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**Display unit for the power detector by  
DJ8ES**

*Alexander Meier, DG6RBP*

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*Thank you to everyone who has renewed their subscription for 2007. This issue sees the return of an article from Franco Rota, he has been working hard to find an interesting topic and I think he has succeeded. I have also been asked to publish the rules for a Spanish 6m contest that will hopefully increase activity on 6m in Spain.*

*73s - Andy*



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Alexander Meier, DG6RBP

# Display unit for the power detector by DJ8ES

**In the article by Wolfgang Schneider on his power detector, published in issue 3/2006 [1], he mentioned a display unit. This indicates the measured power, on five 7 segment displays, in dBm or Watts.**

**With very little extra hardware cost an offset function can be added. This can be used to offset the frequency response of the sensor for more precise measurements or the use of a coupler for measuring higher powers.**

With the latter, for example, the variations in the level response of the detector can be taken into consideration for particularly precise measurement. If an external coupler is used to be able to measure higher power levels in the Watt range, the extra attenuation can be taken into account for the displayed value. Powers up to +60dBm (1000W) can be displayed. The circuit to do all this can be realised using very few components.

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## 1.

### Introduction

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Because the 12 Bit A/D converter is integrated on the detector PCB [1], the measuring signal can be transferred simply with a multi way cable from the sensor to the display unit. The digital signal is fed directly to an AT89C52 micro controller from ATMEL, this computes the measured power in dBm. The use of a micro controller makes it possible to add further, practice oriented features like: relative measurements (display of the level relative to a measured value), the conversion and display of the power in Watts, a PC interface (RS-232 or USB) and the consideration of an offset.

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## 2.

### Description of the circuit

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The circuit diagram of the display unit is shown in Fig 1 and is the same as the circuit of the frequency input module [2] as far as possible. This display unit for the power detector has ESD protection on the detector input, several light emitting diodes and a serial interface. The serial interface is 0/5v level suitable for USB but can have an interface converter added such as the MAX232 [4] for a conventional RS-232 interface. A direct connection between the display unit and an RS-232 interface of a PC without an interface converter is possible.

The technical data of the display unit

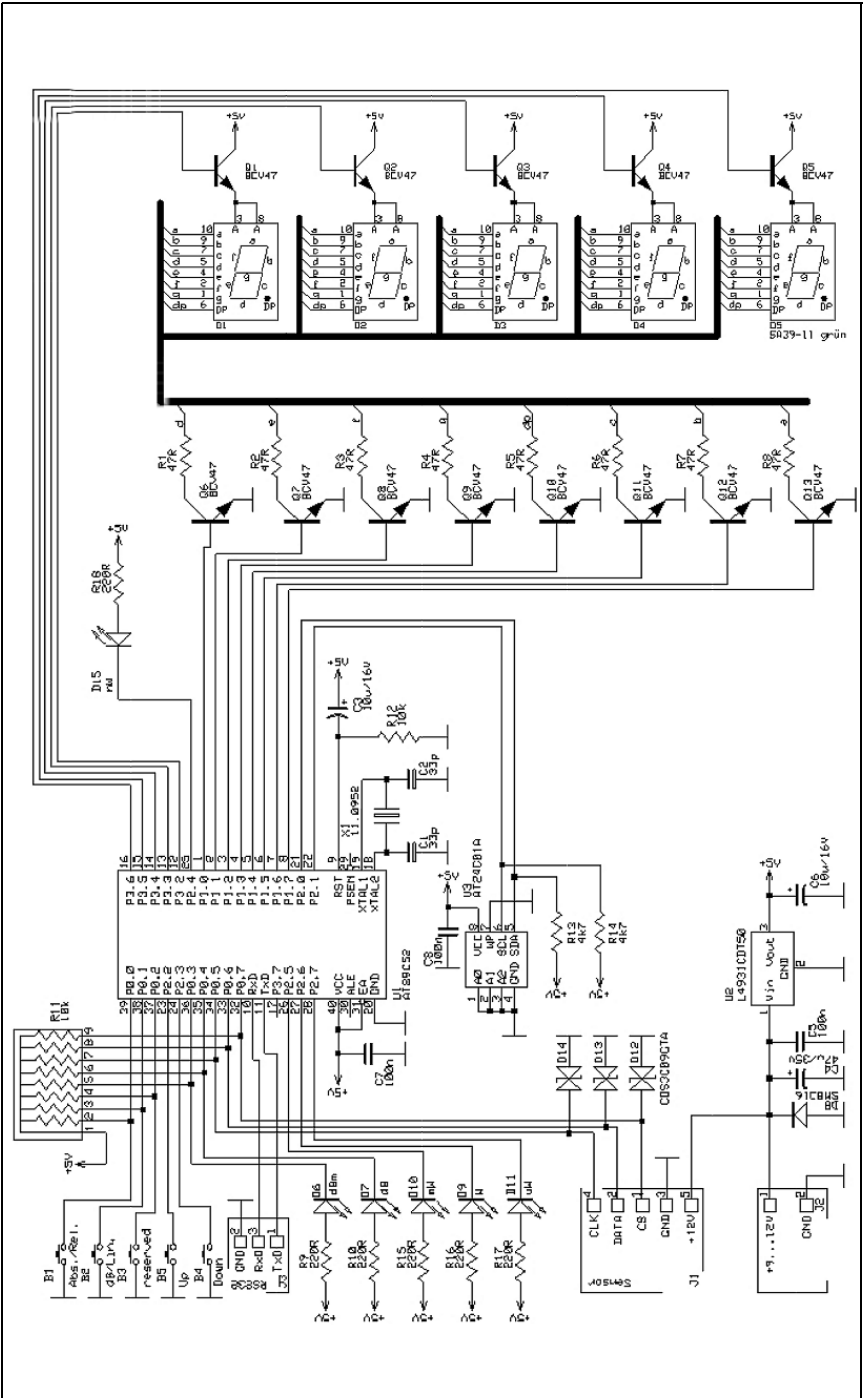


Fig 1 : Circuit diagram of the display unit.

**Table 1 : Technical data for power detector and display unit.****Power Detector (with DG6RBP PCB)<sup>1</sup>**

Frequency response	10MHz to >2.3GHz
Measuring range	-50dBm to +10dBm
Typical resolution	0.02dB
Input impedance	50Ω
Typical level error	±1dB
PCB size	62 x 18 x 0.6mm

**Display Unit**

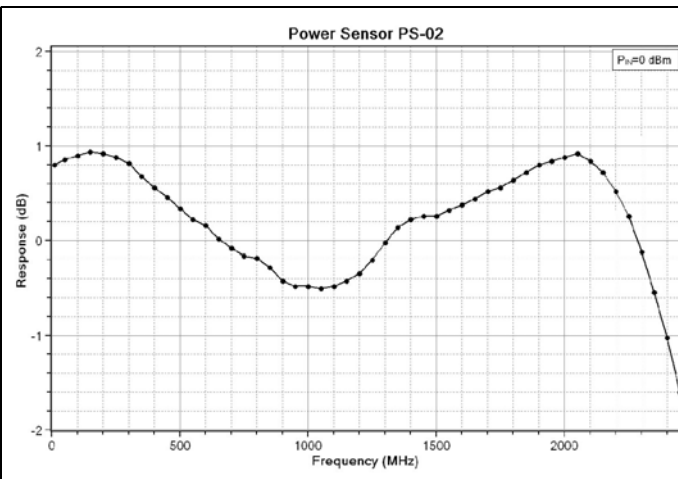
Display range	-55.0dBm to +59.9dBm 3.00nW to 995W (3 digits)
Display resolution	0.1dB (for display in dBm) 0.02dB (for display in watts) <sup>2</sup>
Measurement interval	4 measurements per second
Reference level for relative measurements	Last value used before relative function
Attenuator compensation function	-1.9dB to +49.9dB (adjustable)
Interface <sup>3</sup>	Serial, 19200 Baud, 8 data bits, 1 stop bit, no parity, 0/5V level
Power supply	9 - 12V approximately 130mA
PCB size	47 x 90 x 1.5mm

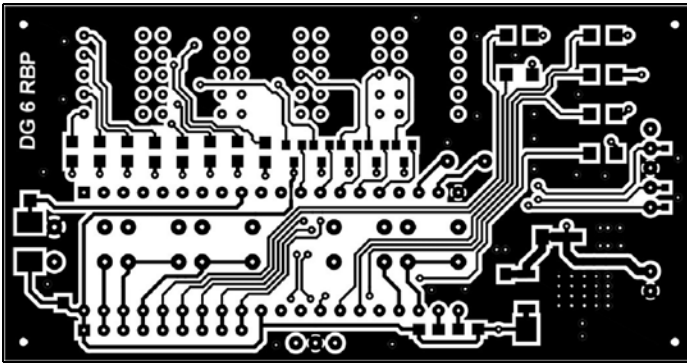
with the power detector is summarised in Table 1.

The main part of the circuit is the micro controller U1 (AT89C52) from ATMEL. This frequently used and proven controller is inexpensive and has an integrated Flash memory to store the software. An EEPROM (U3) is used to permanently

store static data e.g. sensor calibration. The switches and LEDs needed to operate the unit are mounted on the PCB with the micro controller. The use of SMD LEDs and light pipes on the front panel increases the ESD protection.

Port 1 and 3 of the micro controller drive the five 7 Segment displays to show the

**Fig 2 : Frequency response of the power detector using optimised PCB.**



**Fig 3 : Layout for the front side of the display unit PCB.**

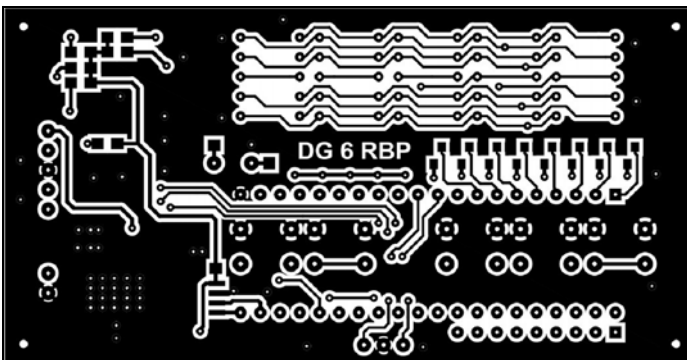
measured values. These displays, D1 to D5, are activated alternately for 1 ms via their common anode, this reduces the current consumption. The power input of the circuit is thus below 150mA.

