximately 1.45µH. A third core (number 3 in Fig 10) was also closely wound but with only 12 turns to get back to 1µH. The quality comparison of the three T37-2 versions measured with the Boonton-RX-meter is interesting and shown in Table 2.

You can see that amazingly the quality of three cases is practically the same. This means that almost the entire magnetic field always runs in the core thus determining the quality with its losses. Only when the windings are slided together (until they touch) the coil quality is decreasing probably due to the mutually induced currents in the adjacent windings. Also the self capacity is increased causing the natural resonance frequency to go down. So the first version with windings spread around the entire perimeter was used for the project.

There is still the problem of the correct method to mount the cores on the board to ensure that:

- Absolute mechanical stability and reproducibility is guaranteed
- The high quality of the coil is not reduced too much

After several experiments and some pondering the following solution was found (Fig 11):

A small disc is cut out from a piece of 1.52mm thick single sided FR4 board (glass fabric + epoxy) using a jigsaw. The 1µH coils are glued onto the non copper side of this disc using UHU-plus

Table 2: Comparison of different types of winding on a T37-2 core.

Number	Winding technique	Turns	Inductance	Loss resistance	Q
1	Turns distributed round the core	15	1.006µH	18.2kΩ	180
2	Turns close wound	15	1.434µH	26kΩ	180
3	Turns close wound	12	1.0015µH	19kΩ	184

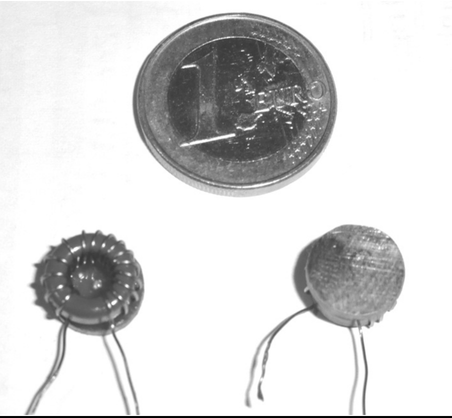


Fig 11: A version of the coil glued onto a piece of PCB material.

(epoxy) glue. The windings are also fixed on the top of the core with glue. (The coil is set to 1µH while gluing by connecting the coil of the inductance meter and moving the windings around the core until the exactly inductance required is reaches). This method produces a component that cannot change its value and is always kept at a distance of 1.5mm from the copper surface. This means that neither the stray capacitance or the quality of the coil can vary during handling and fitting (the tiniest changes of the inductors in the order of parts per million can change the response curve of the narrow bandpass filter). The trick with the copper layer on the bottom means that the coils can be soldered onto the board with the other components and all this does not influence the coil data.

The only anxious question was would the coil quality be significantly reduced by the glue or because it is close to the copper layer at the bottom of the FR4 disc. The inductance and quality of the first three coils made for the filter were measured using the Boonton RX meter. Here the result:

The inductances were as follows:

L = 1.013, 1.01 and 1.0015µH

The Q values were:

Q = 162 , 167 and 164

So the coil problems are regarded to be solved.

5. The first filter prototype with N = 3

Fig 12 shows the first attempt to transfer the circuit provided by the filter design into a practical version. The different

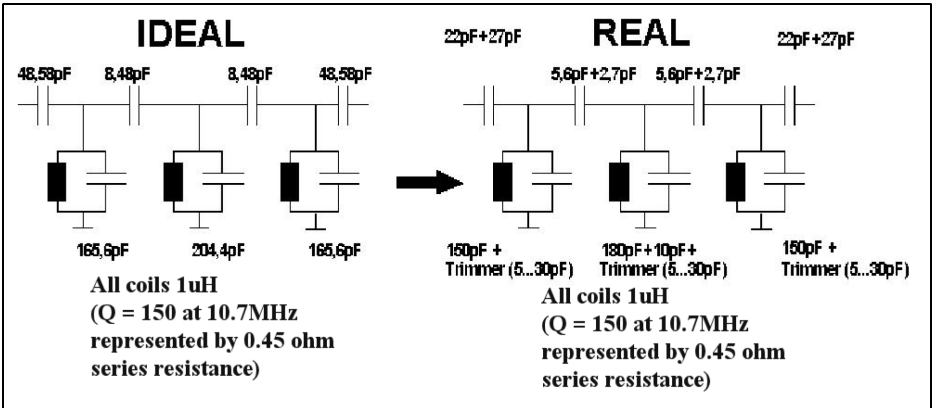


Fig 12: A very important and essential step: The theoretical design is transferred the real circuit.

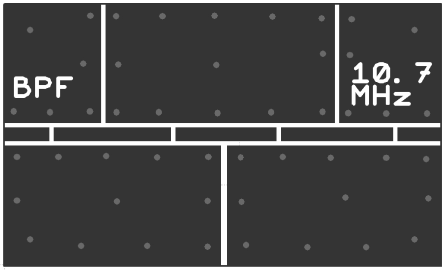


Fig 13: The PCB layout using Rogers RO4003 material 32mil thick. The microstrips line has Z = 50Ω with gaps for the coupling capacitors.

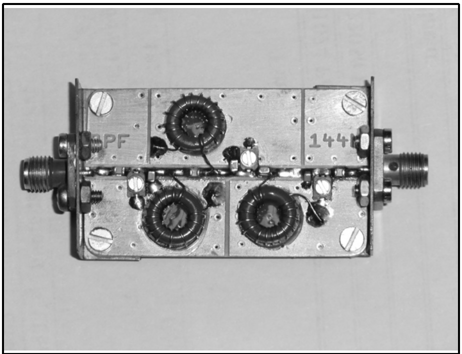


Fig 14: The finished prototype

coupling capacitors have been replaced by standard value SMD components with values as close as possible to the required value (with not more than 3% deviation). The shunt inductors have a trimmer capacitor added across the coils. Thus the required total capacitance value can be convenient set and also the "manufacturing tolerance" of the three torroidal coils can be corrected.

The prototype of the filter circuit was built on a 30 x 50mm board (Fig 13). The different ground islands for the three

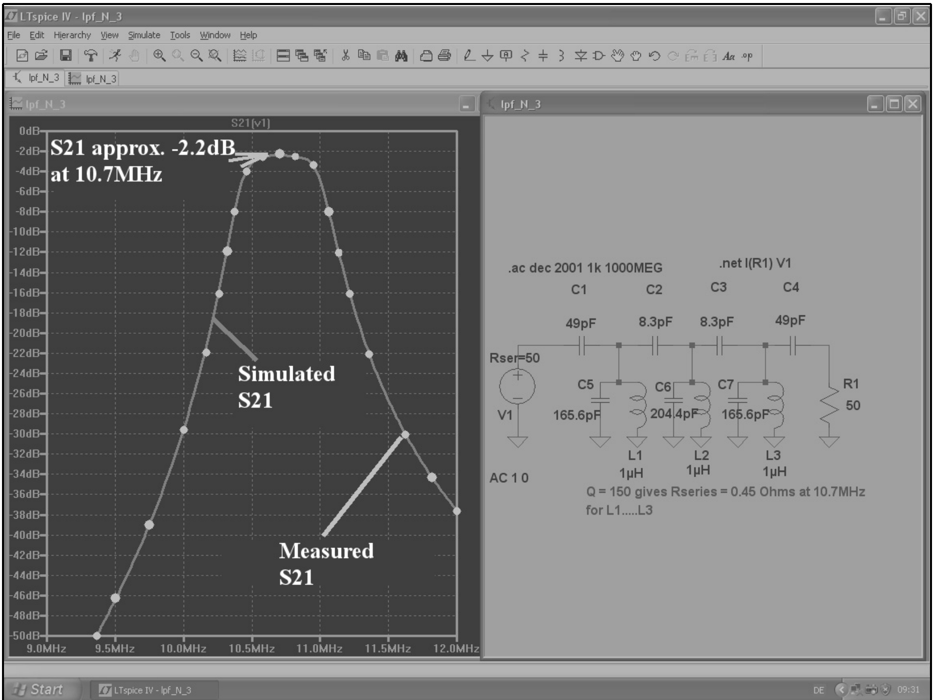


Fig 15: Theory and practice. They are almost identical and so we must say "Thank you" to modern simulation programmes.

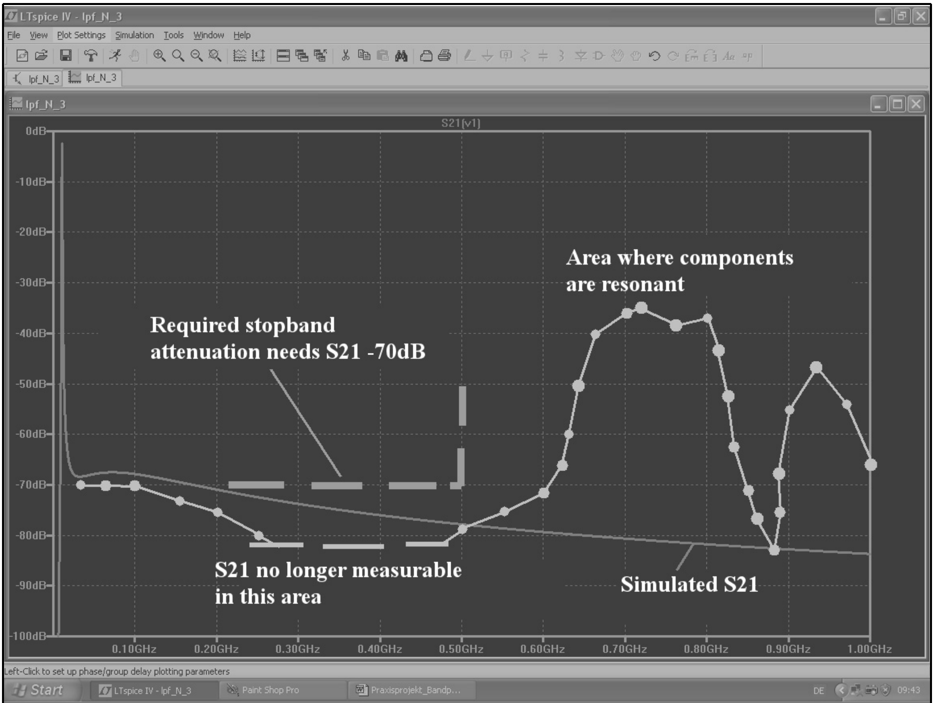


Fig 16: The result is as expected.

circuits can be seen as well as the required vias. The assembled board is shown in Fig 14. One thing we must be highlighted: the rotors of all variable capacitors must be connected to ground otherwise the filter curve will be changed by the adjusting tool touching the trimmers when adjusting the matching. SMD variable capacitors usually have a tiny marker denoting the rotor.

This circuit was examined in an LTspice simulation with the coil quality simulated in the form of series loss resistance of 0.45Ω. The result together with the S21 measurement at 10.7MHz after successful adjustment of the curve to maximum can be seen in Fig 15. The in-band attenuation is about 2.2dB that is confirmed nicely by the simulation. The remaining results between 9 and 12MHz are also in accordance with the simulation.

Now it was interesting to examine the stopband behaviour to see if it can achieve 70dB up to 500MHz. Secondly, it is interesting to see how the circuit behaves up to 1GHz. The answer is shown in Fig 16 and we can be very satisfied. Above 500MHz the natural resonances of the components cause trouble but that was to be expected. These effects are limited and a maximum of about 30dB is reached at 720MHz.

The measurements were made with a good precision signal generator (HP 8640B) as the signal source to power the filter. The signal was measured with a RACAL DANA 9301A 50Ω input impedance true RMS millivoltmeter. With an output level of +10dBm and the lowest measuring range of the millivoltmeters of -50dBm gave a measurement range of 50 + 10dB = 60dB. The -10dB level is clearly marked on the scale of the

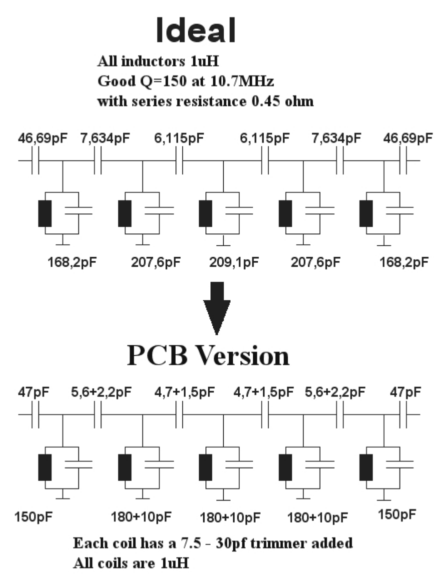


Fig 17:

millivoltmeter so -60dB can be seen and a final measurement range of 10 + 50 + 10 = 70dB is shown in Fig 15. The whole thing was verified with a modern vector analyser (ZVRE from Rohde & Schwarz) in the laboratory of the University of Cooperative Education - this turned out to be unnecessary because the results were completely identical to those of "do-it-yourself" measurements.

The results of measurements over a wide frequency range with the combination of a signal generator and a broadband millivoltmeter should be treated with caution. The signal generator output has harmonics generated in the power amplifier, the strongest being at odd harmonics numbers. As it happened in our example the third harmonic of the frequency f = 240MHz falls in to the maximum shown in Fig 16 at f = 720MHz.

The perfect remedy is of course an additional harmonic filter at the signal generator output or a Spectrum Analyser as a voltmeter - the same as using a Vector Network Analyser. A solution is still possible for wide band measurement: you choose the next higher output voltage range (here +20dBm) and then turn down the output level to +10dBm. By decreasing the control of the final stage by 10dB the amplitudes of the third order harmonics will decrease by -30dB. Thus the results of wide-band measurements again agrees with the measurements from the Vector Network Analyser.

6.

The final objective: Five Pole version

The calculations required were shown in Fig 6 but again the result must be implemented in a practical circuit with SMD capacitors and torroidal coils. The

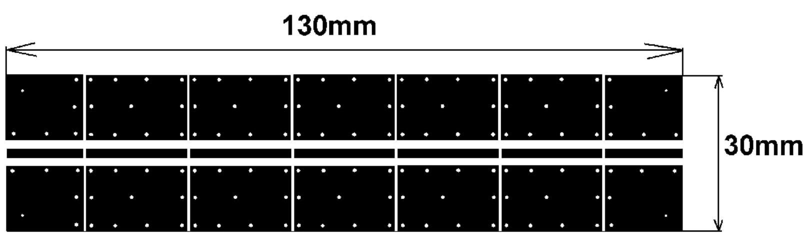


Fig 18: The PCB is now quite long.

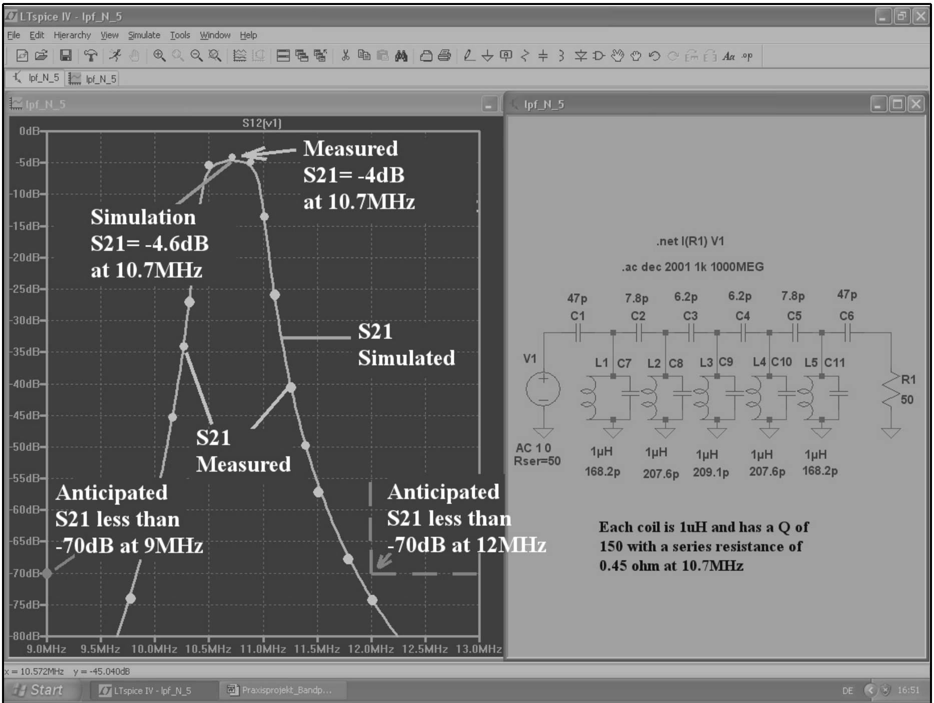


Fig 19: The result is not far from the simulation, a good result after all the work.

result can be seen in Fig 17, Fig 18 shows the circuit board used (material as always: Rogers RO4003 with a thickness of 32MIL = 0.813mm). The concept can be seen with the separate ground islands and many vias with a 50Ω microstrip line running down the centre from the input to the output interrupted by a small air gap for the coupling capacitors. This line gives something of a conflict for the simulation, should these lines be included in the simulation? Below 500MHz they only constitute additional capacitors and that can be balanced in every tuned circuit by the existing trimmer capacitor. If anything, the transformation effects of line is only apparent above of 1GHz. Fig 19 shows a comparison between the S21 simulation and measurements.

The wide frequency range up to 1GHz is of interest, it can be seen in Fig 20 and probably leaves no wish unfulfilled. It was measured with an HP 8410 Network Analyser and this result was confirmed with the Rohde & Schwarz ZVRE analyser.

The finished circuit in the housing is shown in Fig 21 and is an improvement compared with the first three-pole filter version. The torroidal coils were not stable enough when fitted because the smallest changes (such as slight bending of a connection wire when soldering) changed the response curve significantly. So a 2.5mm wide, 0.4mm thick, 15mm long strip of copper plate was soldered to the bottom of the circular FR4 substrate. The torroidal coil is now like an SMD component and can now be repositioned if required (Fig 22). Now it is immovable with absolutely rigid wires on the circuit board. This fact is confirmed by Fig 23 showing one of the five circuits with all associated parts.

