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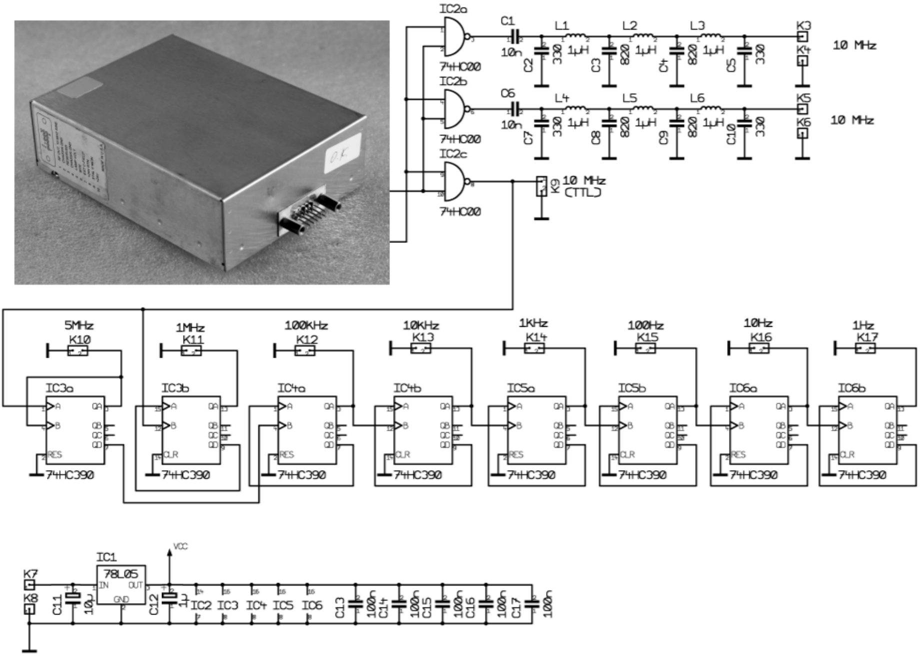

LPRO-101 Rubidium frequency standard with output driver for different frequencies

Wolfgang Schneider, DJ8ES

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I have managed to fill the pages of this issue. At the last minute I was contacted by Mike Scirocco and the article on his Parabola calculator was the result. It seemed a good chance to fill the remaining gap with a reprinted article about a dish feed.

The big issue for me is the postal price increase announced by Royal Mail. Please read the discussion document on page 66 and send me your thoughts. Because the increases took effect on 30th April 2012 I will be subsidising the postage for the rest of 2012 but the increase is so large that I cannot continue in the same way for 2013.

73s - Andy

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

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

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

Royal Mail Price Increases

On April 30th 2012 the Royal Mail postal prices rose dramatically. This will affect the future of VHF Communications Magazine because all magazines are posted to subscribers. I need feedback from subscribers to help me decide the future of the magazine. Please send your comments by email, post or via the special form on the web site. I will reports the outcome in issue 3/2012

The postal increases were:

Desination	Old price per issue	New price per issue	Annual increase	Actual annual price
UK second class post	£0.36	£0.50	£0.56	£2.00
Surface mail	£1.12	£2.20	£4.32	£8.80
Airmail to Europe	£1.49	£2.70	£4.84	£10.80
Airmail outside Europe	£2.07	£3.30	£4.92	£13.20

The worst affected will be surface mail, which is the largest proportion of the current subscribers. As a result of these price increases I think there are three possible outcomes:

1) The magazine could continue as a printed magazine posted to subscribers but the postal prices would have to increase to cover the true postal cost. This would mean a rise in the basic subscription from £21.60 to £22.20 to include the UK postage. This means that £2.00 of the subscription price covers postal costs. The additional prices for overseas postage would have to be:

- Surface mail £6.80
- Airmail to Europe £8.80
- Airmail outside Europe £11.20

I think these postal prices are very high and would result is a significant drop in the circulation that is already at a very low level.

2) The magazine could become electronically delivered only. This would cut out the postal costs and printing costs but I would still have to pay for:

- Royalties to UKW Berichte
- Translation of German articles
- Sourcing non UKW Berichte articles
- Production of the magazine
- Running the web site
- Bank Charges
- Credit card charges

To cover these costs the subscription price for an all electronically delivered magazine would have to be £14.00.

I think that electronic delivery only would prevent quite a few subscribers from being able to receive the magazine, which is a great shame. I have based the above price on an expected loss of about 25% of the current subscribers if this change took place.

3) I could stop publishing the magazine. I would try to find someone to take over the magazine in some form but this would not be an easy task and may not result in the magazine continuing.

Please send me your feedback to help me make this fundamental change to the magazine.

Wolfgang Schneider, DJ8ES

LPRO-101 Rubidium frequency standard with output driver for different frequencies

Rubidium frequency standards from the GSM systems have been available on the surplus market for quite some time, suitable for amateur radio applications, at extremely interesting prices. This highly precise frequency standard can be used for such things as: synchronising frequency counters, signal generators and transverters for the UHF/SHF range. These components usually supply a sinusoidal output signal with a level between 5mW and 10mW on the standard frequency of 10MHz. For universal use an additionally driver circuit with several outputs on different frequencies is needed.

1. Introduction

For many radio amateurs it has been a serious problem to produce really accurate frequencies for measuring purposes or as local oscillators for transverters in the GHz range. For a long time OCXOs (Oven control crystal oscillators) have been used for these applications. However all crystal oscillators have a fundamental deficiency; they age and the frequency varies from the nominal frequency, this increases more and more with time. Regular calibration of the output frequency is the only solution. Synchronising to highly precise (and available) reference frequency can overcome this effect. There are many articles in amateur radio literature about frequency synchronisation to DCF77 or GPS.

2. The Efratom LPRO-101 rubidium frequency standard

The Efratom LPRO-101 Rubidium frequency standard (Fig 1) is a very precise frequency standard used in GSM systems. The main technical characteristics are:

- 10MHz sine wave output
- Output level 0.55Vrms (approximately 5mW into 50Ω)
- Frequency accuracy; <1x 10e^{-10} (1Hz at 10GHz)
- Ageing rate; <5x 10e^{-11}/month
- Operating voltage 24V DC, tolerance range 19 to 32V
- Power requirement approximately 10W (after heating)

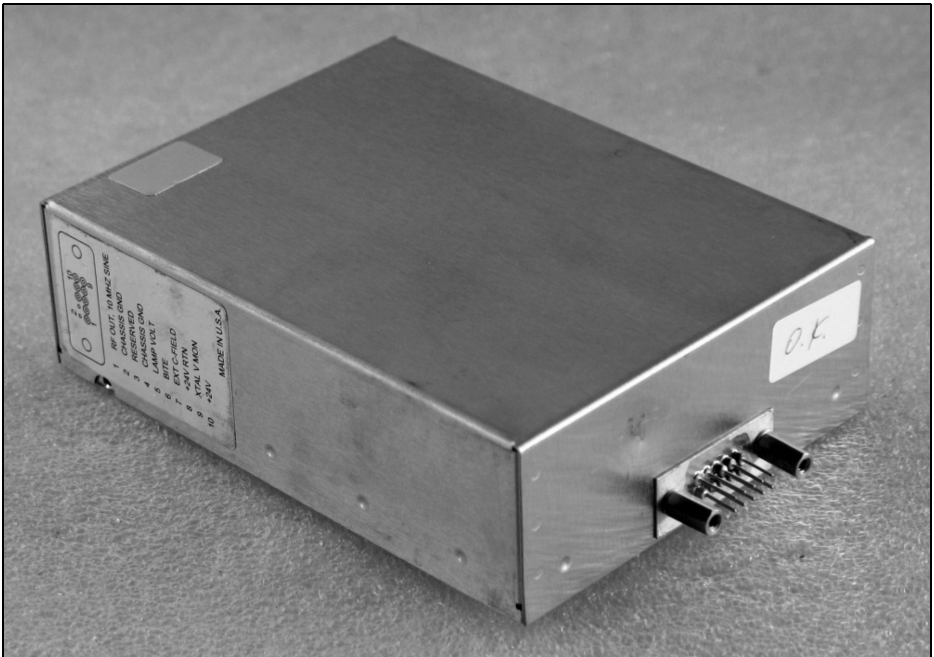
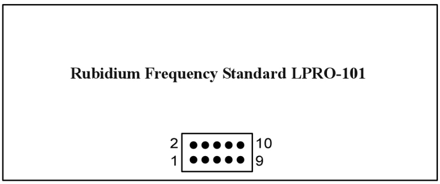


Fig 1: The LPRO-101 Rubidium Frequency Standard module.

- Signal output with Lock display
- Dimensions: 138mm x 95mm x 38mm
- Weight 0.47 kg



Pin	Signal Name
1	Rf Out
2	GND
3	GND
4	GND
5	Lamp Volt
6	BITE Out
7	Ext. C-Field
8	+24V RTN
9	Crystal Volt Monitor
10	+24 V Power

Fig 2: Pin connections for the 10 way connector on the LPRO-101.

This handy module can be built into finished devices or as described in this article it can be used in a universal frequency standard with a power supply and driver circuits for different frequencies from 10MHz down to 1Hz. The rubidium frequency standard requires an operating voltage of 24V. According to data sheet a supply voltage ranging from 19V to a maximum of 32V DC volts can be used. The power requirement is about 10W after the warming up phase. This corresponds to a current of approximately 0.4A. Following switching on the maximum is 1.5A falling slowly as the temperature rises. Depending on the ambient temperature this can last between 3 and 5 minutes until the oscillator locks at its nominal frequency of 10MHz. The BITE (Build in test equipment) output indicates the status with a High/Low signal. The rubidium frequency standard module uses a 10 way connector; the connections are shown in Fig 2. The

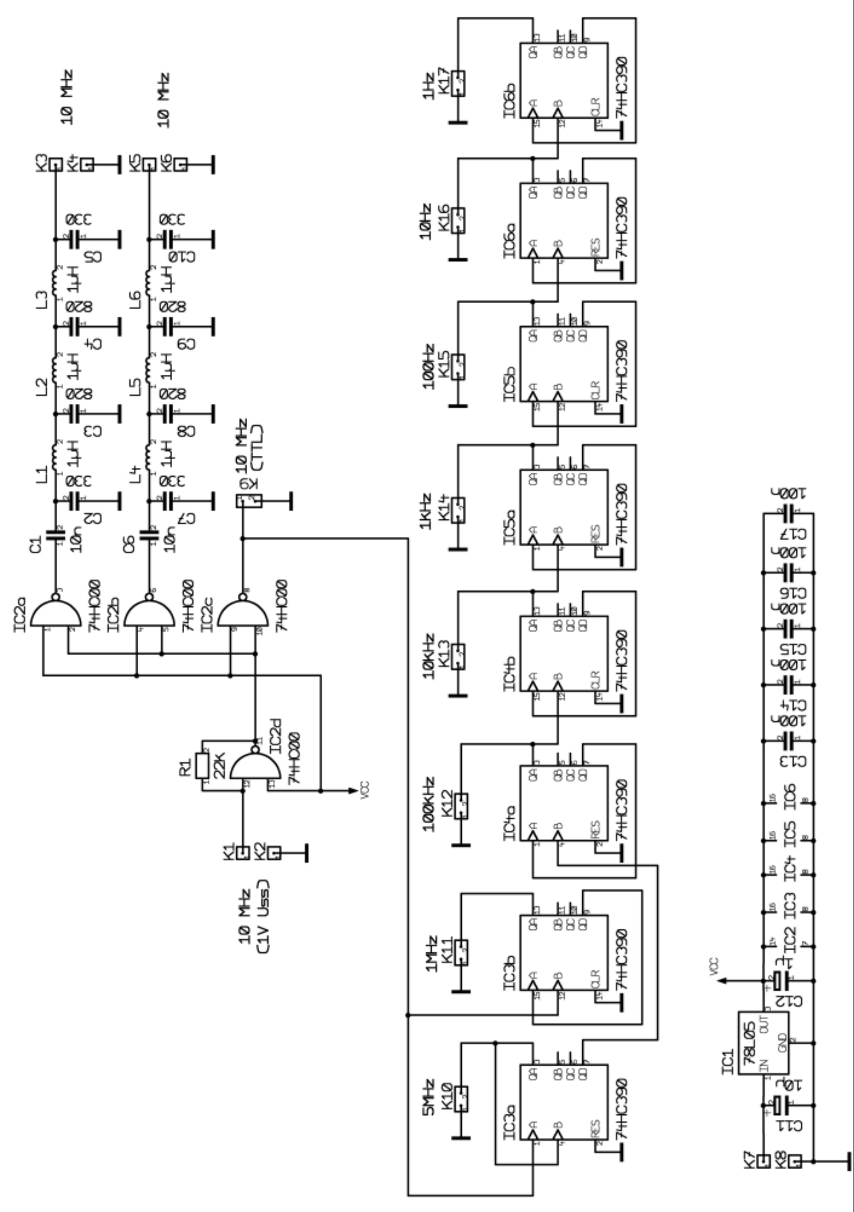


Fig 3: Circuit diagram for the output driver for different frequencies.

frequency accuracy is better than 1x $10e^{-10}$ of the nominal frequency, or if used to synchronise a 10GHz transverter it's output frequency will be accurate to within 1Hz. For those who would like look more deeply into this subject, the publication [1] covers the subject in more detail. The two authors describe the basic function of a rubidium frequency standard in great detail. They deal with the topic of frequency accuracy and relevant measuring methods.

3.

Output driver for different frequencies

The input gate of IC2 (74HC00) works like a schmitt trigger. It changes the approximately 5mW 10MHz sine wave signal to TTL levels. The input pin is connected to the resistor R1 (22 kΩ) to give over half a TTL level. The output of the gate is a clean TTL signal available for the further processing. Three further gates from IC2 serve as drivers. One supplies a TTL level the other two feed 7 pole low pass filters to produce 10MHz sine wave signals at approximately 10mW (+10dBm) to feed external circuits for synchronisation of devices like oscillators, frequency counters or synthesisers. IC3 to IC6 (74HC390 decade counters) produce further frequencies of 5MHz, 1MHz, etc. down to 1Hz from the 10MHz TTL signal. These counters are internally constructed from a divide by 2 counter and a divide by 5 counter and by appropriate interconnection they produce the desired output frequencies. All output frequencies are fed to 2 pin connectors for connection to output sockets using ribbon cable or thin coax cable to BNC sockets. The circuit diagram for the output driver module is shown in Fig 3. It should be operated with a voltage from 9 to 14 V. The voltage regulator IC1 (78L05) converts this to the 5V required.

Note: With higher input voltages, e.g. 24V and depending on the power input of the module the maximum power dissipation of the voltage regulator could be exceeded.

3.1. Parts list for DJ8ES 082:

IC1	78L05, TO-92, 5V regulator
IC2	74HC00, DIL14, nand gate
IC3 - IC6	74HC390, DIL16, counter
R1	22kΩ, SMD 1206
L1 - L6	1µH, SMD inductance, 1806
C11	10µF, SMD electrolytic, 1806
C12	1µF, SMD electrolytic,1806

Ceramic capacitors, SMD, 1206:

4 x	330pF
4 x	820pF
2 x	10nF
5 x	100nF
1 x	DJ8ES 082 double sided PCB, 60mm x 100mm
8 x	1mm solder in pins
9 x	2-pin in line connectors

3.2. Construction

The output driver for different frequencies (DJ8ES 082) is built on a 60mm x 100mm double sided PCB (Fig 4). The lower surface of the PCB (Fig 5) serves as ground surface, the few tracks feed the supply to the TTL ICs (Nand gates and decade counters). It is plated through in various places. All components are soldered on the top side of the PCB according to the component layout shown in Fig 6. All capacitors, resistors and inductors are SMD types, the remaining components are wired. The SMD components should be soldered on first followed by the connections and plug connectors and last the ICs.

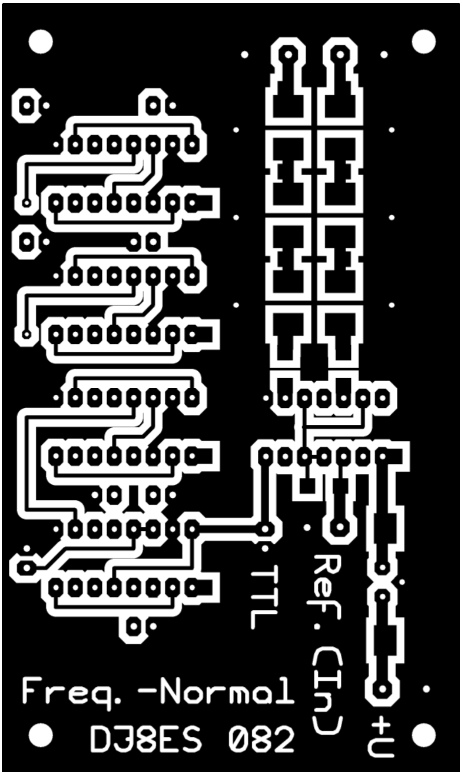


Fig 4: Top side PCB layout for the driver module.

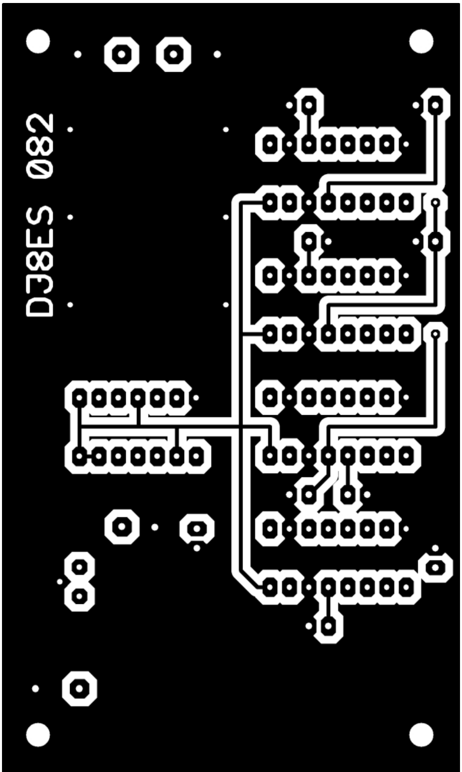


Fig 5: Bottom side PCB layout for the driver module.

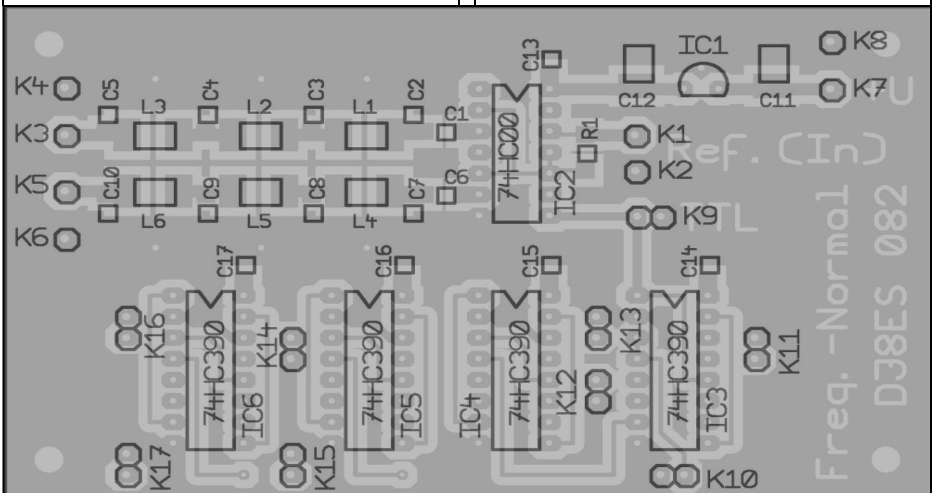


Fig 6: Component layout for the driver module.

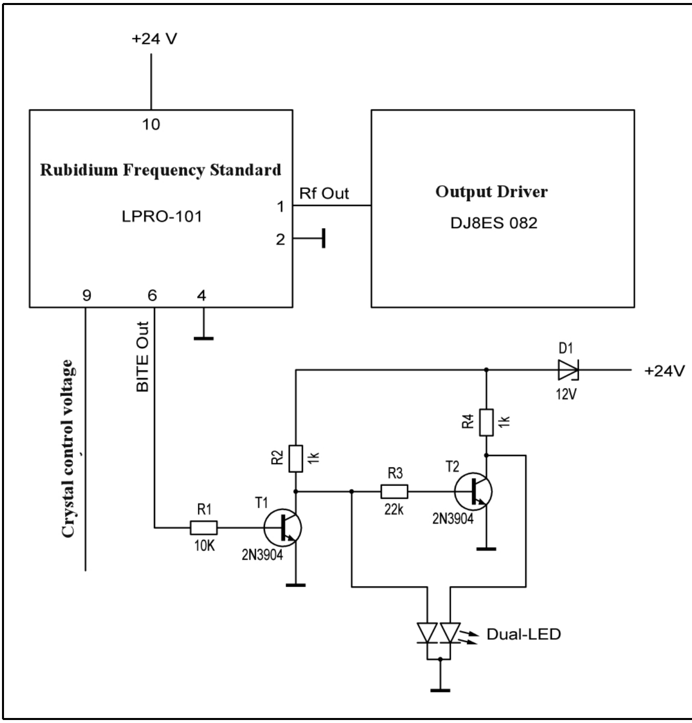


Fig 7: Circuit diagram for a Frequency Lock display from KA7OEI.