The power detector is connected by the plug, J1, directly to the micro controller. All connections are protected against electrostatic discharge using diodes. The supply voltage for the sensor can be fed via the plug connector J1. A suitable connector on the front panel of the prototype wattmeter was a mini DIN socket (with 6 pins) which worked satisfactorily.

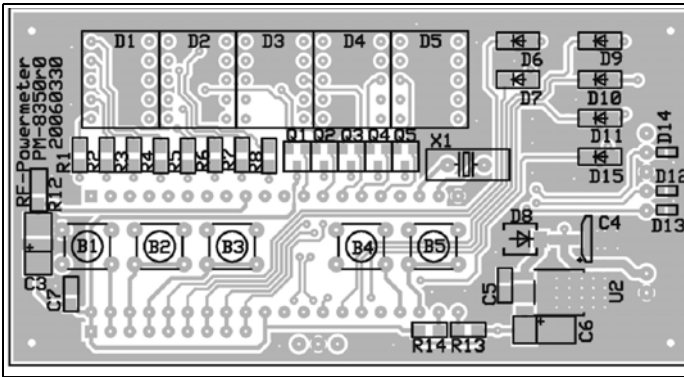
The circuit is powered by an external 9 to 12V DC supply. The voltage regulation consists of an internal 5V regulator for the display unit. The operating voltage for the sensor is stabilised on the sensor printed circuit board. [1]

Explanations for the data from Table 1:

- <sup>1</sup> For the power detector [1] a somewhat smaller printed circuit board with optimised conductors and smaller SMD parts was designed. Also ESD protection was added on the connections. A large capacitor (10 $\mu$ F) is used on port CLPF so that various modulation signals e.g. AM Modulation, can be measured correctly. Fig 2 shows the improved frequency response of the sensor. The small size of the 0603 SMD components used gives an optimised response.
- <sup>2</sup> The power is measured in dBm with a resolution of 0.02dB and converted to display in Watts. The display of power in Watts always takes place to three decimal places.
- <sup>3</sup> The circuit of the display unit can be attached directly to a USB interface [3].



**Fig 4 : Layout for the back side of the display unit PCB.**



**Fig 5 : Component layout for the front side of the display unit PCB.**

### 3. Construction

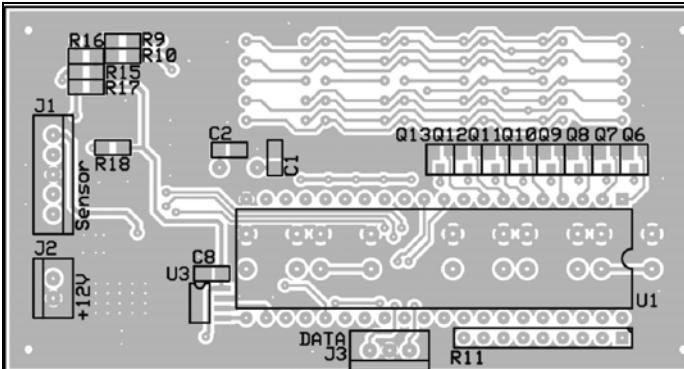
The layout of the double sided printed circuit board for the display unit is shown in Fig 3 and 4. Because of the small size, only 47mm x 90mm, the small “Ultra-mas” enclosure from Bopla [5] can be used. This plastic housing is well designed for measuring instruments with folding feet to raise the front panel.

The component layout is shown in Fig 5 and 6. There are no special requirement for assembling the PCB. An IC socket should be used for U1 to make software updates easier. Note that the micro controller, connectors and some SMD components are on the back of the PCB. The

displays, switches, crystal and remaining SMD components are on the front side of the PCB.

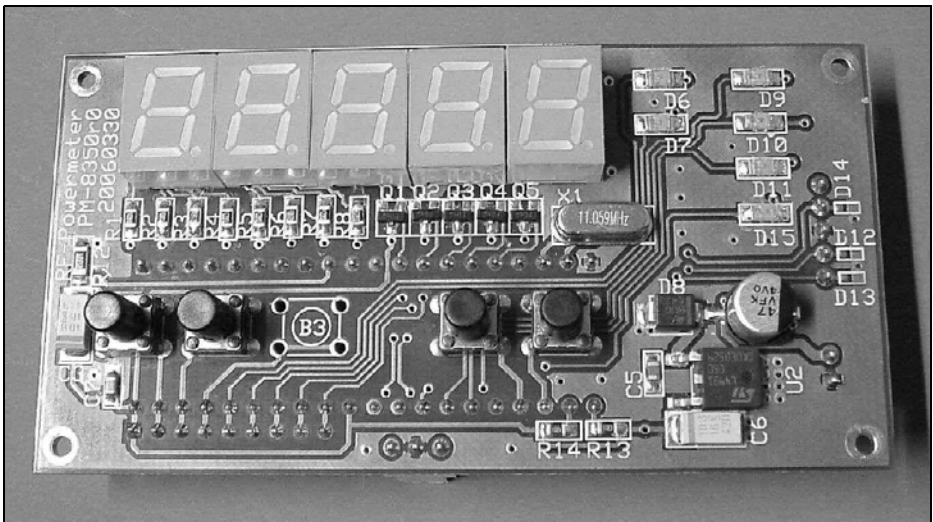
### 4. Parts list

- U1 AT89C52, DIP, programmed
- U2 L4931CDT50, DPAK
- U3 24C01-2.7, SO-8
- Q1-Q13 BCV47
- D1-D5 SA39-11 green
- D6-D7, LED green, SMD 1206
- D9-D11,
- D15
- D8 SMBJ16
- D12-D14 CDS3C09GTA



**Fig 6 : Component layout for the back side of the display unit PCB.**





**Fig 7 : Picture of the front side of the prototype PCB.**

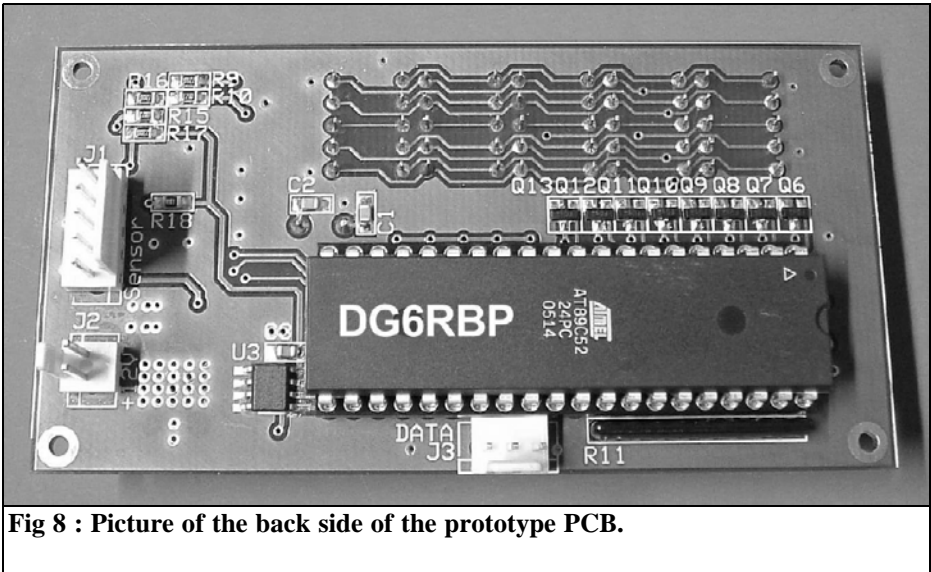
R1-R8	47Ω, SMD 0805
R9-R10,	220Ω, SMD 0805
R15-18	
R13, R14	4.7KΩ, SMD 0805
R12	10K, SMD 1206
R11	8 x 10KΩ, resistor network
C1, C2	33pF, SMD 0805
C5, C7,	100nF, SMD 0805
C8	
C3, C6	10μF/16V, SMD tantalum
C4	47μF/35V, SMD Electrolytic
B1-B54	push switch DTS-64N
J1	LP patch cord. RM 2.54, 5 pin
J2	LP patch cord. RM 2.54, 2 pin
J3	LP patch cord. RM 2.54, 3 pin
1 x	40 pin IC bases 40
1 x	PCB, DG6RBP
1 x	mini DIN socket, with 6 pins
B3	is not fitted it is for future extensions!

## 5.

### Calibration and operation

Before using the display unit it must be calibrated with the power detector. To do this the “Up” and “Down” switches should be pressed while the power is switched on. “CAL” then appears on the display. Subsequently, the input voltage of the A/D converter in millivolts is displayed. Connect a signal generator to the detector set to 0dBm (1mW) and press the “Up” key to store the output of the detector as a reference voltage. This is confirmed with a short flashing of the display. A level of -20dB (10μW) is now set on the signal generator and the “Down” key is pressed. The micro controller calculates the voltage level and the upward gradient of the detector characteristic then stores these. This is confirmed with a short flash of the display.

The reverse sequence using -20dBm first then 0dBm is not possible. An error with calibration is indicated by a “C” on the



**Fig 8 : Picture of the back side of the prototype PCB.**

display e.g. a completely wrong reference voltage or upward gradient, points to an error in the power detector. In the case of an error the calibration must be repeated again after a reset.

The calibration is only valid for the detector used and should be repeated at regular intervals (e.g. every 2 years). The signal generator used must be as accurate as possible, otherwise the measuring accuracy of the wattmeter suffers accordingly. The actual output level of the signal generator should not deviate from the adjusted level by more than  $\pm 0.01$  dB. The frequency used for calibration can be between 10 MHz and 2.4 GHz but should be the same for both calibration steps.