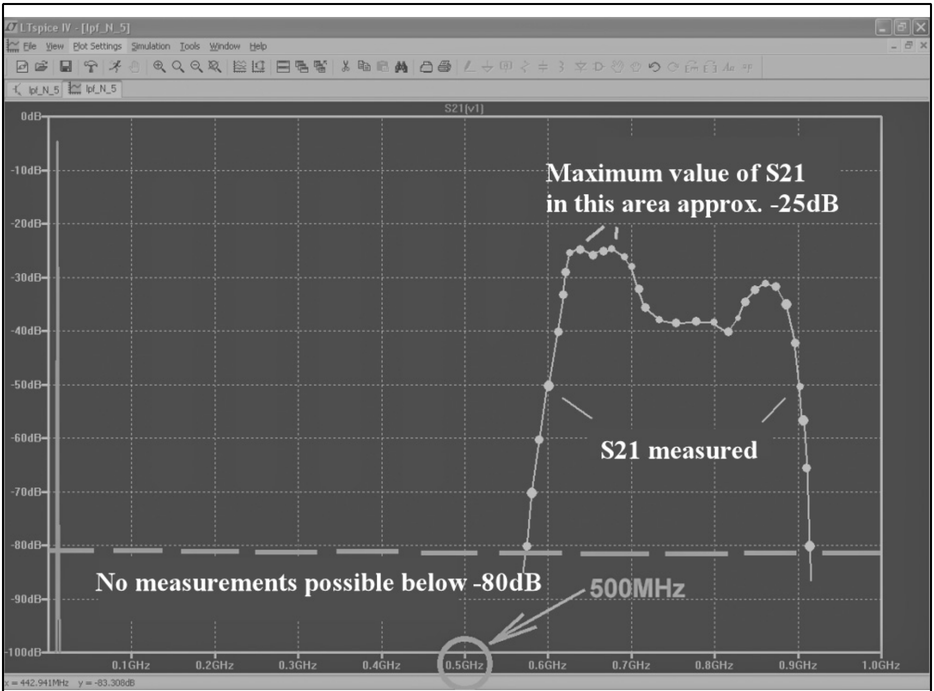


Fig 20: The wide band results fit the requirement.

Fig 19 shows that these experiments with the coil quality have not harmed the S21 characteristic in the passband. The simulation with a Q of 150 showed that a value of S21 = -4.6dB at the centre frequency f_o = 10.7MHz and it was measured as -4dB. That means that the Q is higher than 150 and was measured as Q = 164 from the previous chapter for the "SMD version" of the torroidal coils.

Fig 23 shows the "piggy-back versions" of the coupling capacitors. It is harder than it looks to get this construction technique correct therefore something on the subject of "Soldering of SMD com-

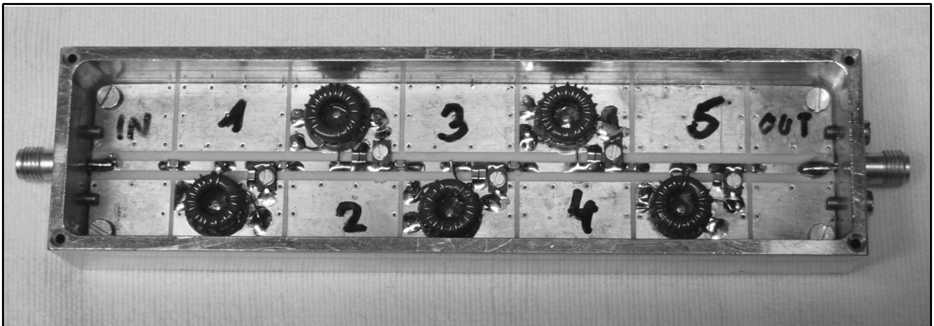


Fig 21: The five pole filter is an impressive sight.

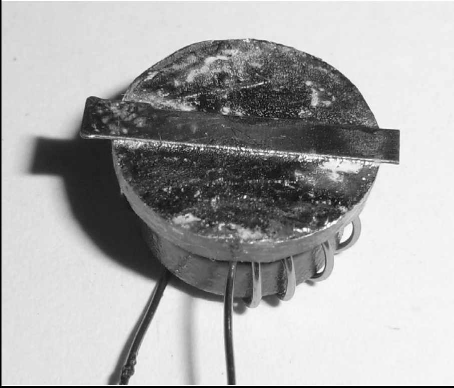


Fig 22: The finished coil is converted into an "SMD component".

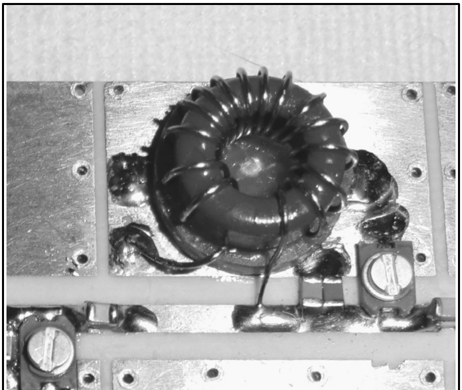


Fig 23: A close-up of the finished component that would withstand a professional vibration test.

ponents" should be noted. Bernd Kaa has already put together all the information in his excellent guide article [1]. There is a problem with the recommended SMD soldering paste. It is not the success or handling - on the contrary it is excellent and the default method in industrial production with convincing results. But, for the small consumers it is quite expensive (my experience) and tends to age so it should be used quickly.

Since it was not always possible to use fresh 0.5mm SMD solder I use cheap solder paste. I have used the same tube for several years without problems soldering SMD components. It came from a company called Bürklin Electronics, please try it some time because you will marvel at how well it works! But I can only confirm every word in Bernd Kaa's article – do work this way with SMD soldering, non-compliance with the recommendations can cause extensive damage.

7.

Summary

Everything has worked out but it was not easy so you have to persevere to get the result. The high Q coils, the small passband losses obtained and the anticipated steep filter edges are fascinating.

Another three-pole version was examined but with the coils reduced from 15 to 14 turns giving an inductance of 850nH and increased in the Q from 183 to nearly 190. The new filter design and implementation on a test board were carried out as described in Chapter 5. The result were a little surprising:

The attenuation was increased by a few dB between 550 and 600MHz and deteriorated above 700MHz from just over 30dB to 25dB. Otherwise the results were all identical to the results in Chapter 5. So you can safely stop here.

You need the appropriate simulation programs, sufficient experience and knowledge, and a good memory for past and errors to undertake such developments.

A creative imagination and patience help (think about the nearly 150 holes and vias fitted with silver plated hollow rivets on the large PCB, each individual rivet must be trimmed with the file to 3mm long). High quality measuring instruments are need for a successful development (beware of old instruments or valve designs!). All that combined with the joy to develop - this gives full success.

8.

Literature

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André Jamet F9HX

The use of electrolytic capacitors in DC power supplies

1. Introduction

DC power supplies are a very important part of most electronic equipment. And yet, they are considered with a certain disdain. The result is that the Achilles' heel of measurements on much sophisticated equipment from very well known brands is located in their DC power circuits. Poorly designed and poorly made, they are the causes of many failures. For installations of high reliability, power supplies are the subject of studies and very careful manufacture (aviation, nuclear, medical). The choice and mounting of components, such as electrolytic capacitors, are important points of design and implementation.

Today, the Solid State Power Amplifier (SSPA) can deliver power up to a kilowatt. They require linear or switched-mode power supplies with electrolytic capacitors. The migration from valves to transistors in power circuits was not the object of updates to the rules on the use of electrolytic capacitors. However, the conditions are sufficiently different to warrant a review. Indeed, the voltages that can reach thousands of volts had required safety precautions such as discharge resistors, and features such as series capacitors with splitting resistors. On the contrary, the use of low voltage with high currents requires different implementations.

This study is intended to advise the designers of 230V AC input power supplies with DC output voltages of 50V that can supply currents up to several tens of amperes.

It also concerns boost and buck converters whose purpose is to generate lower or higher voltages for large currents from a 12V battery. Two examples of these: A 13.8V 10A booster to get this voltage even when the battery voltage drops to 10V [1] and a good efficiency buck converter delivering 9V at 20A from a 12V battery [2].

2. Capacitors

It should be noted that the capacity of a capacitor is given by the formula:

$$C = \varepsilon_r \, . \, \varepsilon_0 . \, S \, / \, e$$

where:

C = capacity (F)

ε_0 = permittivity of vacuum $(1/36.10)^9$

ε_r = permittivity of the dielectric (dielectric constant, specific inducing power)

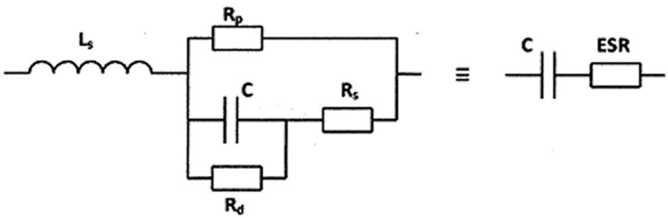

Fig 1: Equivalent circuit of a capacitor.

S = surface of electrodes in contact (m)^2

e = distance between the electrodes (m)

The permittivity can be very high for some ceramics that can reach values up to 10µF at a few volts working voltage [3].

Beyond that, one must use electrolytic capacitors that use the principle of obtaining very high capacity using an extremely small distance between electrodes.

Another family, recently introduced, is the super capacitors. These can realise values reaching hundreds of farads. The voltage is limited to approximately 3V requiring series connection that can be very disadvantageous. Their use is very promising for energy storage and the possibility to supply very useful currents during their discharge [4].

3.

Electrolytic capacitors

Electrolytic capacitors are constructed from two metallic foils, one is coated with an insulating oxide layer, and a paper spacer soaked in liquid or gel electrolyte. A formation process produces the oxide layer.

The oxide ε_r is not very high; it is the thinness of the dielectric layer that ensures very high capacity, the impregnated paper only playing the role of a conductor during the formation process. The oxide must survive during the life of the

Fig 2: Examples of electrolytic capacitors.

capacitor to keep the initial capacity.

These capacitors are known as polarised because they are sensitive to the polarity of the voltage applied to them. This voltage must not exceed a value limit defined by the nature and the thickness of the oxide.

Two families are used routinely; one uses metal aluminium electrodes, the other tantalum.

3.1 Tantalum capacitors

Tantalum capacitors are confined to values not exceeding tens of microfarads. Their use is usually reserved for by-pass, inter-stages and timers applications. They are therefore not discussed further in this article, but they require a very good knowledge of their characteristics to use them and avoid the setbacks often encountered.

3.2 The aluminium electrolytic capacitor

These are the ones that are used to filter the rectified DC in linear power supplies. In switched mode power supplies they are used for filtering but also as switched reservoirs (Fig 2).

It is necessary to know the various parameters of such capacitors. Fig 1 shows the equivalent circuit of a capacitor with parasitic elements. These can be defined as:

- Ideal capacitor C
- Equivalent series resistance R_s (resistance of the connections and the gel or liquid)
- Equivalent parallel resistance (leakage of the dielectric and the package) R_p
- Loss due to dielectric (tg δ of the material) R_d
- Equivalent series inductance (inductance of connections) L_s

The parasitic inductance does not have an effect given the frequencies involved, except for switching circuits reaching the megahertz.

For applications relating to this article, it is customary to define the equivalent resistance ESR that represents the set of losses.

The ESR is very influential. It reduces the efficiency of a ripple voltage filter. In a switched mode power supply, it results in losses that cause a decrease in the conversion efficiency.

The power lost in the capacitor is:

$$P = (ESR).\ I_{rms}^2$$

where:

I_{rms}^2 = RMS current drawn by the capacitor

It causes capacitor heating. The permissible maximum current is given in the data published by the manufacturers.

3.3 Choosing an electrolytic capacitor.

Catalogues from suppliers and especially those from manufacturers have all the characteristics of the proposed products. The design of a DC power supply gives the specifications of the required capacitors.

- Voltage of the DC supply (V_{DC})
- Maximum voltage (V_{max})
- Capacity (F)
- Current: frequency (f) and value (A_{eff} or A_{peak})
- ESR (mΩ)
- Ambient temperature (°C)

From these values, you must choose the most appropriate capacitor model in accordance with the following rules:

- It is not necessary to overrate the voltage of an electrolytic capacitor; a margin of 20% to the maximum predicted voltage is sufficient. Loss of capacity during their life is less because the oxide formation is still a maximum.
- The capacity tolerance can be -20 to +100%. Moreover a margin must be made because the capacity may de-

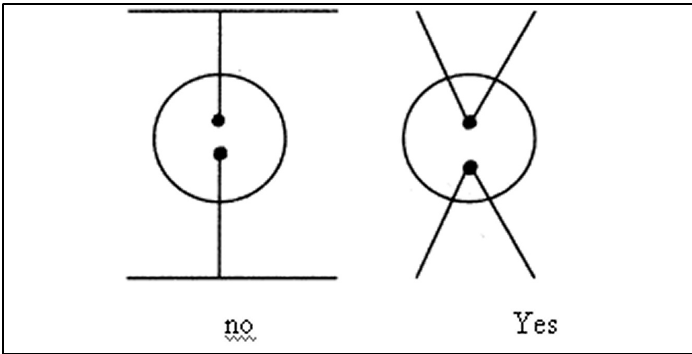

Fig 3: Path of tracks on a printed circuit board avoiding ESR increase.

crease over time. Also consider the value standards 1 2.2 4.7 6.3 and their multiples.

- The current parameter must be as large as possible, given acceptable capacitor size.
- The ESR must be less than the value required; choose low or very low ESR models
- A low temperature decreases the capacity and increases ESR.
- A high temperature leads to a progressive decrease in capacity and causes failures. Life follows more less the Arrhenius's law: reduced by half for 10°C increase in temperature. Avoid placing an electrolytic capacitor near a heat source, and above all, not to the top of a heat source.

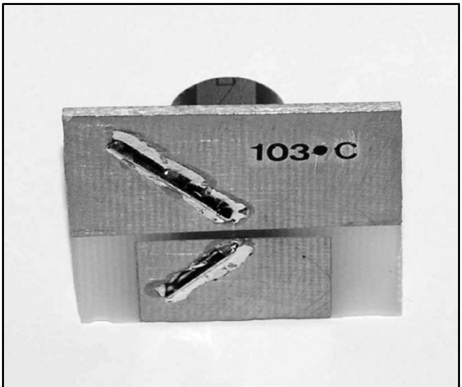

Fig 4: PCB soldering.

- The life is superior for liquid electrolyte capacitors.

The type of package depends on the mounting to be used for the capacitor. For PCBs: SMD, axial or single ended wired or snap-in types can be used up to a few tens of amperes. Screws terminals types can be used for wiring by cable up to tens of amperes. Bus bar types can be used up to hundreds of amperes.

3.4 Mounting and connecting an electrolytic capacitor on a printed circuit board

It is not sufficient just to select a capacitor it must be mounted and connected correctly. Its mechanical attachment must be secure to prevent any stress on its connections. Wire capacitors must be fitted against the PCB and glued before or after soldering. This makes them difficult to dismantle but reliability is ensured. Electrically, avoid adding any resistance in series so that the ESR is not increased as shown in Fig. 3.

The PCB should be FR4 1.6 or 2.4 mm preferably with a 70 or 140µm copper cladding. Fig 4 shows a trick that ensures the minimum increase in ESR for "stand up" fixed capacitors; the wires are not cut flush with the PCB but folded against it prior to soldering. Again this will make it more difficult to disassemble but this is not a priority.

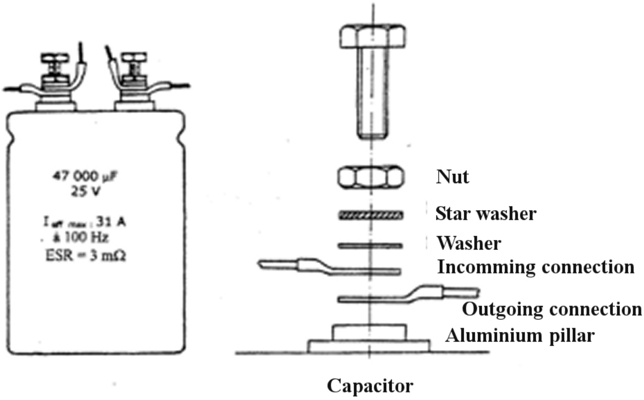

Fig 5: Connection of a screw terminal capacitor.

3.5 Mounting and connecting electrolytic capacitors in high power circuits

Capacitors with screw connection on flanges. Links between capacitors, if several are in parallel, and to associated circuits are carried out using stranded cable with fast-on connectors (well made!). Strips of copper (10 mm x 1 mm) or bars of copper or aluminium can also be used.

They should be connected as shown in Fig 5. This method has been proved up to thousands of amps [6]. It allows the maximum number of connections to be made to the aluminium pillar without having to choose a specific length of the screw. A threaded stud can replace the screw but this is less common.

To avoid two common problems when several capacitors are in parallel:

- Do not bring them too closer because this reduces the effective area for cooling.
- Do not cause unequal AC currents through them.

Fig 6 shows a "do - or do not" for parallel capacitors so that their currents are equal. Instead of a high value capacitor it may be interesting to parallel several units of lower capacity. Like most things when capacitors dissipate heat the dissipation increases as their surface (square of average size) while their losses grow as

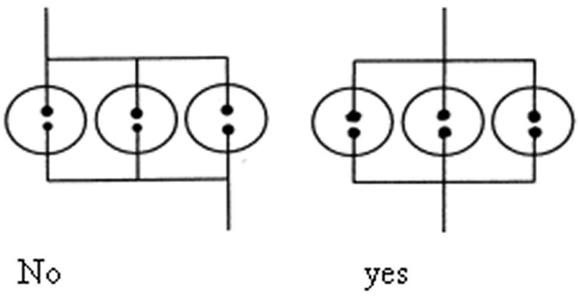

Fig 6: Parallelling capacitors.

Fig 7: Transformers.

their volume (cube of average size). In addition the specified RMS current and ESR are not a direct function of capacity. To find the optimal solution, read the data from the chosen manufacturer.