4.

Connection and testing

The output driver for different frequencies (DJ8ES 082) is connected directly to the RF output of the LPRO-101 rubidium frequency standard. With a supply voltage of 12V the current input to approximately 70mA. The individual outputs of the module can be measured with a frequency counter or oscilloscope. The accuracy is defined by frequency standard. The rubidium frequency standard requires a supply voltage of 24V but the technical data permits a certain tolerance. After the heating phase, that is dependant on the ambient temperature, of between 3 and 5 minutes the current input is approximately 0.4A. Directly after switch on this is a maximum of 1.5A. The control voltage for the internal crystal oscillator is fed out for control purposes on pin 9 of the module. During the warm up phase this varies in a saw tooth waveform between 0 and 12V that can be monitored with a suitable meter. During the second phase the oscillator locks onto it's exact frequency of 10.0MHz. This lock status is indicated by the BITE output (pin 6, TTL levels). A high level corresponds to the Unlock state and a low level to the Locked state. Fig 7 shows a simple but interesting circuit by Clint Turner, KA7OEI, that operates an LED to display the Locked state. A description of this circuit is shown on the Internet in [2].

5.

Proposed addition

It is useful to have multiple 10MHz

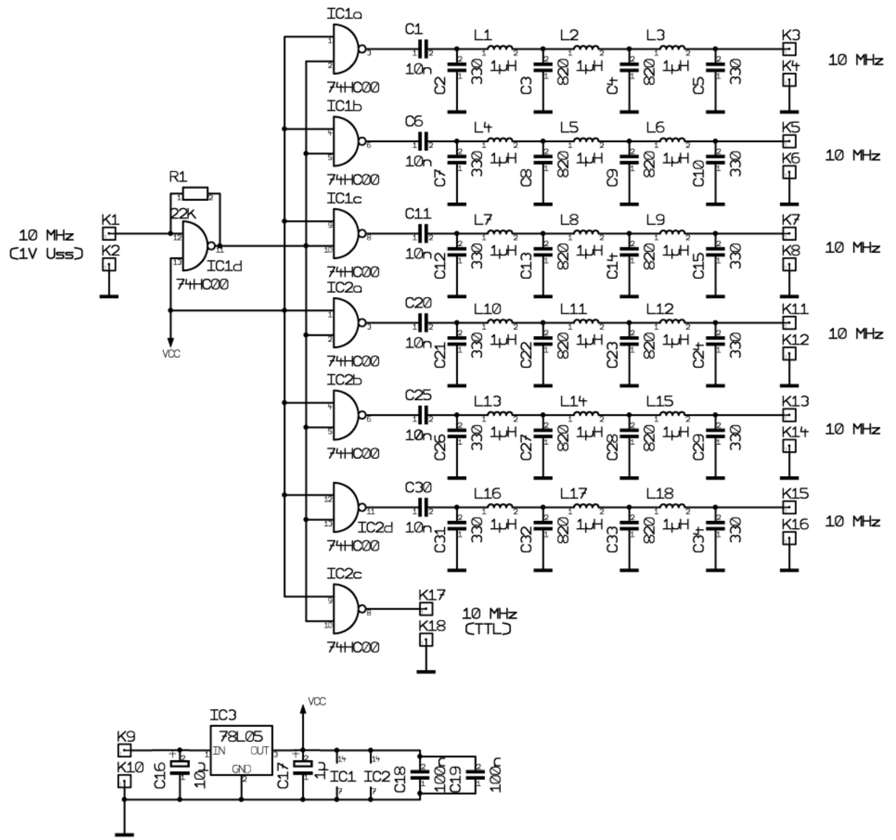


Fig 8: Circuit diagram of the additional module for multiple 10MHz outputs.

outputs particularly for synchronising measuring instruments or transverters at the same time. A circuit with six 10MHz outputs of approximately +10dBm is shown in Fig 8. Unlike the previous circuit for different frequencies a 74HC00 driver replaces the decade counters. Each output has its own 7 pole low-pass filter to suppress the higher harmonics. It is important that unused outputs are terminated with 50Ω; this prevents a signal reflection and distortion at the active outputs. This module is built on a 100mm x 60mm PCB. Fig 9 shows the layout for the top side of the PCB. Fig 10 shows the layout for the bottom side of the PCB. Fig 11 shows the component layout for the six 10MHz output PCB. No detailed description is required to set up this module since it is similar to the previous module.

Fig 12 shows a photograph of the prototype. The prototype was built before the six 10MHz output PCB was designed.

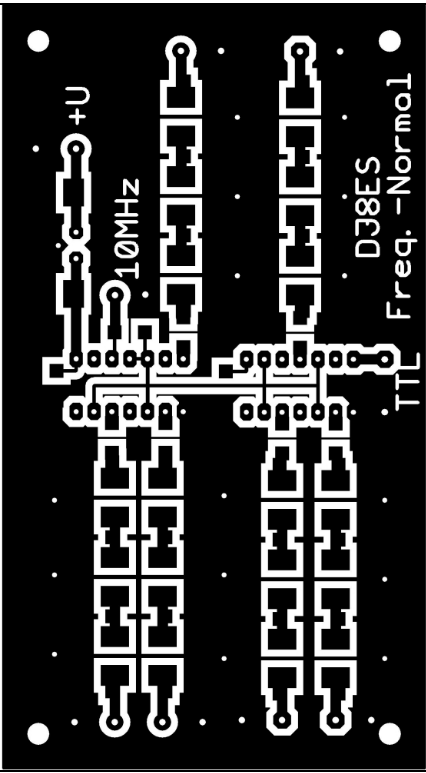


Fig 9: The layout for the top side of the six 10MHz output PCB.

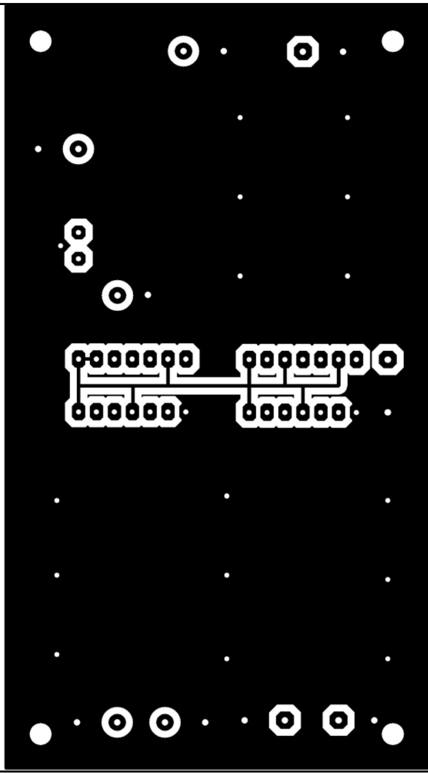


Fig 10: The layout for the bottom side of the six 10MHz output PCB.

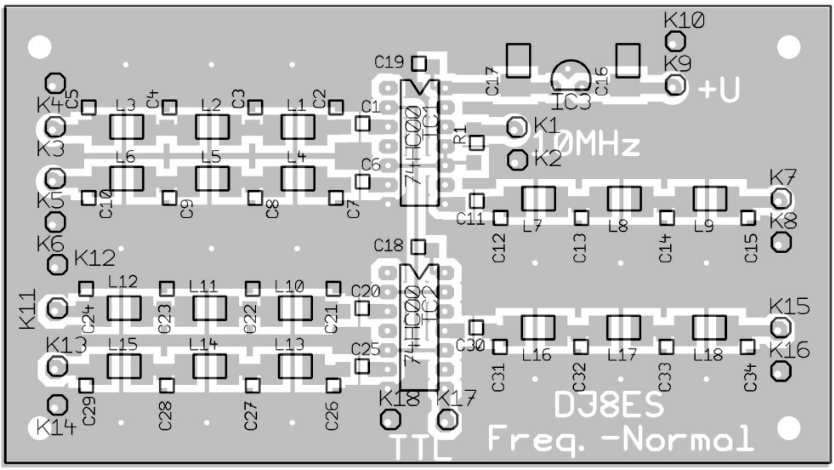


Fig 11: The component layout for the six 10MHz output PCB.

Fig 12: Photograph of the completed frequency reference.

6. References

[1] The Surplus 10MHz rubidium frequency standard LPRO-101 by Datum/Efratom in the Ham Shack, Hans, DL2MDQ, and Jürgen, DD6UJS, 20.10.09, www.openhpsdr.org

[2] A portable 10MHz rubidium frequency reference, Clint Turner, KA7OEI, www.ka7oei.com

[3] LPRO-101 Rubidum Frequency standard LPRO-101 datasheet, Symmetricom 2003, www.tenmhz.com

André Jamet, F9HX

The harmful effects of image frequency in microwave transverters

1. Introduction

For the most common microwave bands such as 5.7, 10, 24 and 47GHz, the equipment used includes a transverter SHF/VHF or UHF, followed by a 144 or 432MHz transceiver.

Transverters, especially for the highest bands, may have their performance degraded as a result of the adverse effects of the image frequency in the conversion that they perform.

1.1 Reminder of the fundamentals

Any device adds its own input signal noise and degrades the signal/noise ratio. A noise factor F is defined as:

$$F = \frac{(S/N)_{input}}{(S/N)_{output}}$$

(expressed in a linear manner, and not in dB)

Where: S/N are, respectively, the signal-to-noise ratio at the input and the ratio at the output device.

For an active device, we have:

$$F = \frac{N_{out}}{G \cdot N_{input}}$$

Where: G is the device gain.

The noise figure is routinely used that is the expression of noise factor in decibels:

$NF = 10 \log_{10} (F)$

1.2 The mixer

An SHF transverter uses a mixer to operate at a higher frequency than the associated VHF/UHF transceiver. A mixer produces two frequencies equal to the sum and the difference between a local oscillator (LO) and one wanted frequency.

In a whisper:

A radio mixer might be more accurately called "frequency mixer" to distinguish it from devices such as a console for audio signal mixing. It is the passage of two signals in a non-linear circuit that performs a mathematical operation more complex than a simple addition. It multiplies the amplitudes together to produce the sum and difference of input frequencies.

The mixed frequency (called intermediate frequency) is:

$f_{if} = f_h - f_{lo} = f_l + f_{lo}$

with :

f_{if} = intermediate frequency

f_h = high input frequency

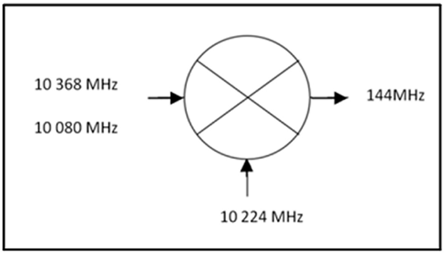


Fig 1: 3cm band receiver.

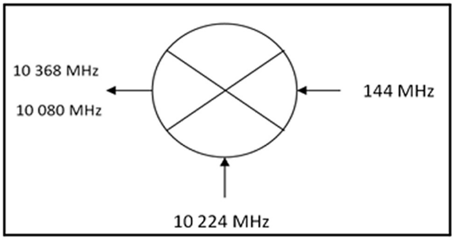


Fig 2: 3cm band transmitter.

f_l = low input frequency

f_{lo} = local oscillator frequency

From these two (upper and lower) input frequencies, one is useful and the other is undesirable and called image frequency. Fig 1 shows an example for a 3 cm band receiver.

Conversely, for transmitting, a given intermediate frequency produces two frequencies in the mixer where one is undesirable. Fig 2 gives an example of a 3 cm band transmitter.

In fact, there are many other frequencies due to the imperfections of the mixer and harmonics of the LO. If no provision is made, any unwanted frequency will cause defects, both for reception and transmission.

Fig 3 shows an example of an SBL1 mixer output spectrum: input frequency 14MHz, LO 10MHz, desired frequency 4MHz, 24MHz image. There is also the LO, input and combinations due to the LO harmonics.

2. Mixers

In the panoply of the many types of mixers two large families can be distinguished, passive diode mixers that introduce a conversion loss, and active mixers that can bring gain.

In the first category, there are simple mixers and balanced mixers giving cleaner signals, such as the popular SBL1. They are ideal up to 1GHz with interesting medium performance. For microwaves mixers diodes can be inserted on microstrips.

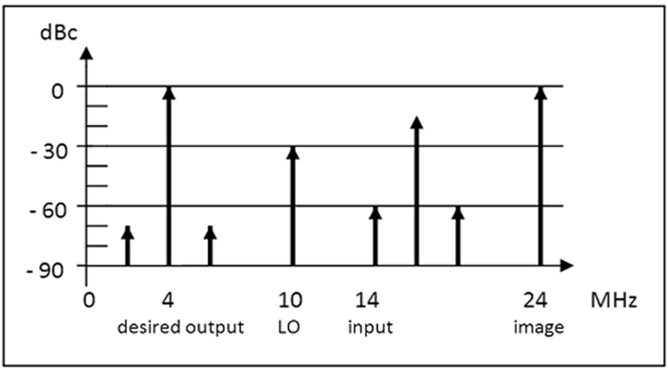


Fig 3: Sample of a mixer output.

Fig 4: Amplifying system.

The second category, with a substantial conversion gain, includes dual gate FET transistors such as the BF998 that are usable at VHF/UHF and GaAsFET for microwaves. Gilbert cells are a symmetric variant providing the same benefits as balanced diode mixers.

2.1 Transmitter defects

The mixer can produce unwanted frequencies. In addition to the frequency image and those due to the above imperfections, parasitic coupling between the ports of the mixer can let the LO frequency pass.

The emissions on frequencies not allocated to the amateur are prohibited. This can remain hidden if there is no interference to users; but its power is subtracted from that produced by the PA and it is lost.

Fortunately a transmitter usually includes signal amplifiers so selective circuits filter the signal. The LO and image frequencies attenuation is often sufficient. However, the addition of filters may be necessary to eliminate any discrete frequency that remain. Indeed, our regulations specify the relative level of spurious permissable.

2.2 Defects at the receiver

First of all, a principle to keep in mind:

Principle N°1:

a receiving system noise factor is determined solely by one amplifier (LNA) input and the preceding antenna and coaxial cable, if it's gain is sufficient to hide the noise of the stages that follow.

Fig 4 shows a system with a very low noise preamplifier, followed of a low noise amplifier stage and, finally, a mixer.

The formula for obtaining this system noise factor is as follows:

$$F_{system} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + ...$$

F_{system} = system noise factor

$F_{1,2,3,n}$ = noise factor of each stage

$G_{1,2,3,n}$ = gain of each stage

All these noise and gain values are linear, not in decibels. They include potential losses between stages.

Table 1 shows an example of two 10GHz/144MHz transverters. They involve a two stage preamplifier (LNA), an amplifier and an active transistors GaAsFET mixer.

From the calculations the noise factor of the system used since 2007 the preamplifier noise factor is not degraded.

But, is it true? We will see …

Table 1: Example of a 10GHz/144MHz transverter.

	10GHz preamp		10GHz/144MHz transverter		Performance at 10GHz
In operation	NF_{1+2}	G_{1+2}	NF_3	G_3	$NF_{cascade}$
1997	2dB	10dB	2dB	10dB	≈4.3dB
2007	0.76dB	23.7dB	2dB	23dB	≈0.76dB

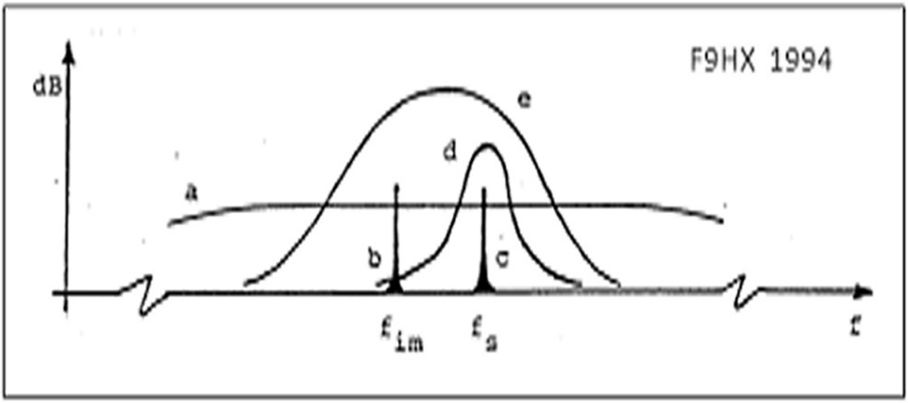


Fig 5: Noise and selectivity.

Principle N°2:

change in frequency by a mixer fails to, by itself, make a difference between the signal at the useful frequency and any signal that can produce a signal at the intermediate frequency at the output of the mixer.

Any unwanted frequency at the mixer input is taken into account to produce the intermediate frequency, such as the useful frequency, if no circuit attenuation occurs. Because the noise before the mixer is almost constant in the range of frequencies including signal and image, if there is no selectivity at the mixer input, the noise power is therefore "used twice" as shown in Fig 5. We can see:

(a) noise before the mixer

(b) image frequency noise

(c) signal frequency noise

(d) proper selectivity curve at mixer input

(e) insufficient selectivity curve

This is why this feature is particularly "vicious" because the input selectivity of the system before the mixer will affect the system noise. Only the selectivity at the mixer input can attenuate the noise produced at the image frequency of the system.

To avoid this increase in noise due to the image frequency it must be prevented from reaching the mixer or it must be neutralised. A narrow filter before the mixer meets the first requirement. An image frequency suppression mixer is the second. The latter requires an LO with two IQ outputs. The best solution will be chosen depending on the operating frequency.

This failure is very common if the gap between the signal frequency and the image frequency is low. Microwave transverters do not always use an intermediate frequency high enough and the image frequency is too close to the signal. Filtering is essential to obtain the minimum noise.

An important note: this frequency image filter should not make a signal loss. Otherwise, this loss would obviously correspond to an equivalent increase in noise.

If you stop there, it seems that without the image frequency noise attenuation applied to the mixer, the noise will be doubled at the output, therefore the factor is increased by 3dB. But this is not as simple as we shall see.

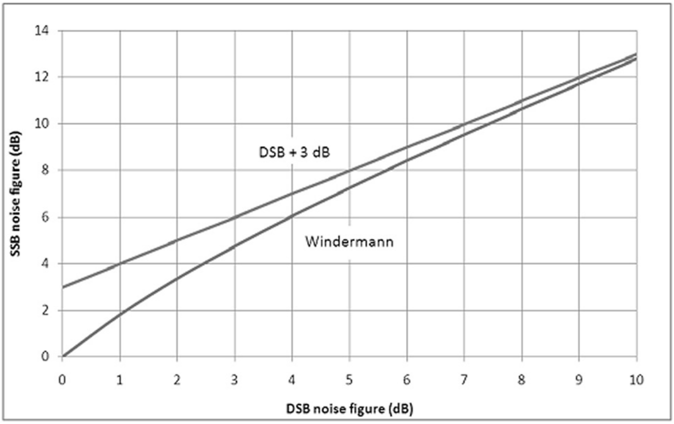


Fig 6: True SSB figure and DSB +3dB versus DSB figure.

3.

DSB and SSB noise factors

In amplitude modulation, AM, the signal includes a carrier and two sidebands. It is common to use the term DSB, Double Side Band, for this mode and SSB for single sideband mode.

Similarly, using a mixer, we can distinguish two factors of noise, DSB that takes into account the signal frequency and the image frequency and SSB that only takes into account the signal frequency.

SSB and DSB were used in the radar range equation calculations. A mixer receives the signal after amplification. When measuring the output of the mixer there is a power that is proportional to the signal associated with its noise, but in addition, the same noise of the image frequency. The result is that it is optimistic because the noise of the image made us believe that the noise of the signal was half of the real noise! Therefore, the rule was to add 3dB to the measured noise for the true signal/noise ratio at the input.

Fortunately if the image frequency is eliminated the measurement is exact. This is the case for the PANFI (Precision Automatic Noise Figure Indicator) with a YIG filter or a mixer able to eliminate image frequency.

This method of calculation is found in almost all the majority of the articles and old books on the image frequency influence on the signal/noise ratio. The IEEE (Institute of Electrical and Electronics Engineers) provides definitions that are sufficiently obscure so that prominent experts make completely opposite interpretations.

At the outset, it is easy to refute this - Actual +3dB - by considering the case of a signal without noise attacking a mixer without noise. It is clear that there is no noise at the output and a 3dB noise factor cannot be imposed to input while it is zero by hypothesis!