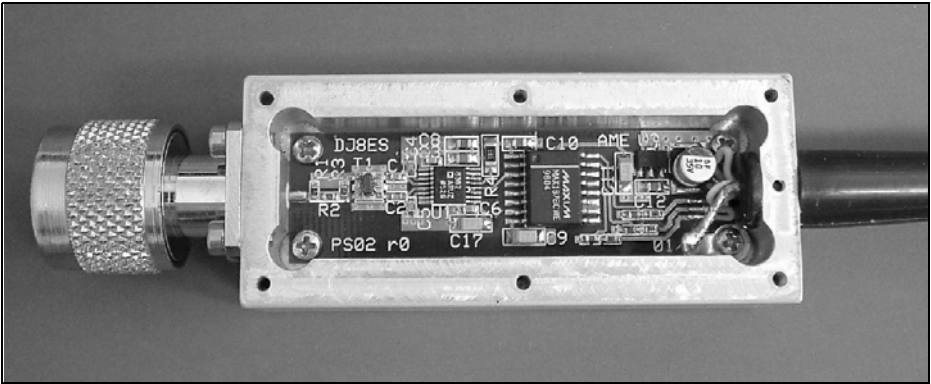
It is advisable to determine the frequency response of the detector before calibration and to use a calibration frequency that corresponds to the average value. In the frequency response of the sample (Fig 2) this would be approximately 650 MHz or 1.3 GHz. Thus the measuring error of the power measuring instrument over the entire frequency range becomes as symmetrical as possible.

After final calibration the power measuring instrument can be used immediately.

The display power is indicated in dBm with a resolution of 0.1 dB (the measured value with 0.02 dB resolution is taken internally and rounded on 0.1 dB). Pressing the key "Abs./Rel." switches to relative mode, the measured value is stored from now as a reference level and the difference between measured value and reference level in dB is indicated. Pressing the key again changes the display back to dBm. This function useful for transmission measurements.

To measure the gain of an amplifier first the input power is measured and the "Abs./Rel." changed to relative mode. Subsequently measures at the output of the amplifier. The instrument now indicates the gain directly in dB on the display.

Pressing the "dBm/Watt" key the display is switched between dBm and Watts. The appropriate unit is indicated with the light emitting diodes. The measured value in Watts is calculated from the measured value in dBm. Here the micro



**Fig 9 : Picture of the optimised power detector.**

controller uses the maximum resolution of the sensor (0.02dB from the sensor circuit with the 12 Bit A/D converter). The display resolution for a linear display (Watt) is 0.02dB. That means, for example, following 955 Watts (29.80dBm) the next step is 959 Watts (29.82dBm), or following 50.1mW (17.00dBm) the next step 50.4mW (17.02dBm).

The keys “UP” and “Down” can be used to adjust the offset level. If one of these keys is pressed briefly, the current offset in dB is indicated. An “F” in the first display place marks the offset. Pressing one of the two keys the offset value can be changed between  $-1.9$  and  $+49.9$ dB. If the key is pressed for a longer time, the adjusting speed is increased by large changes of the offset level. If neither of the two keys is pressed for a long time (approximately 2 seconds), the indicated offset level is removed and to the display of the measured value returns. The offset value remains stored, in EEPROM, after switching the equipment off. In order to remind the user after switching on the offset is indicated briefly in the display.

The offset level is added to each measurement. If the power detector is operated, for example, at a frequency where the deviation in level amounts to  $-0.3$ dB (i.e. the result of measurement would be too small around  $0.3$ dB), an offset can be

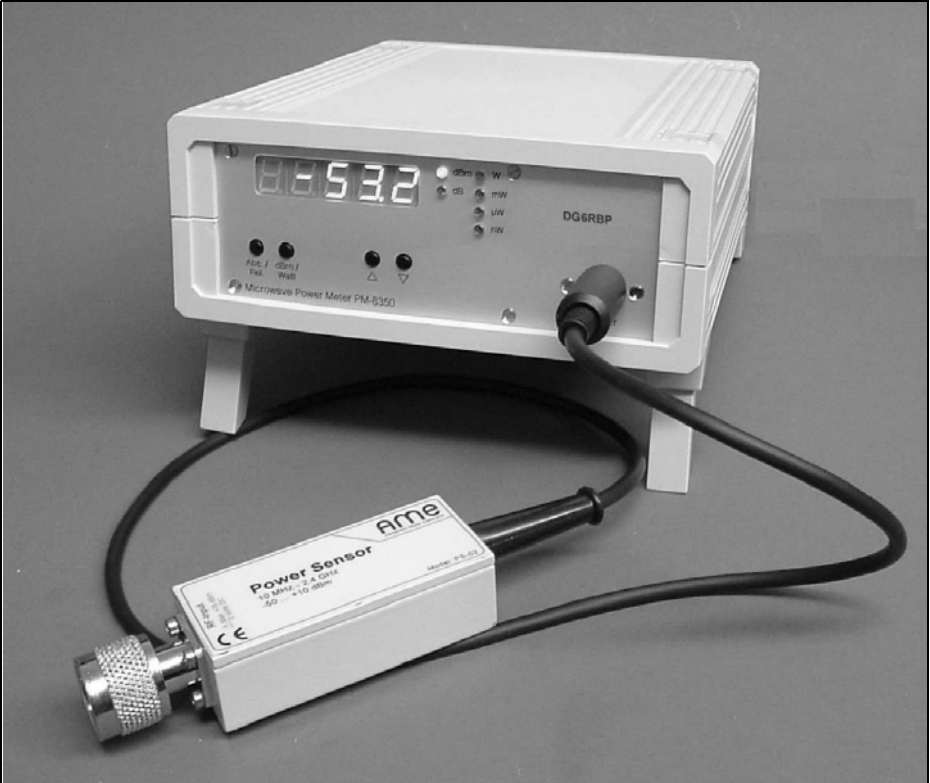
set of  $+0.3$ dB for precise measurements. The measuring instrument will automatically add  $0.3$ dB to make the measurement accurate.

A still more interesting application is the measurement of high powers using a coupler. Since the power detector can only measure up to approximately  $10$ mW, a coupler can be inserted between the transmitters and the antenna or dummy load. The uncoupled power is measured and the coupled attenuation is used as an offset levels to display the actual transmitter power.

The indicated value in Watts or dBm is also constantly sent over the serial interface once it is activated. To activate the transmission the ASCII value “B”, for beginning, is sent. In order to terminate the transmission the value “S”, for stop, is sent. Further interface functions are not implemented yet.

Printed circuit boards as well as printed front panels will be available from VHF Communications.

The front and back of the completed PCB are shown in Fig 7 and 8, the optimised power detector is shown in Fig 9 The complete power measuring instrument in the housing and with power detector is



**Fig 10 : Picture of the completed display unit with power detector.**

shown in Fig 10.

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## 6. Literature

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[1] Wolfgang Schneider: Power detector covering up to 2.7GHz, VHF Communications Magazine, 3/2006 pp179 - 183

[2] Alexander Meier: Frequency input module for 10GHz the ATV transmitter module, VHF Communications Magazine, 4/2005, pp 217 - 221

[3] Alexander Meier: VX-2-Speicher komfortabel füllen, CQ-DL 10/04, DARC publishing house, Baunatal (2004)

[4] Data sheet MAX-232, Maxim Integrated Products, Inc., 120 San Gabriel drive, Sunnyvale, CA. 94086, USA, <http://www.maxim-ic.com>

[5] Bopla Enclosures, Phoenix Mecano Ltd, Unit 6/7, Faraday Road, Aylesbury, Buckinghamshire, HP19 8TX, [http://www.bopla-enclosures.co.uk/products/enc\\_tech.htm](http://www.bopla-enclosures.co.uk/products/enc_tech.htm)



Franco Rota, I2FHW

# Noise source diodes

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## 1.

### Noise

---

#### 1.1 Introduction

All materials generate noise and the noise is proportional to its temperature starting from 0°K (-273°C). The noise depends on the chaotic movements of the electrons, the thermal noise is known as white noise (from optical physics) as it fills the whole spectrum.

From an electronics point of view the noise causes big limitations to our devices for example amplifiers, instruments, radars, receivers, electro-medical, etc... A very simple example is the sensitivity limitation of receivers caused by the noise.

Although I have said that noise causes problems and limitations, I want to explain how in some cases, if it is artificially generated, it can even improve our electronic devices (see dithering in Table 1) or help to do some tests, a calibrated noise source is a very important tool in our labs.

#### 1.2 Output level

For noise source applications the output level cannot be indicated as for other signal generators. Signal generators, transmitters etc... have the output level

indication in mV, dBm, W etc.... If you have a 100 to 200MHz sweep signal generator we say that the output level is, for example -10dBm, the amplitude of -10dBm is swept from 100 to 200MHz but it is not simultaneously in the whole frequency range.

In the case of noise sources the amplitude is simultaneously on the entire frequency range, this means that the amplitude is defined in dBm/Hz power spectral density, or in ENR excess noise ratio. ENR means the ratio in decibel of the output noise between the ON and OFF state of the diode, in the OFF state the diode has only -174dBm/Hz which is the output level generated by a resistor at 290°K.

For example, if you have a power spectral density of -142dBm/Hz it means that (174 - 142 = 32) the ENR is 32 dB. If the bandwidth is 10Hz the noise power is -132dBm/10Hz if the bandwidth is 10KHz the noise power is -102dBm/10KHz.

---

## 2.

### Noise generator diode

---

#### 2.1 Diode selection

The first noise generators (in the 1940's) used noble gas such as Argon with

**Table 1: Some applications regarding the generation of noise, it can improve electronic devices or help to do some tests on them.**

- **Dithering**  
In an A/D converter for example digital receivers, the noise injected improves the quantisation error, the sensitivity will be improved (this method is also used in audio and video).
- **Spectrum Analyser Calibration**  
With a calibrated noise source devices it is very easy to verify the amplitude calibration of a spectrum analyser, the real advantage is the RF generation simultaneously on all the band.
- **Noise Figure Measurement**  
Test instruments for noise figure measurement in low noise amplifiers, converters, receivers, mixers and front-ends.
- **Gain-bandwidth measurements**  
A flat noise source can be used as a “tracking generator” combined with a spectrum analyser to ease measurements of gain and bandwidth.
- **Encryption**
- **Audio And Ultrasonic Test**
- **Test On Receiver**  
The noise is useful to measure the sensitivity in some complex receivers like radars, base stations, radiometers etc...  
A noise source can substitute for a more complex RF generator, moreover it can generate noise in a broad band spectrum simultaneously.
- **NPR Distortion**  
This is a complex intermodulation measurement very often made on multichannels FDM, MMDS, CATV, cellular base stations, etc..  
Injecting noise and measuring the distortion with special notch filters is used to obtain the measurement.
- **Fading Simulator**  
By modulating an RF signal with noise it is possible to simulate a signal affected by fading, this is very useful in mobile radio testing.
- **Radio Astronomy**
- **EMI Testing**

15.3dBENR, Neon with 18.5dBENR, Helium with 21dBENR and were born in order to test the first radar systems.