3.6 Inrush current

In a 230V AC powered linear power supply a transformer supplies a voltage that is rectified by a diode bridge. Filtering is often done by an electrolytic capacitor of sufficient value to obtain the desired ripple. The RMS current through the capacitor is calculated and leads to a good choice. There is another parameter that can affect reliability; the inrush current at power on. The capacitor is discharged and behaves like a short circuit. The current supplied by the transformer and diodes can reach a very high value. It is the internal impedance, resistance and inductance of the transformer that determines the inrush current. This is approximately:

$$I_{inrush} = 100 / U_{cc\ \%}$$

where:

$U_{cc\ \%}$ = short circuit transformer voltage voltage drop in % at rated power

If this voltage is 10%, the inrush current will be about 10 times the rated value. The short-circuit voltage of a transformer depends on its design. Two very different types are used:

- A primary and secondary winding on a laminated magnetic core. Often, they are wound side by side and not superimposed (Fig 7). This gives very good isolation between the 230V and the secondary winding. In addition it increases the

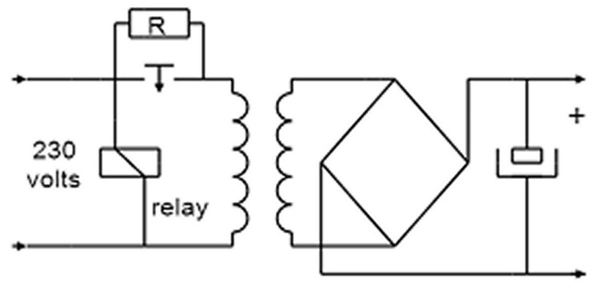

Fig 8: Inrush current limiting with a resistor.

primary/secondary leakage so increasing the transformer voltage drop and decreasing the short circuit current.

- Two windings either side-by-side or superimposed on a torroidal core (Fig 7). In both cases, the primary/secondary coupling is tighter than for a laminated core transformer so the leakage is lower and the short circuit current is higher.

When power up reaches hundreds of watts it is essential to ensure that the inrush current will not damage the diodes and the filtering capacitor. This can be achieved using a large distance between the primary and secondary of a transformer. Most often a temporary series resistor switched with a relay (or triac) provides a two-step power up of the transformer. The resistor limits the current supplied by the transformer allowing the capacitor to charge progressively (Fig 8).

Another solution is to insert permanent or temporary Negative Temperature Resistor (NTC). This method has a problem if the on/off/on sequence is too brief because the NTC does not have time to cool and return to its initial value.

Limiting circuits have another advantage; they limit the transformer inrush current itself. Indeed, the magnetising current can be very high when the transformer is designed to a very high specification (for example, up to 50 times the rated value, for a grain-oriented steel core at 1.5 Tesla). Without limiting the inrush current can trip a circuit breaker feeding the power supply.

Likewise for a Switched Mode Power Supply (SMPS) connected to the 230V mains the inrush current must be limited to preserve diodes and electrolytic capacitors. It is still possible to insert a resistor or a NTC. A better solution is to combine PFC (Power Factor Correction) and a current limiter [7, 8].

3.7 Storage and commissioning

It is risky and dangerous to turn on an electrolytic capacitor that has been powered off for years. The oxide layer could partially disappear and the capacitor is more or less in short circuit. Powering up by connecting directly to a high current DC voltage source causes a fatal breakdown of the capacitor. The dielectric layer must be restored with a current limited voltage source that gradually increases the voltage as the current decreases. This may seem tedious and some may have switched on and old piece of equipment that has been out of service for a long time without incident. The chance is not always certain, more often Murphy's Law is right!

3.8 The danger of electrolytic capacitors

In applications limited to 50V the danger of electrocution is virtually non-existent. However other risks are present. As already mentioned there is a risk of leakage of electrolyte by a badly used or faulty capacitor. Induced corrosion can cause a deterioration of the corroded components and cause failures.

The high energy stored in capacitors can cause very high short circuit currents during power on or in case of faults. Severe overheating can occur with gassing, explosion and fire. Pay attention to screwdrivers and test probes!

4. Conclusion

No, DC power supplies are not a part without great interest for high and very high frequencies users. Well, they show their importance when they fail!

5. References

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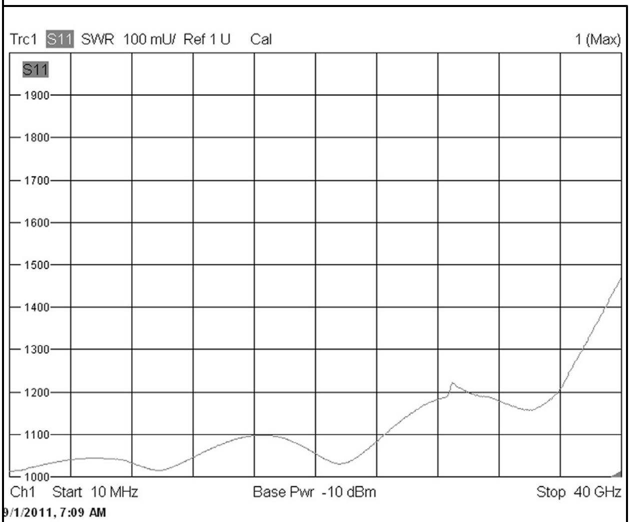
[8] PFC Circuit Halts Inrush Currents, John Bottrill, Texas Instruments, Manchester, N.H., June 1, 2008

Correction to the article by André Jamet in issue 2/2012 on SMD resistors

The left hand pane of Fig 15 was the same as Fig 14. It should have been the plot shown below



Carsten Vieland, DJ4GC

Microwave oscillators using cavity resonators

Revised submission to the 35th Microwave meeting in Dorsten 2012

A signal generator was needed to complete an extensive measurement system for 5.000GHz that should have as accurate a frequency as possible, low drift, low noise and free from spurious signals. Tests with existing modular 50Ω components such as cavity Resonators, amplifier blocks, directional couplers and pieces of line are discussed. If your requirement is not exactly my requirement of the 5GHz oscillator described here, the building instructions and results should motivate your own project. If necessary and depending on the frequency the circuitry can be transferred to PCB construction.

1.

Laboratory construction using coaxial components

In the old days, when the interest in microwave technology was already large, ordinary spectrum analysers were available but at the price of a house. It was common to use a tunable cavity resonator with a downstream detector. Despite their simple construction they often had a scale marked on the filter giving useful accuracy and an amazingly wide frequency range, see Fig 1.

An active oscillator will be described using modular 50Ω technology first as a test circuit with a passive cavity filter, see Fig 2 and 3. A compact PCB construction with modern broadband amplifiers (gain block / MMICs) will be developed from this test oscillator. Transistorised cavity oscillators have already appeared in old HP measurement systems. You can find ready made oscillators manufactured for specified frequencies, for example at [2].

At first, generating a signal from a single coaxial cavity appears somewhat precarious. RF technology such as the directional coupler helps offering a wonderful solution. The actual cavity has little impact, both the phase and the amplitude of the signal are only slightly affected. The so-called "pulling effects" remain largely absent. An isolation amplifier is not required for signal investigation.

Instead of a directional coupler a broadband power splitter using a transformer can be used at frequencies below about 1GHz as described in [1]. When waveguides were still signal processing components, you would simply have drilled a hole in the side wall of the cavity. Then any frequency would have been “captured” in the waveguide and transmitted. This type of cavity may use less elements but gives reduced operating

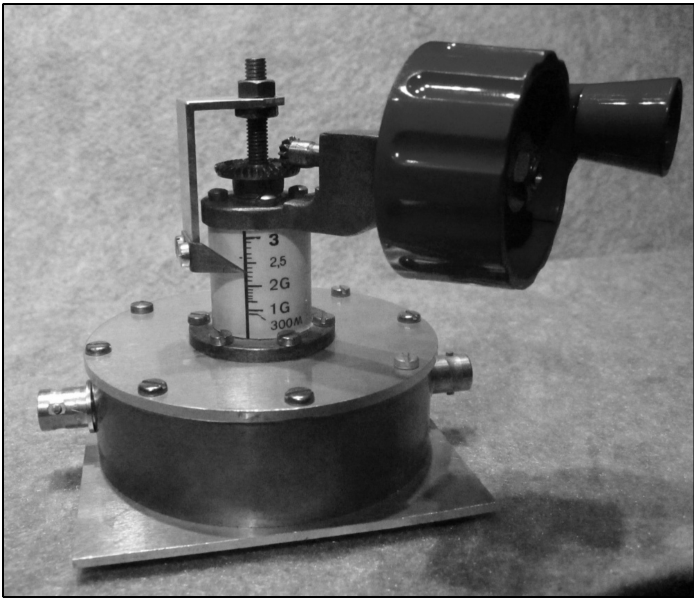


Fig 1: Poor man's spectrum analyser from ancient times. This tunable cavity resonator has a scale from 300MHz to 3GHz and is the frequency determining component of a modular test setup.

quality of the resonator.

The signal fed back to the amplifier must arrive with the correct phase shift otherwise the circuit will not produce continuous oscillation. The propagation speed of signals in a line can produce the phase shift. Coax cables can be used at lower frequencies, also printed striplines can be considered from a Gigaherz.

There is an interesting tool in the form of the so-called "Air Line" for experimenting with precision adjustment of the optimum length of the delay line, in the laboratory a "Trombone" can be used. Their geometric length can be adjusted and thus in the delay can be adjusted to a given amount.

A larger model of delay line with N type connectors (Fig 4) has six parallel air lines with a spindle drive, giving a continuously adjustable maximum travel of 24cm. Thus, any arbitrary phase shift between 0 and 360° is adjustable at 1.25GHz. In addition, the delay line of the oscillator circuit can be stretched to

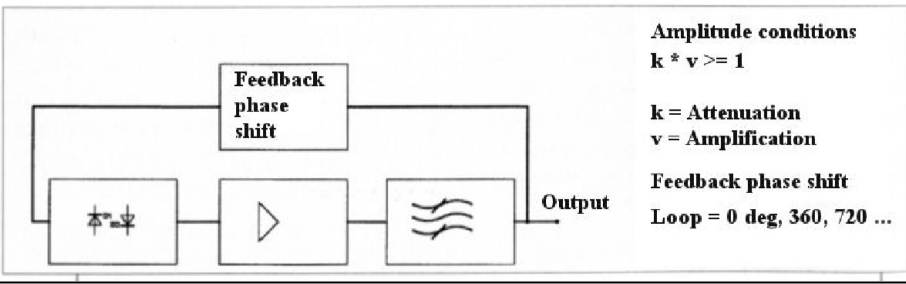


Fig 2: Block diagram of a feedback coupled oscillator with the following elements: Amplifier, Frequency determination (resonator), Feedback (negative feedback) and Amplitude limitation (reduction of increase to balance the levels).

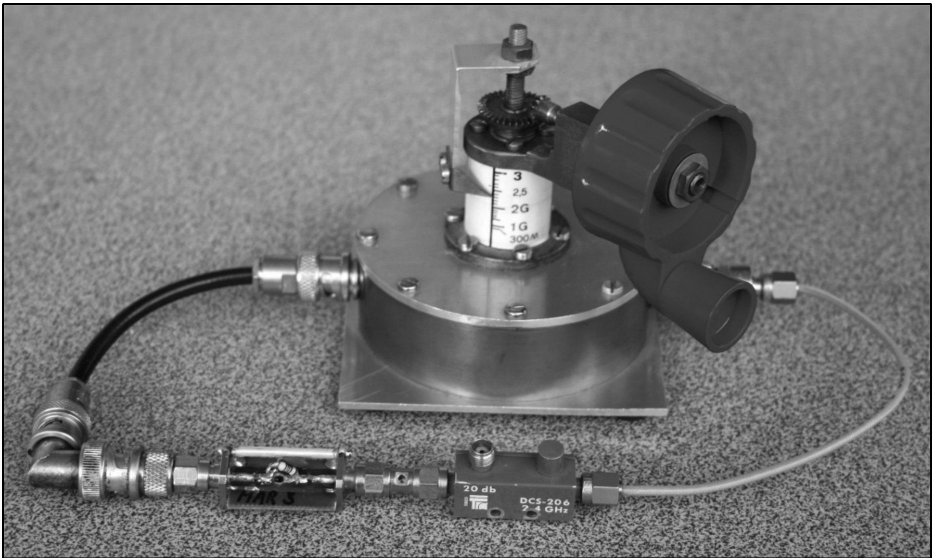


Fig 3: With only two additional modular components (Mar 3 amplifier and a directional coupler) the tuneable cavity is turned into a stable resonant oscillator,

one or more whole wavelengths. However, the steepness of the phase change with the frequency of the output increases. This effect restricts the possible frequency adjustment. At very high frequencies the natural damping of line also sets a limit to any increase in length of the phase line.

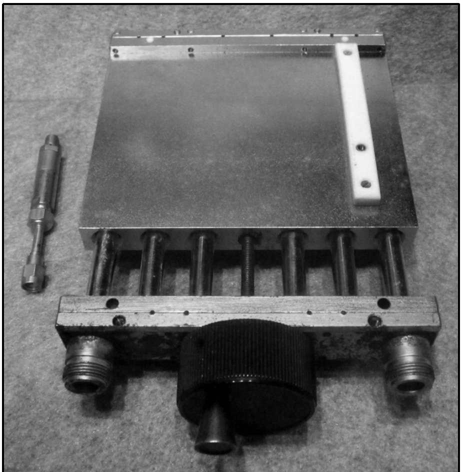


Fig 4: Two so-called "air lines" for adjustment of the phase angle; the path length can be adjusted to 16mm for the SMA version or 24cm for the N type version.

The result of the experiment exceeded expectations. The lowest frequency that oscillation could be achieved was 400MHz, Fig 5. At lower frequencies, up to the end of cavity at 300MHz the attenuation increased so that the amplification of the MMICs was no longer sufficient to comply with the amplitude condition. About 10% above the start point of oscillation it stops then tuning through it restarts 20% above the last cut off frequency. This interaction continues up to the top of the tuning range of resonator at 3GHz (also see the similarly effect in Fig 14).

An extension of the coaxial delay line moves operation to lower frequencies but reduces the higher frequencies. Oscillation is possible at all frequencies between 400MHz and 3GHz with an adjustable delay line or different cable lengths.

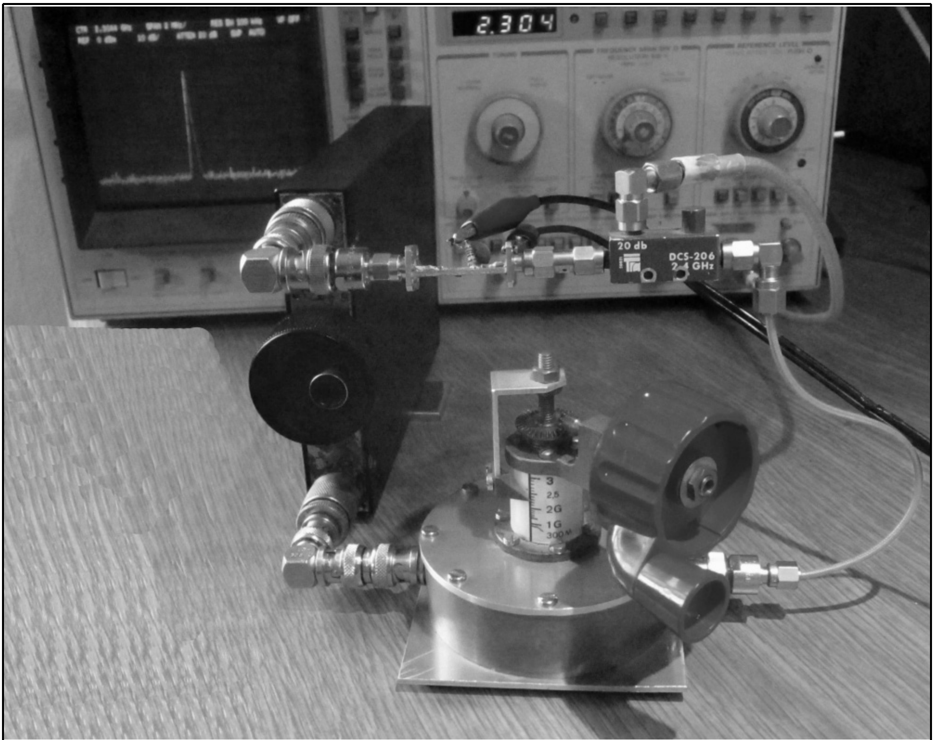


Fig 5: Prototype construction using modular components. It can be adjusted using the delay line length to oscillate between 1GHz and 3GHz.

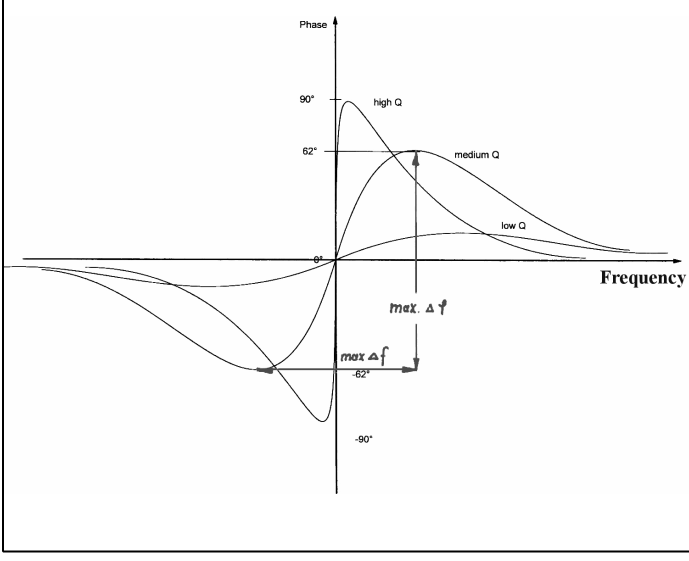


Fig 6: A chart of phase vs frequency (discriminator curve) for a single cavity frequency filter. The resonator produces some 120° phase shift over the range shown, this is a measure of it's "goodness". High quality resonators such as a crystal have a much reduced range.