Wedermann [3] gave a solution, confirmed by other publications [4,5,6] that shows the increase depends on the initial noise factor. The increase is gradually asymptotically from 0 to 3dB when the initial noise factor is stronger.

Here is the demonstration of this formula. First of all, recall the relative to the active device formula and develop it:

Table 2: Increase in overall noise by the image frequency factor.

In operation	system	Performance at 10GHz system + 3dB	system + Windermann
1997	≈4.3dB	≈7.3dB	≈6.4dB
2007	≈0.76dB	≈3.76dB	≈1.41dB

$$F = \frac{N_{out}}{G \cdot N_{in}} = 1 + \frac{N_{out} - G \cdot N_{in}}{G \cdot N_{in}} = 1 + K$$

K is the fractional increase in output noise relative to input noise.

As the image frequency response makes a noise that can be considered equal to that of the signal, the increase is therefore 2 K.

Then:

$F_{DSB} = 1 + K$ and $F_{SSB} = 1 + 2K$

These two noise factors are related by:

$F_{SSB} = 2F_{DSB} - 1$

Expressed in decibels:

$NF_{SSB} = 10 \log [(2 \times 10^{NFDSB/10}) - 1]$

This formula is reproduced by Fig 6.

3.1 Our case study

We do not seek to calculate SSB noise knowing a DSB noise. We know what noise is issued by the system amplifying that applied to the mixer. We want to know the output noise, given the image frequency action.

With the assumption of +3dB, it would be impossible to obtain less than 3dB noise figure! But, If we consider that the most sensitive receivers, for example for radio astronomy, using cryogenic LNA (nitrogen or liquid helium at -269°C), reaching a factor of 0.07dB noise figure at 10GHz, you can be certain that effective noise factor is not increased by 3dB at the output of the mixer of these receivers which show actually a sensitivity that the calculation of noise factor has provided.

In this case, the increase in the noise factor is defined by a reciprocal formula from that established by Windermann.

Not to be penalised by the ambiguous terms SSB and DSB, we will talk about mixer input noise factor and mixer output noise factor. As already shown above, if the gain between the LNA input and the input mixer is 30dB, the mixer noise factor does not make a difference

It was then:

$F_{in} = 1 + K$

Thus: $K = F_{in} - 1$

$F_{out} = 1 + K + K = 1 + 2K$

$F_{out} = 2F_{in} - 1$

Expressed in decibels:

$NF_{out} = 10 \log [(2 \times 10^{-NFin/10}) - 1]$

Taking the above example with the display of performance calculated showing the increase in the overall noise by the image frequency factor according to the two studied methods is shown in Table 2.

3.2 And if the image frequency is reduced?

In practice, the image frequency attenuation can be total. It is therefore interesting to know the output noise versus the image frequency attenuation. For a long time the curve shown in Fig 7 has been proposed. It is wrong because it corresponds to the application of the principle +3dB excluded above.

From the Windermann formula, we have:

$F_{out} = 1 + K + K/\alpha$

with: $K = F_e - 1$

Then:

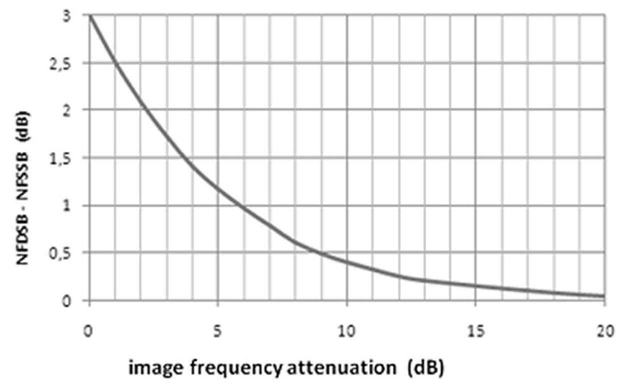


Fig 7: Erroneous curve.

$F_{out} = 1 + (F_{in} - 1) + (F_e - 1) / \alpha =$
$1 + [\alpha (F_{in} - 1) + (F_{in} - 1)] / \alpha$
then:

$$F_{out} = 1 + \frac{((\alpha + 1) \times (F_{in}) - 1)}{\alpha}$$

With:

F_{output} the signal noise factor from the output of the mixer and F_{input} that of the system.

In decibels:

$NF_{out} = 10\log[(\alpha/\alpha + 1) \times 10^{NFin-1/10}]) + 1]$

The curves shown in Fig 8 give the output noise obtained for various image frequency attenuations and the system output noise (also see similar data curves for noise increase instead of out noise [4]).

It is interesting to note that an attenuation of 10dB is enough so that the effect of the image frequency can be neglected for lowest noise.

Table 3 shows the same two examples and perform calculations show the influence of the image frequency attenuation.

4.

Measurements on a bench

Noise measurement benches, such as

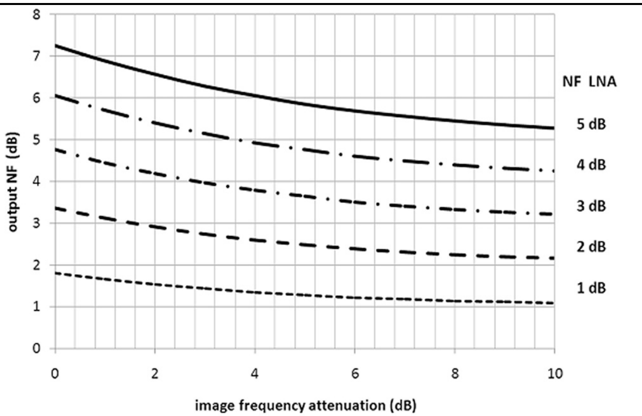


Fig 8: NF_{out} vs NF_{LNA} for various image frequency attenuations.

Table 3: The examples showing the effect of image frequency attenuation.

In operation	Total attenuation	Performance at 10GHz α= 10dB attenuation	α= 3dB attenuation	attenuation = 0
1977	≈4.3dB	≈4.5dB	≈5.5dB	≈6.4dB
2007	≈0.76dB	≈0.83dB	≈1.1dB	≈1.41dB

PANFI, have a mixer producing a signal at about 30MHz, regardless the operating frequency of the device in test (DUT). This mixer introduced a systematic error if it is sensitive to the image frequency and if it is present in the noise created by the DUT. Some have a correction to get the SSB noise taking account of the level of measured DSB noise and their internal frequency image attenuation.

5. Conclusion

Reading of this article, some might think that worrying about tenths of decibels is pointless. However those working on microwave bands are well aware that the success of a link depends closely on the chain of elements in the system. Each gain or loss is significant: see [10].

It is clear that the image frequency can degrade the transverter performance. This is especially true for microwaves because the frequency change is often done with an intermediate frequency too low to obtain sufficient attenuation of the image frequency. A number of stations at 24 and 47GHz, and of course beyond, may be affected by this defect that penalises the real sensitivity at the receiver.

6. References

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[4] Effect of an interstage filter on the sensitivity of analogue radio receivers, Nils Nazoa & Paul Heleine, Microwave Engineering Europe, 8/9 1996. The page 45 formula contains a typo, should read:

$$\frac{P_{o\gamma}}{P_o} = \frac{1+(\alpha_m+1)\times(F-1)}{F}$$

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Gunthard Kraus, DG8GB

An Interesting Program: Development of a Circularly Polarised Patch Antenna for 2.45GHz with Sonnet Lite

This article is the revised and extended version of a lecture to the 2011 VHF conference in Bensheim. The article is an "Interesting Program" dealing with the basics of Patch antennas and then the production of an antenna for circularly polarised waves.

The free software "Sonnet Lite" was used to develop the antenna. To verify the characteristics of the antenna a prototype was produced to compare it with the simulation.

1. The project

The 2.45GHz band is familiar not only for the WLAN allocation but also for other uses for approved low power transmission. A friend who is an enthusiastic model flier complained about the uncertain and varying 2.45GHz connection to his aeroplane. An examination of the expensive remote control transmitter showed that it is fitted with an SMA connector but only a simple small tilting blade antenna. The blade antenna could be tilted and pointed at the aeroplane. Those of us who know about antennas realise that unfortunately the minimum radiation occurs exactly where the antenna is pointed, which is absurd for this application. Thus the idea was born to develop a circularly polarised patch antenna with an SMA connection for the frequency range from 2420 to 2480MHz. It can be connected to the remote control transmitter using a short piece of semi rigid cable so that the main radiation direction can be pointed exactly toward the aeroplane. Thus a stable radio link should be possible.

2. The Basics

A patch antenna consists of a piece printed circuit board material (PCB) covered on both sides with copper. The lower surface forms a continuous ground surface and on the top side there is a simple square or rectangle made from copper making the "patch". Obviously the PCB must be larger than the patch for correct operation (approximately 3 to 5 times - and more is better).

The patch is designed in such a way to exhibit an electrical length of $\lambda/2$. A better choice:

Antenna length = 0.49 x (λ)

It can be regarded as an open circuit microstrip line. Feeding this line at the input with a signal at a frequency corre-

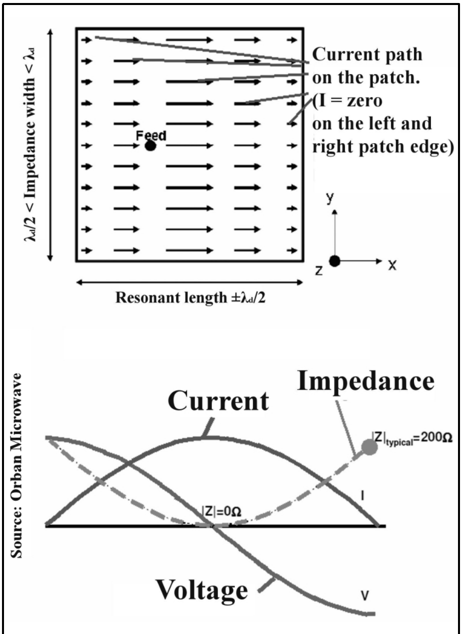

Fig 1: A patch antenna showing the current and voltage distribution along a λ/2 open circuit microstrip line.

sponding to the resonant frequency produces the current and voltage distribution shown in Fig 1 (the feed point is on the symmetry axis of the patch).

The voltage at the ends of the line is equal and in anti phase (λ/2 means 180 degrees of phase shift) but at the centre of the patch the voltage is zero. For a lossless line the current is zero at the ends and a maximum in the centre. This is shown in Fig 1.

There is a high radiation resistance at the beginning and end of the line representing the radiation. In the diagram a typical total resistance of 200Ω is shown; thus there is 400Ω at the patch edge because the two load resistances can be thought of as being in parallel.

The voltage goes through zero at the centre of the patch and thus the input impedance equals zero. Thus there is a point somewhere between the centre and the edge of the patch where the input impedance will be 50Ω - that is marked as the feed point.

The patch width (line width) only affects the self-resonant frequency slightly with this construction method. In practice the square patch is always used. If the width is increased (the patch made wider than it is long), the bandwidth increases and the radiation resistance becomes smaller.

Software is always used to design such antennas nowadays. For a long time the DOS program “patch16.zip” was used that was developed in form of a recipe for rectangular and square patch antennas (it can be found on the Internet). Modern designs use one of the “EM simulators for surface structures” and the most popular is Sonnet. For the private user there is an extremely efficient and free LITE version of this program enjoying worldwide popularity on the Internet. A Sonnet Tutorial in English or German can be downloaded from [1]. This has a complete practical design of a patch antenna for 5.8GHz described in detail.

A question remains to be answered:

Why and how does such piece of copper radiate at all?

For this, take a look at Fig 2. Taking a careful look there is an electrical scattering field at the outer edges of the patch (fringe fields) and this gives the solution! At the left and right edge the voltage is in anti phase (see Fig 1), interestingly the lines of flux at the edges both point in the same direction and are in phase! Thus these two patch edges with their scattering fields work as two parallel connected “slot antennas” (a slot antenna is similar to a dipole antenna supported on a mast. That consists of a wire and has air as an environment. The slot antenna has that, but exchanged: the mast antenna is replaced by air and in place of surrounding air, now there is a copper surface. Thus the directions of the electrical and magnetic fields exchange themselves).

Thus the question of the polarisation of

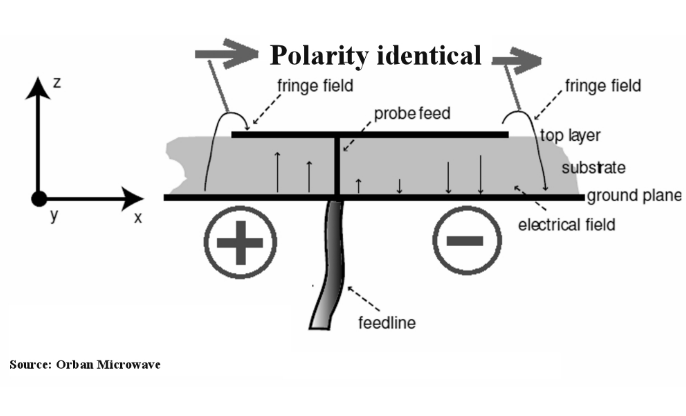

Fig 2: Despite the anti phase polarity of the voltage at the two line ends, the lines of flux are in the same direction as shown. See text.

the radiated electrical field is answered because this corresponds to the two arrows at the top of Fig 2. If the metallisation of the PCB lower surface projects sufficient far past the patch, it works as a screen and prevents the backward radiation (in Fig 2 the antenna will only radiate upwards). The radiation pattern of a simple dipole antenna is the familiar figure eight. The ideal patch antenna is missing one half of this eight, giving a simple circle without backward radiation.

3. Circularly polarised patch antenna

3.1. Circular polarisation

The principle of circular polarisation is to work with two antennas:

- Radiating individual fields at 90 degrees to each other.
- Additionally the two antennas are fed with signals that have a 90 degrees phase difference.

That has astonishing consequences, viewed from in front of the antenna; the polarisation of the radiated field rotates nonstop in the circle like a propeller. Since the fields move away with the speed of light the radiation vector describes a spiral (more descriptively; a corkscrew).

A linear polarised antenna can be used for reception because no matter how it is aligned it always receives this propeller movement of the radiation vector giving the same antenna voltage. Twice per revolution this vector is aligned with the linear polarised receiving antenna and results in maximum antenna voltage (with opposite sign). Likewise the vector is perpendicular to the antenna twice per revolution and produces no antenna signal. The drawback is that only one of the two radiated EM field components is received and therefore 3dB less signal than a circularly polarised antenna (with the correct direction of rotation).

3.2. Production of the circularly polarised patch antenna

This is not difficult at all because both antennas at 90 degrees to each other are already available by the length and width of the patches. As soon as they are fed with two signals at 90 degrees to each other (a 90 degree shift is produced using a power splitter with an additional quarter wave line), the problem is solved, see Fig 3.

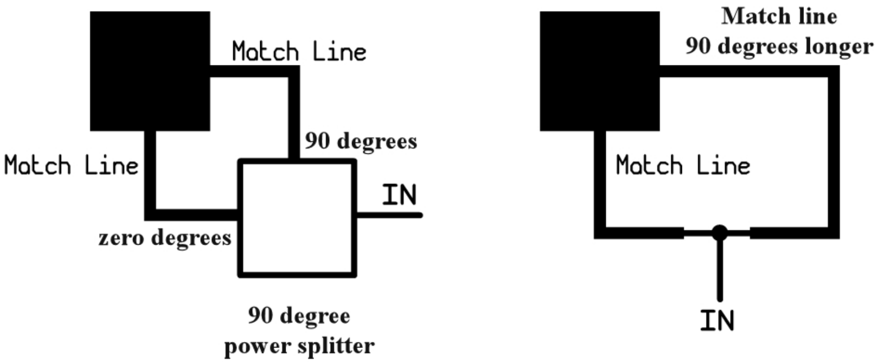

Fig 3: The two possibilities to produce circular polarisation with a square patch antenna.

There is another much simpler, but highly interesting way:

By using different dimensions for the patch length and patch width the resonant frequencies for the two antennas are easily shifted. In an equivalent circuit each antenna can be represented by a tank circuit producing an interesting effect when fed at the centre frequency.

The shorter antenna (e.g. formed by the patch length) is operated below its resonant frequency and behaves like an inductance. The current lags the voltage. The other longer antenna (in this case the patch width) is operated above its resonant frequency and behaves like a capacitor with leading current.

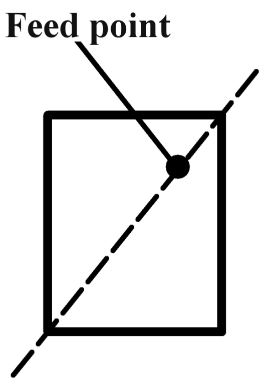

Fig 4: First possibility: Length and width of the patch are different, the feed point lies on the diagonal. See text

When the same voltage feeds both antennas within, for instance, 1 to 3% of the resonant frequency it gives the desired 90-degree phase difference between the two antennas (responsible for the radiation).

Naturally the feed point must be selected correctly so that both radiation directions become excited. Usually moving diagonally across the patch face finds the feed point.

The voltage is always zero at the patch centre (with maximum current) giving an impedance of zero. At the patch edges the radiation resistance measures more than 100Ω. Therefore between them a feed point of 50Ω input impedance can be found. In practice there are one of these three possibilities:

- In Fig 4 length and width have different dimensions, the feed point lies on the diagonal. This version is unfortunately very sensitive to tolerances. Differences of a hundredth

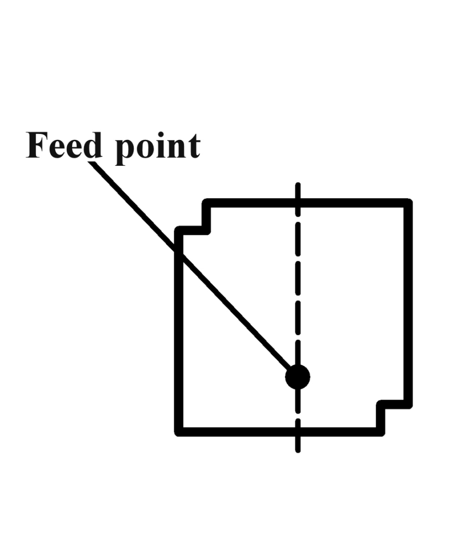

Fig 5: Second possibility: the patch is square with the corners punched out, the feed point is on the centre line.

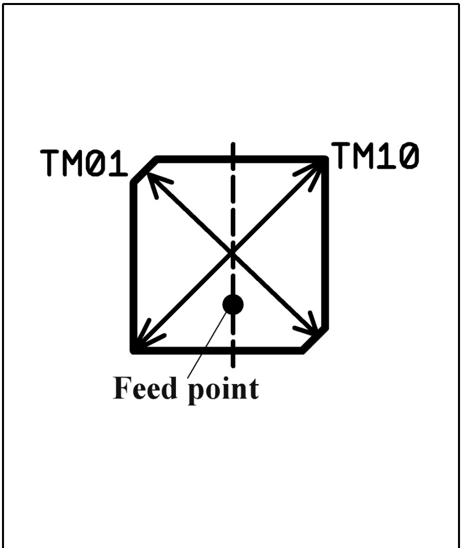

Fig 6: Third and most popular possibility: the corners are cut at 45 degrees, otherwise the same as Fig 5.

of a millimetre in the length or width greatly affect the feed point.

- Fig 5 shows a square patch with punched out corners. The feed point is on the centre line. The size of the punched out corners determines the difference of the two resonant frequencies (and works like different coupling with a double-tuned bandpass filter). Now the patch diagonals are different lengths and form the two antennas. This version is easy to design.
- The most frequently used variant is shown in Fig 6. It is a square patch with corners tapered at 45 degrees. This gives diagonals of different lengths (as with B) to ensure circular polarisation. It develops a TM01 and a TM10 mode with the individual waves. It gives the same behaviour as B), it is more good-natured and error tolerant. Such an antenna is to be described now.

4.

Antenna design with Sonnet Lite

(Preface: The design process to produce the final version that can be converted into a PCB antenna naturally takes innumerable simulations to arrive at the optimum. These will be skipped over here but are assumed to have taken place.)