Another system to generate noise is to use hot and cold resistors, mainly used in research labs with very high precision.

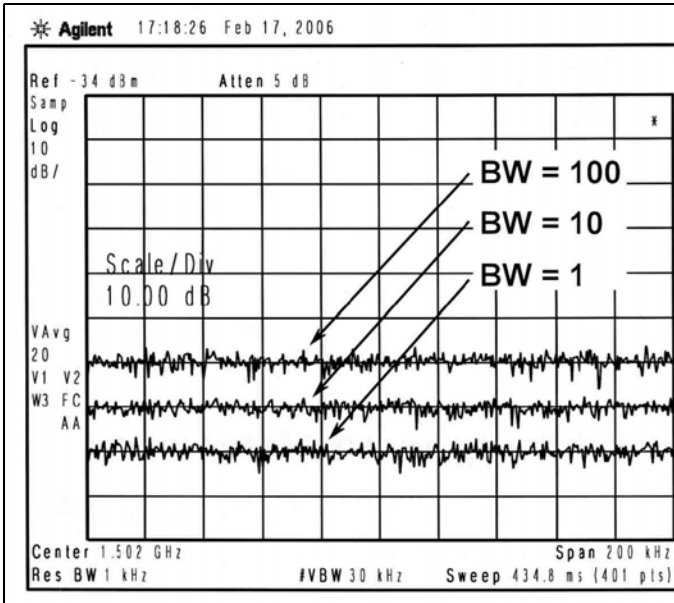
Zener diodes can be used to generate noise but the output level is not constant, not predictable and used only for HF frequencies, even some bipolar transistors like BFR34 have been used in the past for amateur applications using the reverse biased base-emitter diode, the output level is definitely not constant.

For our applications the right selections are:

- NS-301 SMD sod323 case, up to 3.5GHz
- NS-303 ceramic gold plated case, up to 10GHz

Both types are silicon avalanche diodes that provide 30-35dBENR with a broad-band spectrum starting from 10Hz. In this article I will focus on the 3.5GHz type and in a second article I will also describe the 10GHz type which is more complicated.

At the beginning I tested the glass case type but this case was not suitable because the maximum frequency can be around 1.5 - 2GHz, for the same price we can have 3.5GHz with a flatter output



**Fig 1: The same noise level related to 3 different bandwidths.**

level.

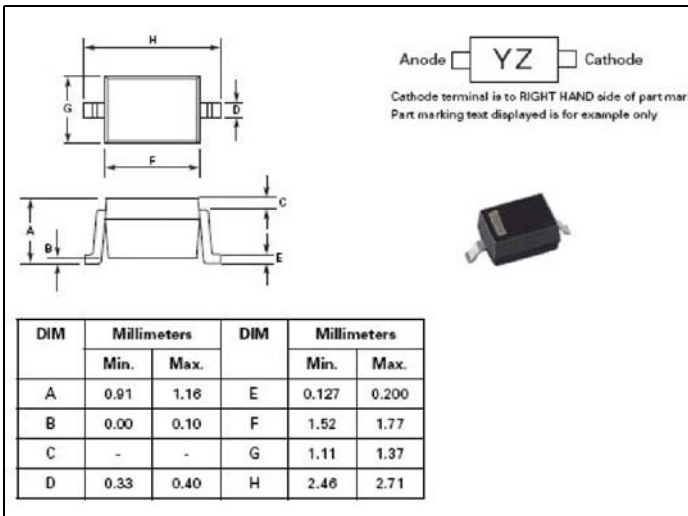
The SMD sod323 case has a very low series inductance typically 1 - 1.5nH which is reasonable for a 3GHz application. Fig 2 shows the SMD case sod323, the body is about 1.9mm long, it is useful for many applications in the lower microwave frequency range.

## 2.2 Schematic diagram

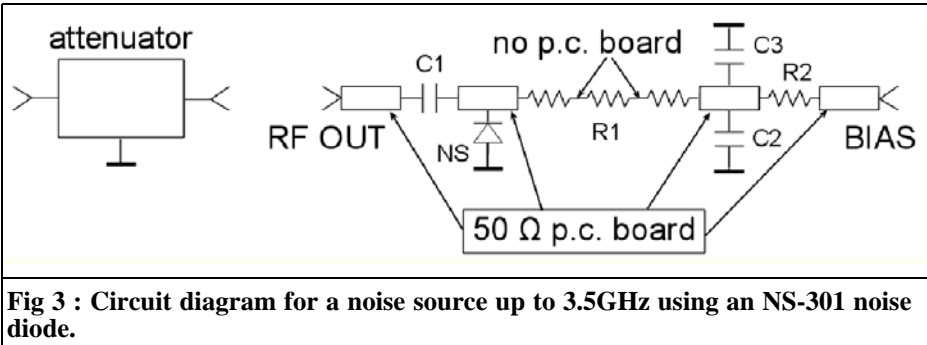
Fig 3 shows the circuit diagram of a NS-301 noise source diode up to 3.5 GHz.

### C1 – dc blocking capacitor

The selection of this capacitor is extremely important to flatten the output level. I spent much time testing several



**Fig 2 : The SMD SOD323 case.**



**Fig 3 : Circuit diagram for a noise source up to 3.5GHz using an NS-301 noise diode.**

types of capacitors, ATC porcelain capacitors have less insertion loss at microwave frequencies but they can't be used because their Q increase the self resonance dip.

For this purpose I selected a special capacitor case, 0805 COG, which can be used up to 12GHz (about 1.5nF), with this capacitor the minimum frequency is about 10MHz.

In the next article about the 10GHz noise source diode I will describe these capacitors in more details.

For 3GHz application the C1 capacitor isn't a crucial component, case 0805 or 0603 and values form 1nF to 10nF are good anyway.

### **C2, C3 – bypass capacitors**

These capacitors are not critical; they can be 1nF and 10nF.

### **R1 - RF load resistors**

This resistor is the sum of 3 resistors in series in order to keep the stray capacity as low as possible, the total value can be around 30 to 40Ω

The manufacturer of noise diodes says that the diode impedance is about 20 to 40Ω, I noticed that by assigning to R1 a lower resistance (20Ω), the output noise level is flatter, on the contrary with an increased resistance (40Ω) the output noise level is a little higher.

If possible, it is better to solder the resistors without using copper track on

the PCB.

### **R2 - bias resistor**

For the noise diode NS-301 at about 5mA, +8/+12 V, the correct value is 3.3KΩ if you use the diode for noise figure measurements with a classic +28V pulse available from all the noise figure meters. If the diode is used as a general purpose noise generator to test a filter, for example with a spectrum analyser, you can connect directly to a +8/+12Vdc without the R2 resistor.

### **NS - Noise diode**

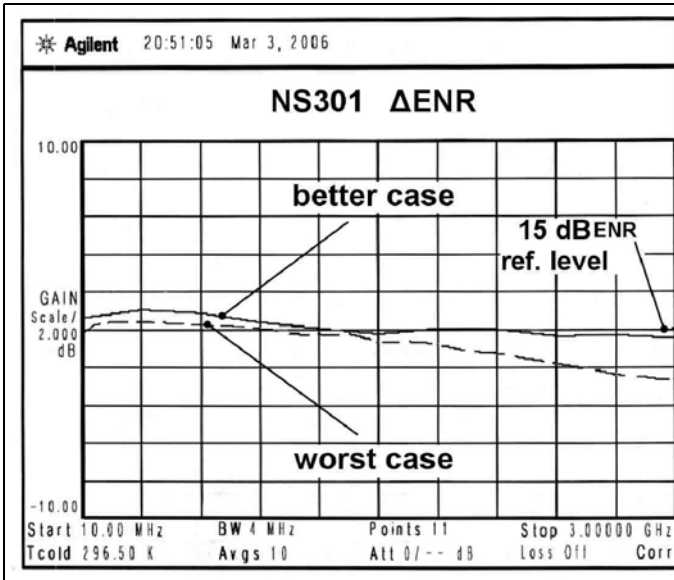
As described above the NS-301 sod323 diode is a good selection for the 3GHz frequency range, it is important to remember to keep the pins as short as possible! The diode must be mounted very close to the output connector.

### **P.C. board**

The FR4 fibreglass p.c. board is ok, the insertion loss is so little that it isn't worth a teflon laminate, vice versa it is very important the noise diode ground connection that has to be as short as possible (see the above explanation).

I tested several noise source diodes in my lab with sod323 case, Fig 4 shows the best and the worst result, in the frequency range 10MHz to 3GHz with 2dB/step and 300MHz/step, the centre reference level is 15dBENR and the noise source diode is connected with a 16dB pad attenuator.





**Fig 4 : The best and the worst test results on different noise source diodes.**

### 2.3 Output attenuator

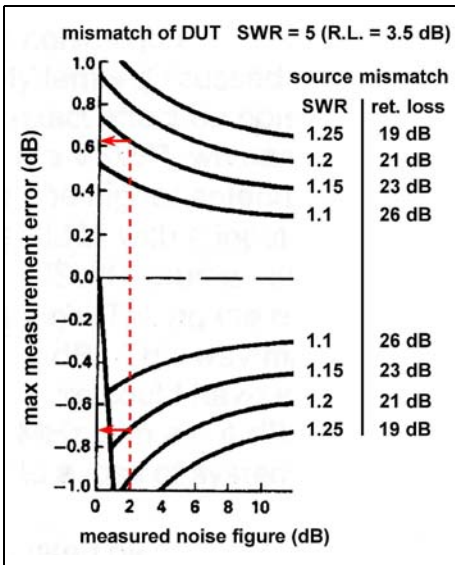
The purpose of this attenuator is two fold, the first one is to obtain the

15dBENR which is the right noise level accepted by a lot on noise figure meters. The output noise of the NS-301 diode is about 30-35dBENR this means that with a 16dB attenuator you can have about 15dBENR. Any other attenuation values can be used to get other ENR values.