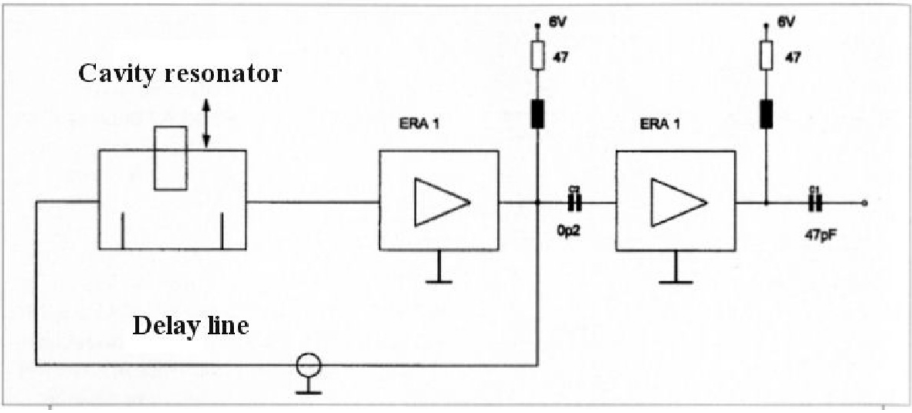


Fig 7: Circuit of the PCB oscillator using a cavity resonator. The very small value of coupling capacitor between the amplifier stages can made from a copper flag. The value can be adjusted by bending the flag.

The main frequency is determined by the cavity resonator. The correct length of the phase or delay line moves the feedback into the discrimination curve of the cavity resonator. The start and cut off frequencies are within the 120° phase shift window of the resonator, see Fig 6.

2.
The 5GHz oscillator

A resonator made from brass or copper pipe (from the hardware store) with 23mm internal diameter and 17mm high was used for the PCB version of the 5.000GHz cavity oscillator (Fig 7). The exact dimensions are quite uncritical because an M6 adjusting screw allows for

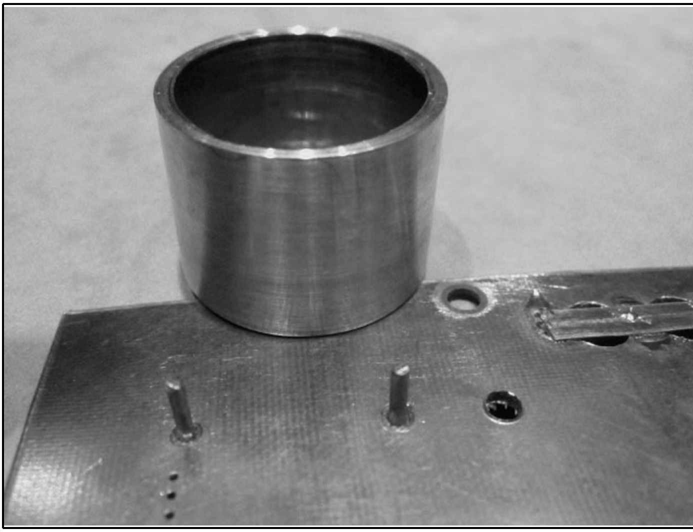


Fig 8: The PCB cavity oscillator being assembled. The cylindrical resonator is soldered with hot air onto the ground side of the RT Duroid PCB. The component side only contains a 50Ω stripline with breaks for the MMICs.

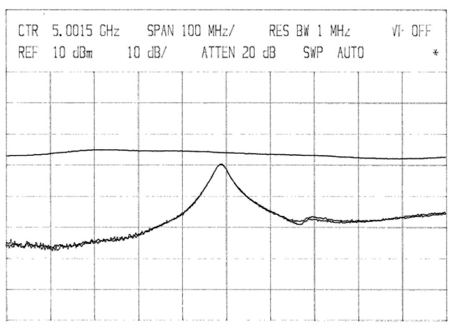


Fig 9: Response of the cavity resonator on the PCB. Top: reference signal (0dB). Bottom: signal showing 4dB attenuation at 5GHz.

an adjustment range of several GHz. Silver plating the inside slightly increases the quality of resonance. The coupling pins protrude from the head into the PCB and approximately 6mm into the Interior of the cavity (Fig. 8).

The best total length of the phase shifter line was experimentally determined to be approximately 10cm. This was approximately 8cm (outside the stripline) of thin solid sheath cable (D = 2.1mm). The required frequency of 5.000GHz is approximately in the centre of the adjustment window. The choice of gain block was between a GaAs FET or HEMT. The low cost SiGeT gain block from mobile

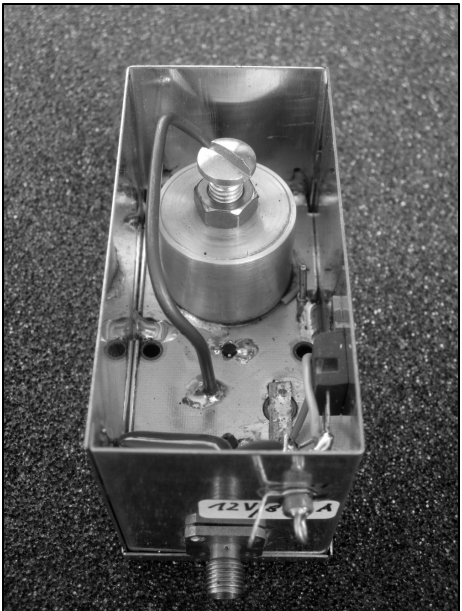


Fig 10: Functional 5GHz cavity oscillator. The thin solid coax phasing line can be seen running along the "cold" side of the PCB.

technology such as the SGA5289 (Sirenza) or BGA 616 (Infineon) should be well suited to about 5GHz.

A SiGeT semiconductor noise performance is better than MOS or GaAs FET therefore in the oscillator because they show less flicker noise at the frequency used. As a result of this noise the signal gets noise sidebands. In a normal driver this effect of the FETs doesn't usually matter.

Because they were available ERA1 MMICs from Mini-Circuits were used for the oscillator and output amplifier. For different amplifiers the phase angle between the input and output can vary considerably so the optimum length of the delay line may need to be changed.

The supply voltage of +5V is controlled by an LT1963A (low drop, low noise) regulator. Tests have shown that the affect of the supply voltage on the oscil-

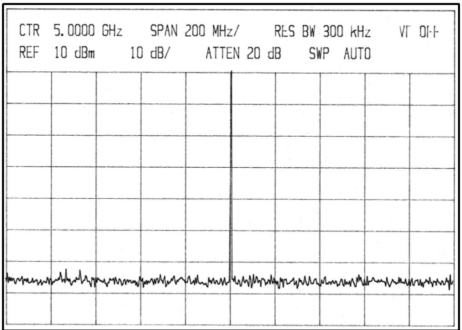


Fig 11: The output spectrum of the 5GHz oscillator showing that there are no parasitic oscillations in the range from 4 to 6GHz.

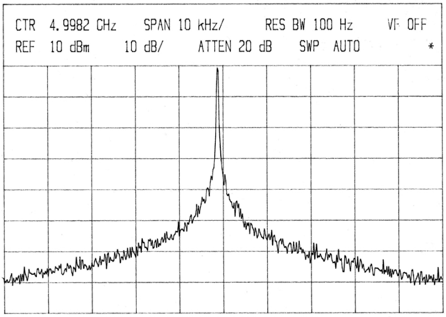


Fig 12: The spectrum of the 5GHz oscillator magnified by 2000 times showing the low noise signal with a 100Hz scan range.

lation frequency is rather low. Nevertheless, a certain FM sensitivity of about 1kHz per mV is generated by ripple and noise.

Fig 9 shows the response curve of the 5GHz cavity and Fig 10 shows the prototype 5GHz oscillator.

Fig 11 shows the output spectrum of the 5GHz oscillator and Fig 12 is a close in view showing that the signal is low noise.

3.

Tuning accuracy

A PLL can be used to regulate the frequency to a reference frequency, this requires an electronic adjustment of the frequency.

The phase shifter line offers such a VCO function. Modifying the delay time of this line using an L/C phase shifter so that the resonator responds with a corresponding change in frequency Δf to compensate for $\Delta\varphi$ (see Fig 6). The higher the Q of resonator, the smaller the possible change in frequency using the phase shifter line. Practical tests showed that only low frequency shifts in the range of about a tenth of a percent of the oscillating frequency are possible. The 5GHz oscillator can be moved about 6MHz. Further attempts to increase the frequency controls are not complete.

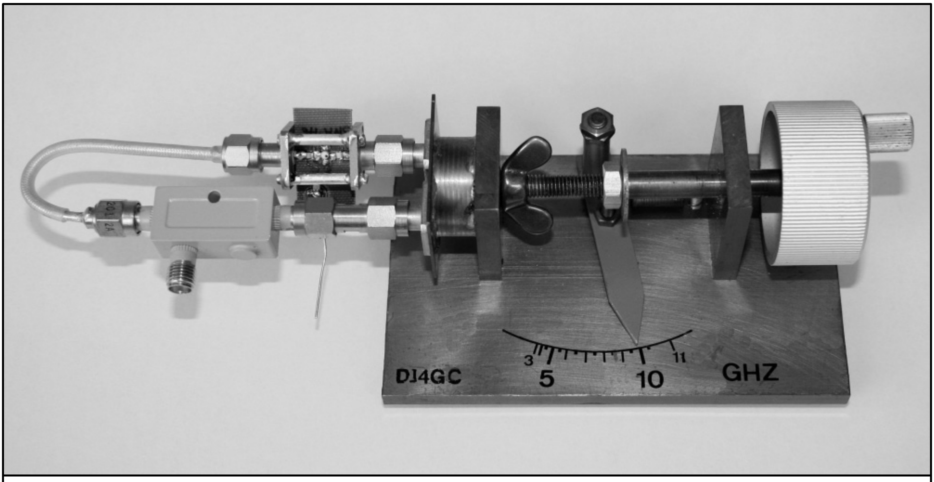


Fig 13: The laboratory 4 to 11GHz oscillator using coaxial parts.

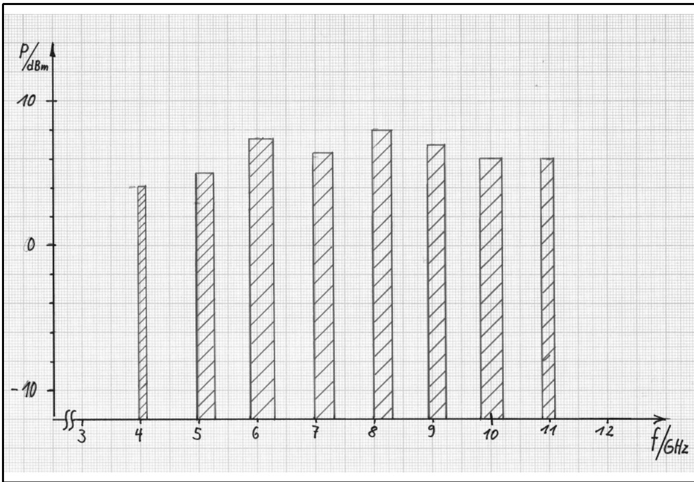


Fig 14: The windows where oscillation occurs for the 4 to 11GHz oscillator. All other frequencies are possible by lengthening the delay line.

4.

Test oscillator in the X band

An X-band resonator filter was found as part of a so-called "drag indicator" with numerous other craft projects. A clean microwave oscillator was possible with this incorporated in a closed gain loop. A small custom amplifier module using an ageing SNA176 from Stanford Micro Device has more than 10dB gain at 10GHz and proved an ideal gain block. The common and technically similar ERA1 from Mini Circuits should also work. The cheapest current MMIC successor models to the SNA176 are the NLB300 (plastic) and NBB310 (ceramic) from RFMD have the highest frequency limit. The NBB310 achieves a gain of 8dB at 14GHz.

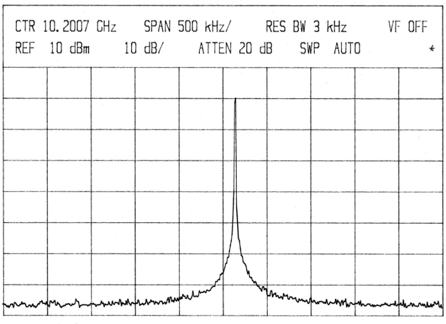


Fig 15: The output spectrum of the oscillator at 10GHz.

The laboratory prototype shown in Fig 13 is assembled from 50Ω coaxial components. The frequency determining cavity resonator with the two SMA connectors is tuned by using a M5 threaded rod. The oscillation is possible in narrow areas between 4 and 11GHz. Any frequency between these limits is adjustable depending on the length of the delay line. Since the construction of clean microwave oscillators is a stony area, these results are baffling given the almost spartan circuit design.

5.

Conclusion

Microwave oscillators from approximately 1GHz are easy to build with a cavity resonators and a modern gain block. The geometry of the resonators (including the question whether round or square) is quite uncritical, because they have a very wide range using the tuning screw. An upper frequency limit is so far

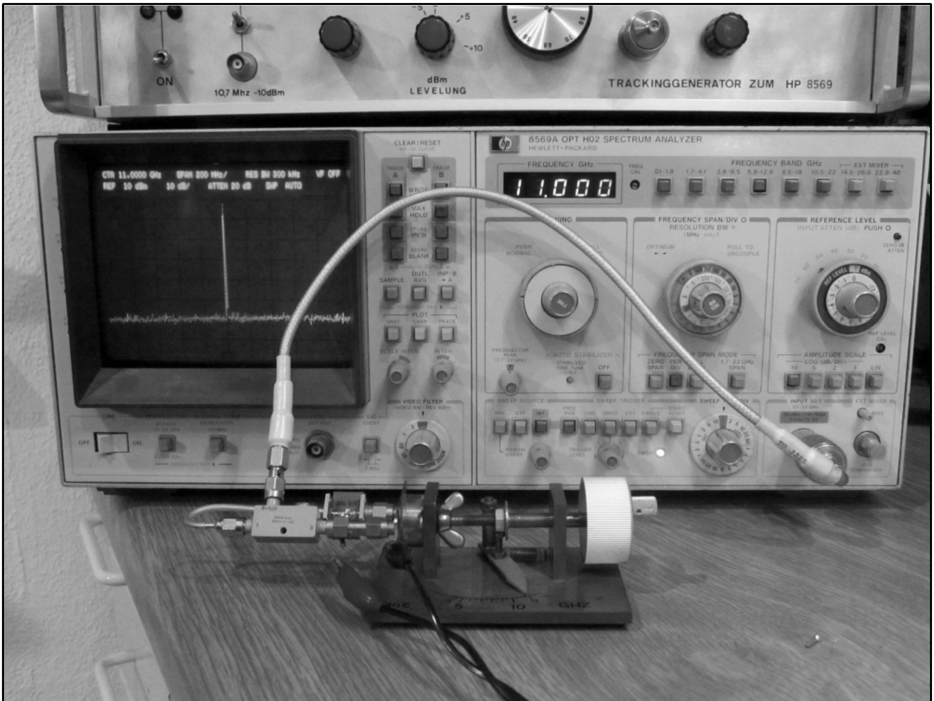


Fig 16: The x band oscillator being measured with a spectrum analyser.

not discernible. With the help of microwave semiconductors frequencies above the X band can be achieved. Approaches to improve the VCO function to lock such oscillators using a PLL systems are still being investigated.

6.

References

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[3] www.minicircuits.com

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[5] www.rfmd.com

Konrad Hupfer, DJ1EE

A Directional Coupler for the HF and UHF range

1. General

This article describes a directional coupler using operating principles known since about 1939 [1].

The mechanical and electrical construction is designed so that it can be constructed by experienced radio amateurs using only "normal" standard components.

A directional coupler can be used to measure the frequency and matching performance between a generator (transmitter) and consumer (load resistance, a following stage or an antenna). It enables the forward and backward RF energy to be measured to give information about the RF power consumed in the load and the determination of reflection coefficient or standing wave ratio.

Although most amateur radio antennas do not normally present a large standing wave ratio (VSWR) at the transmitter output with their feeder cables, it has proven itself for controlling the load VSWR.

A VSWR of 1: 1.3 "will" ensure that the power transistor in an amplifier "sees” a higher working resistance by a factor of 1.3 than for a standing wave ratio of 1: 1. At full power the load line, Id/Vd characteristic, can go outside the linear area: the consequence would be rising modulation products.

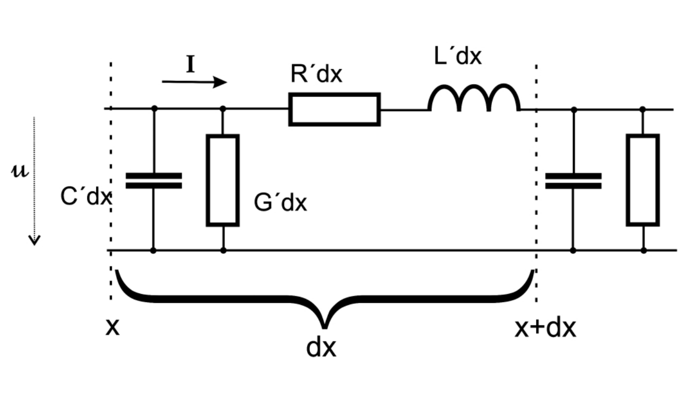

Fig 1: A small section of a transmission line.

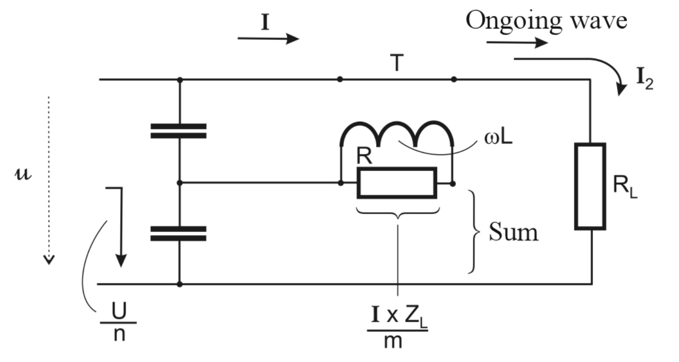

Fig 2: Measuring the forward wave current (Sum of currents).

2.

Waves on the feeder

There will be no introduction into the non-trivial theory of transmission lines, that is beyond the scope of the article. But some overview results from the theory will be stated.

Cleaver people have created a description of the voltage and power distribution using differential calculus (Fig 1) for the equivalent circuit of a very small section of the line. From this the direction of the line voltage and current values that apply at the input or output of point x can be determined.