First a few words about Sonnet (for those who have no experience):

Sonnet is a marvellous thing, an EM simulator (using the moment method) for all conceivable planar structures. These are mainly Microstrip, Stripline or Coplanar circuits including; couplers over transformation lines, filters, Gaps and Stubs can be examined and patch antennas etc. For those coming from SPICE simulation or S parameter simulation there are some introductory problems because an EM simulator contains every-

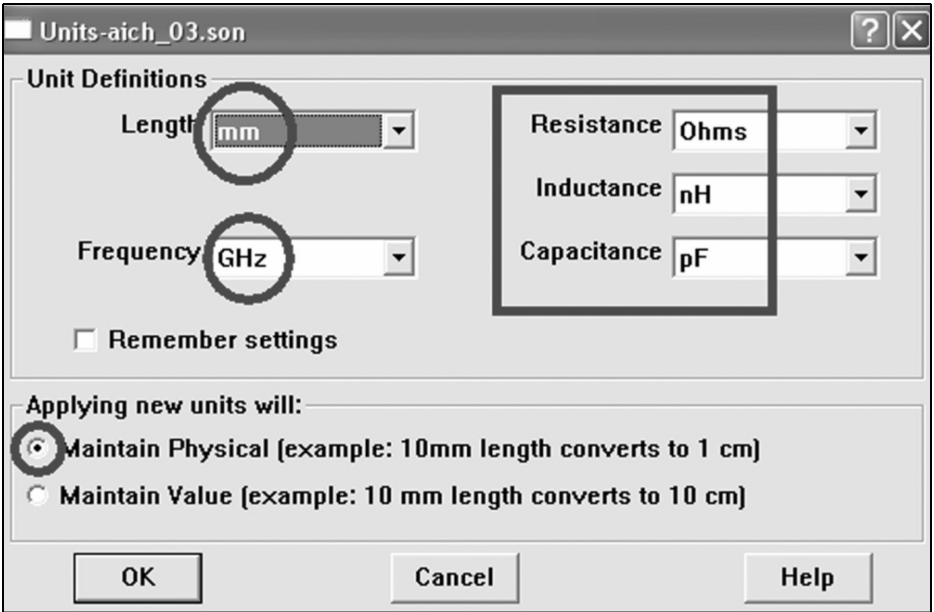

Fig 7: The basic parameters to be set when starting Sonnet.

thing but everything is different. First the structure to be examined must be divided into small cells that are valid for consideration. The characteristic of each individual cell is examined, computed, stored and then the program combines them by integration into the total effect. Each item under test by Sonnet must be placed in a cubical metal box where several basic adjustments to are made.

The walls of the box are always made of loss-free metal therefore they work as mirrors. The cover or the ground looks different; if e.g. an antenna is simulated the energy must escape from the box somewhere (in this case via the cover).

Therefore there are different options, i.e. loss-free, WG load, free space and naturally a metal that can be selected, e.g. copper.

The field behaviour in such a box is well-known and calculably, producing simulations with quite high accuracy. It depends on, the cell size selected normally somewhere between 1% and 4% (the smaller, the more exact, however more computing time is required) and attention to the rules, particularly with the restrictions of the Lite version. Sometimes the resonant frequency is simulated a little too high particularly with resonant objects (e.g. patch antennas) - that must be known and manually adjusted.

4.1. Setting attributes

When the program is started the Sonnet task bar appears. Start your own project from “Project/new Geometry”.

Open the “units” window from the “Circuit” menu and set the values: mm, GHz, ohms, nH, and pF as shown in Fig 7.

Open the “Metal Types” window from the “Circuits” menu and set the values as shown in Fig 8. If only “Lossless” is in the list, click on “ADD” and then on “Select metal from Library” and choose copper. Click OK to open the Property menu of the copper layer; this should then look like Fig 8. The thickness of the

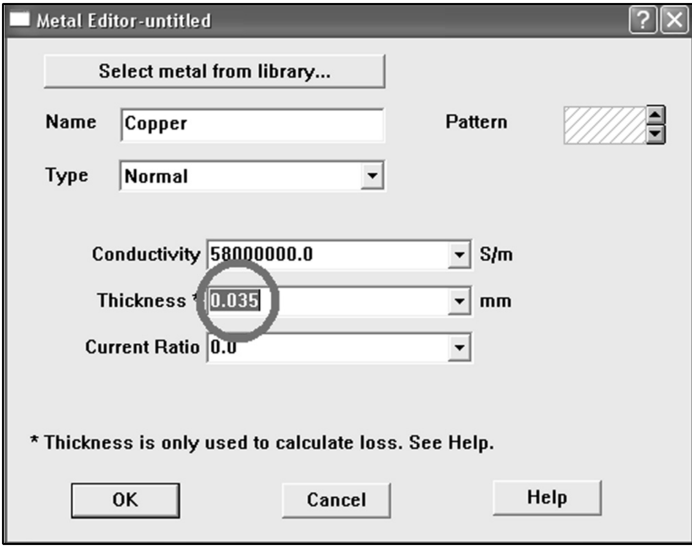

Fig 8: This is important: The PCB is 0.035mm thick and coated with copper.

copper layer is set to 35µm = 0.035mm and confirmed by pressing OK. The overview menu ensures that everything drawn later is realised in copper (Fig 9).

Fig 9: Do not forget this: All objects in the structure will be made from copper.

This can be identified easily in the drawn structures because copper surfaces appear in green and lossless surfaces in red.

Now select the "Dielectric Layers" option from the "Circuits" menu and select the "Edit" option on the upper level for the air cushion in the box to be simulated by Sonnet (Fig 10). This air cushion (according to Sonnet manual) should be about half a wavelength thick for a patch antenna - that would be about 60mm. To improve the overview add "Air" to the names.

This is confirmed with OK then final select the "Edit" option for the lower level. The PCB will be made from Rogers RO4003 material with a thickness of 32MIL (0.813mm) with a dielectric constant of ε_r = 3.38 and a dissipation

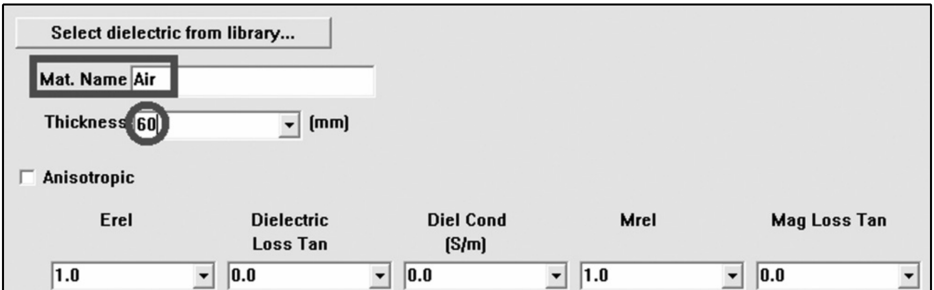

Fig 10: The area above the PCB in the box is filled with a thick air layer. See text

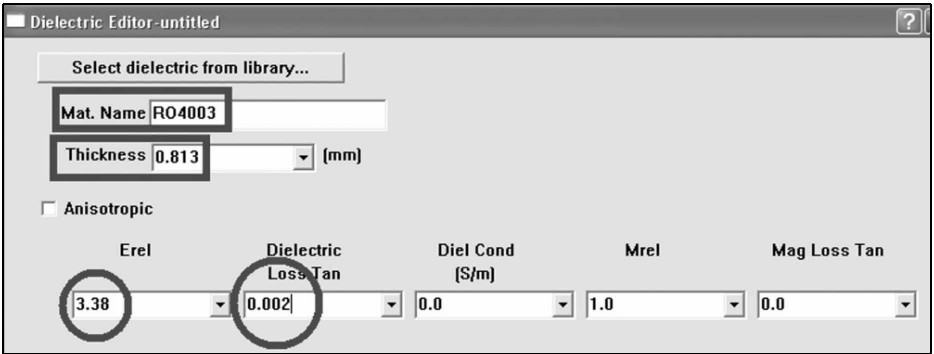

Fig 11: The characteristics of the PCB material (Rogers RO4003) must be registered accurately. See text.

factor - "loss tangent" = 0.002 at 2.5GHz. This data is entered in sequence in accordance with Fig 11.

Now the most important thing, i.e. the "Box" that Sonnet will use for the simulation is shown in Fig 12. The dimensions of the "cells" that Sonnet will use to divide the structure must be specified. A good approximation is 1% of the wavelength, this would a bit more than 1mm x 1mm in this case. But: smaller cells give higher simulation accuracy, however in the Sonnet Lite version 12.53 the memory is limited to 16Mb. Therefore 0.5mm x 0.5mm was selected.

Sonnet recommends that the structure to be simulated should be kept away from the box wall by between one and three wavelengths. For the patch antenna the maximum should be used if possible if the Lite version does not terminate with an error message. The cover of the box

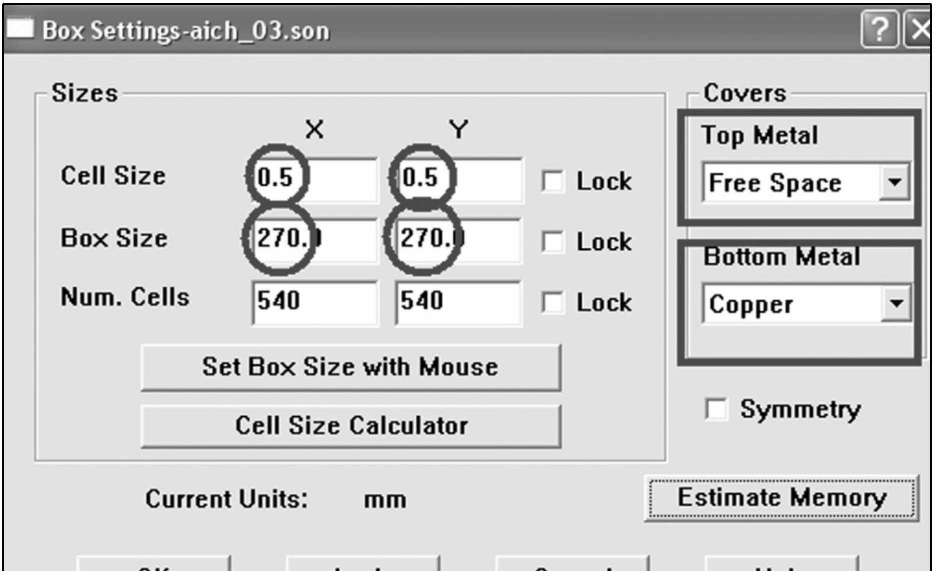

Fig 12: Very important: the entries for the cell size and the box dimensions as well as "free space" for the cover. See text.

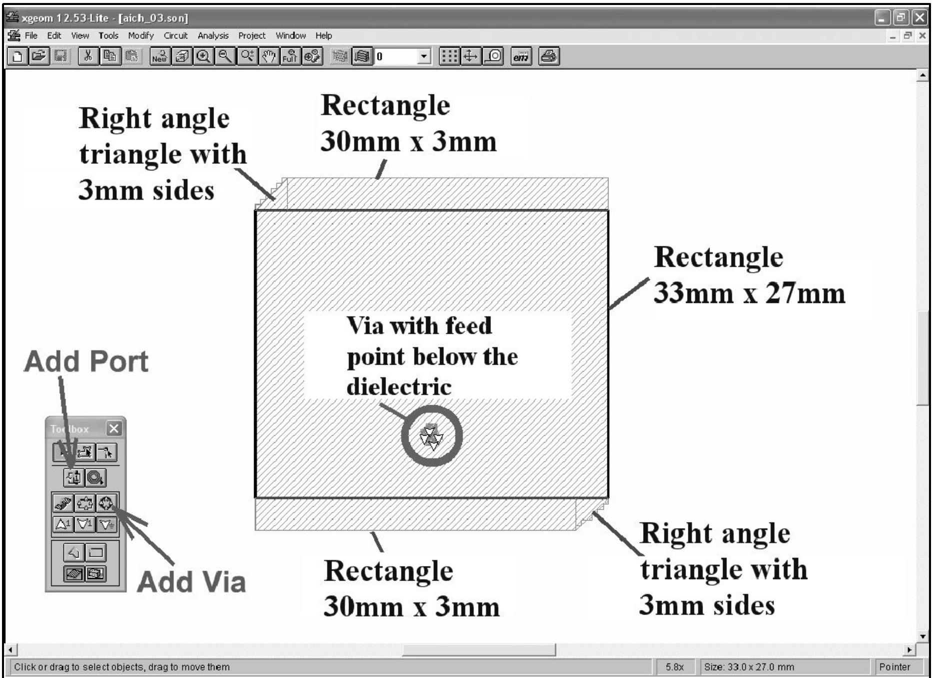

Fig 13: After some work: The complete antenna structure is centred provided in the box.

may not consist of metal. That must be replaced with "Free space" because we want it to radiate. Copper is used for the ground of the box.

The box size should be:

2 x wavelength + 1 x Patch edge length,

therefore approximately:

2 x 12cm + 3cm = 27cm.

That fits into the approved 16Mb of main memory for the Sonnet Lite version.

The preparations are complete and it time to draw the antenna structure. It consists of a square with 33mm long edges with the corners shortened by around 3mm each. This can be constructed from three rectangles and two triangles (Polygons) see Fig 13. There are some useful tools

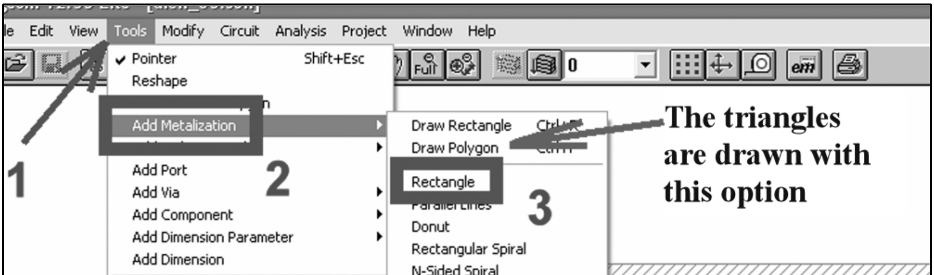

Fig 14: The tools to draw the rectangles and triangles.

Fig 15: The feed port can only be seen if you switch to "ground" on the menu.

for drawing the structure that make drawing the three rectangles as well as the two triangles child's play. Select the path:

"Tools/Metalization/Rectangle

and

"Draw Polygon" (see Fig 14).

After the puzzle has been joined together and positioned in the centre of the box a plated through hole (via) with a diameter of 1.27mm must be inserted at:

Position X = 16.5mm Y = 9mm

(relative to the lower left corner of the antenna)

Before adding the "via" set the zero point (origin) on lower left corner of the patch. This can be done easily, right click on the patch to open the menu and select "Place origin". The "via" diameter (1.27mm) corresponds to the pin of the SMA socket used, this can be adjusted in the menu. Another port is required that is attached automatically after the placement on the "via" by the program on the lower surface of the dielectric. This port can only be seen if the "Ground" level is selected (Fig 15), it presents itself as a simple rectangle with the number "1" inside. Double clicking on this number opens the Properties menu to correct the attributes (type = Standard/Resistance = 50Ω). Now the simulation can be started.

4.2. Simulation

First go to "Analysis" on the main menu and complete the Setup as shown in Fig 16. Set: Start, 2.4GHz; Stop, 2.5GHz and "Adaptive Band Sweep ABS" that results in shorter computing time, the current

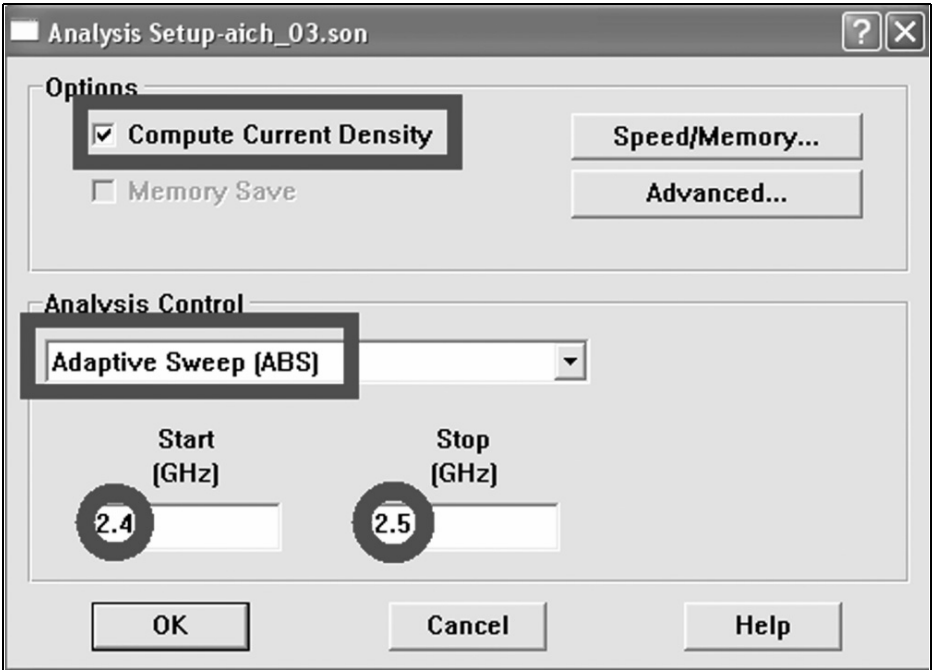

Fig 16: The last job before starting the simulation is to programme the sweeps.

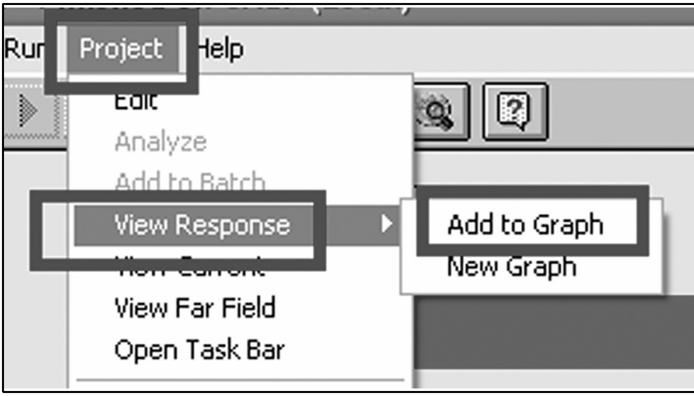

Fig 17: As is the case for most modern programs, nothing is displayed after a successful simulation. The output results must be programmed.

density on the patch can also be calculated.

Start the computation by pressing the "EM" button on the right above in the menu border. If it ends successfully the results shown in Fig 18 can be selected as shown in Fig 17. The resonant frequencies of both single antennas can be recognised and the desired operating frequency should lie in the centre. The resolution of Sonnet Lite is limited but we must be content with the result. Click the "Graph/Type/Smith" option in the menu on the border to display the Smith

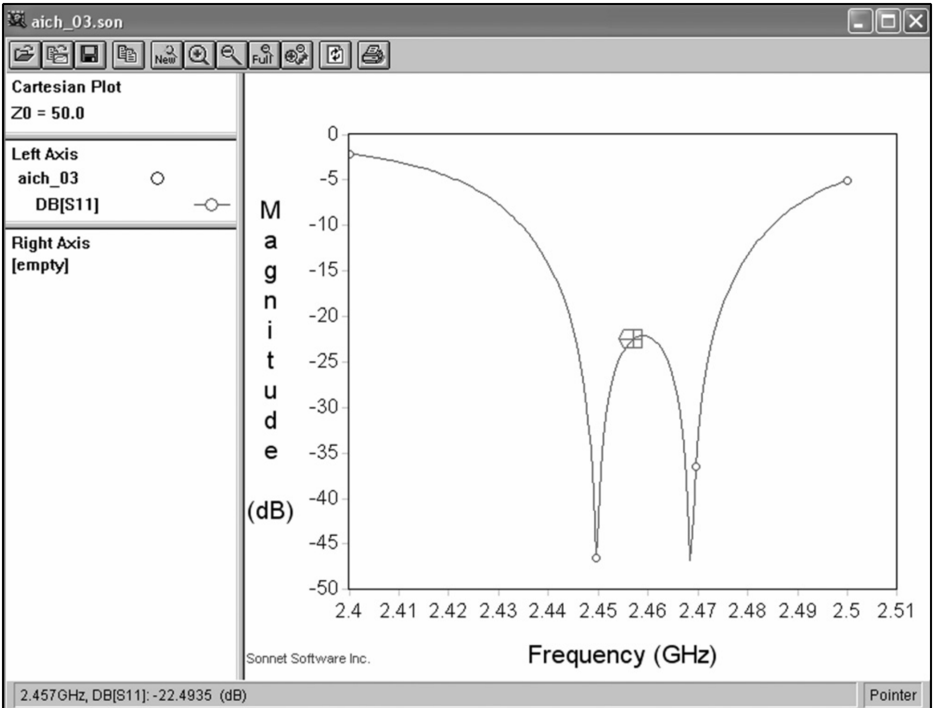

Fig 18: The results are pleasing and show that the design is correct.