The second and most important purpose of this attenuator is to match the output impedance to 50Ω. In noise source devices used for noise figure measurement, one of the most important condition is to match the output impedance as near as possible the 50Ω resistive load, the easiest way is to insert an attenuator to the output connector.

Normally the ultra low noise GaAsFet preamplifiers have a very bad input return loss, typically a VSWR from 20 to 2 (return loss from 1 to 9.5dB), so if we test this kind of preamplifier with a noise source with an high return loss the total error is unacceptable.

Fig 5 shows a simple explanation of the mismatch due to the noise figure measurement, we can assume that the preamplifier input return loss is 3.5dB, SWR = 5 (it can seem too high but it is a realistic



**Fig 5 : A simple explanation of the mismatch due to the noise figure measurement.**

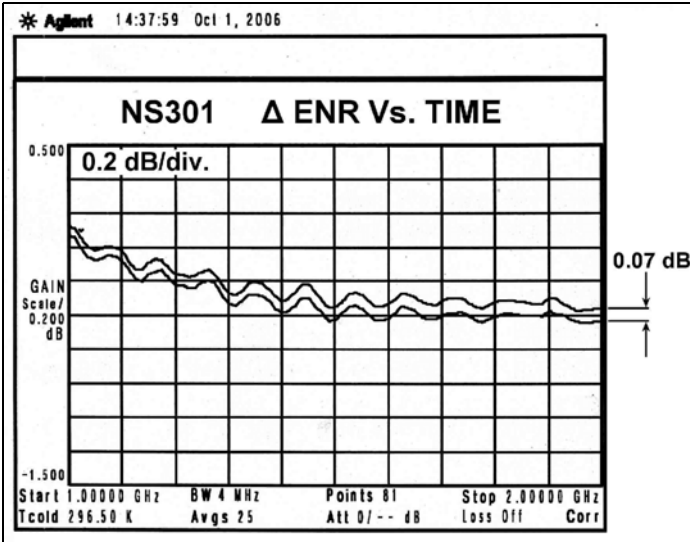


Fig 6 : Instability of noise source diode output.

value).

If our noise figure meter measures 2dBNF and we assume also that the noise source output return loss is 23dB (SWR 1.15), the true noise figure can be between +0.63dB/ - 0.7dB for a 2dB measured value.

In conclusion we should keep the SWR of a noise source as low as possible in

order to do more accurate noise figure measurements.

Fig 6 shows the instability that it is quite good for amateur applications, for 8 hours of continuous operation it is only 0.07dB of output level but there is also a 0.03dB of testing instrument instability to consider.

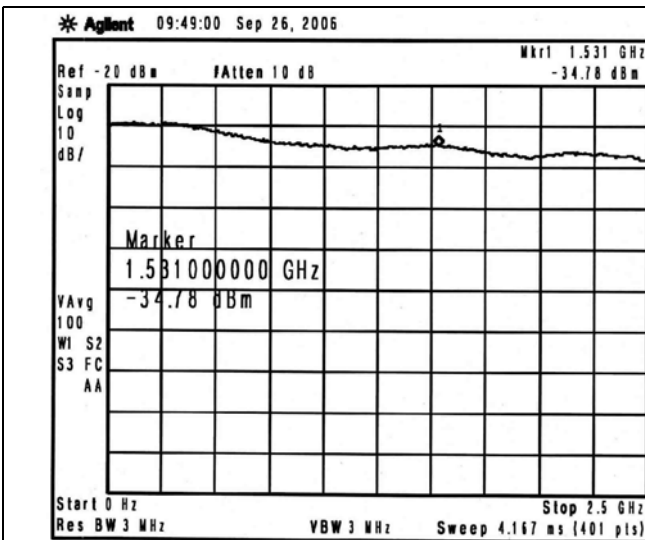
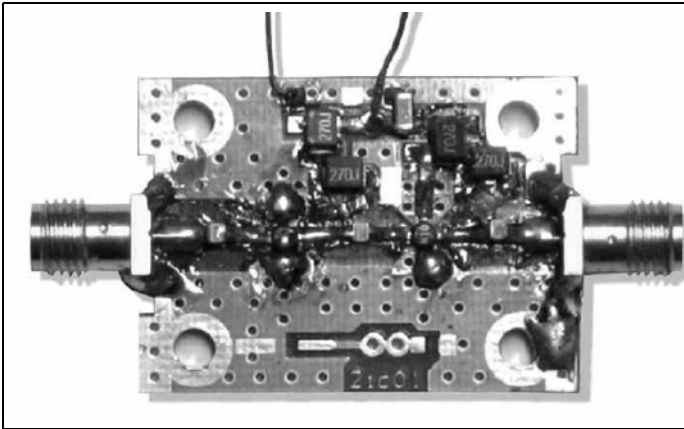


Fig 7 : Response with a 45dB amplifier.



**Fig 8 : Picture of the broadband amplifier.**

**3.**

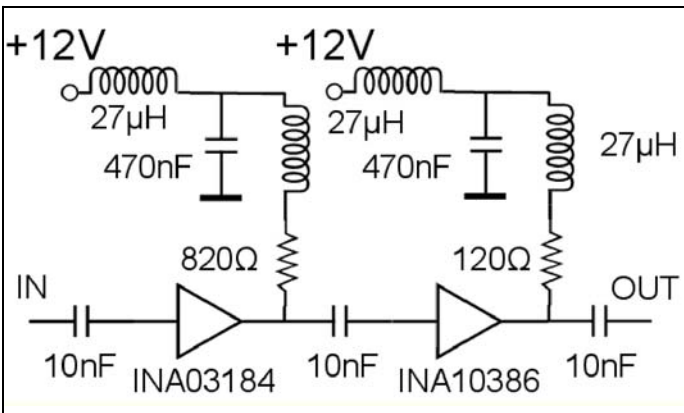
**General purpose noise generator**

As shown in Table 1 a diode noise source can be used successfully in a broadband noise generator combined with a spectrum analyser like a “tracking generator”. This is not a true tracking generator because it works in a different way. As I said above the tracking generator is like a sweep generator so its frequency moves from start to end but it is not simultaneous in all the frequency range.

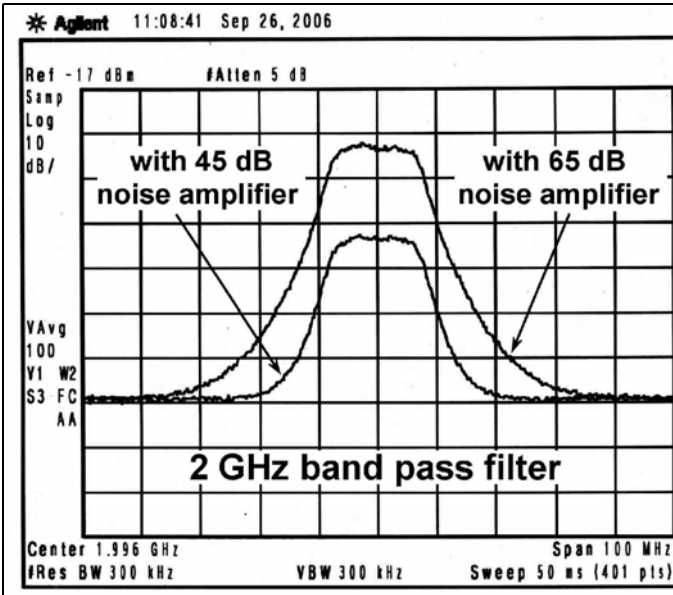
If we combine a broadband noise generator with a spectrum analyser we can do a

measurement of band pass filters, return loss etc. The signal coming from the noise generator diode is very low so we need at least 45dB of amplification, however 65dB is better. The real difficulty is to obtain a reasonable flat amplifier response. For this purpose I made an amplifier using INA03184 and INA10386 MMICs, the result is shown in Fig 7 and the total response is given by the noise source diode combined with the 45dB amplifier.

Figs 8 and 9 show the 45dB broadband amplifier from few MHz to 2.5GHz used as noise amplifier in order to test the 2GHz band pass filter. This circuit is not difficult to build and it can be used in any lab as general purpose broadband amplifier.



**Fig 9 : Circuit diagram of the broadband amplifier.**



**Fig 10 : Dynamic range of a 2GHz band pass filter.**

Fig 10 shows the dynamic range of a typical 2GHz band pass filter with a noise amplification of 45 dB and 65dB. The dynamic range improves with more amplification, but it is more difficult to achieve a flat output level.

Fig 11 shows the equipment setup used for the filter measurement.

It is demonstrated that with a simple noise generator and a good amplifier it is possible to build an instrument very close to a tracking generator to use with any kind of spectrum analyser. It means that we can “upgrade” an old spectrum ana-

lyser, typically the HP 141 series or any other type, with an option that works like a tracking generator.