After several mathematical "actions" the equations for current and voltage distribution in a loss-free homogeneous line (such as a 50Ω coaxial cable with negligible attenuation) can be derived:

$$U = \frac{U_2 - I_2 \cdot Z_L}{2} \cdot e^{-j\beta x} + \frac{U_2 - I_2 \cdot Z_L}{2} \cdot e^{-j\beta x} \quad (1)$$

$$I \cdot Z_L = \frac{U_2 - I_2 \cdot Z_L}{2} \cdot e^{-j\beta x} + \frac{U_2 + I_2 \cdot Z_L}{2} \cdot e^{+j\beta x} \quad (2)$$

Where:

U_1 and I_1: Voltage an Current at point x on the line.

Z_L: Characteristic impedance of the line

U_2 and I_2: Voltage and Current at the end of the line

ß: phase constant 2π/λ

The second term in equations (1) and (2) represents the forward wave and the first term correspond to the reflected waves [2].

3.

Measurement of forward and reverse waves

To measure these two waves a method must be devised from (1) and (2) for a simple voltage and current measurement.

Adding (1) and (2) gives:

$$U + I \cdot Z_L = (U_2 + I_2 \cdot Z_L) \cdot e^{j\beta x} \quad (3)$$

This corresponds to the forward wave. U and I x Z_L at point x can be measured at any point on the line, see Fig 2.

Here the line voltage U is measured as proportional U/n by a capacitive voltage divider. A voltage proportional to the line current, I x Z_L, is generated across R using a transformer. In the case of a correct match these two vector voltages are equal; their total U_S equals the wave current to the end of the line.

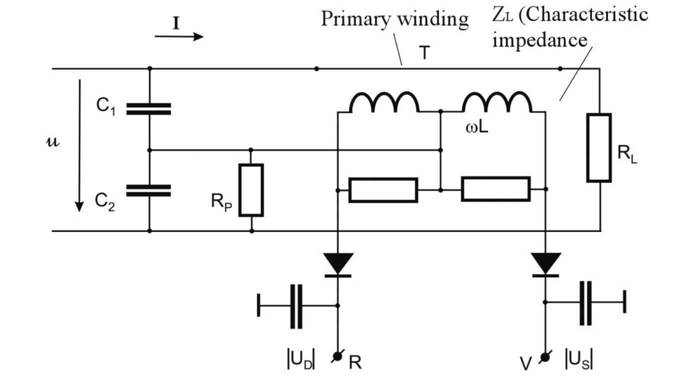

Fig 3: Circuit showing the principles for measuring the forward a reverse waves.

In the same way by subtracting equations (1) and (2) gives the reverse current wave:

$$U_S = prop\left(\frac{U}{n} + \frac{I \cdot Z_L}{m}\right) \qquad (4)$$

Where:

Us: total voltage

$$U_D = prop\left(\frac{U}{n} - \frac{I \cdot Z_L}{m}\right) \qquad (4a)$$

Where:

U_D: differential voltage

(In the case of a correct match U_D will be 0)

The principle of a circuit for the measurement of forward and reverse waves is shown in Fig 3.

The ports R and V are the rectified values of the vector sum and difference voltage U/n and I x Z_L/m and can be used to evaluate the forward and reverse waves.

If both diodes operate only in the square part of their function, we obtain a linear display of power (in watts) of the forward and reverse waves.

Similar indicators can be found in numerous amateur radio publications and radio literature [3], [4].

The principles of the circuit shown in Fig 3 have been used particularly in the kW range. A narrowing of the frequency range is mainly due to the current transformer, whose secondary inductance at low frequencies gives a significant shift in the phase measurement of:

$$\frac{I}{n} \cdot Z_L$$

This influence of smaller ωL can be compensated by a counter phase shift of the voltage U/n using a parallel resistor R_P. At low frequencies to keep this large enough (5 to 10 x Z_L), the permeability of the ring core must be very high and and more turns used. This increases the unwanted stray inductance and there is a disturbance of the phase and the amplitude of the higher frequencies.

In practice it has proved that such a directional coupler can be used with sufficient accuracy and directional focus (approximately -20dB) in the frequency range from 1MHz to about 50MHz measuring up to 1kW.

4.

R-bridge coupler

Broadband directional couplers are possible, if the required voltage (3), (4) are available in a purely resistive network. This is the famous bridge circuit shown in Fig 4.

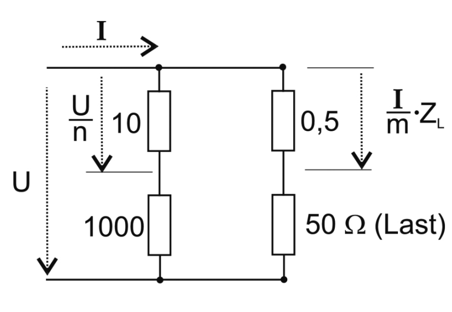

Fig 4: Resistance bridge to measure U_D and U_S.

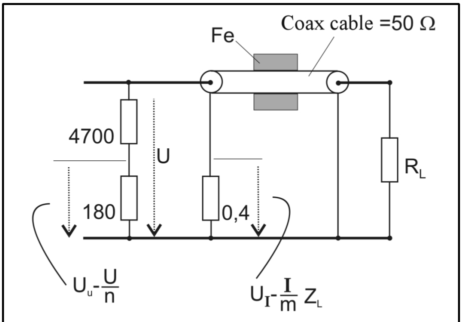

Fig 5: Design of the bridge circuit.

The vectors U/n and I/m x Z_L can be presented but their practical measurement is almost impossible! (High buried Baluns, etc.) Fig 5 shows how the circuit would look with transformers. The voltages U_u and U_1 must be measured with the correct phase as shown in Fig 6.

The line current is proportional to voltage U_I measured across a very low-impedance series resistance (0.41Ω in the rebuilt directional coupler). In parallel is the inductive resistance, X_L, consisting of a piece of coaxial cable sheath with ferrite rings. This is practically negligible over a very large frequency range! (Without the ferrite a 50mm length of coaxial cable would be sufficient to operate the arrangement down to about 5MHz with enough accuracy. With the additional ferrite the frequency range is expanded to < 0.1MHz).

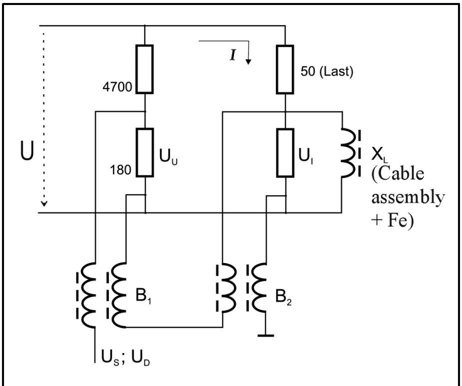

Fig 6: A simple bridge to measure U_D and U_S.

The return measurement comes from the values of U_u and U_I.

Despite using "disturbing" Baluns B_1, B_2, a measuring instruments with the following performance could be realised:

- Forwad coupling: 41dB;
- Frequency response: +/-0.4dB
- Directional focus: -25dB to -30dB;
- Frequency range: 0.1MHz to 500MHz

The two Baluns B_1 and B_2 (Fig 6) can be omitted if the voltages U_u and U_I are measured using an a common device (here a 50Ω RF Wattmeter measuring head) [5] (Fig 7). To keep the ohmic losses low in this arrangement, the voltage divider R_1, R_2 must have as low as possible impedance.

The "power resistor R_3", where the voltage U_I is measured must naturally be as low impedance as possible because the entire load current flows through it. R_L is set at 50Ω and the bridge condition (U_u = U_I) for adjustment

$$\frac{R_1}{R_2} = \frac{R_L}{R_3} \tag{5}$$

For the directional coupler shown in Fig

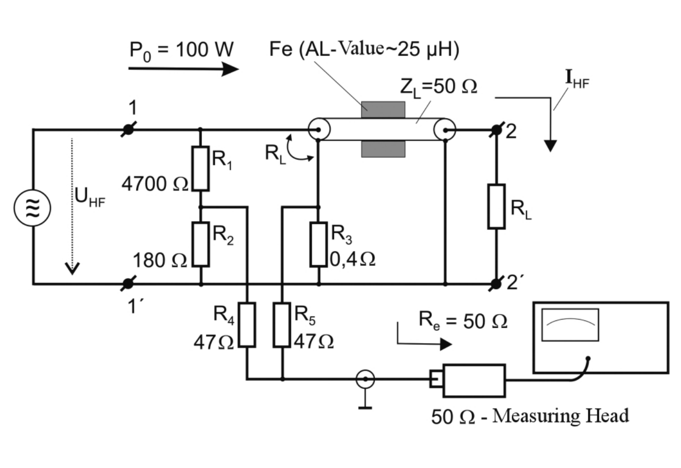

Fig 7: Bridge based sensor without Baluns to make broadband measurements of the values U_S and U_D.

7 it is necessary to swap the connections to the directional coupler. 1-1' to 2-2' to show the measurement of forward power and when swapped it is a measurement of the reverse power (R_L not equal to Z_L). When $R_L = Z_L$ the display on the Watt-meter should be zero. However in practice this is not the case; it is a measurement of "residual return power" because of the non-ideal construction. The ratio of this residual return power to the previously measured forward power is known as directional focus.

An example of a directional coupler constructed as Fig 7 is shown in Fig 8.

The ohmic losses expected at 100W (70.7V = UHF) can be determined. To simplify, the resistors R_4, R_5, and R_6 should not be considered:

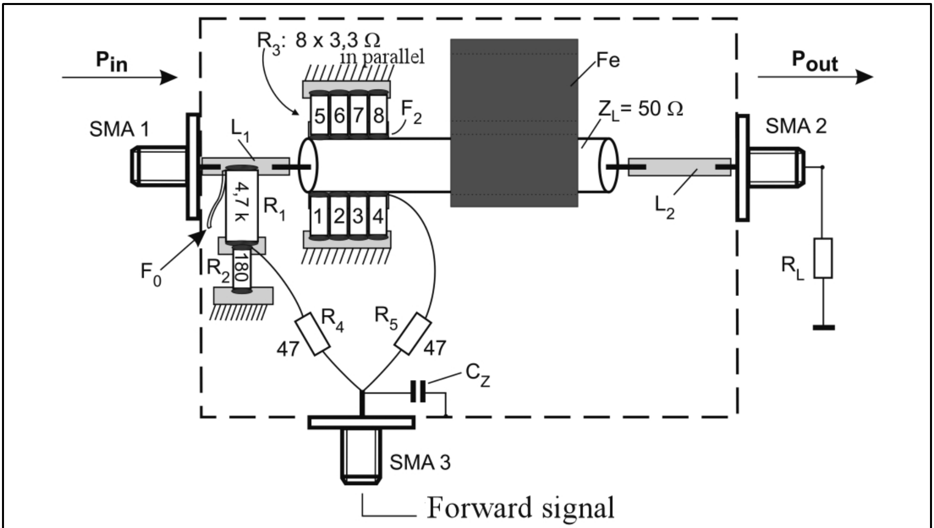

Fig 8: Diagram showing the construction of the coupler.

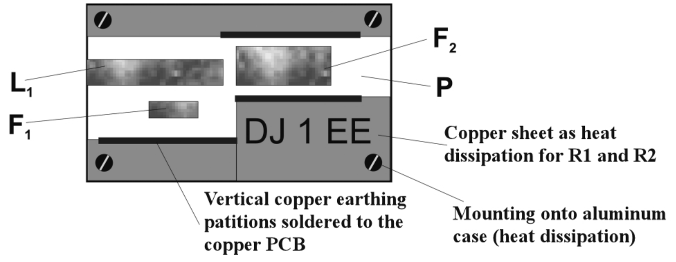

Fig 9: See legends below.

RF loss in the U_U voltage divider R_1/R_2:

$$P_{R_1/R_2} = \frac{U^2}{R_1 + R_2} = \frac{(70.7V)^2}{4700 + 180\Omega} = 1.02W$$

RF loss in the current shunt resistor R_3:

$$I_{HF} = \frac{U_{HF}}{R_L} = \frac{70.7V}{50\Omega} = 1.4A$$

$$P_{R_3} = I_{HF}{}^2 \cdot R_3 = (1.4A)^2 \cdot 0.41\Omega = 0.82W$$

Total loss = 1.84W = 0.1dB.

The forward coupling is selected as 42dB ±0.3dB in the frequency range from 0.1MHz to 500MHz.

With the coupler reversed (feed on 2-2, load on 1-1 shows an attenuation of about 70dB (R_L at 1-1 ' = 50Ω!))

For the indication range a value 28dB is the difference.

5.

Mechanical and electrical construction and matching

The possible and also proven mechanical and electrical construction is shown in Figs 8, 9 and 10.

To match the bridge, replace R_2 with a small pot of about 500Ω. Feed the generator energy (for example from a scalar Analyser) to port 2 and terminate port 1 with 50Ω and set the RF energy measured at port 3 (return performance) to a minimum. Because a potentiometer is being used, high frequencies are rather inappropriate. Determination the value for R_2 only at low frequencies (2MHz to 30MHz) should be OK. For several examples a value of 180Ω was found for R_2. Thus a directional focus of about

Legends for Figs 8 and 9.

L1, L2: 50Ω stripline
P: PCB with microstrip islands with F_1 and F_2 copper sheet metal soldered on
R_1: SMD resistor of 4.7kΩ 1W, 6.2 x 3.1mm
R_2: SMD resistor 180Ω 0.5 W, 3.1 x 1.5mm
R_3: SMD resistor 8 x 3.3Ω 0.5W each
$R_4/_5$: 47Ω axial resistor 3.5 x 1.5mm Ø

F_1: continuous copper area approximately 4mm x 2mm
F_2: continuous copper area approximately 6mm x 3.5mm
Fe: double hole ferrite core approximately 14 x 12 x 9mm; MU approximately 250; or VAC Vitroperm-Ring 6-L 2009 (AL giving a value approx. 25µH)
F_0: "compensation" flag for measurement levelling

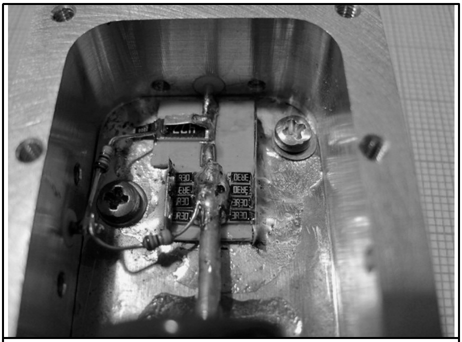

Fig 10: A photograph of the prototype showing the SMD resistors, the SMA connectors and the coaxial cable.

-20dB should be achievable in the frequency range from 1MHz to 100MHz.

As mentioned, this resistor bridge is full of stray inductance and capacitance. The measurements at high frequencies will fall off without additional compensation. With an additional capacitor (flag F_0 in Fig 8) relatively broadband matching in the frequency range from 1MHz to 500MHz is achievable.

A minimum value for the directional focus of about -20dB should be achievable over the above frequency range. A good forward coupling of approximately 41dB depends mostly on the line performance (Construction at port 1 to port 2) and a deviation of of ±0.3dB up to 0.5dB over the whole frequency range.

The "fine balance" relies on the sources U_u and U_I not “seeing” one another. A small metal partition between R_1 and R_3 can be useful both for the frequency response, also for best directional focus.

The position of the resistors R_4 and R_5 influences the forward frequency response. C_Z (2 to 8pF) at the connection of the measuring head is one of the possibly frequency improvements.

Details of the construction of the resistor network are shown in Fig 10. Fig 11 shows a double directional coupler with two bridge circuits. Thus the forward and reverse measurements can be determined at the same time. The further processing of the forward and return measurements to give a direct display of the real power transmitted to the load plus the display of the standing wave ratio (VSWR) and reflection factor in R_I Z_L will not be described in this article. References [3] [4] can be used for information.

For queries about the practical construction the author can be contacted by email: konrad.hupfer@t-online.de

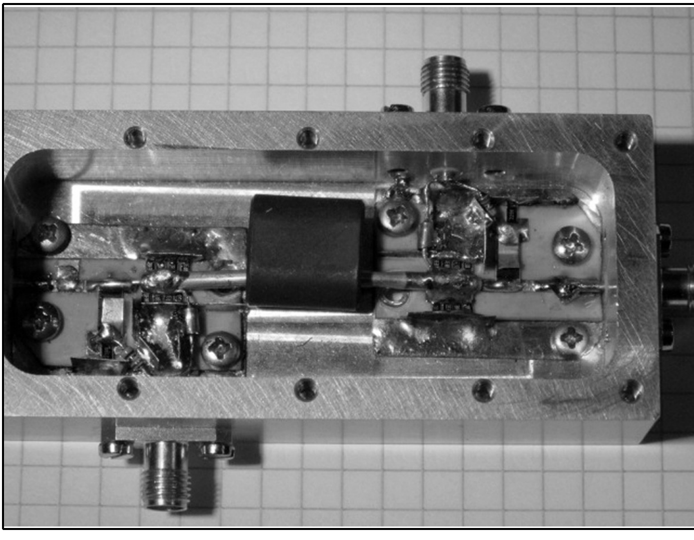

Fig 11: Prototype of a dual coupler design.

6. Literature

[1] "Arrangement for the monitoring of radio frequency energy services", The Reich Patent Office, patent 737877, July 28, 1943, Werner Buschbeck, Berlin.

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[3] "The original form after Buschbeck" CQ-DL 3-2004, pages 188 to 190

[4] Digital PEP Watt-meters and SWR calculator; ARRL Handbook 1989, pp. 3410-3415

[5] Patent application US 5926076, Werlatone, U.S.A.