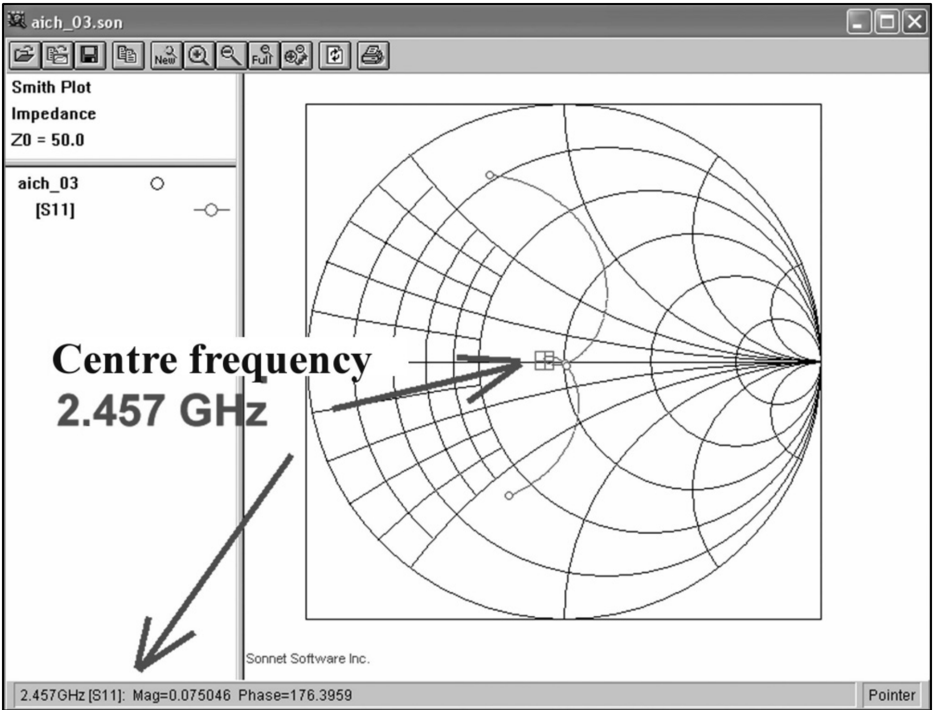

Fig 19: The Smith chart can be produced for the specialists after providing substantial additional information.

chart shown in Fig 19 after some computation.

Having reached this point simulations were stopped and a prototype PCB was made and measured. A centre frequency of 2393MHz was measured (approximately 2.5% lower than desired).

In relation to the required value of 2.45GHz it was too low by the factor (2393/2450) = 0.9767346. The patch measurements were made smaller by this amount and a new PCB made.

The external dimensions of the patch were therefore:

0.9767346 x 33mm = 32.23mm

and the corners:

0.9767346 x 3mm = 2.93mm

Now the new feed point is (measured from the left lower corner):

X = 0.97673 x 16.5mm = 16.12mm

and

Y = 0.97673 x 9mm = 8.79mm.

The results were most interesting and can be seen in Fig 20.

The centre frequency was approximately 8MHz too high in the last simulation and then it was too low when measured. When the new PCB was measured it was approximately 10MHz too low and the two resonant frequencies of the antennas were somewhat further from each other. That means that the corners would have to be filed to cut off the corners because these are responsible for the frequency difference of the antenna resonances (this is reminiscent of the coupling of a double-tuned band-pass filter). But this

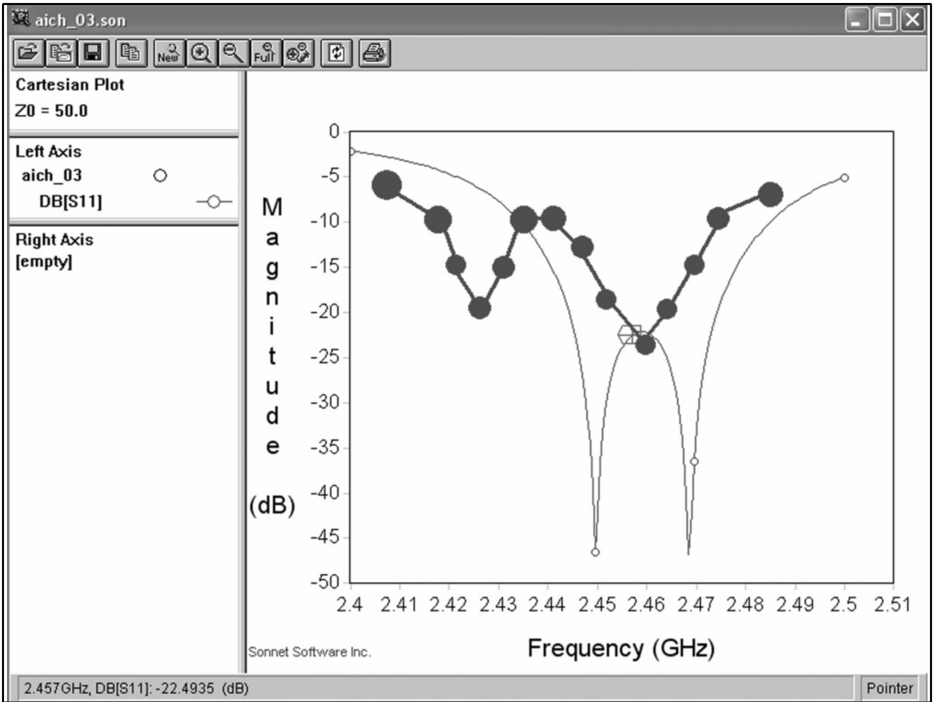

Fig 20: That is the difference between theory and practice! The result is 10MHz below the required centre frequency of 2.45GHz.

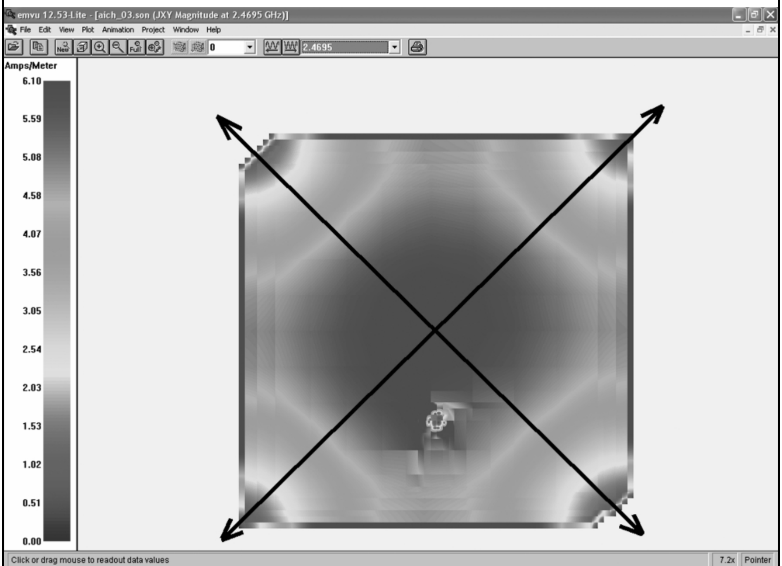

Fig 21: The current distribution at the required centre frequency is fascinating. It is easy to see the two active antennas, even in black and white. It looks much better in colour - Andy.

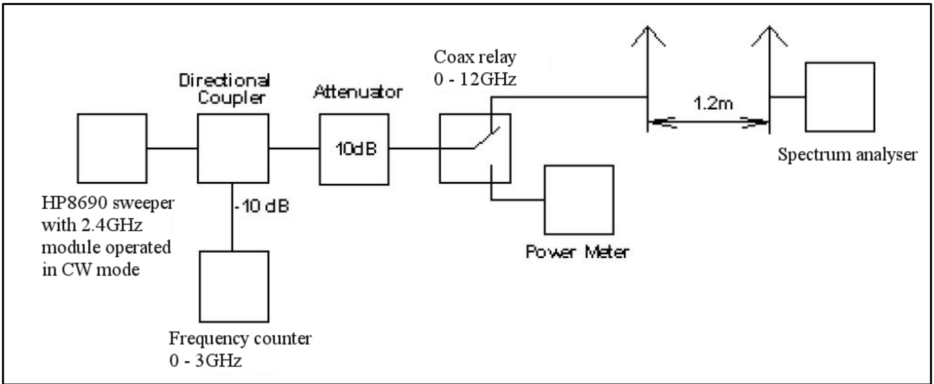

Fig 22: This is the measurement setup (here for 2.45GHz) that has been used successfully several times to perform gain measurements on antennas in the far field. See text.

could be continued eternally…

4.3. The current density

In the menu on the border when the patch is drawn there is an option "View Current" under "Project". Selecting the frequency f = 2450MHz in this option shows visual confirmation of the theory and simulation. It is very beautiful (much better in colour - Andy) showing the two radiating diagonals because the current is zero at the four corners and maximum at the patch centre (Fig 21).

5. Radiation pattern and gain

The measurement setup shown in Fig 22 was used with the first two identical linear polarised patch antennas that were perfectly adjusted (input impedance accurately = 50Ω). The left antenna is fed with exactly 1mW (zero dBm). A frequency counter supplies information over the measuring frequency. The gain measurement takes place according to the method shown in full in [1]. Here is an edited version:

The distance between the two antennas was 120cm and this corresponds to approximately 10 wavelengths at 2.45GHz. Thus one is in the far field giving a receive level so that the famous "Friis formula" is valid:

$$P_{receiver} = P_{sender} \cdot G_{senderantenna} \cdot G_{receiverantenna} \cdot \left(\frac{\lambda}{4\pi \cdot d}\right)^2$$

If this is converted into logarithmic representation and remembering that both antennas are identical, the formula takes the following form for "dB" representation:

$$\mathrm{Re}\,ceiverlevel = Senderlevel + 2 \cdot antennagain + 20 \cdot \log\left(\frac{\lambda}{4\pi \cdot d}\right)$$

With a transmit level of 0dBm, an antenna distance of 120cm, a wavelength of 12.24cm and the two antennas with identical behaviour a receive level of –29dBm results. Changing the above formula round for the antenna gain the result is:

$$Antenngain = \frac{\mathrm{Re}\,ceiverlevel - 20 \cdot \log\left(\frac{12.24cm}{4\pi \cdot 120cm}\right)}{2} = \frac{129dBm + 41.8dB}{2} = 6.4dB$$

A condition for a correct result is that the measurement is in the genuine far field. Increasing the antenna distance the antenna gain determined according to this

Fig 23: The final result, the prototype antenna, measured and given to the model flier.

formula and method remains constant. For antennas with very high gain and too small distance to the transmitter an error can occur and the computed gain can suddenly become negative. The gain produced can be interpreted in such a way that it is produced between the transmitting and receiving antennas or that the transmitting power of the generator was raised. In each case these antennas form the near field (Fresnel zone) or the zone up to the far field, called the Fraunhofer-Region.

Therefore the default must be that the distance between the two antennas is greater rather than too small (the receive level should still be determined correctly and with changes in distance the gain determined should remain constant). This information and considerations originate from the publication [3] that can be found on the Internet.

Back to the current project. The control antenna connected to the analyser input was replaced by the new circularly polarised version. It was then turned in steps of approximately 30 degrees in a circle and at each angle the change the level from the spectrum analyser was read.

Ideally (according to the theory) a constant level decreased by 3dB would be to be expected in relation to linear polarisation with the radiation pattern that should be a circle (logically with circular polarisation the transmitting power is split in two equal components shifted by 90 degrees not only spatially but also in the phase. The linear polarised receiving antenna can only receive one component - thus one antenna receives only half the maximum power possible and that is the 3dB mentioned). The fact is that everything is not so simple as it sounds as shown in the appendix of the article written for the specialists. In the literature [2] it was suggested that the ideal case is rarely reached in practice and an ellipse instead of the circle is usually observed. But less gain in the two minima by up to 3dB can normally be tolerated. Unfortunately there was between 4 and 5dB!? Fortunately this effect was not as strong in practice as feared and the model flier was overjoyed (original quotation from his mails: “I tested the antenna in the model airfield and determined at ground level a range increase of 30 - 50%”). Fig 23 shows the prototype supplied including an SMA semi rigid cable fitted that has been given an acid test in the field.

P.S.: The best message last - in the time between giving the lecture and the publication of this article the maximum usable main memory for the newest Sonnet Lite version 13.51 has increased from 16 to 32Mb. Therefore after the update the above simulations were immediately repeated with a cell size of only 0.1mm x 0.1mm and afterwards even 0.05mm x 0.05mm (0.5mm x 0.5mm in the article). The result reads: only minimum and insignificant deviations and therefore all simulations, results, corrections, measurements and statements are valid. Thank God….

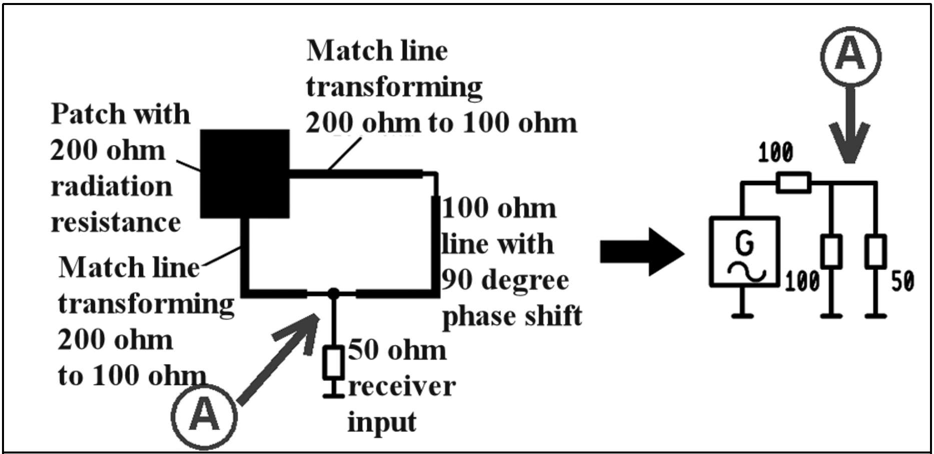

Fig 24: The equivalent circuit produced from the antenna on the left makes it possible to understand the behaviour of the antenna. See text.

6. Literature:

[1] An interesting program: Simulation and construction of a Helix antenna for 2.45GHz using 4NEC2, Gunthard Kraus, DG8GB: VHF Communications Magazine, 2.2011 pp 89 - 101; (www.elektronikschule.de/~krausg)

[2] Orban Microwave Application note: The Basics of Patch Antennas, Updated. To found on the Internet

[3] Paul Shuch, N6TX: Far Field Fallacy. Appeared in QEX December 1987, find page 10 on the Internet.

7. Appendix

The level decrease of approximately 3dB when receiving a circularly polarised signal with a linear polarised antenna was justified logically and the simulation confirmed the considerations. This stimulated a discussion among friends and an easy confusion about the procedures with different possible antenna combinations. A list of possibilities is shown below (those that occur in practice) with the signal conditions that can be observed.

A) We begin with the simplest case: Transmitting with a linear polarised patch antenna and using the same linear polarised model for reception. Both antennas are optimally aligned and thus accurately form the case from chapter 5 where the measuring position is calibrated with propagation in the far field. The received level of the antenna can be determined according to the Friis formula and simply named “P1”.

B) Both linear polarised antennas replaced by two identical circular polarised antennas (see chapter 3.2, Fig 3). The power supplying the transmitting antenna is divided into two portions out of phase by 90 degrees and then supplied to the two individual antennas at 90 degrees to each other. The two radiated signals are then only half transmitting power. On the receive side two individual antennas at 90 degrees to each other pick up the signals, that is half the size of P1, to produce a received level. These two halves are phased together; the phase difference is correctly compensated by 90 degrees with the phase shifter in one arm of the circular polarised antenna. That results in the level P1 at the receiver input. But be careful: “phased” means that the receiving antenna must exhibit exactly the opposite circular polarisation to the transmitting antenna…

C) The transmitting antenna is circular polarised and the receiving antenna is linear polarised. Thus only one of the two radiated components is used and the linear polarised receiving antenna supplies only half of the level P1. That results in the famous 3dB less gain in relation to case A), but the advantage is that the receiving antenna can be rotated at will.

D) This is getting exciting with the possibility of a linear polarised transmitting antenna and a circular polarised receiving antenna. The linear polarised transmitting antenna radiates the full power only in one polarisation plane. If the circular polarised receiving antenna is aligned accurately to the transmitting antenna, the same receiving level P1 is developed as in case A). The path of this energy in the phasing circuit is shown in Fig 24.

The received signal (e.g. supplied from the lower right antenna) goes through only a λ/4 matching line to bring the radiation resistance of the patches from 200Ω to 100Ω. The equivalent circuit on the right can be used; this antenna is represented by 100Ω including the matching line as a supply with an internal resistance. This supply now sees a load at point “A” consisting of a parallel connection of the receiver input (50Ω) and the second antenna including its matching line and its phasing line - thus again a resistance of 100Ω. The total resistance of this parallel connection is approximately 33.33Ω. If level P1, coming from the antenna is considered as a incident wave the reflection factor that the supply terminal sees can be calculated:

$$r = \frac{33.33\Omega - 100\Omega}{33.33\Omega + 100\Omega} = \frac{-66.66\Omega}{133.33\Omega} = -0.5$$

Thus the reflected level becomes a portion of:

$$P_{reflected} = r^2 \cdot P1 = [-0.5]^2 \cdot P1 = 0.25 \cdot P1$$

This level is reflected to the feeding antenna and radiated again.

The remaining part is three quarters of P1 and it is feeding two load resistances. But the receiver input of 50Ω absorbs twice of the part which enters the 100Ω transmission line leading to the second antenna.

The result is: The receiver input “swallows” two quarters, thus half of P1 (the familiar 3dB reduction). The remaining quarter of P1 goes to the other transmission line and feeds to the other patch antenna after a 90-degree phase shift.

Summary:

Only one antenna in the circular polarised receiving antenna supplies the level P1 in this example. Half is supplied to the receiver and results in the familiar level difference of –3dB in the case of a circular polarised antenna receiving a linear polarised signal. A quarter of P1 is reflected back to the receiving antenna. The missing last quarter is radiated from the other antenna with a 90-degree phase shift giving a “circular polarised reflection radiation”.

What an effort - it cost a lot of sweat to come up with the details of the procedure. But honestly: The effort was worthwhile?!

André Jamet F9HX

Become familiar with SMD resistors for good use from DC to microwaves.

An application : A homemade 10GHz 50Ω load

1.
Introduction

Components for surface mount, SMD, have very specific characteristics that should be known for a good job. The subject of an article for ceramic capacitors [1] has already done. Now here is information on SMD resistors.

2.
SMD resistor technology

A conductive layer is placed on a ceramic support. At the ends of the ceramic support two metal connections are taken out.

The most popular SMD resistors usually have a thick resistive layer. These are made by printing a ceramic material base (Cermet) or glass (glaze), containing a metal or oxide.

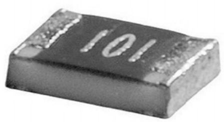

Fig 1: Magnified photograph of a Chip SMD resistor.

Some have a thin layer whose thickness may be thousand times smaller than that of a thick layer. It is a film of a metal or oxide deposited under vacuum.

In both cases, thin or thick layer, the precision of the manufacturing process does not produce resistors with values in the desired tolerance range. A trimming process is then carried out using mechanical abrasion or a laser.

Finally the resistive layer is covered with a colourful protective layer with the marking.

3.
SMD resistor packages

There are two standards for SMD packages, parallel sided packages (chip) that are the most common and the cylindrical package (MELF = Metal Electrode Leadless Face).

3.1 Chip

These resistors can be made thick or thin, see Fig 1. Table 1 gives the characteristics of chips for the most common models. Standardisation is much broader as

Table 1: Chip resistor dimensions and ratings.

Type of package SMD	Dimensions L x W x D inches	mm	Rated dissipation mW @ 70°C	Maximum voltage V AC/DC
1206	0.126 x 0.063	3.00 x 1.60 x 0.60	125	200
0805	0.08 x 0.05	2.00 x 1.30 x 0.06	100	150
0603	0.063 x 0.031	1.50 x 0.08 x 0.45	63	50
0402	0.04 x 0.02	1.00 x 0.54	31 to 63	

the packages range from is 1005 to 2512. The thickness is given for information only as it only comes from a single manufacturer. The maximum power is still given for information only and depends on the mounting position of the resistor.

For marking, three systems can be found with 3 or 4 digits, and the EIA marking. The first, the most common, is that used for standard tolerance resistors. The first two digits are significant and the third, the multiplier.

Examples:

220 → 22Ω
221 → 220Ω
223 → 22kΩ

For values less than 10Ω, R is used to indicate the position of the delimiting comma.

Examples:

0R1 → 0.1Ω
2R2 → 2.2Ω

The four-digit system is used for tight tolerance resistors. The first three have significant values and the fourth, the multiplier.

Examples:

1000 → 100Ω
4702 → 47kΩ

For values less than 100Ω, an R is used to replace the comma.

Example:

0R56 → 0.56Ω

Marking according to EIA96 is used for 1% tolerance resistors. This code requires a development that is too long for this article (see [3]).

In practice, it is advisable to measure the resistor if there is the slightest doubt on its value.

3.2 MELF

These resistors are made thin-layer. There are three different cylindrical packages (Fig 2). The marking is consistent with the colour code used for conventional resistors.