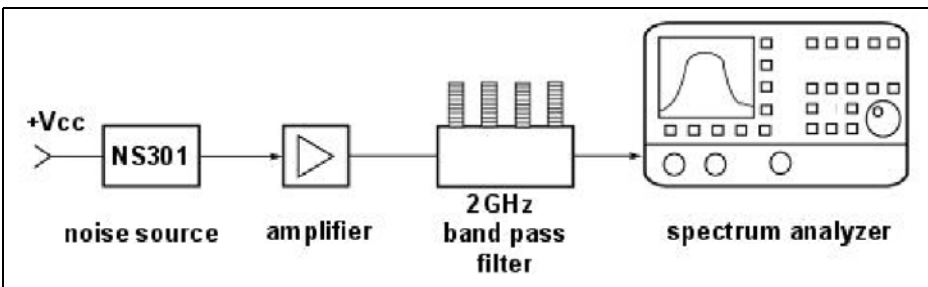
NS301 noise source specifications are:

Frequency range: 10Hz - 3.5GHz

Output level: 30/35dBENR (-144/-139dBm/Hz)

Bias: +8/+12V, 5mA

It is available from R F Elettronica - [www.rfmicrowave.it](http://www.rfmicrowave.it)



**Fig 11 : Equipment setup used for the filter measurement.**



Henning C. Weddig, DC5LV

# Extending the 50MHz converter into a Transverter

Some time ago the author described a 50MHz converter. The positive feedback and requests to extend this into a transverter have persuaded the author make this possible. The following article takes the converter already described and adds the necessary transmitter driver.

- Intermodulation products at full output ( $d_3 < -60\text{dB}$ )
- Sideband suppression ( $< -60\text{dB}$  especially at converter sidebands)
- Harmonic suppression ( $> 60\text{dB}$ )

By choosing the input level in consideration to the sidebands (harmonics of the 22MHz crystal oscillator and other mixing products) the signal-to-intermodulation ratio can be kept to a minimum.

---

## 1.

### Concept

---

The basic circuit simulations presented in [12] pointed out the extensions required to the 50MHz converters already described in [11] to realise a transverter. These extensions are:

- Input levels of the transceiver into the converter (+6dBm single tone; 0dBm for two tone rejection for a Signal-to-intermodulation ratio of 60dB)
- Power output of the driver (dependant on the output stage used)
- Power output of the output stage (10W = +40dBm for use with a directional antenna with a gain of 6dB)
- Amplification of the driver (dependent on the output stage used)

The specification did not specify the minimum signal-to-intermodulation ratios at the transmitter output however third order products should be less than 25dB. Because under unfavourable conditions, the intermodulation products add together, the stages before the output should have more than the required 25dB rejection. This must be considered when choosing the devices for the output stages.

The output stage should be protected against bad connections (e.g. open or short circuit of the antenna connection) effectively and without affecting the ALC of the transceiver. This is achieved using pin diode control using a circuit that is activated at a fixed value of VSWR. A manual control of output power is required. The pin diode has an estimated insertion loss of 4dB.



### 1.1 Signal-to-intermodulation ratio; Input level

In transmit mode the converter is operated “backwards”, i.e. the input signal (28MHz) of the transceiver is fed into the IF output of the converter, the 50MHz output signal is coupled from the RF input of the converter.

From the Intercept point of the converter (+30dBm) it can be reckoned back that with two input signals of 0dBm at the IF output, a signal-to-intermodulation ratio of 60dB is obtained. This signal-to-intermodulation ratio is high enough. Also the stages of the driver should keep to this level. Under these conditions the output signal at the RF input of the converter will be -8dBm because the transmission loss of the converter is 8dB. Thinking about the rejection levels, 6db can be added, i.e. with an input signal (of the transceiver) of +6dBm the output signal amounts to -2dBm.

### 1.2 Transmitting power

From [1] a transmitting power of 10W and a 6dBd directional antenna is sufficient, in order to keep to the permitted maximum output of 25W ERP (41W EIRP).

---

## 2.

### Choice of the transmitter output stage

---

The transmitting power of 10W is easily attainable with one the remaining Mitsubishi [2] M57735 hybrid modules. This module gives approximately 10W (+40dBm PEP) SSB, using an input power of 20mW (13dBm). The amplification of the module is 27dB. Third order intermodulation products of 32dB PEP

are 6dB more than related to a single tone, thus “only” 26dB. Thus the intermodulation products of the output stage determine the output signal.

With the first harmonic at 25dB a second attenuation of 30dB is required in order to reach the > 60dB harmonic suppression. At least a seven pole low pass filter is required (Cutoff 60MHz with 0.05dB ripples in the pass band and around 20dB return loss).

The intermodulation values of the hybrid module are not a problem. If the hybrid module is selected as an output stage, the 50MHz signal output of the converter must be amplified by about 15dB.

An alternative to the hybrid module as output stage is to use discrete transistors. Unfortunately the number of manufacturers in the RF power transistor market is very small. Motorola started the manufacture of transistors in the portable radio market but they have given up, so transistors like 2N6080 and 2N6081 are no longer made. Also the silicon transistors BLY87 and BLY88 from Philips are not recommended for new developments. These are possibly still available at flea markets. The popular types 2SC1970, 2SC1971 and 2SC1972 are no longer manufactured by Mitsubishi so there is not an alternative.

The trend is clearly toward MOS transistors. In [3] and [4] a broadband output stage with two BLF244 MOS transistors is described. It reaches a third order signal-to-intermodulation ratio of 40dB at 10W output with a supply voltage of 28V and 125mA quiescent current. The amplification of the output stage is approximately 15db. A BLF242 could be used as a driver. Unfortunately these transistors are expensive. An alternative could be the MOS transistor RD16HFF1 in a TO220 housings from Mitsubishi, see [5]. This Transistor is relatively inexpensive.

**Table 1: List of mixing products.**

Frequency in MHz	Level in dBm	Mix product	Level of mix product with no signal in dBm
16	-81	2*LO - RF	73
22	-65	LO	58
28	-59	RF	51
38	-36	3*LO - RF	28
<b>44</b>	<b>-51</b>	<b>2*LO</b>	<b>43</b>
50	-8	IF	0
54	-74	5*LO - 2*RF	68
<b>60</b>	<b>-36</b>	<b>4*LO - RF</b>	<b>28</b>
<b>62</b>	<b>-50</b>	<b>3*RF - LO</b>	<b>42</b>
<b>72</b>	<b>-43</b>	<b>2*LO + 2*RF</b>	<b>35</b>
82	-57	5*LO - RF	49
88	-78	4*LO	70
84	-52	3*LO + 3*RF	44
100	-71	2*RF + 2*LO	63
106	-77	3*RF + LO	69

### 3.

## Mixing products of the Transverters

If a signal of 0dBm at 28MHz is fed into the IF output the desired 50MHz signal appears at the RF input as well as other mixing products, the frequencies and levels are listed in Table 1. The mixing products should as far as is technically possible be suppressed to a value of >60dB. With the values from this table the filter required can be measured. All mixing products that are a small distance away from the carrier are in bold. The following mixing products are noticeable:

- The first harmonic of the quartz oscillator (44MHz; -43dB)
- The mixing product 4\*LO-RF (60MHz; -28dB)
- The mixing product 3\*RF-LO (62MHz; -42dB)
- The mixing product 2\*LO+2\*RF (72MHz; -35dB).

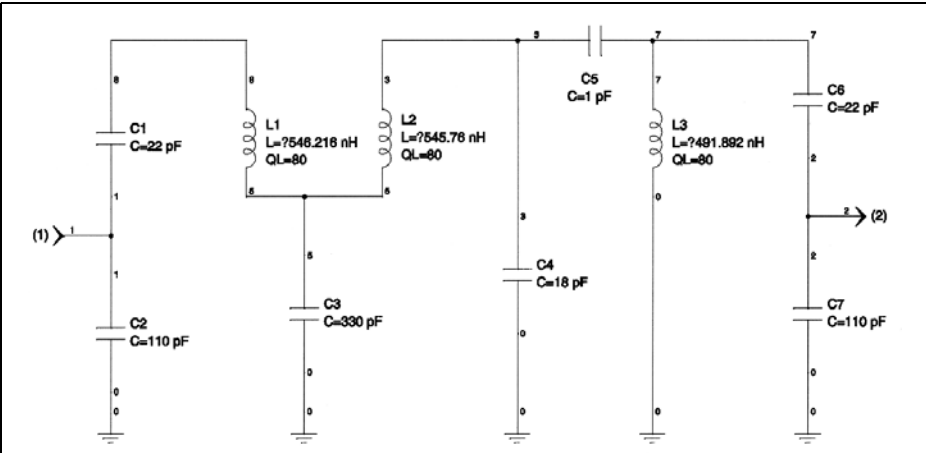
Only the first mixing product is level independent at -43dB, the filter would have an attenuation of 17dB.

The mixing product at 60MHz is at -28dB. In order to reach 60dB a filter with 32dB attenuation is required. The results are about 6dB better if the mixing products are measured for PEP rather than single tone.

A relatively large range low insertion loss and low noise figure 50MHz filter is required at the input of the converter as a band pass filter in the transmit path. An insertion loss of <3dB is quite acceptable. A three stage band pass filter was computed with the program "nrnbpas.bas", having the following data:

- Centre frequency 50MHz
- Bandwidth 5MHz
- Tschebychev characteristics
- Pass band ripples 0.1dB

Since the program "nrnbpas.bas" gave non-standard capacitor values, the filter was simulated using "RFSim99". Unfortunately this program does not offer



**Fig 1: The circuit diagram of the optimised band pass filter.**

automatic optimisation. Therefore the optimisation of transmission loss to < 3dB and return loss -20dB in the pass band was carried out with “Genesys” from Eagle Technology. The non-standard capacitors were replaced where possible with normal E12 components or using several capacitors together plus changes in the inductor values. This is what would happen with a manual alignment. The optimised filter is shown in Fig 1 with the response curves in Figs 2 and 3. At 44MHz the attenuation is 32dB and at 60MHz 42dB is achieved. Also the insertion loss is kept to 3dB. With these values the filter fulfils the requirement to suppress the sidebands by >60dB.