Andy Barter, G8ATD

An Electronic Compass

1. Introduction

When I go out with a portable station on 23cm I often use a 55-element Tonna antenna that has a very narrow beamwidth. I prefer to manually turn the antenna because it is quicker and easier to line up on a distant station. There are two problems with that:

- It is impossible to see the antenna at night so it is difficult to see where the antenna is pointing
- If it is bad weather, with rain, it is an unpleasant experience looking out to see where the antenna is pointing

For a long time I wanted some form of electronic compass fixed to the antenna with a display in the shack (tent or car). With the increase in home made robots various sensors have become available at a reasonable price. This compass is based around a small unit available from robot component suppliers with two magnetic sensors at right angles and a PIC to do

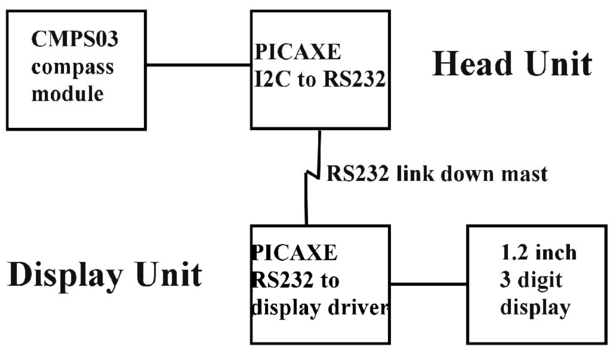

Fig 1: Block diagram of the electronic compass.

Fig 2: The Head Unit mounted on the mast.

the maths to work out the direction that it is pointing.

I have recently found the PICAXE microcontroller family and for someone like me who can write a simple program but is not a programmer they are ideal to turn the output of the compass board into something that can display remotely in the shack. The PICAXE is programmed in a Basic like language with some very useful statements that make talking to I2C buses and RS232 links very easy. The PICAXE development software that

Fig 3: The Display Unit in the shack (tent).

runs on a PC is free to download from [1]. Once a program is written it is easily transferred to the PICAXE microcontroller via the in circuit programming connection using a cable from either a serial port on your PC or using a special USB cable.

2.

Description of the compass

Fig 1 shows a block diagram of the compass. It is split into two units; the Head Unit mounted at the top of the mast (Fig 2) and the Display Unit (Fig 3) that sits in the shack with a digital display of the bearing.

The Head Unit contains the CMPS03 compass board [2] that has several outputs to send the direction that it is pointing to the outside world. Because the PICAXE has a very simple way to communicate with an I2C bus, I chose that method to pick up the output. The PICAXE microcontroller does a simple job of taking the direction information that is output as 0 – 3599 and sending it down to the shack over an RS232 link.

The Display Unit contains a second PICAXE microcontroller to receive the RS232 data sent by the Head Unit and displaying it on 3 large seven-segment displays.

I built this project to check if this type electronic compass would be accurate enough and useful enough to be worth developing further.

3.

The Head Unit

Fig 4 shows the circuit diagram of the Head Unit. It is fairly straightforward with a 5v regulator generating the internal 5v supplies from a 12v supply sent up the cable from the shack. The CMPS03 compass board requires a 5v supply and the I2C bus (SDA and SCL) are connected to the PICAXE microcontroller, the bus connections have 4k7Ω pull up resistors.

The CMPS03 board comes pre-calibrated and I have not recalibrated mine but there is provision to calibrate it for your chosen location. The calibration process is described in the literature on [2], it involves connecting the calibrate pin to 0v and accurately pointing the unit at the compass cardinal points. The circuit includes header pins to connect the calibrate pin to 0v and an LED to indicate that the calibration process is being carried out.

The PICAXE program (Table 1) reads the I2C bus and picks out the bearing data sent as 0000 to 3599. This is converted from the two-byte word into five ASCII characters then sent via a MAX3223 RS232 driver to the Display Unit. Only the most significant four characters are sent because the fifth is the decimal digit of the bearing and not thought worth displaying. The message sent is “BRGnnnn” this enables the PICAXE microcontroller in the Display Unit to synchronise with the data being sent by detecting the “BRG” in the string sent.

The PCB and component layouts are shown in Figs 5 and 6. The unit is housed in a plastic box using all nylon screws for fixings to minimise the local magnetic fields (Fig 7). The box is fixed to a short piece of square aluminium tube that is attached to the mast with a standard mast clamp (Fig 8). The Head Unit is connected to the Display Unit with a 3-core cable.

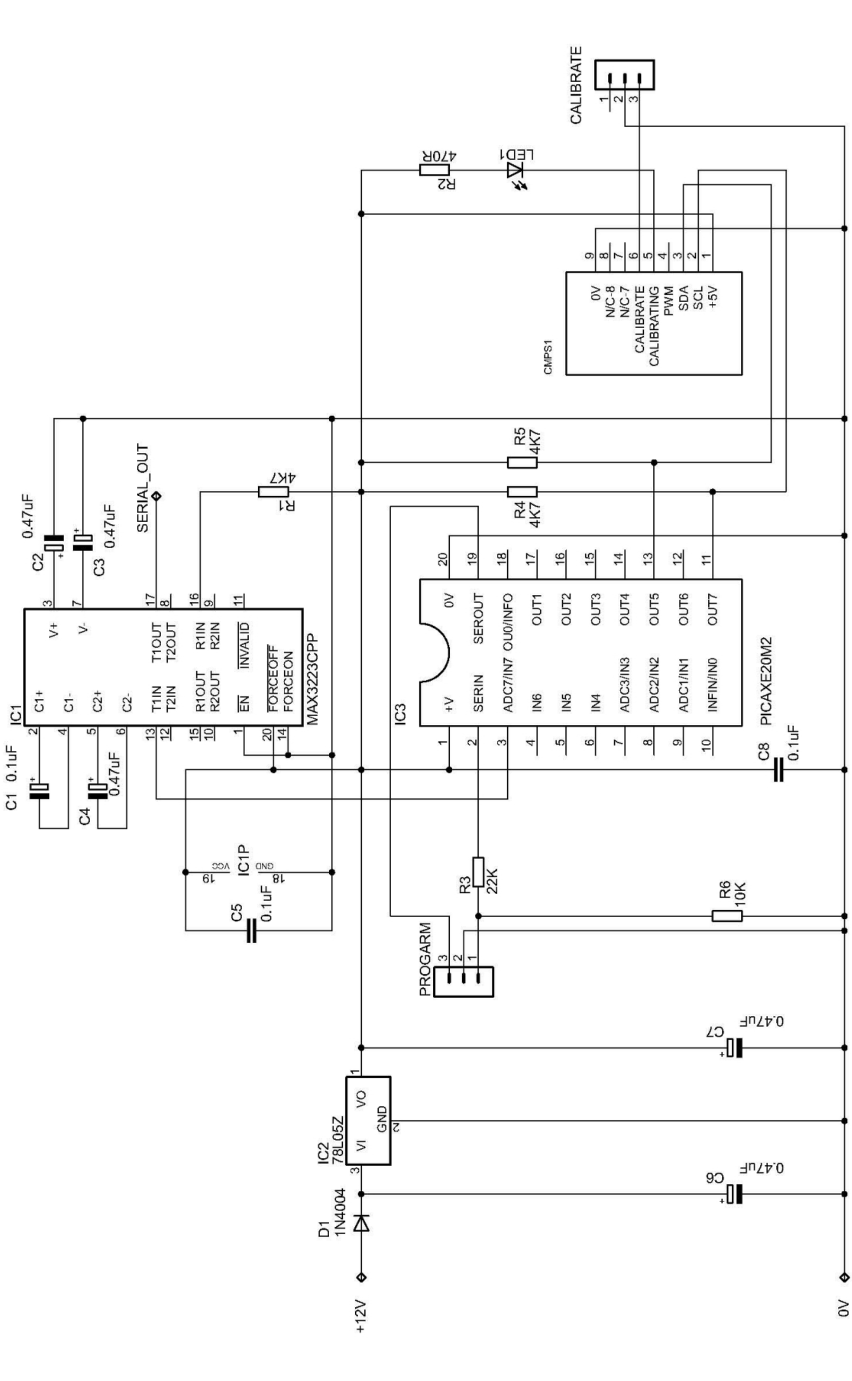

Fig 4: Circuit diagram of the Head Unit.

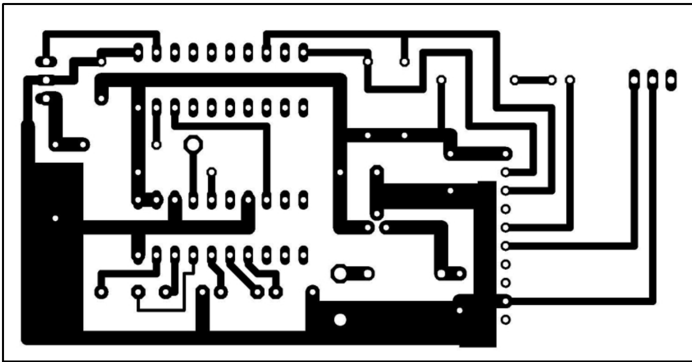

Fig 5: PCB layout for the Head Unit.

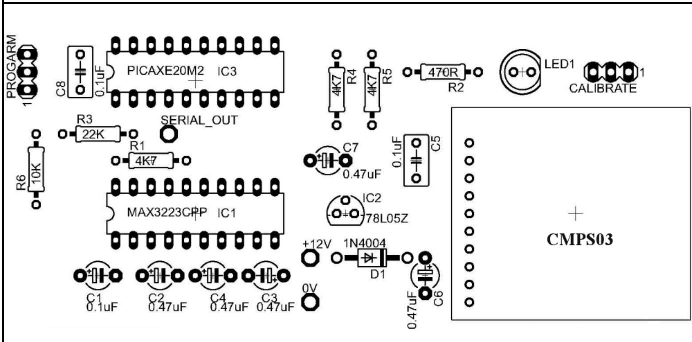

Fig 6: Component layout for the Head Unit.

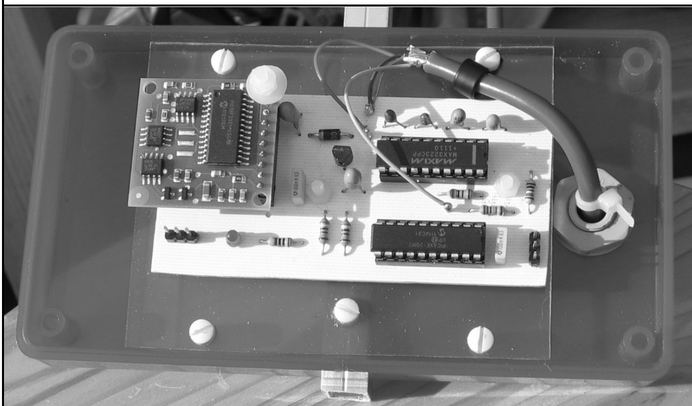

Fig 7: The Head Unit mounted in its plastic case. All nylon screws used to reduce the "local" magnetic field.

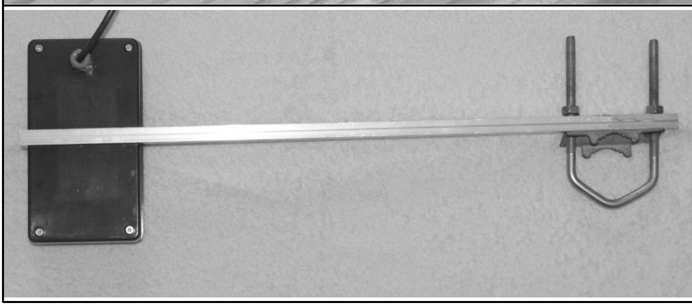

Fig 8: The Head Unit mounted on an aluminium bar to attach to the mast with a standard U bolt.

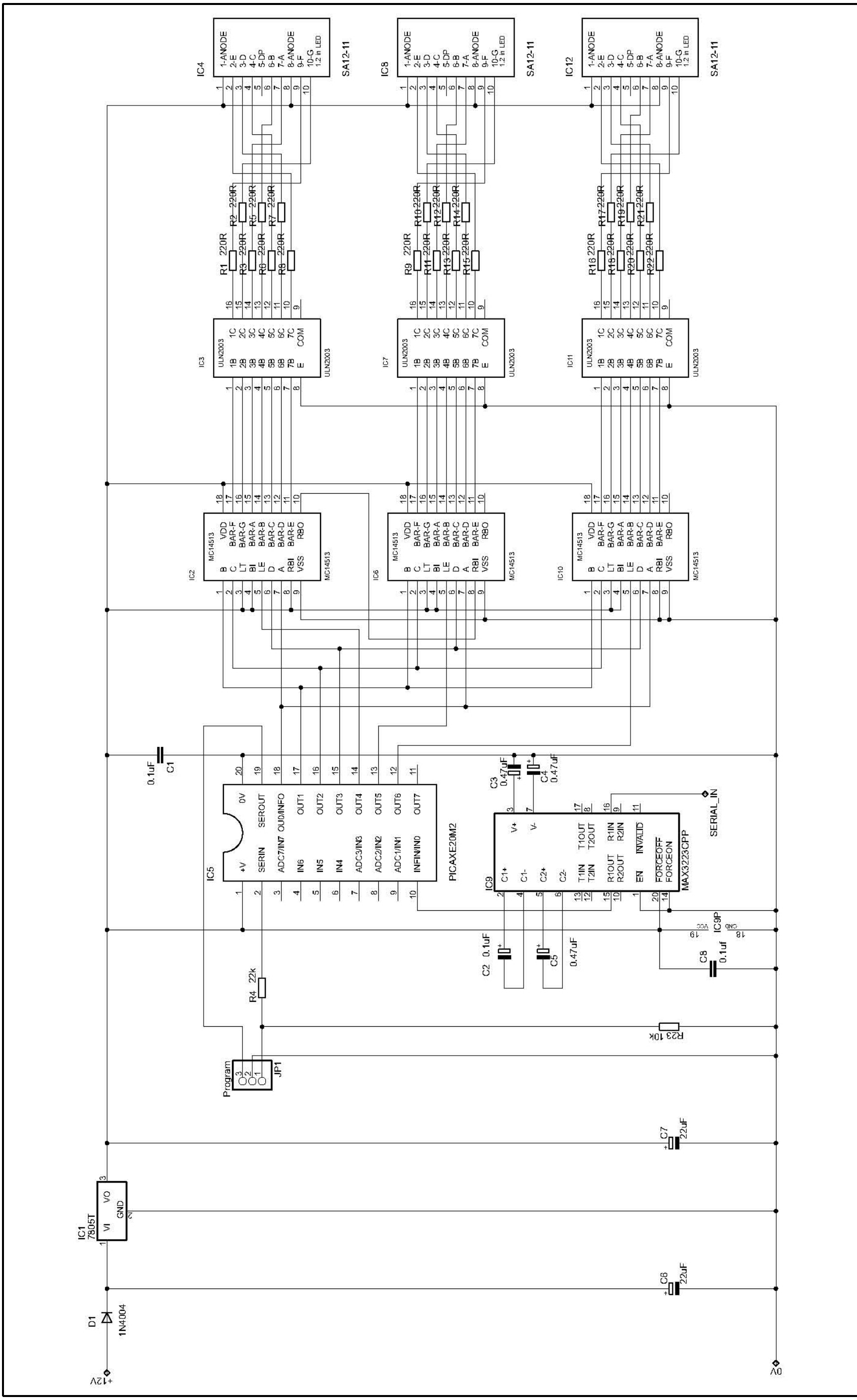

Fig 9: Circuit diagram of the Display Unit.

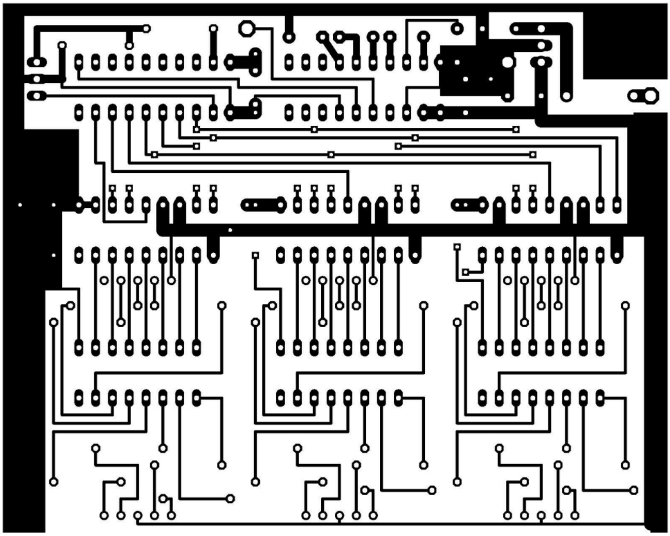

Fig 10: PCB for the Display Unit.

4. Display Unit

Fig 9 shows the circuit diagram of the Display Unit. It is fairly straightforward with a 5v regulator generating the internal 5v supplies from a 12v supply. The RS232 data coming from the Head Unit is fed to a MAX3223 for conversion to TTL levels and then fed to pin IN0 of the PICAXE microcontroller. The data required is taken out of the data stream and fed to the lower four bits of the output port (OUT0 – OUT3). These are connected to three MC14513 7-segment latch/display drivers, the latch inputs of each one is fed from a different pin on the PICAXE output port so that the correct data can be stored in the latch when it is available. The segment outputs are connected to ULN2003 buffers to feed the SA12-11 1.2 inch 7-segment displays.

The PICAXE program (Table 2) uses the “serin” statement of the PICAXE that reads the serial input and looks for the string “BRG” sent by the Head Unit. When that is received the following 4 bytes of the message are read and saved. The first byte received is the ten thousands digit and will always be zero so that is not used. The remaining bytes are presented to the latch/display drivers and the correct latch line operated to store the byte for display.

The PCB and component layouts are shown in Figs 10 and 11. The 7 segment displays are plugged into header sockets so that the buffer ICs and limiting resistors fit under them allowing the PCB to be laid out using the free version of EAGLE CAD [3]. The free version of EAGLE CAD has a few limitations over the full version, one of those restrictions is the size of the PCB that can be produced. I am still on the learning curve with this PCB design software so some of the features in the circuit diagrams and PCB layout may not be ideal. The PCB has some wire links because it was impossible to track as a single sided board. I did in fact track it as a double sided board and the links shown are

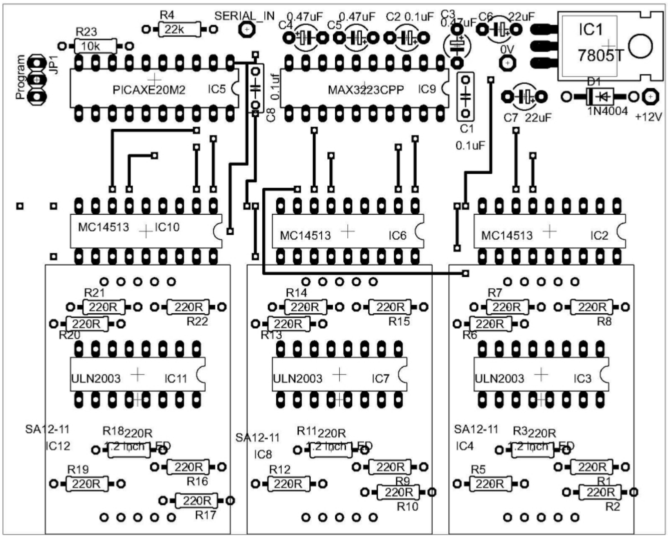

Fig 11: Component layout for the Display Unit. The 1.2 inch LED displays are mounted on header sockets so that they sit above the buffer ICs and limiting resistors.

tracks that were placed on the upper layer with vias where the wire links connect. The unit is housed in a plastic box with the PCB mounted on the front panel and the displays showing through a cutout. A 3-pin din socket is used to connect the cable to the Head Unit.