Footprints for the soldering of the micromelf and minimelf on a printed circuit are the same as those for chip types 0805 and 1206.

4. Tolerances

Even for thick layer resistors, it is possible to obtain tolerances to ±1% and the E96 series is available for the non-professional users for a cost very little more than ± 5% tolerance resistors. Thin layer can achieve much tighter tolerances, up to 0.1%.

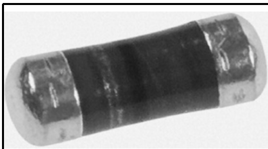

Fig 2: Magnified photograph of a MELF resistor.

Table 2: MELF resistor dimensions and ratings.

Type of package Cylidrical	Dimensions L x W x D inches	mm	Rated dissipation mW @ 70°C	Maximum voltage V AC/DC
Micromelf	0.045 x 0.09	1.1 x 2.2	200	100
Minimelf	0.06 x 0.14	1.4 x 3.6	250	200
Melf	0.09 x 0.23	2.2 x 5.8	1000	500/350

5.

Influence of temperature

The temperature coefficient of thick is much higher than that of thin films. With a coefficient of ±250 x 10^{-6}/°C, a variation in temperature of 40°C causes a variation by 1%. Thin-layer can be ±25 x 10^{-6}/°C.

6.

Overload

Overload may be due to voltage or power exceeding the limits given in the Table 2.

Voltage overload is possible without exceeding the power for high ohmic values. On the contrary, it is easier to exceed the maximum power without exceeding voltage for low ohmic values.

Voltage overload may be due to an external cause such as electrostatic discharge (ESD) or the device using the resistor. ESD susceptibility is good enough to be neglected in general. Planned voltage surges can be obtained in pulse operation. Fig 3 shows an example of maximum permissible peak pulse voltage for a 1206 SMD.

As with any electronic component a resistor can be overloaded, accidentally or intentionally. In the first case, it is difficult to determine all the possible causes and prevent their effects. On the contrary, certain overloads are planned and their effects can be calculated.

A very common case is overload when a device is powered on. One or more components may be subject to a much higher than normal current. This is the case of a resistor supplying a high-value capacitor.

The use of resistors in pulsed operation can give a planned overload.

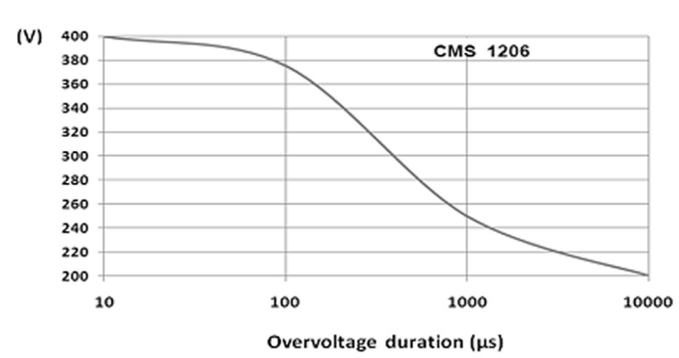

Fig 3: Example of maximum permissible peak pulse voltage as a function of its duration.

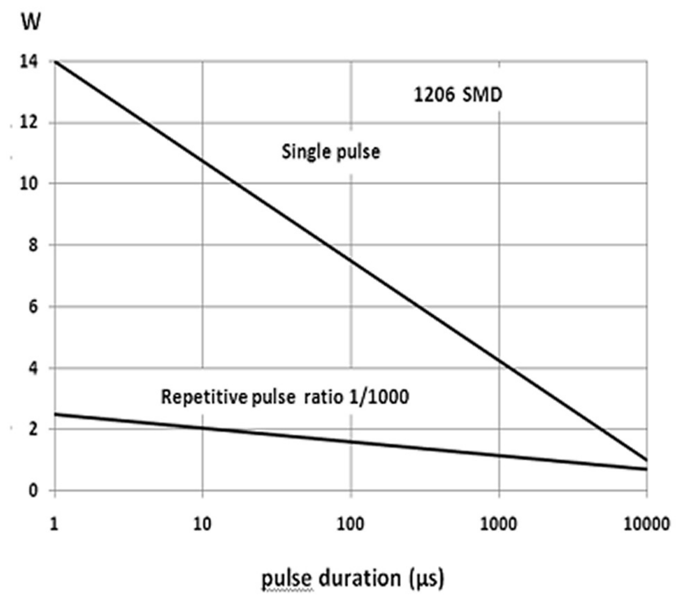

Fig 4: Example of maximum permissible power overload vs. duration.

In the case of planned overloads, it is essential to ensure that resistor chosen will be able to withstand it without damage. In view of the small dimensions of the SMD, it is wise to be cautious and know their acceptable limits.

Contrary to what one might think, thick film devices withstand fewer overloads and pulses of power than thin layers.

In the case of overload at power on, it is recommended to provide a limited number (approximately 1500 - switching a TV set on/off/on/off twice a day for 2 years is 365 x 4 = 1460) and a higher one-hour rest period [2] between two overloads. This seems harsh, but non-compliance is probably the cause of many failures of consumer electronics.

The curves in Fig 4 show the permissible values for a 1206 SMD for a single and repetitive overload pulse. The overloads will also cause accelerated ageing, so some variation of ohmic value must be expected.

7.

The noise level

The noise of thermal agitation caused by a resistor is normally given by:

$$E_{eff} = (4 \times 1.38 \times 10^{-23} \times R \times T \times \Delta f)^{0.5}$$

with:

T = temperature in K ($T_{°C}$ + 273)

R = resistance in Ω

Δf = bandwidth in Hz

For example a 100kΩ resistor causes 2µV_{RMS} noise in 2.7kHz bandwidth at 17°C.

This is true for metal resistors. But, as for semiconductors, the noise is higher than when the conductive material is not a pure metal or alloy. The excess noise spectral distribution follows a 1/f law and depends on the voltage applied to the resistance. This is specified in µV of

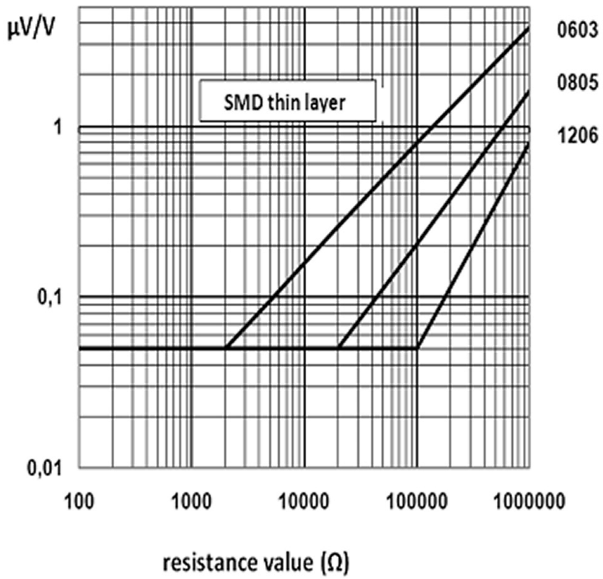

Fig 5: Excess noise of SMD thin layer resistor.

noise by voltage applied to the resistor for a decade of frequency: µV/V/decade. This value is often expressed in decibels. For a resistor with an index equal to 0dB, the excess noise will be $1\mu V_{RMS}$ for each volt applied to it, in each decade of frequency.

The thick layer resistors are 10 to 100 times noisier than thin metallic film resistors. The noise decreases with the size of the resistor and increases as the

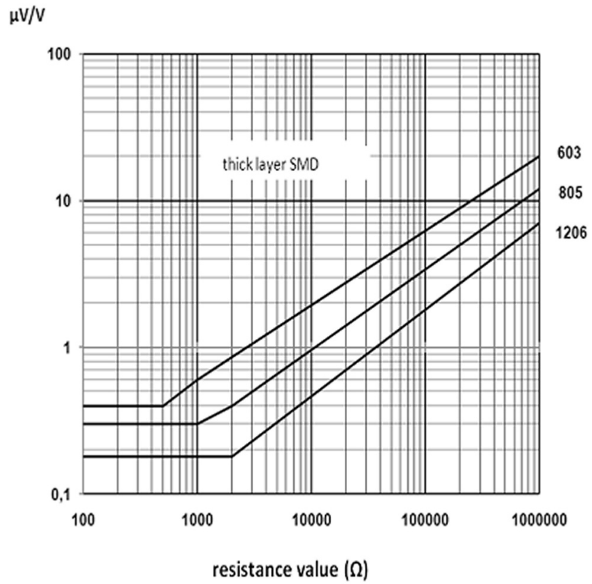

Fig 6: Excess noise of SMD thick layer resistor.

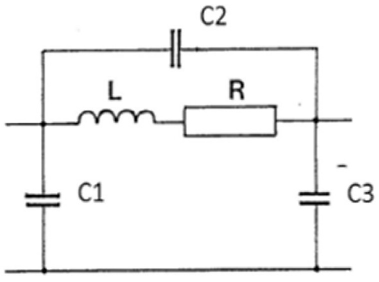

Fig 7: High frequency equivalent circuit of a resistor.

applied voltage.

For example, a thick layer of 100kΩ, 0805 package resistor, may cause a 15µV_{RMS} noise if it is subject to a 5V_{dc}.

This noise can be very harmful if the resistor is intended to feed a varicap diode associated with the oscillator in a PLL. Phase noise will be greatly increased.

It is recommended to choose a resistor with a low ohmic value and the largest size possible. Fig 5 and 6 give an example for thick and thin film resistors.

8. Non-linearity

SMD resistors have some non-linearity that results in distortion of the signal showing essentially the third harmonic. For high ohmic values, this term can reach –40dB for thick film and remains below –80dB for thin films. It must be taken into account in Hi-Fi and some measurement devices.

9. Reliability

General, the reliability of these resistors is good if they are well used. This is particularly true for SMD components, when soldering it is important to prevent their deterioration by heating or, on the contrary, to obtain a good solder joint.

The quality of manufacturing can guarantee the absence of corrosion or ageing of components. Good manufacturer can give a good MTBF (Mean Time Between Failures).

The Arrhenius law provides a reduction of a component life by half when its temperature increases by 10°C. Its application is questionable in the case of resistors but however gives the trend. The drift of the ohmic value and lethal defects are directly related to the operating temperature.

A lethal defect results in an open circuit. Using two resistors in parallel, or better four resistors in series/parallel, to obviate such an eventuality, if the variation of the ohmic value on failure is acceptable.

10. Behaviour at high frequencies

It is imperative to know the high frequency behaviour of resistors. Indeed, like all component resistors, SMD or not, comprises parasitic elements that have significant affects with frequency that are not negligible or be can ignored. This is the case for any frequency above the hundred of kilohertz and must be taken into account in the design of loads and attenuators.

The equivalent circuit of a resistor is shown in Fig 7. It shows the inductance

Table 3: SMD approximate parameters.

Package	L (nH)	C (fF)	$(L/C)^{0.5}\Omega$
402	0.2	30	80
603	0.4	50	110
805	1	50	180
1206	2	50	200
1210	3	50	250
Micromelf	2	43	220
Minimelf	3	50	245
Melf	4	60	260

of the resistor and its connections, and the capacity between the resistor ends and ground. In reality, the parasitic reactances are distributed along the resistor body. Calculations carried out if they were lumped constants remains valid up to several gigahertz in view of the small dimensions in use. Beyond that they are more trivial because they are too large a fraction of the wavelength. Reference [4] gives all formulae for the equivalent circuit. To avoid any repetition which would lengthen this article, it will be accepted that an approximate formula is:

$$|Z| \approx R\left[\frac{1+(X_L/R)^2}{1+\left(\frac{R}{X_C}\right)^2}\right]^{0.5}$$

This formula shows that:

- When $f \to 0$, $X_L \to 0$ therefore: $|Z| \to R$ that is that as the frequency decreases, the resistor impedance tends to a pure resistance.
- When $R \to 0$, $|Z| \to X_L$ that is that a low value resistor tends towards its parasitic inductance therefore the impedance increases with frequency.
- When $R \to$, $|R| \to X_C$ that is that high value resistor tends to its parasitic capacity so the impedance decreases with frequency.

Another important note:

- When $R \approx (L/C)^{0.5}$ $|R| \approx R$, that is, the behaviour is neutral, non-reactive, whatever the frequency. This is valid only up to a few gigahertz for the SMD resistors in view of their size. Above, the behaviour is more complex.

Parasitic inductance and capacity values are given for information only in Table 3 for the low ohmic value resistors. It is clear that these values should match only the resistor. In reality, it is difficult to

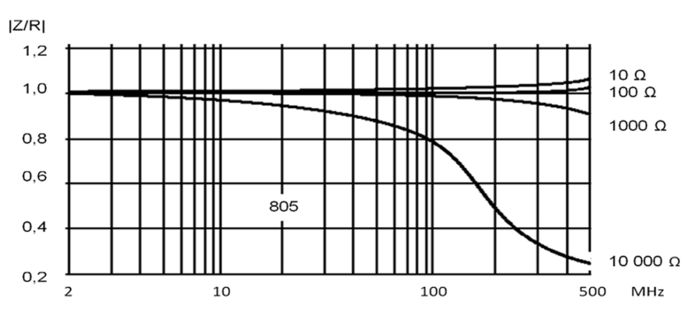

Fig 8: Example of SMD thin layer 0805 behaviour vs. frequency.

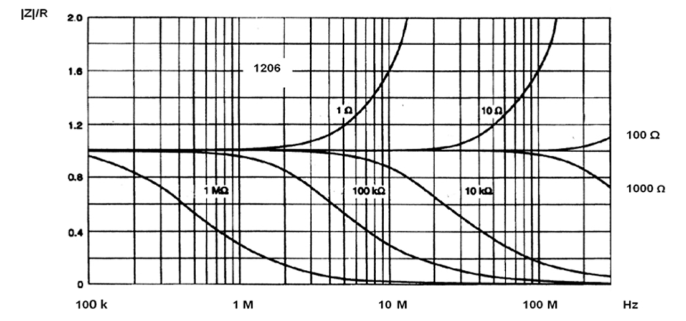

Fig 9: Example of SMD thick layer 1206 behaviour vs. frequency.

escape additional parasitics introduced by the measurement bench. This is the reason why the values given by manufacturers for the same package are widely dispersed.

It is expected the same is true when mounting SMD resistor on a printed circuit board.

More easily and more effectively, manufacturers publish curves giving the impedance and phase shift of their resistors vs. frequency. See examples in Fig 8 and 9.

As stated earlier resistors are trimmed during manufacture to adjust the value. But, by extending the active effective length by a cut for chips or a spiral for the cylindrical, this also increases the inductance (Fig 10 and 11). There are chip and MELF resistors specifically designed for microwaves (≤20GHz) with well-specified and reduced parasitic elements [7,8]. They are available from distributors for non-professionals for 10, 50, 100, 500Ω and 1kΩ values in 0402 packages. Their cost is of the order of 3 to €5 each! The cylindrical MELF resistors can be housed in a coaxial structure that may result in very good behaviour (50Ω load).

Another effect tends to increase the resistor value with the frequency: the skin effect. This especially penalises thick film and reaches 20µm, less for thin films of the order of a µm. This means that the increase in value is negligible to 5GHz only thin film layers.

Fig 12 gives the curve of a special 0402 resistor usable to 20GHz.

11.

Application: A homemade 50Ω 10GHz load

Thanks to F1EER, here is a valuable homemade load for our 10GHz band. It is made with a standard 18GHz bulkhead SMA and two 0805 SMD 100Ω thick layer resistors (Fig 13). "Stubs" are placed carefully for best results. The

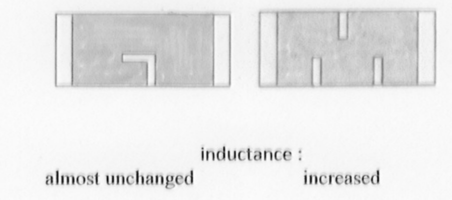

Fig 10: Chip resistor trimming.

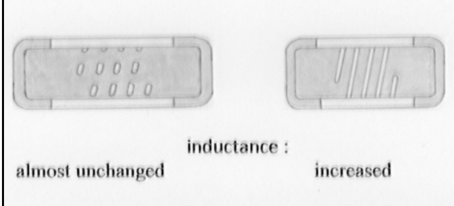

Fig 11: Trimming of cylindrical resistor by sanding or laser.

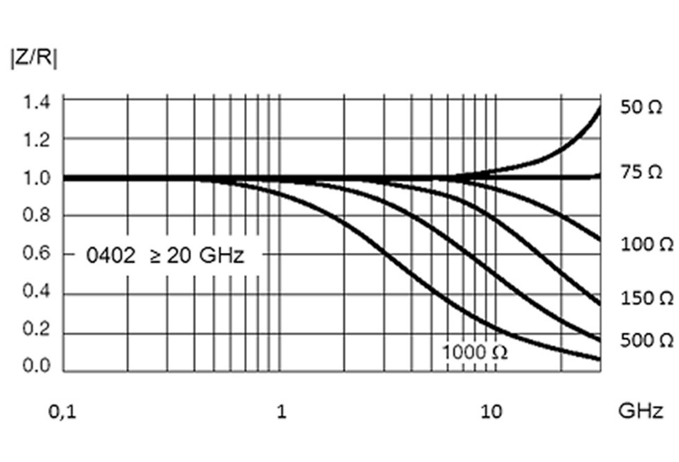

Fig 12: Frequency behaviour of a special 20GHz SMD.

ROS is 1.15 to 10GHz. Tests with a single 50Ω or three 150Ω in parallel have given worse results.

Fig 14 curves have been drawn with an R & S Vector Network Analyser ZVA67 67GHz, cables and calibration in SMA 2.9 kit.

For comparison, an unidentified professional SMA load is shown in Fig 15. In both cases, permanent maximum power is 0.5 Watts and can reach 1W for a few minutes.

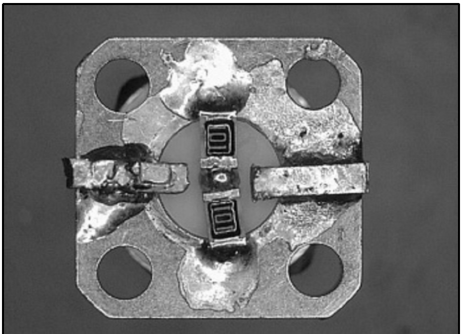

Fig 13: Homemade 50Ω load.

12. Conclusion

In electronics devices, a resistor is a component that seems very banal but can give problems if used without knowing the rules.

In practice, as radio amateur, I use 1% chips without worrying whether it is thick or thin layer. SMD recovery is not to advise given their very low prices and the risk of weakening.

In the professional field, a more cautious approach must be taken, study the manufacturer data for certified equipment choose the appropriate type of resistor.

13. References

[1] Become familiar with ceramic capacitors for good use from DC to microwaves, F9HX, VHF Communications Magazine, 4/2011 pp237 - 241

[2] Fixed Resistors, Data Handbook, Philips, 1994

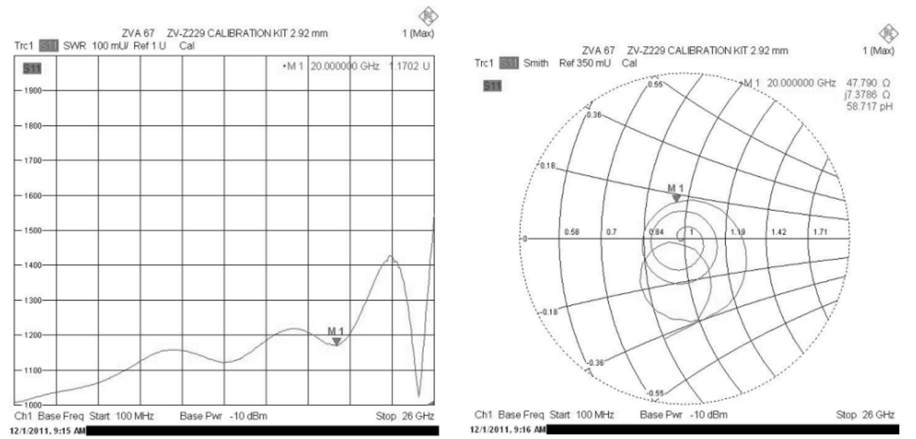

Fig 14: Homemade 50Ω load.