#### 4. Level plan

The amount of amplification between the output of the converter and the transmit amplifier can be determined from the above. For a single tone of 4mW (+6dBm) with the transmission loss of the converter of 8dB this would give a

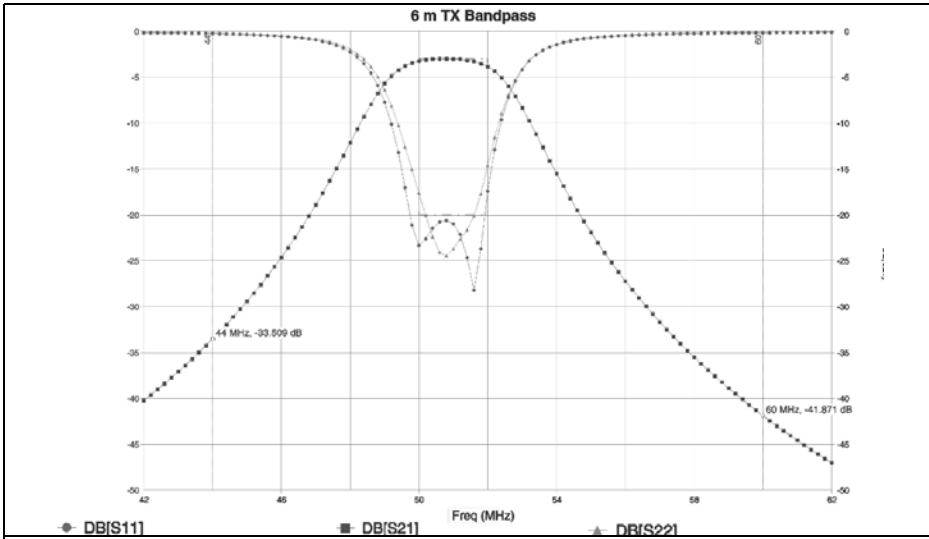
signal of -2dBm at 50MHz. To drive the hybrid amplifier fully +13dBm is required so an overall gain for the driver of 15dB is necessary.

For two tones the levels per tone must be lowered around 6dB, i.e. with two tones -8dBm each at the input of the driver. Considering the attenuation of the band pass filter of 3dB and the attenuation of the pin diode of 4dB the gain required is: 15dB + 3dB + 4dB = 22dB.

It would seem sensible to arrange the band pass filter after the amplifier stage in order to avoid any reaction between the transverter band pass filter and the converter band pass filter. On the other hand operating the band pass filter after the amplifier with relatively high gain, non-linearity due to the ferrite cores used in the coils can contribute to intermodulation.

The amplifier stage could use an RF2360 IC like the one used in the LNA (G = 20dB; OIP3 = +37dBm, IIP3 = +17dBm; P1dB = +25dBm). It would be operated 1dB below the compression point (-2dBm + 20dB = +18dBm), however it is 38dB from the third order





**Fig 2: Frequency response of the optimised band pass filter.**

intermodulation point. Thus the RF2360 would severely degrade the intermodulation behaviour of the transverter. The amplifiers type AH3 and AH101 from Watkin Johnson are therefore used. The data sheets are downloadable from the manufacturer, see [7] and [8]. The relevant data from the data sheet is:

**AH3**

- G = 12dB min., (12.9dB typical)
- OPIP3 = 37dBm min., (+41dBm typical)
- P1dB = +27dBm
- Noise figure (NF) = 4.5dB (typical)
- S11 -10dB, S22 -20dB
- Supply: 5V; 120 - 180mA

The value of (S11) is not acceptable, it must be optimised using a narrow band filter to <-20dB.

**AH101**

- G = 12dB min. (13.5dB typical)
- OPIP3 = 43dBm min. (+47dBm typical)

- P1dB = +21dBm
- Noise figure (NF) = 3.5dB (typical)
- S11 -20dB; S22 -15dB
- Supply: 9V; 170 - 230mA

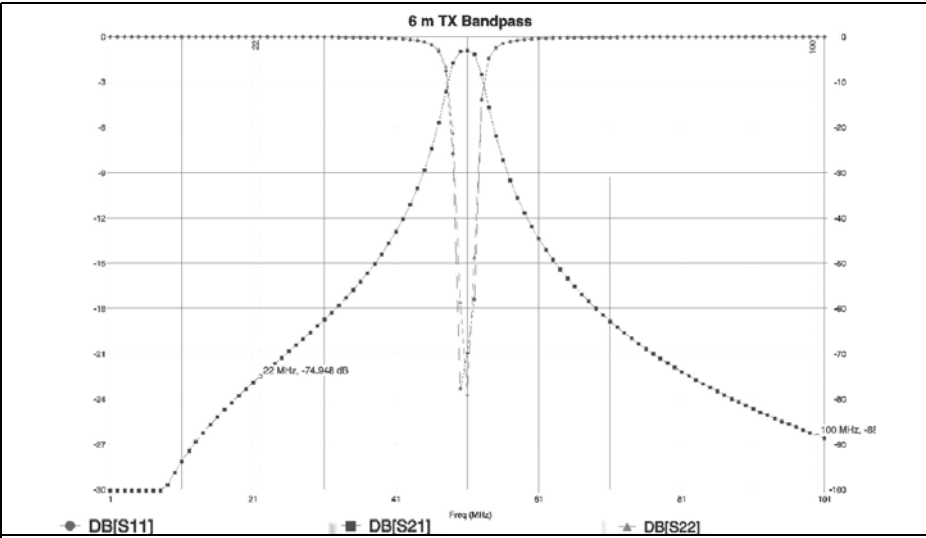
The input matching for this IC is acceptable but some improvement could be made to give the ideal value of <-20dB.

Both amplifiers together result in a minimum gain of 12dB + 12dB = 24dB, typically 26.4dB which is sufficient to achieve the level plan.

**5.**

**Test of filter construction**

A test of the three-stage filter in a tinplate housing with the AH3 amplifier on a printed circuit board showed that the 44MHz signal is reduced still further. Therefore a further three-stage filter is planned at the output of the second



**Fig 3: Damping factor of the optimised band pass filter.**

amplifier. The additional attenuation of another filter can be tolerated, since there is sufficient gain in reserve.

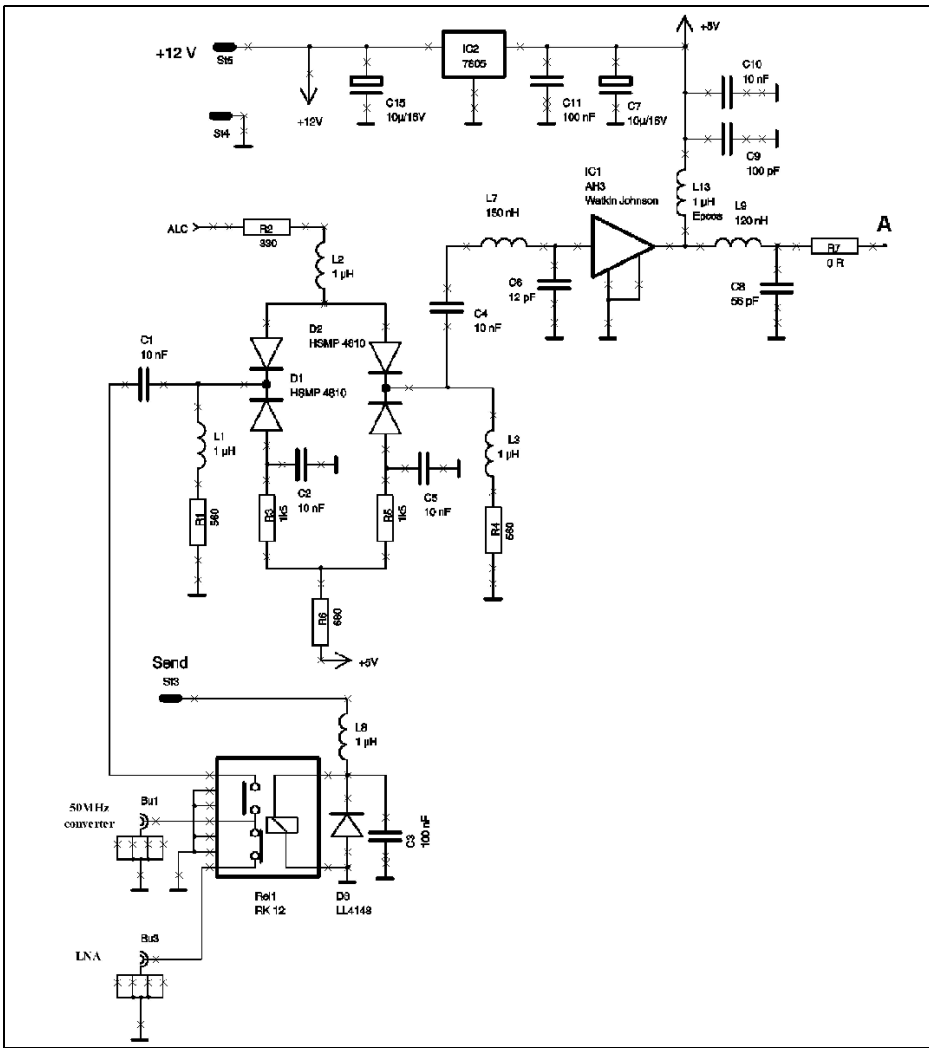
**6.**

**System simulation**

The result of the system simulation can be seen in Fig 4. The power outputs are

	PIN Diode attenuator	Watkin Johnson AH3	Bandpass filter	Watkin Johnson AH101	Bandpass filter	Total
NF (dB)	2.00	4.50	3.00	4.00	3.00	6.81
Gain (dB)	-2.00	13.00	-3.00	13.50	-3.00	18.50
OIP3 (dBm)	40.00	41.00	50.00	47.00	50.00	41.82
OP1dB (dBm)	30.00	21.00	30.00	27.00	30.00	24.00
Po (dBm)	-10.00	3.00	0.00	13.50	10.50	
IP3+ (dBm)	0.06	1.05	0.06	4.04	0.72	
NF+ (dB)	0.57	3.99		0.22		
Input Pwr (dBm)	-8.00	System Temp (K)		290.00		
Modulation: Custom						
System BW (MHz)	0.00	MDS (dBm)		-132.40	Input IP3 (dBm)	23.32
S/N (dB, Actual)	124.40	S/N (dB, Req'd)		20.00	Output IP3 (dBm)	41.82
Src Temp (K)	290.00	Sens. Loss (dB)		0.00	OIM3 (dBm)	-52.15
Te Eff. (K)	1099.84	Sensitivity (dBm)		-112.40	ORR3 (dB)	62.65
SFDR3 (dB)	103.82	G/T (dB/K)		-21.43	IRR3 (dB)	20.88

**Fig 4: The results of the simulation.**



**Fig 5: Circuit diagram of the transverter driver.**

appropriate for a single tone during a two-tone test. There are approximately 2dB in reserve, ( $G = 16.9\text{dB}$ , 15dB are required). The cable attenuation could be overcome if the output stage is at the base of the antenna like the LNA. The required intermodulation values are achieved.