5. Performance

The electronic compass was used in anger for the first time for the RSGB VHF NFD at the beginning of July. It was found to be accurate to within ±3 degrees when peaking up on a weak signal and comparing the displayed heading with that calculated using the QRA locator of the distant station. There was an unexpected benefit that meant that any weak stations heard but not worked could be found again accurately by noting down the heading displayed and the frequency. Stations tend to stay on the same calling frequency during a contest so this technique worked well especially since conditions were pretty poor. The other benefit was that only my lower arm got wet in the atrocious weather.

6. Conclusion

The project was fairly easy and the end results were better that expected. It is perfectly useable in its present state and is certainly worth developing further. The only problem discovered was that the 1.2-inch LEDs were not bright enough for strong daylight conditions. Perhaps increasing the display current will overcome that shortcoming. I am now planning to write a Windows program to display the bearing and a compass pointer in a window on the logging computer. My Windows programming skills are less than my PICAXE skills so that project may take some time. Whilst searching for the right programming tool to do the job I came across a commercial unit that does the same job [4]. If you can’t wait for the Windows program, it will not be a good as the Sierra Radio version, then there is a ready made offering available.

7. References

[1] PICAXE - http://www.picaxe.com/

[2] CMPS03 - http://www.robot-electronics.co.uk/htm/cmps3tech.htm

[3] Eagle CAD - http://www.cadsoftusa.com/

[4] Sierra Radio compass - http://www.hamstack.com/ss_compass.html

Table 1: Head Unit PICAXE program.

```
; PICAXE electronic compass 20/5/2012
; Reads the I2C output from a CMPS03 compass module
; The output is a single byte showing the revision level - not used
; A single byte with the direction as 0 - 255 - not used
; A word (two bytes) with the direction as 0 - 3599
; The direction word is converted to an ASCII string  and transmitted at 600 baud
; over a serial link with a header of "BRG"


symbol BEARING_1 = b7           ; CMPS03 Bearing word high byte
symbol BEARING_2 = b6           ; CMPS03 Bearing word low byte
                                ; b6 and b7 are referred to as w3 - word 3
symbol BEARING = b27            ; CMPS03 Bearing byte 0 - 255 - not used
symbol RE = b26                 ; CMPS03 revision byte - not used
symbol BTTH = b14               ; Bearing ten thousands
symbol BTH = b13                ; Bearing thousands
symbol BHND = b12               ; Bearing hundreds
symbol BTEN = b11               ; Bearing tens
symbol BUNIT = b10              ; Bearing units


main:                           ; main program loop
    gosub rdcompass             ; read the compass and output serial data
    pause 1000                  ; wait a while
    goto main                   ; go round again

rdcompass:                      ; read compass
    hi2csetup i2cmaster,$C0, i2cslow, i2cbyte  ; initialise compass on I2C bus
    hi2cin 0,(RE, BEARING, BEARING_1, BEARING_2)  ; read compass via I2C bus
    bintoascii w3,BTTH,BTH,BHND,BTEN,BUNIT  ; convert bearing word - word 3 - to ASCII
    serout C.7, T600_4, ("BRG",BTTH, BTH, BHND, BTEN) ; send data at 600 baud to pin C.7
return
```

Table 2: Display Unit PICAXE program.

```
; PICAXE electronic compass display 26/5/2012
; Receives bearing data string from Head Unit and displays on 7 segment LEDs
; Bearing data in string is: Ten Thousands digit - not requires
; Thousands digit - MSD of display
; Hundreds digit - middle digit of display
; Tens digit - LSD of display

symbol BEARING_1 = b9          ; Bearing hundreds
symbol BEARING_2 = b7          ; Bearing tens
symbol BEARING = b5            ; Bearing thousands;

start:  let dirsB = 255        ; Set pins on port B to outputs
        let BEARING = 0        ; initialise bearing
        high b.4               ; Set thousands display digit latch line high
        high b.5               ; Set tens display digit latch line high
        high b.6               ; Set units display digit latch line high

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

                               ; b2 holds ten thousands digit - not required
                               ; BEARING holds thousands digit - MSD of display
                               ; BEARING_1 holds hundreds digit
                               ; BEARING-2 holds tens digit - LSD of display
                               ; Display shows 0 - 359 from 0 - 3599
        let b1 = BEARING       ; Set b1 = thousands digit
        let b1 = b1 & 15       ; Mask off top 4 bits
        let b1 = b1 + 240      ; Make top 4 bits high - they are used for display latch
        let pinsB = b1         ; Write b1 to the B port - sends digit to display
        low b.4                ; Set thousands digit latch low to store data in display driver
        pause 10               ; Wait a while
        high b.4               ; Set latch high
        let b2 = BEARING_1     ; Set b2 = hundreds digit
        let b2 = b2 & 15       ; Mask off top 4 bits
        let b2 = b2 + 240      ; Make top 4 bits high - they are used for display latches
        let pinsB = b2         ; Write b2 to the B port - sends digit to display
        low b.5                ; Set hundreds digit latch low to store data in display driver
        pause 10               ; Wait a while
        high b.5               ; Set latch high
        let b3 = BEARING_2     ; Set b3 = tens digit
        let b3 = b3 & 15       ; Mask off top 4 bits
        let b3 = b3 + 240      ; Make top 4 bits high - they are used for display latches
        let pinsB = b3         ; Write b3 to the B port - sends digit to display
        low b.6                ; Set tens digit latch low to store data in display driver
        pause 10               ; Wait a while
        high b.6               ; Set latch high
        goto main              ; Go round again
```

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Using Smith Diagrams

Reprint from issue 1/1984

In RF-technology, one often displays real and reactive impedances as a function of frequency in the form of a “locus”. Fig 1 shows a typical example of this with an RF transistor TDA1087, and Fig 2 with a HB9CV antenna. This article is to describe how to read Smith diagrams, and how to use them to advantage.

Every electronics technician knows that a series connection of real and reactive impedances can be shown graphically as two arrows that are perpendicular to one another (Fig 3). If this is to be displayed in a more exact manner mathematically, these two vectors should be inserted into a system of coordinates where the real impedances are inserted to the right, and the reactive impedances vertically. A vector pointing upwards (positive sign) represents an “inductive reactive impedance”, and a vector pointing downwards (negative sign) shows a "capacitive reactive impedance" (Fig 4).

This impedance plane has the disadvantage that it must be extended infinitely if it is to display all possible impedance conditions, such as the infinite impedance present under non-load conditions.

This problem can be solved by “bending”

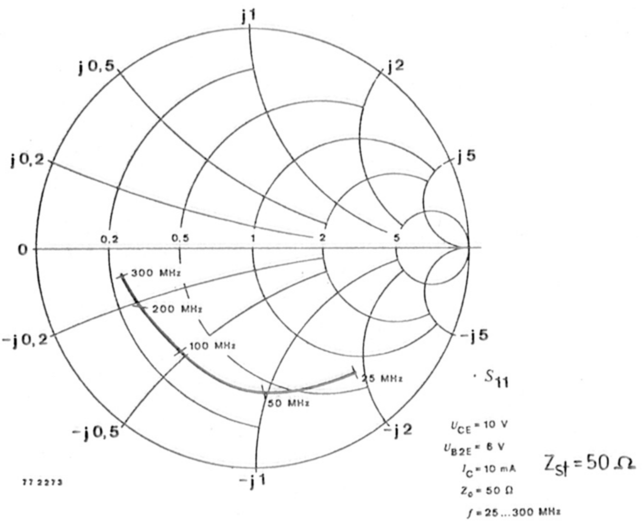

Fig 1: Input impedance of a TDA1087 as locus in a Smith diagram.

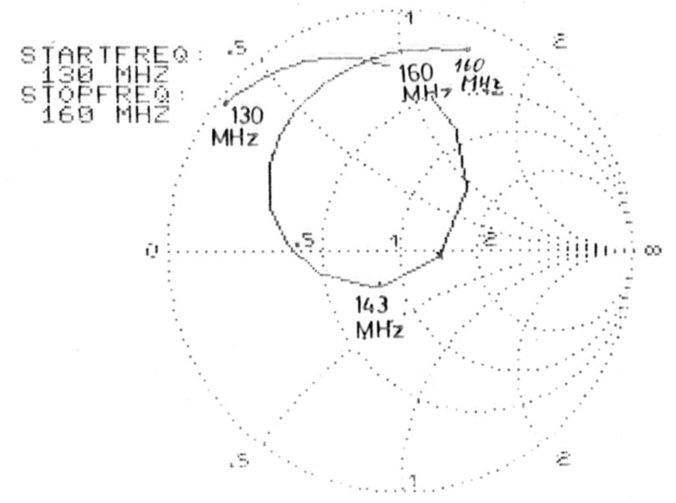

Fig 2: Impedance locus of a HB9CV antenna for the 2m band.

the vertical coordinates in the impedance plane (Fig 5, left) to form circles. Of course, the vertical axis must form the outer circle (Fig 5, centre); the reactive impedance values +∞ and -∞ are in the "finite" then.

This is where the outer circle (Fig 5, right) intersects the horizontal axis. The consequence of this is that the real impedance axis is no longer linearly scaled. Here, also the impedance value ∞ comes into finite values and meets the other two infinite points at the right.

The bending process also has an effect on the horizontal coordinates which are in the form of sectors. With increasing inductive or capacitive reactive impedance they become continuously shorter and are shifted more and more to the infinite point. The whole rectangular system of coordinates in the impedance plane with its straight coordinates is thus transposed to a curved coordinate system

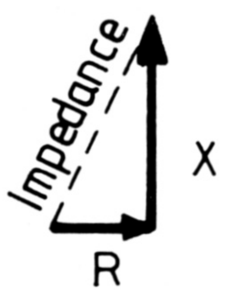

Fig 3: Real and reactive impedance.

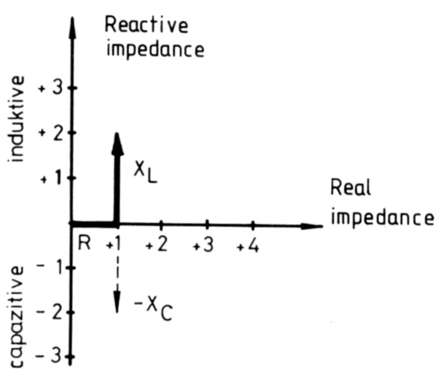

Fig 4: Real and reactive impedances in a coordinate system.

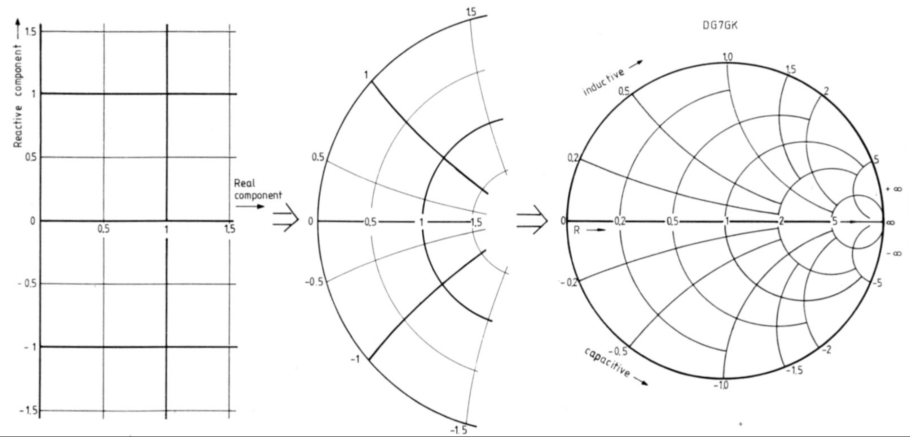

Fig 5: Transition from rectangular coordinate system to Smith diagram.

that fills the inside of a circle. Only the right angles (intersections of originally horizontal and vertical coordinates) remain intact during the "bending" process.

It is advisable to practice by inserting the real and reactive impedances in the left half of the Smith diagram, before attempting to enter into the area where the originally vertical coordinates transpose into the horizontal and finally reverse the

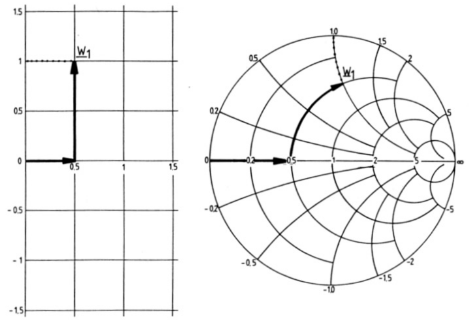

Fig 6: Series connection of real and reactive components: Real component = 0.5; reactive component = 1.0. "+" means "inductive".

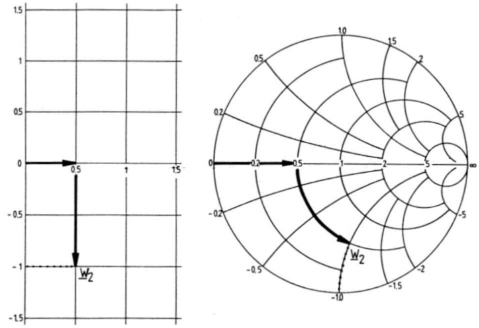

Fig 7: Series connection: Real component = 0.5' reactive component = -1. "-1" means "capitative".

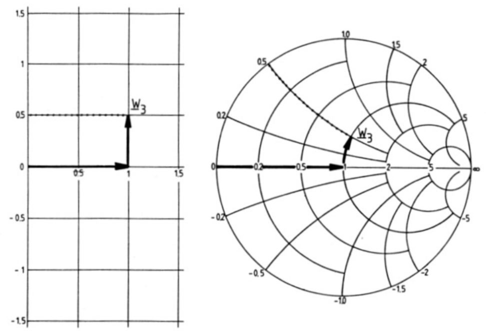

Fig 8: Series connection: Real component = 1.0; reactiev component = +0.5.

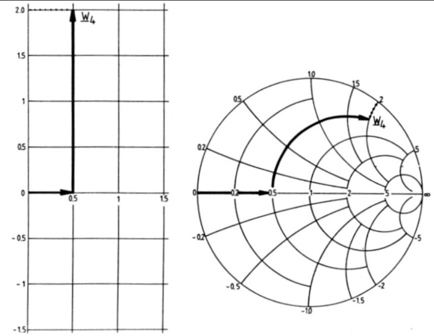

Fig 9: Series connection: Real component = 0.5; reactive component = +2.

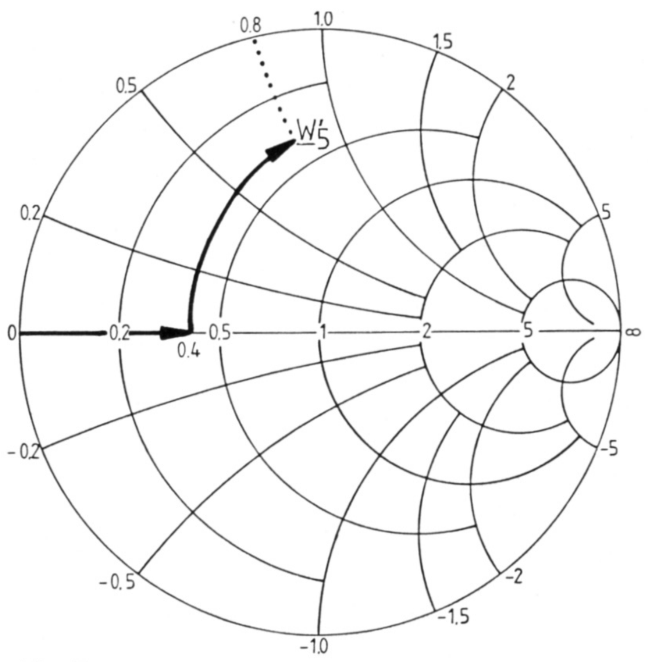

Fig 10: Real impedance = 20Ω; reactive impedance = 40Ω. Standard-ised: {Real component = 0.4; Reactive component = 0.8} = W'$_5$

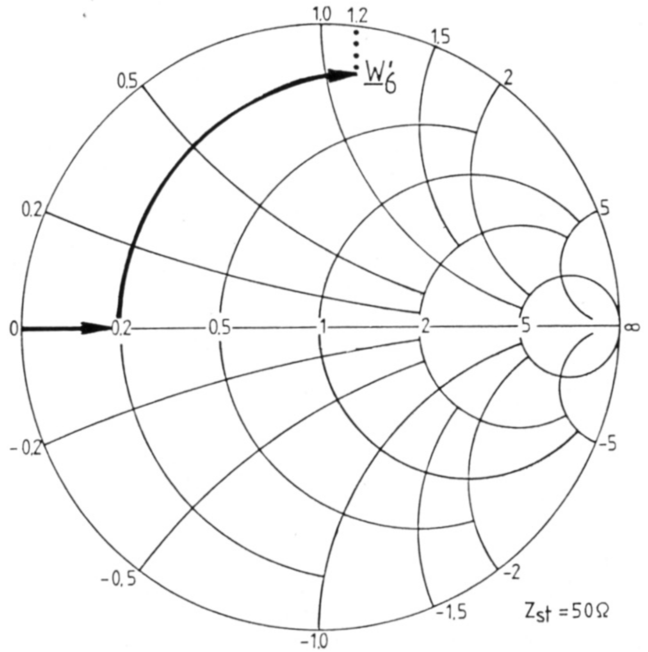

Fig 11: W'$_6$ is standardised: {Real component = 0.2; Reactive component = 1.2} After de-standardisation { Real component = 10Ω; Reactive component = 60Ω.