[3] http://talkingelectronics.com/ChipDataEbook-1d/html/SM-Resistors.html

[4] Résistances en VHF et UHF, F9HX, Radio-REF 1 et 2/1995

[5] Fundamentals Passive Components and their EMC Characteristics, Lars-Olov Johansson, Compliance Engineering European Edition, September/October 1998

[6] SMD Resistors Beyond UHF, W. Laurich, BEYSCHLAG GmbH

[7] High Frequency (up to 20 GHz) Chip Resistors, VISHAY, www.vishey.com

[8] Frequency Response of Thin Film Chip Resistors, Vishay, 4/02/2009

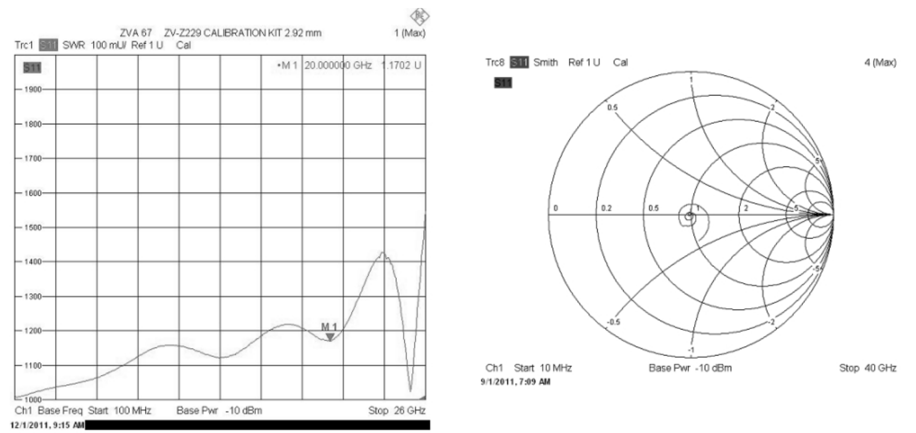

Fig 15: Professional 50Ω SMA load.

Andy Barter, G8ATD

Parabola Calculator

During April Mike Scirocco contacted me to see if I could help him locate Jiri Otypka, who wrote an article called "Calculating the Focal Point of an Offset Dish Antenna" that was published in issue 1/1995 of VHF Communications Magazine. Unfortunately that was well before I took over the publication of the magazine so I gave him another contact to try. While having an email exchange with Mike he told me about the programs that he had already written and they sounded useful to anyone investigating the construction of a parabolic dish. I asked Mike if he would write an article for the magazine but he has left it up to me to try my best to present his programs, so here goes.

1. Introduction

There are two programs:

- The first program is Parabola Calculator version 2.0 [1]. This is a Freeware program written to help with the design solar collectors or Wi-Fi projects using parabolic reflectors. Whether you're improving the signal strength of a Wi-Fi antenna, or designing a satellite antenna or solar trough, this program calculates the focal length and (x, y) coordinates for a parabola of any diameter and depth. It can help you determine what size and shape to make your parabola very quickly. Version 2 includes Wi-Fi calculations for centred or offset feedhorn dishes. If you would like modifications to the program to make it more useful Mike is willing to take suggestions [2]. This may be particularly useful for any readers interested in modifications to help add facilities that would be more useful to radio amateurs.
- The second program is Parabolic Leaf Calculator. This is a Freeware program that Mike has allowed me to put on the VHF Communications web site [3]. It calculates the dimensions of leaves that can be fitted together to form a parabolic reflector. This reminded me of some of the "fold up" dishes that I have seen used by amateurs on EME Dxpeditions, so I thought it would be a useful addition to the downloads from the site.

2. Parabola Calculator 2.0

The program is easy to install on a

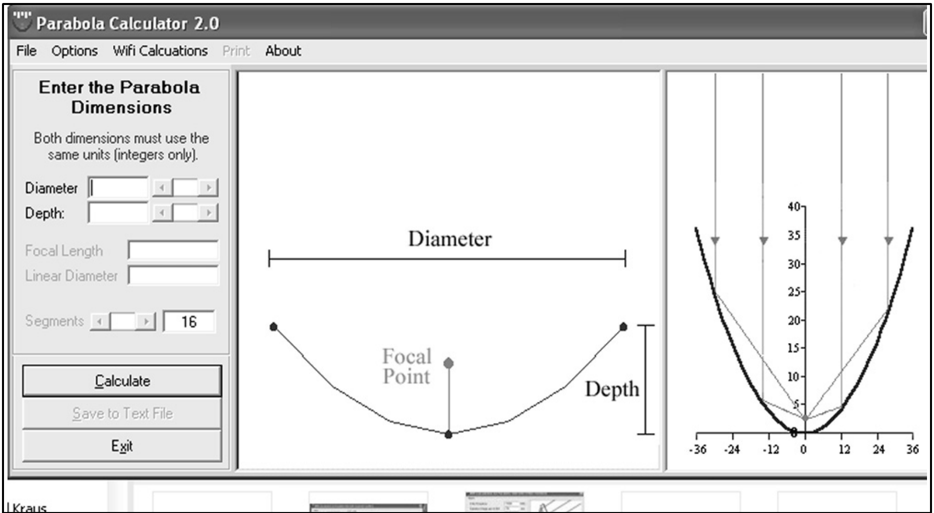

Fig 1: The opening screen of Parabola Calculator version 2.0.

Windows PC, just download the Windows installer from the web site [1], unzip the files and run the setup program. When you load the program the window shown in Fig 1 appears.

The default mode is to specify the diameter and depth of the dish. It is probably more useful to radio amateurs to use the alternative of diameter and f/D ratio. To switch to this mode click on "Options" in the menu bar and choose the "Select Inputs" option followed by the "Diameter and f/D" option.

The window shown in Fig 2 will now be displayed. After entering the Diameter required and the f/D ratio, press the calculate button to display the parabola. The Diameter and the f/D ratio can be

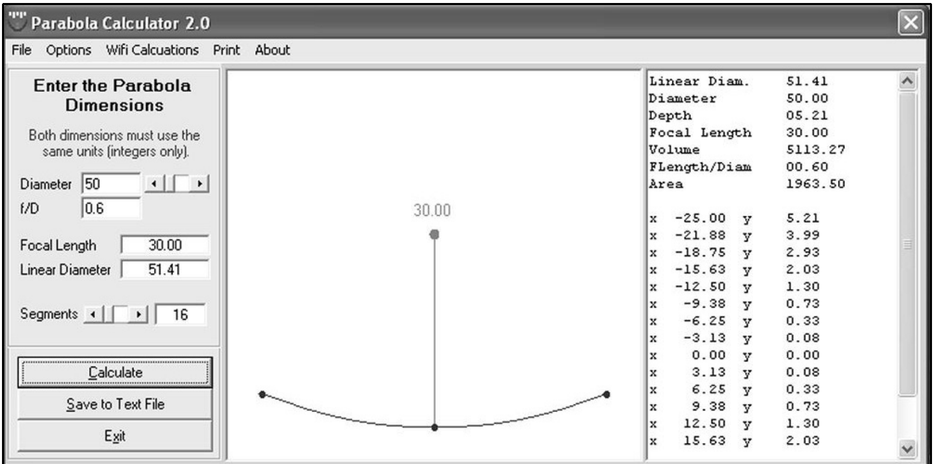

Fig 2: Parabola Calculator version 2.0 in f/D calculation mode.

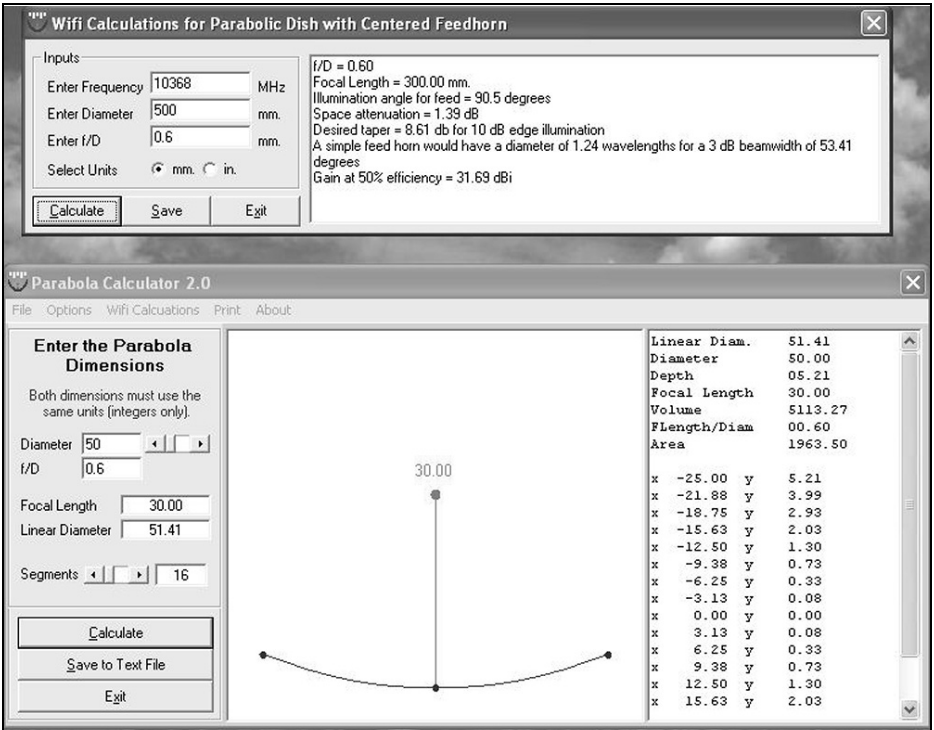

Fig 3: Parabola Calculator version 2.0 with one of the Wi-Fi calculator window open.

incremented or decremented using the arrow buttons next to the data, the image of the parabola changes at the same time.

The number of segment used to draw the parabola can also be changed to increase or decrease the accuracy of the image. The coordinates of the parabola are displayed on the right had side of the window. These can be saved as a text file or an input file that can be used by a graphics or CAD program.

If you want to know the calculated gain of the dish select the "Wifi Calculations" option from the menu bar and then the "Centred Feedhorn: Calculate Gain, f/D" option. This will display a new window with the diameter and f/D ratio selected previously as defaults and a box to enter the frequency. Enter the frequency and select the appropriate dimensions and the program will calculate the gain etc. as shown in Fig 3.

There is a second option on the "Wifi Calculations" menu. That is "Offset Feedhorn: Calculate Gain, f/D". Selecting this option opens a new window with a stand-alone program that does calculations for an oval-shaped offset-fed parabolic reflector. This routine uses a curve-fitting algorithm to find the focal point and tilt angle for aiming the dish. The input data required are the frequency to use, the dimensions of the large and small axis of the oval and the depth and location of the deepest point in the reflector measured along a straight edge placed across the rim on the large axis. The WiFi calculations and output text are copied directly from the source code of the hdl_ant program written by Paul

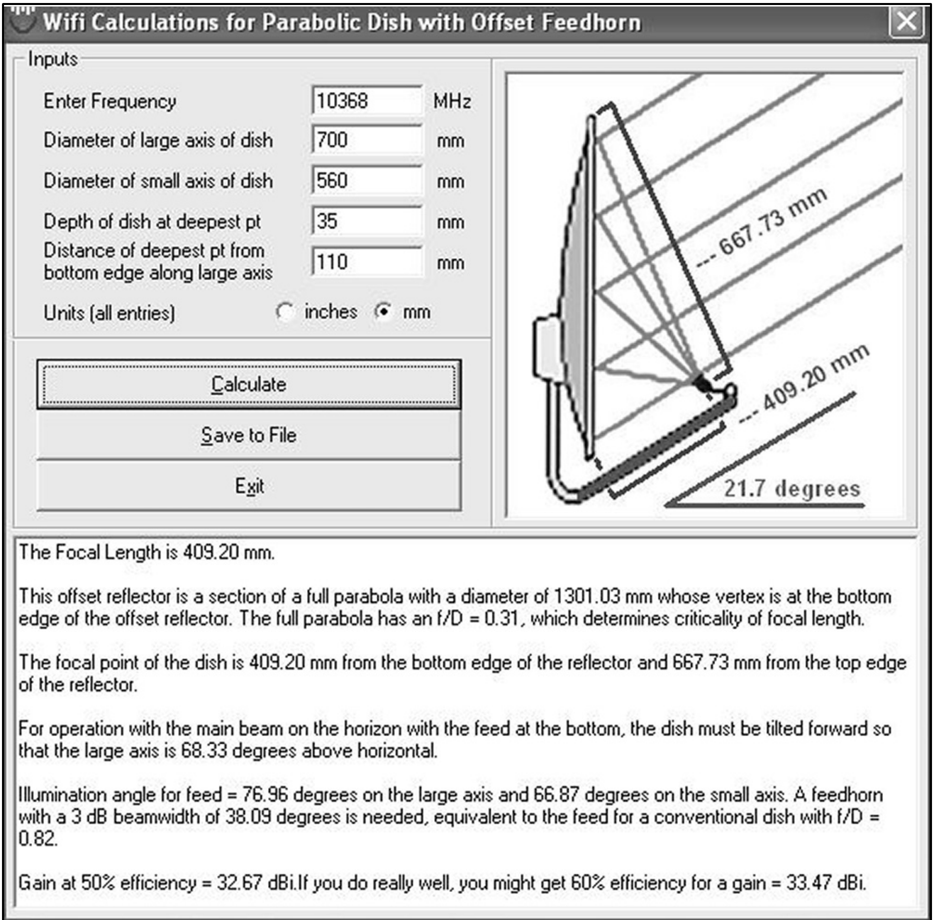

Fig 4: Parabola Calculator version 2.0 showing the offset dish calculator window.

Wade [4]. The result is shown graphically on the right hand side of the window as shown in Fig 4. The values that I entered were taken from my Freesat dish.

3. Parabolic Leaf Calculator

The program is easy to install on a Windows PC, just download the zip file from the web site [3], unzip the files and run the setup program. When you load the program the window shown in Fig 5 appears.

The diagram on the right hand side of the window show exactly what the program does. It calculates the dimensions of the triangular leaves that are needed to manufacture a parabola shaped reflector. The number of leaves to be used is one of the parameters that can be set when the calculation is performed.

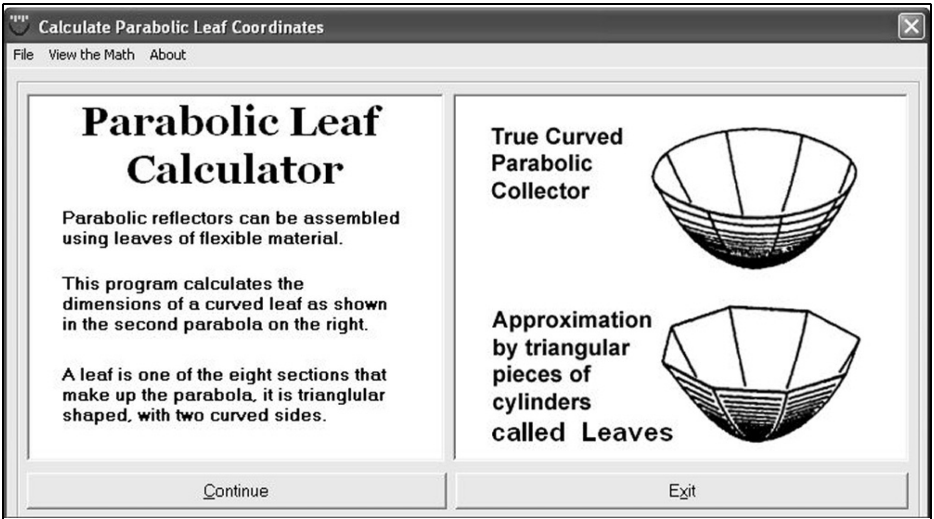

Fig 5: The opening screen of Parabola Leaf Calculator.

To show the calculation screen (Fig 6) from the initial window, click on “Continue”. Enter the focal length, radius and select the number of segments. Press “Calculate” to show the result graphically and as data on the right hand side of the window. The data can be saved as a text file or an input file that can be used by a graphics or CAD program.

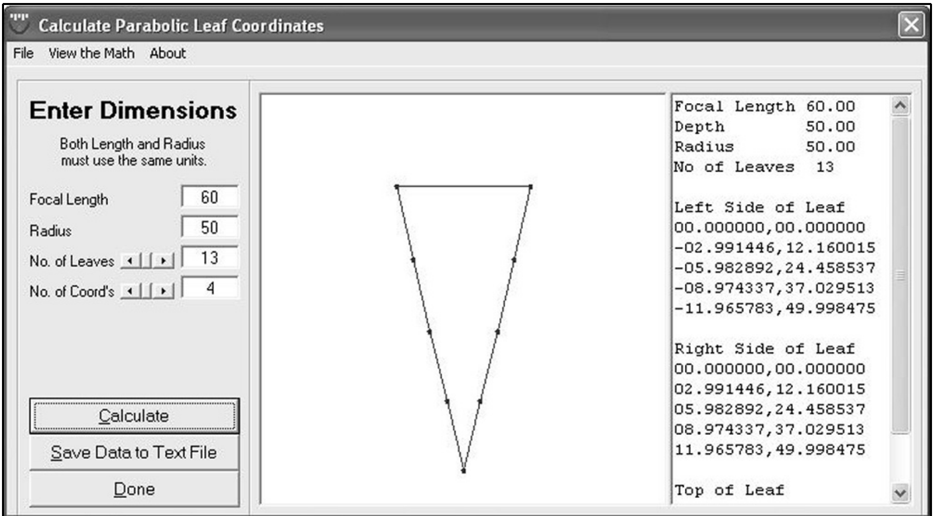

Fig 6: Parabola Leaf calculator results screen.

4. Conclusion

The two programs that Mike Scirocco has produced are easy to use and perform valuable calculations for anyone designing a parabolic dish. The programs were written in response to a requirement for those interested in designing Wi-Fi dishes or Solar Troughs. If Mike can get some feedback from amateur radio users of the program he is happy to make modifications or enhancements to provide facilities that are of use in the design of dishes for use on the amateur radio bands.

5. References

[1] Parabola Calculator version 2.0 is available to download from - http://mscir.tripod.com/parabola/

[2] You can send any comments about the programs or request for enhancements to Mike Scirocco, email: mscir@yahoo.com

[3] The Parabolic Leaf Calculator can be downloaded from the VHF Communications web site: http://www.vhfcomm.co.uk/index/download/parabolaleaves.zip

[4] Paul Wade, W1GHZ web site: http://www.w1ghz.org/10g/software.htm

H.J. Griem, DJ1SL

Tubular radiator for parabolic antennas on the 13cm band

(reprint from issue 4/1976)

1. Application

A parabolic reflector is to be illuminated with the aid of a round waveguide. A coaxial waveguide interface is required if this round waveguide is to be energised with the aid of a coaxial cable. In studying American literature [1 to 6], it was found that such arrangements had been constructed and used but that no general design details were provided.

For this reason, a basic research was carried out to find out the basic radiation geometrics of parabolic antennas and to find suitable feed systems. This was followed by designing, an actual tubular radiator for 13cm to fit an available parabolic reflector of 120cm dia. with a focal point / diameter (F/D) = 0.375.

The result of these measurements and experiments allow design details to be calculated for all other parabolic antennas, and microwave amateur bands. This means that it is possible with a minimum of mechanical work to illuminate parabolic reflectors for amateur radio applications.

2. Radiation Geometrics

2.1. Parabolic reflectors

The beamwidth or focal angle of a parabolic reflector is mainly dependent on the ratio F/D. This relationship is shown in Figure 1 where F is the focal point of the parabolic, and D is the diameter of its aperture. The focal angle is 2φ, since φ is measured from the centre point to one side. If the focal point F is not known, it can be calculated from the diameter D and the depth d according to the following formula 1:

$$F = \frac{D^2}{16d} \qquad (1)$$

The actual positions where these values can be measured are shown in the drawing given in Figure 1.

From the dimensions of the given parabolic reflector, it will be seen that the focal angle amounts to 135°. This angle should be illuminated as completely as possible by the primary antenna. The 10dB beamwidth (not the 3dB beamwidth as is usually used) of the primary (or exciter) antenna should

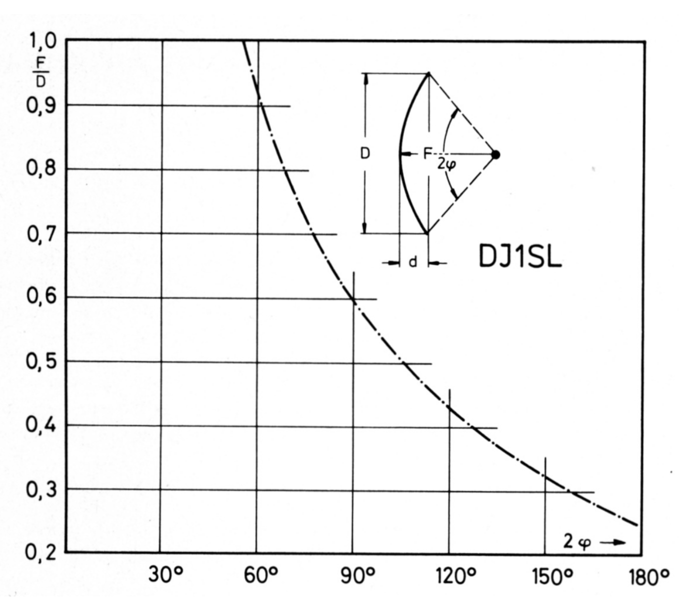

Fig 1: Focal angle of parabolic antennas as a function of F/D.