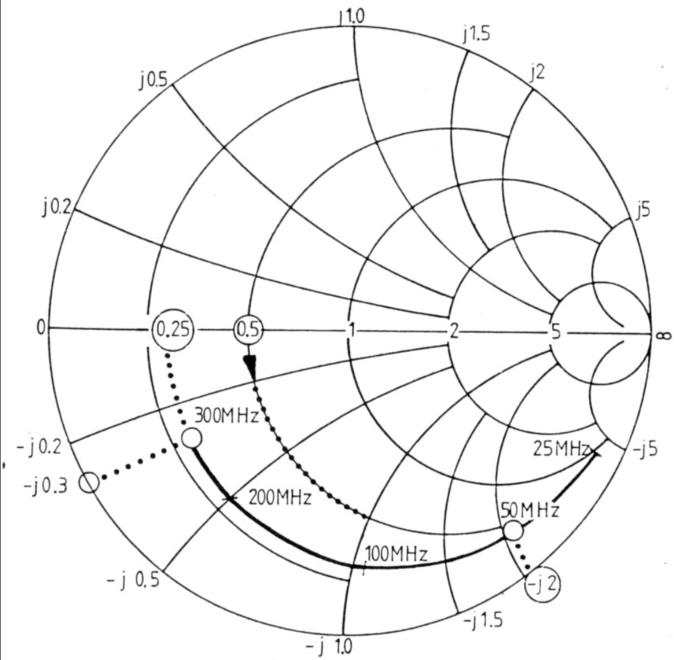

Fig 12: Determining the input impedance of a transistor with respect to real and reactive component at 50 and 300MHz.

direction, or where the originally horizontal coordinates reverse their direction. Several points are given in Fig 6 to 9 (W_1, W_2, W_3 and W_4). The associated reactive and real impedance components must be taken from the bent coordinates. W_4 has an extremely high reactive component, which means that this point has exceeded the vertex of the circle. The insertion and extraction of data in this area is very complicated. In this range in the vicinity of infinite values, the coordinates are very close together, which means that only a few can still be inserted. Intermediate values must be read off or inserted by interpolation.

On studying Figure 6 to 9 as well as 1 and 2, one could assume that only impedances in the order of 0.1Ω to approximately 2Ω are suitable for the diagram. This is, however, not the case in practice, since only standard values are inserted into the diagram. This standardisation allows the same diagram to be used universally.

The standardisation process is now to be shown with the aid of an example: A series circuit exhibiting a real impedance of 20Ω and a reactive impedance of 40Ω is to be inserted into the diagram. If standardisation were not used, it would be necessary to insert the value 20 or 40 virtually at ∞, which would be useless due to the inaccuracies. Standardisation on the other hand means that it is necessary to divide the given impedance values by a suitable, so-called standardising impedance Z_{st}.

In our case, we select a Z_{st} of 50Ω. This results in a standardised real component of 20Ω/50Ω = 0.4; and a standardised reactive component of 40Ω/50Ω = 0.8.

The standardised values 0.4 and 0.8 are inserted into the diagram (Fig 10), and it can be seen that the resulting point W'_5 is located very favourably. In our example, 0.4 and 0.8 are numerical values representing the type of an impedance, with 1 as unit.

If one reads such numerical values out of the diagram, one must know the value of the standardising impedance applied before the entry into the diagram. Another example: In Fig 11 one will see a point W'_6 which is to be found on the real impedance coordinate 0.2, and on the reactive impedance coordinate 1.2. Since the standardising impedance Z_{st} = 50Ω is marked on the diagram, one knows that these values are based on a standardised impedance of 50Ω. If one wishes to know the actual impedance values, it is necessary to “de-standardise” them.

This is carried out by multiplying them with the standardising impedance values: The resulting real impedance value is

$$Z = 0.2 \times 50\Omega = 10\Omega,$$

and the actual reactive impedance is

$$X = 1.2 \times 50\Omega = 60\Omega.$$

The values also receive their actual unit “ohm”.

In principle, the selection of the standardising impedance value can be made freely, however, in practice one is dealing with real and reactive impedances that are to be matched to a certain impedance, such as a source or load impedance. Another application would be to match the characteristic impedance of a transmit antenna to a certain impedance of the feeder cable. If one wishes to use the advantages of the diagram - that is, reading out the value of the return loss of the load or antenna impedance from the diagram, as a function of the source or load impedance - it will be necessary to standardise the diagram to the impedance of the source, that is the source impedance or the impedance of the line. Since an impedance of 50Ω is usually used in communication technology, it is favourable to select this impedance as standardising impedance. In Fig 1, an impedance Z_o = 50Ω was given by the manufacturer. This Z_o is the standardising impedance which we have designated Z_{st} in this article. Fundamentally speaking, it is always necessary to give the value of

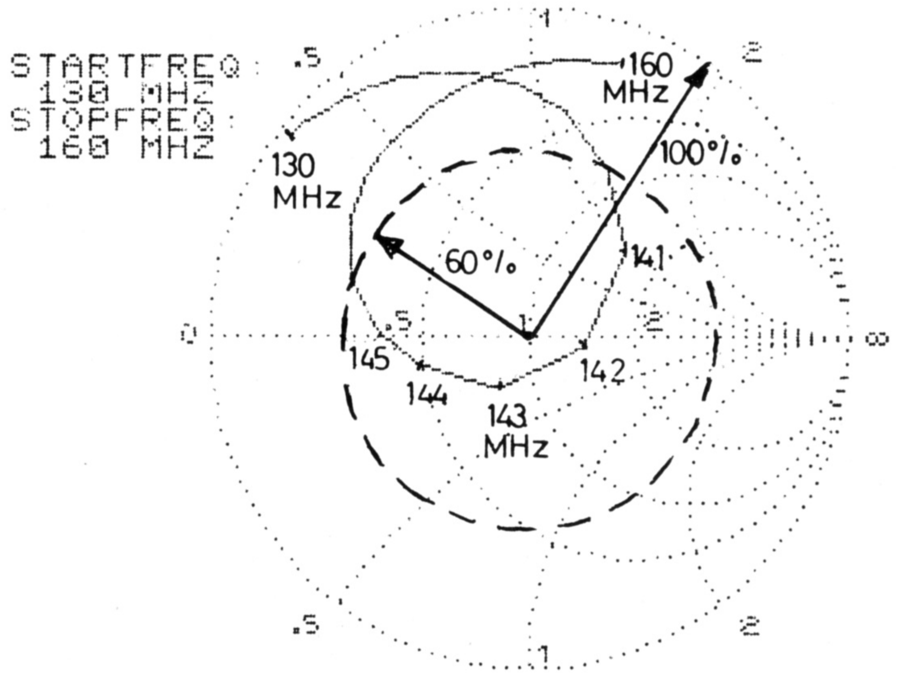

Fig 13: Locus of an antenna impedance: Return loss circle |r| = 60%.

the standardising impedance used when displaying impedances in a Smith diagram!

Finally, a further example is to demonstrate once again the reading of Smith diagrams and the de-standardisation process in conjunction with the RF transistor TDA1087 (Fig 1). This diagram is repeated in Fig 12. The question is:

- What is the impedance (real and reactive components) of the transistor at 50MHz and 300MHz?

Solution: The 50MHz point is to be found on the locus at a real component of "0.5", and a reactive component of approximately "-2" (the factor "j" only shows that it is a reactive and not a real component). The minus sign means that it is a capacitive reactive component.

After de-standardising, the following results at 50MHz:

- Real impedance R = 0.5 x 50Ω = 25Ω;
- Reactive impedance Xc _-2 x 50Ω = -100Ω ("-" means "capacitive").

The 300MHz point is to be found on the locus at a real component of "0.25" (estimated) and a reactive component of approx. "-0.3".

After de-standardisation, this results at 30MHz in the following:

- Real impedance R = 0.25 x 50Ω = 12.5Ω;
- Reactive imp. Xc = -0.3 x 50Ω = -15Ω.

The advantage of the return loss diagram is to be shown together with a HB9CV antenna. The locus of the impedance of this antenna (Fig 2) is repeated in Fig 13. The diagram is standardised to 50Ω and the antenna is to be connected to a feeder cable of also 50Ω, which means that it is not necessary to re-standardise the locus. It is now possible to simply determine in which range the antenna does not exceed a certain return loss! However, this requires one more curve in addition to the many other curves already existing in the diagram.

The return loss limits are indicated by the radial length of a circle whose centre is the geometric centre of the diagram. The radius of the outer circle is used as a reference value, since all impedance values on the outer circle possess a return loss of 100% with respect to the standardising impedance. If one wishes to mark the range in which the locus contains impedances having a certain return loss with respect to the standardising impedance of 50Ω, for instance a return loss for 60%, it is only necessary to form a circle around point 1 (centre of the diagram) which has a radius of 60% that of the outer circle.

All impedances that are to be found within this circle will have a return loss of less than 60% with respect to the standardising impedance, and thus with respect to the impedance of the feeder cable. It will be seen that the antenna is matched to the cable with a return loss of less than 60% in a frequency range of 141MHz to 145MHz. If the voltage standing wave ratio VSWR is to be read off, this is possible by reading it out of the diagram between 1 and ∞ at the intersection between the appropriate return loss curve and the horizontal axis.

It will be seen that one is able to solve many problems encountered with return loss, standing wave ratio, and matching, without having to carry out complex calculation

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Internet Treasure Trove

San Bernardino microwave society

This web site is for the GHz enthusiasts: a page with hours and hours of interesting projects to study with some really interesting things.

Address: http://www.ham-radio.com/sbms/sd/projindx.htm

A CMOS sub-harmonic mixer for WCDMA

Anyone who wants to learn more about the tub harmonic mixer should take a look at this pdf article from Berkeley University. It describes a circuit that converts a 2GHz RF signal directly to baseband using a 1GHz local oscillator.

Address:
http://RFIC.eecs.Berkeley.edu/files/scrose_thesis.pdf

Satcoms UK

The home page provides a detailed look at the transmission and communications via satellite by this company. The extensive "Satellite Communication Tutorials" are particularly interesting

Address: http://www.satcoms.org.UK/satellite/Forum/satellite-communication-tutorials_forums_cat3.html

Q-par Angus Ltd

A company that deals specifically with antennas and providing an incredible wealth of information for their customers. On the listed website has an historical perspective of wideband antennas.

Address: http://www.q-par.com/corporate/marketing/wideband-antennas-an-historical-perspective-poster.pdf/view

SETI

SETI means "search for extraterrestrial intelligence” and the organization has opened a new chapter. They are looking worldwide for interested people to help with their PCs to search the vast amounts of data available. Data comes from the radio telescopes distributed over all the World.

Address: http://www.seti.org/

Harmonic converters for Spectrum Analysers

An interesting article from this familiar author, Matjaz Vidmar. It is always a pleasure to read his articles and to see his ideas.

Address: http://www.s5tech.net/s53mv/spectana/hc.html

Image rejection receiver and techniques

There are several interesting publications for training on the Internet. These are recommended:

Address 1: http://EE.Sharif.edu/~comcir/readings/mixers/image%20reject.pdf

Address 2: http://www.eecg.utoronto.ca/~kphang/papers/2002/jchow_imagereject.pdf

Q factor measurement

Something for the RF Circuit developer because it is important to have good quality coils in resonant circiuits. These links are interesting.

Address: http://dsp-book.Narod.ru/MISH/CH52.PDF

The following sources are also good:

Address 1: http://www.sensorsportal.com/HTML/DIGEST/june_09/P_438.pdf

Address 2: http://www.ee.olemiss.edu/Darko/rfqmeas2b.pdf

LC filter calculator

As Windows 7 is becoming more and more common suddenly the beloved DOS programs no longer work. It is useful to switch to a modern "online filter Calculator". This is a short list of tested links:

Address 1: http://www-users.cs.York.AC.UK/~fisher/lcfilter/

Address 2: http://www.calculatoredge.com/electronics/ch%20pi%20low%20pass.htm

Address 3: http://www.calculatoredge.com/electronics/ch%20pi%20high%20pass.htm

Address 4: http://www.wa4dsy.NET/filter/hp_lp_filter.html

Address 5: http://www.odav-online.com/calculator-electrical-Butterworth-Pi-LC-Low-Pass-Filter-online-engineering-solutions.html

Address 6: http://www.circuitsage.com/filter_design

Note: Owing to the fact that Internet content changes very fast, it is not always possible to list the most recent developments. We therefore apologise for any inconvenience if Internet addresses are no longer accessible or have recently been altered by the operators in question.

We wish to point out that neither the compiler nor the publisher has any liability for the correctness of any details listed or for the contents of the sites referred to!

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

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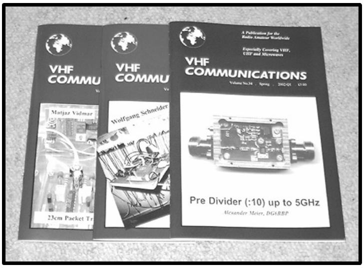
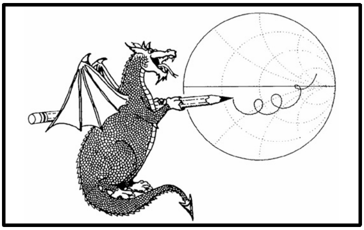


Back Issues

Available either as photocopies or actual magazines. Issues from 1/1969 to 4/2010 are £1.00 each + postage. Issues from 2011 are £5.35 each or £19.60 for all 4 issues + postage. See web site or page 34 of issue 1/2012 for back issue list. There are two back issue sets that contain the available "real" magazines at a reduced price, see web site for details.

Blue Binders

A new style binder was introduced in 2010 with an embossed spine. These binders hold 12 issues (3 years) and keep your library of VHF Communications neat and tidy. You will be able to find the issue that you want easily. Binders are £6.50 each + postage. (UK £2.20, Surface mail £3.25, Air mail to Europe £3.40, Air mail outside Europe £5.10)



PUFF Version 2.1 Microwave CAD Software

This software is used by many authors of articles in VHF Communications. It is supplied on 3.5 inch floppy disc or CD with a full English handbook. PUFF is £20.00 + postage. (UK £2.20, Surface mail £2.90, Air mail to Europe £3.20, Air mail outside Europe £4.50)

Compilation CDs

Two CDs containing compilations of VHF Communications magazine articles are available. CD-1 contains 21 articles on measuring techniques published in the past few years. CD-2 contains 32 articles on transmitters, receivers, amplifiers and ancillaries published over the past few years. The articles are in pdf format.

Each CD is £10.00 + postage. (UK £0.50, Surface mail £2.20, Air mail to Europe £2.70, Air mail outside Europe £3.30)

VHF Communications Web Site www.vhfcomm.co.uk

Visit the web site for more information on previous articles. There is a full index from 1969 to the present issue, it can be searched on line or downloaded to your own PC to search at your leisure. If you want to purchase back issues, kits or PUFF there is a secure order form or full details of how to contact us. The web site also contains a very useful list of site links, and downloads of some previous articles and supporting information.

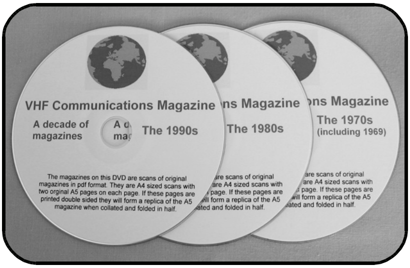
K M Publications, 503 Northdown Road, Margate, Kent, CT9 3HD, UK

Tel / Fax +44 (0) 1843 220080, Email: andy@vhfcomm.co.uk

Back Issues on DVD

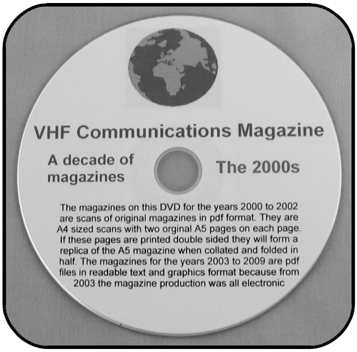
VHF Communications Magazine has been published since 1969. Up to 2002 it was produced by traditional printing methods. All these back issue have been scanned and converted to pdf files containing images of the A4 sheets that formed the A5 magazine when folded in half. These have been put together on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

From 2002 the magazine has been produced electronically therefore pdf files are available of the text and images. These have been used to produce the 2000s decade DVD.



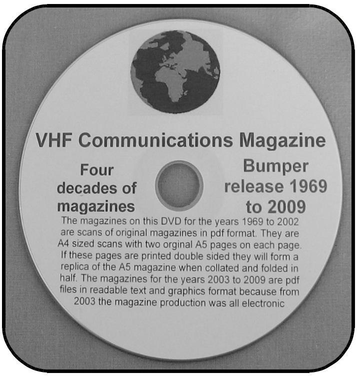
1970s - 1980s - 1990s

These three DVDs cover the first 3 decades of the magazine. The 1970s DVD contains all magazines from 1969 to 1979 (44 magazines) the 1980s and 1990s DVDs contain 40 magazines for the decade. The DVDs are £20.00 each + postage



2000s

This DVD contains magazines from 1/2000 to 4/2001 in scanned image format and from 1/2002 to 4/2009 in text and image format. This DVD is £35.00 + postage.



Bumper 4 decade DVD

This DVD contains all magazines from 1969 to 2009. That is 164 magazines. It also contains the full index for those 41 years in pdf and Excel format so that you can search for that illusive article easily. This DVD is just £85.00 (just 52 pence per magazine). + postage.

To order, use one of the following:

- Use the order form on the web site - www.vhfcomm.co.uk
- Send an order by fax or post stating the DVD required (1970s, 1980s, 1990s, 2000s, Bumper)
- Send the correct amount via PayPal - vhfcomms@aol.com - stating the DVD required (1970s, 1980s, 1990s, 2000s, Bumper) and your postal address

Postage: UK £0.50, Surface mail £2.20, Airmail in Europe £2.70, Airmail outside Europe £3.30