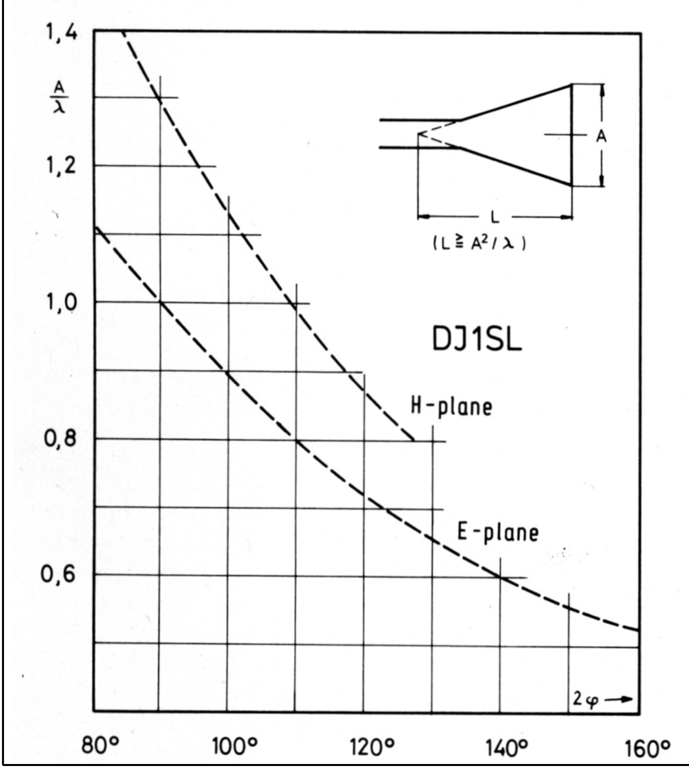

Fig 2: 10dB beamwidth 2φ of horn radiators.

amount to just 135° in the case of our example. If the beamwidth is less than this, the parabolic reflector will not be completely illuminated, and the efficiency and antenna gain will be less than the maximum available. If, on the other hand, the beamwidth of the primary radiator is greater, radiation will be made past the edge of the parabolic reflector, which will deteriorate the front-to-back ratio, and also decrease the antenna gain. A favourable compromise is found between antenna gain and clean polar diagram when the 10dB beamwidth of the primary radiator responds to the focal angle 2φ of the reflector. If these prerequisites are fulfilled, the gain formulas given for parabolic antennas will normally be valid.

A dipole with reflector is often found as primary antenna for parabolic reflectors. It is not easy when using such an arrangement to obtain the required matching of the 10dB beamwidth to the parabolic reflector to be illuminated as when using radiating surfaces such as given in [8]. For this reason, waveguide radiators such as horn antennas should be used, since the beamwidth can be easily determined with the aid of the, dimensions of the aperture.

2.2. Horn radiators

Figure 2 gives the 10dB beamwidth of a so-called horn radiator as a function of its aperture. It will be seen that the beamwidth decreases on increasing the aperture. This diagram, which is also given in [7], is valid both for round (A = diameter) and square (A = side length) cross-sections. In the case of a parabolic antenna with a focal angle of 135°, a horn radiator also having a beamwidth of 135° will be required; this corresponds to a horn radiator having A/λ corresponding to 0.63. The differing values for the H and the E-planes which were mentioned in [8] are not of interest here. With λ = 130mm, "A" will correspond to: 0.63 x 130 = 82mm.

3.

Tubular radiator

Due to its ease of construction, and since tubes are easier to obtain than rectangular profiles, a round waveguide a so-called tubular radiator is to be used. Brass tubes are usually available in diameters with steps of 5mm. It has been seen in the previous section that a diameter of approx. 82mm is required. The dimensions of the round waveguide are now to be established.

3.1. Diameter of a round waveguide

Without going too deeply into the theoretical relationships in waveguides, the permissible diameters of a round waveguide are to be given which is to be operated in a single-mode range. This range is between the limit wavelengths of the H_{11} and the E_{01} wave. The diameter range for this is given in equation: 2, and the first number of the nominator is valid for the H_{11}-wave and the second number for the E_{01}-wave:

$$D = \frac{\lambda}{1.71 to 1.31} \qquad (2)$$

It will be seen that the diameter in which only the H_{11}-wave is possible, is in the range of 76 to 99mm for L = 130mm (f - 2.31GHz).

Due to the increasing loss in the vicinity of the wavelength limit, the diameter of the waveguide should be as large as possible, without going into the range of the next higher oscillation mode, in our case the E_{01}-wave. For this reason, a diameter of 88mm (outer diameter 90mm, wall thickness 1mm) was selected for the, 13cm band. The H_{11}-wavelength limit then amounts to 151mm corresponding to 1.99GHz.

3.2. Waveguide wavelength

The wavelength of a waveguide is not

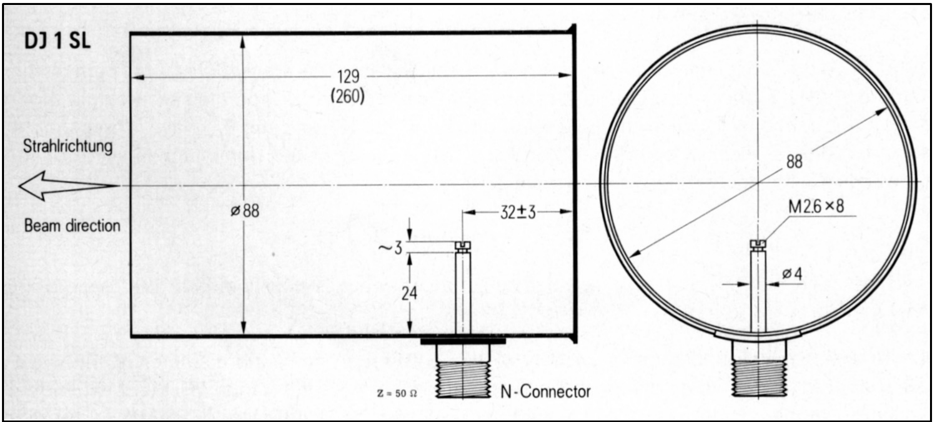

Fig 3: Dimensions of a tubular radiator for the 13cm band.

identical to the wavelength of a RF signal in free space. The waveguide wavelength λ_W is greater than in free space. The relationship between these two values is given in equation 3:

$$\lambda_W = \frac{\lambda}{\sqrt{1-\left(\frac{\lambda}{\lambda_l}\right)^2}} \qquad (3)$$

Where λ_1 is the limit wavelength of the type of wave under consideration.

A value of 151mm was obtained from equation 2 as the limit wavelength for the H_{11}-wave. In the case of the selected diameter of 88mm for the tube, the calculated wavelength of the waveguide amounts to 258mm (free-space wavelength. 130mm). This value is required later for calculating the length of the tubular radiator.

3.3 Construction of the tubular radiator

It was mentioned in section 2.2. that the horn radiator should have an aperture of 82mm. According to section 3.1., a waveguide diameter of 88mm would be favourable. A diameter of 88mm is within the permissible range between 76 and 99mm, and a horn shaped aperture of the radiator will not he required. For this reason, the radiator has been called: tubular radiator.

The question was now which length should be selected for the radiator since various conflicting information was to be found in the various sources. The various lengths varied from 2λ, down to considerably lower values, which meant that the most favourable length had to be established with the aid of measurements. A length of 2λ, was chosen for the start of the measurements which corresponded to a length of 260mm. As will be seen in section 4, the measurements showed that the optimum length is λ/2, when referred to the waveguide wavelength λ_W. This results in a mechanical length of 129mm for the 13cm band.

Figure 3 gives the dimensions of the final tubular radiator. The interface between the coaxial cable (N-connector) and the waveguide mode is clearly shown. This can be classed as a type of vertical antenna (ground plane), however, its length is far shorter than λ/4. The spacing of this radiating element to the rear panel of the tubular radiator, and the length of the actual element are adjustable. The given dimensions have been found to be most favourable for the 13cm band in the following measurements.

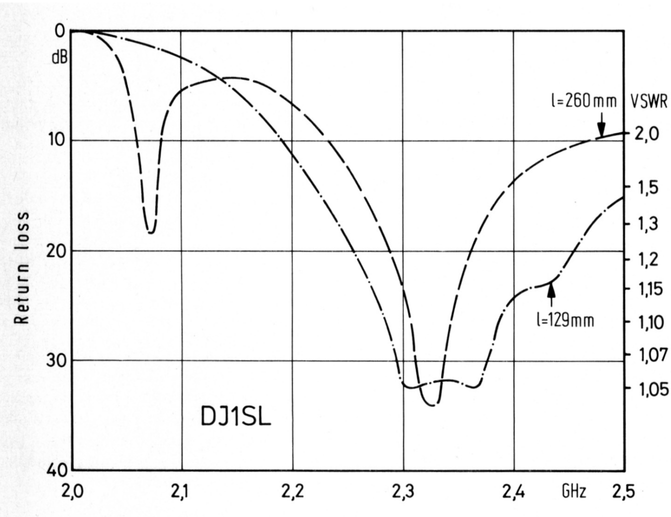

Fig 4: Matching of a tubular radiator in a frequency range from 2.0 to 2.5GHz.

4. Measurements

The following measurements were made to establish the two most important characteristics of the tubular radiator: the radiation diagram and the matching to the coaxial cable. The radiation diagram should fulfil the demand of illuminating the parabolic reflector so that 10dB power reduction is present at the edge of the reflector.

These measurements were carried out using an amplitude reflection measuring system HP8755 having an impedance of 50Ω. This measuring system allowed both the insertion loss and the return loss to be measured simultaneously. The return loss could be read off directly, whereas the insertion loss was measured by measuring the attenuation from the tubular radiator to an auxiliary dipole placed several meters away. If the spacing between these two antennas is kept constant, the radiation diagram can be obtained by rotating the tubular radiator in its axis, and giving the loss as a function of the various angles.

4.1. Matching

Figure 4 gives the diagram of the return loss over the frequency range of 2GHz to 2.5GHz. The amateur band is from 2.3 to 2.35GHz. The measuring range for the return loss amounted to 0 to 40dB which corresponds to the VSWR values given on the right side of the diagram. The reference impedance is 50Ω.

The "vertical antenna" was shifted with respect to the rear panel of the tubular radiator and the length of the element varied with the aid of a screw in order to obtain the lowest VSWR within the amateur band. It was interesting to find that a high return loss could be obtained at 2.3GHz that could be shifted in frequency somewhat using the described means; however, a peak in the return loss was also found at 2.08GHz that could not be affected by these adjustments. By calculation, it was found that this resonance was caused by the length of the tubular radiator; since the length of 260mm corresponds to a waveguide

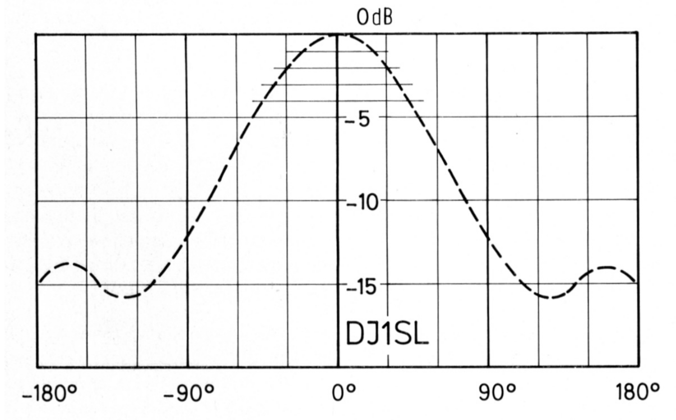

Fig 5: Radiation diagram (E plane) of the tubular radiator.

wavelength of λ/2 at a frequency of 2.08MHz. This dimension indicated a natural matching of the tubular radiator to space, and further experiments were made to establish further details.

In order to do this, the length of the tubular radiator was calculated so that it corresponded to exactly λ/2 of the waveguide wavelength in the 2.3GHz amateur band. This resulted in a mechanical length of 129mm. The measurements made on this radiator, and the results were as expected. A resonance condition existed at 2.3GHz which was not affected by alignment of the energising element, whereas a second resonance point could be varied. By combining these two matching points, it was possible to obtain a bandpass characteristic which allowed matched conditions to be obtained over a far wider frequency range than was the case with the 260mm long tube. The return loss amounted to approximately 32dB over the whole amateur band, which corresponds to a VSWR of 1.05 (Figure 4). This is a very good value when one considers that the tolerance in the impedance of coaxial cables amounts to 5%, and thus is in the same order of magnitude.

The described experiments were made in a large room where the radiator was either able to radiate freely through a window, or against the ceiling (2m). In order to establish what effect the parabolic reflector would have on the matching, a metal plate of approximately one square meter was moved towards the radiator. It was found that a slight fluctuation of the return loss was indicated, which increased the nearer the plate came to the aperture of the tubular radiator. These fluctuations remained in the order of 30dB (min. 25dB) up to a spacing of approximately 40cm. This corresponds to a VSWR of approximately 1.1, which is still an excellent value that would not be measurable with amateur means. For those readers that have access to such a measuring system, it is possible for the complete antenna including parabolic reflector to be measured and the optimum matching obtained by varying the coaxial element.

4.2. Radiation diagram

The second task was to measure the radiation diagram of the tubular radiator, which of course could not be carried out indoors due to reflections which were found to considerably distort the radiation diagram. On the other hand, a small spacing between the tubular radiator and the auxiliary dipole could cause adjacent effects. For this reason, the measurements were made out-of-doors.

The measuring dipole was mounted 4 to 5 meters away from the tubular radiator

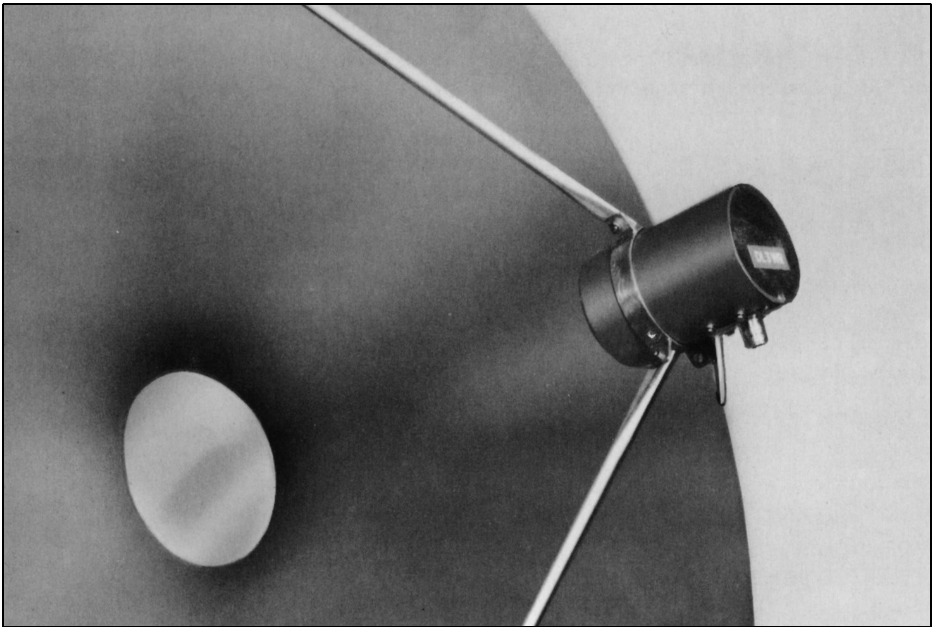

Fig 6: 1.2 meter parabolic with tubular radiator for the 13cm band (G_{dipol} approx. 24 - 25dB).

and the loss between the tubular radiator and auxiliary dipole was measured with the measuring system. The tubular radiator was rotated in its vertical axis. A second series of measurements was made after the tubular radiator had been rotated by 90°. The vertical plane diagram was obtained finally by rotating the tubular antenna by 360°.

The result of these measurements is given in Figure 5. It will be seen that the signal drop of 10dB with respect to the maximum value is obtained at an angle of approximately 80°. This corresponds to a total beamwidth of 160° which is somewhat more than the ideal value required for the parabolic reflector of 135°. At 135°, this value amounts to 9dB which is tolerable. It should also be taken into consideration that this measurement of the radiation diagram was not made at highest precision since the measuring arrangement was too simple for this. Due to a lack of time, it was not possible to repeat these measurements under laboratory conditions.

It was found during these measurements that considerably more time would have been required without sweep generator in order to obtain the same results. For instance, if the measurements were only made within the amateur band, the resonance position at 2.08GHz would not be discovered, and thus the cause of any matching difficulties.

5.

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Gunthard Kraus, DG8GB

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Bama

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For people who deal with high frequencies at high powers and therefore are dependent on waveguides, here is a set of useful on-line calculators.

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Publishers
KM PUBLICATIONS,
503 Nothdown Road, Margate,
Kent, CT9 3HD, United Kingdom

Tel: +44 (0) 1843 220080
Fax: +44 (0) 1843 220080

Email:
andy@vhfcomm.co.uk

Editor
Andy Barter G8ATD

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AUSTRALIA - Mark Spooner c/o, W.I.A SA/NT Division, GPO Box 1234, Adelaide, SA 5001, Australia Tel/Fax 08 8261 1998

BELGIUM - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133 7798-0. Fax: 09133 779833.
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Fax: +44 1843 220080. Email: andy@vhfcomm.co.uk

FRANCE - Christiane Michel F5SM, Les Pillets, 89240 PARLY, France
Tel: (33) 03 86 44 06 91, email christiane.michel.s5sm@orange.fr

FINLAND - KM PUBLICATIONS , 503 Northdown Road, Margate, Kent, CT9 3HD, UK. Tel: +44 1843 220080.
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SWEDEN - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: +46 705 164266; GSM: 0705 16 42 66,
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SWITZERLAND - KM PUBLICATIONS , 503 Northdown Road, Margate, Kent, CT9 3HD, UK. Tel: +44 1843 220080.
Fax: +44 1843 220080. Email: andy@vhfcomm.co.uk

UNITED KINGDOM - KM PUBLICATIONS , 503 Northdown Road, Margate, Kent, CT9 3HD, UK. Tel: +44 1843 220080.
Fax: +44 1843 220080. Email: andy@vhfcomm.co.uk

U.S.A. - ATVQ Magazine, Mike Collis, WA6SVT, P.O. Box 1594, Crestline, CA, 92325, USA,
Tel: (909) 338-6887, email: wa6svt@atvquarterly.com

ELSEWHERE - KM PUBLICATIONS, address as for the U.K.

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

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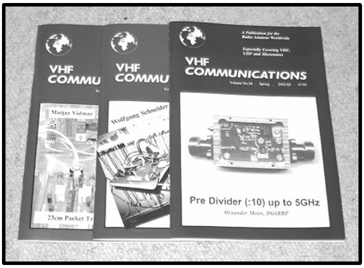
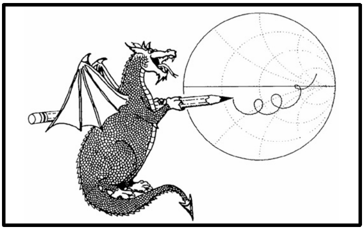


Back Issues

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.

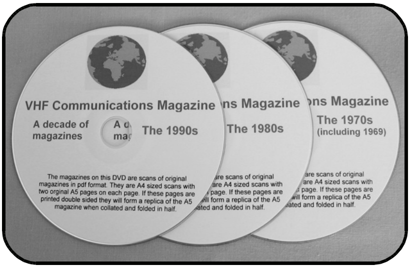
K M Publications, 503 Northdown Road, Margate, Kent, CT9 3HD, UK

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

Back Issues on DVD

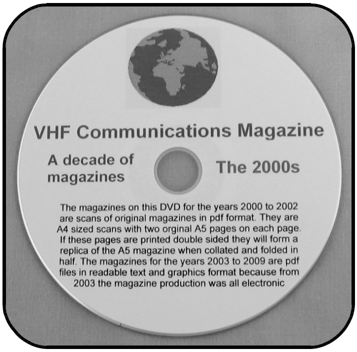
VHF Communications Magazine has been published since 1969. Up to 2002 it was produced by traditional printing methods. All these back issue have been scanned and converted to pdf files containing images of the A4 sheets that formed the A5 magazine when folded in half. These have been put together on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



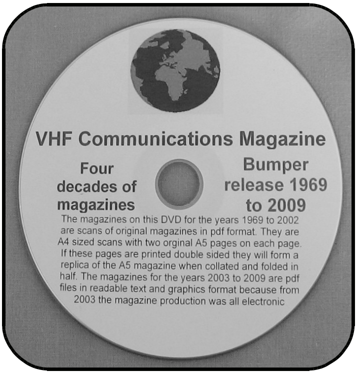
1970s - 1980s - 1990s

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