**7.**

**Circuit description**

The circuit diagram of the transverter driver is shown in Fig 5 and 6. The pin diode PCB is taken from the circuit in [6]. The attenuation is greater than 70dB,

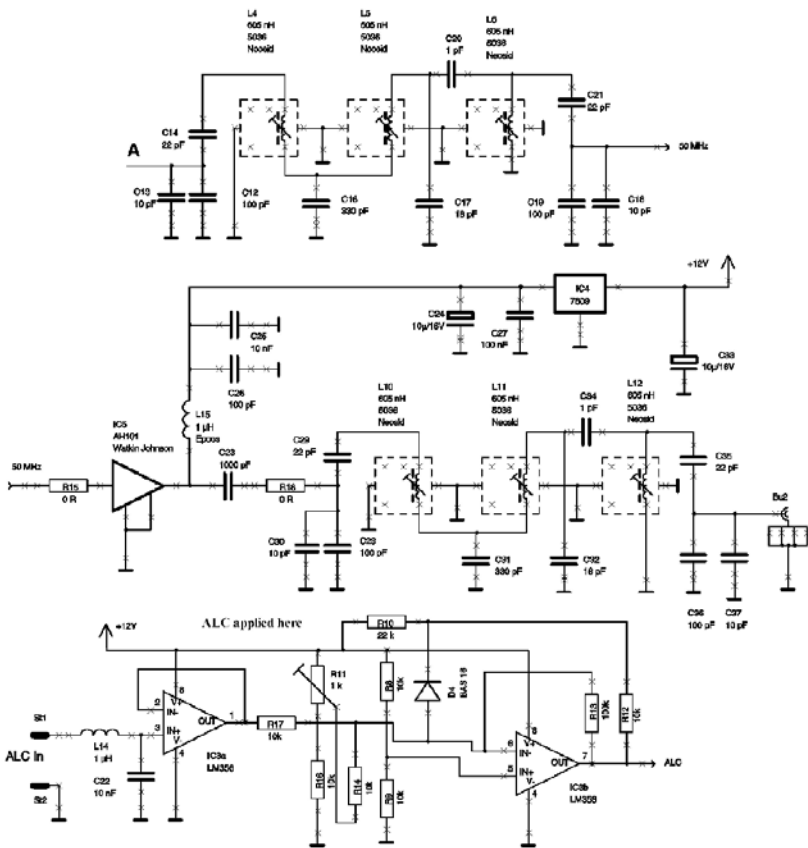


Fig 6: Circuit diagram of the transverter driver.

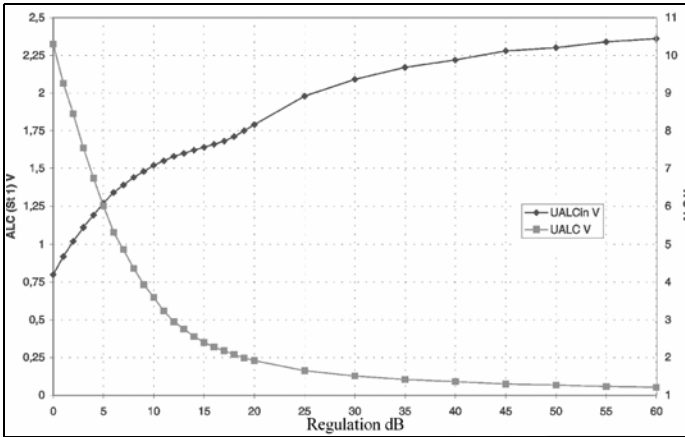
the control characteristic is however strongly nonlinear. The attenuation starts with less than 8.9V at the output of IC3b, pin 7. At 3.5V the attenuation is 10dB and 1.93V gives 20dB.

The two operational amplifiers in IC3 are used as a buffer, amplifier and an inverter for the control signal which can be supplied externally. The control level can be changed with R11. In the prototype the control voltage is 10.93V with regulation starting at 0.8V (at St1).

Tests on the finished circuit showed that the regulation start point became sensi-

tive to interference because of the high DC voltage amplification of IC3b. Therefore D4, R10 and R12 were changed on later units, the values were determined experimentally. The control characteristic is shown in Fig 7. It is linear within the range from 0 to 20 dB, it could be made more linear by adding additional diodes and resistances. Unlike the AGC in a receiver, this application only uses a small range (0 to 20dB) of the control characteristic therefore this is acceptable.

The AH3 amplifier does not have good output matching (S11; S22 < -20dB) at 50MHz, a matching circuit was com-



**Fig 7: The regulation of the PIN diode attenuator.**

puted with GENESIS.

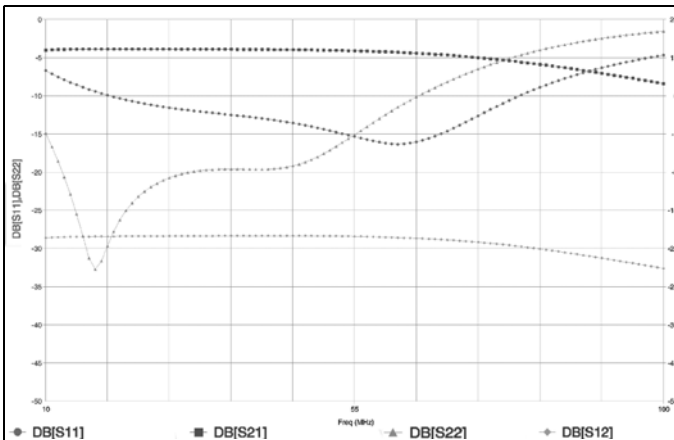
Unfortunately the downloadable S parameters from the manufacturer, Watkin Johnsn, only start from 50MHz, therefore approximate S parameters were used for GENESYS. The simulation results are shown in Fig 8 and 9. There are discrepancies between simulation and measurement. Differences can be explained by the fact that the manufacturer does not provide the measuring conditions, under which the S parameters are provided.

The AH101 has a good input match at 50MHz ( $S_{11} = -20\text{dB}$ ), so that only the output needs to be matched. Although  $S_{12}$  is quite small, the input matching

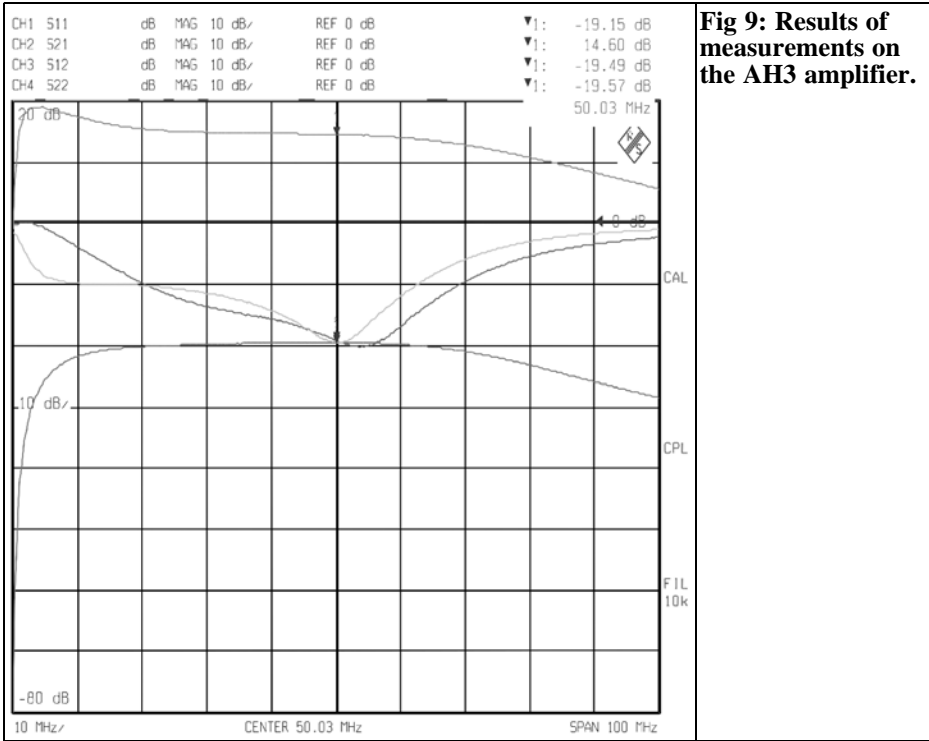
gets worse if the output matching is improved. Finally no matching was used. The results of measurement are shown in Fig 10.

Supply voltages of +5V for the AH3 and +9V for the AH101 are generated using regulators from the 12V system supply voltage. The voltage regulators are mounted directly on the printed circuit board, in order to dissipate the heat.

The chokes planned ( $1\mu\text{H}$ , size 1206, from Stettner) for supply voltage decoupling have a high DC resistance and therefore become too hot with the working currents (150mA and 200mA). Further the voltage drop across these chokes



**Fig 8: The simulation results for the AH3 amplifier.**



**Fig 9: Results of measurements on the AH3 amplifier.**

appeared too high. Therefore chokes from Coilcraft, type 1008 were used, they have a DC resistance of 1.75 ohms and a maximum permissible direct current of 350mA. The voltage drop is  $V = 1.75 * 150mA = 0.265V$ , in the prototype a voltage drop was measured as 0.45V, it is however higher for the lower supply voltage of 5V. Therefore a choke with a 0805 ferrite core was used in the final version, type B82498-B1102 from Epcos. The DC resistance is a maximum of 0.5, with a self resonance of 350MHz which is far enough away from the working frequency, see also [10].

Both amplifiers have a relatively high power dissipation (AH3: 0.7W; AH101 1.8W) so it is not sufficient to use the surface of the printed circuit board to dissipate the heat. In [9] it is pointed out that for a long life span good heat dissipation must be ensured. That can be done on the printed circuit board using

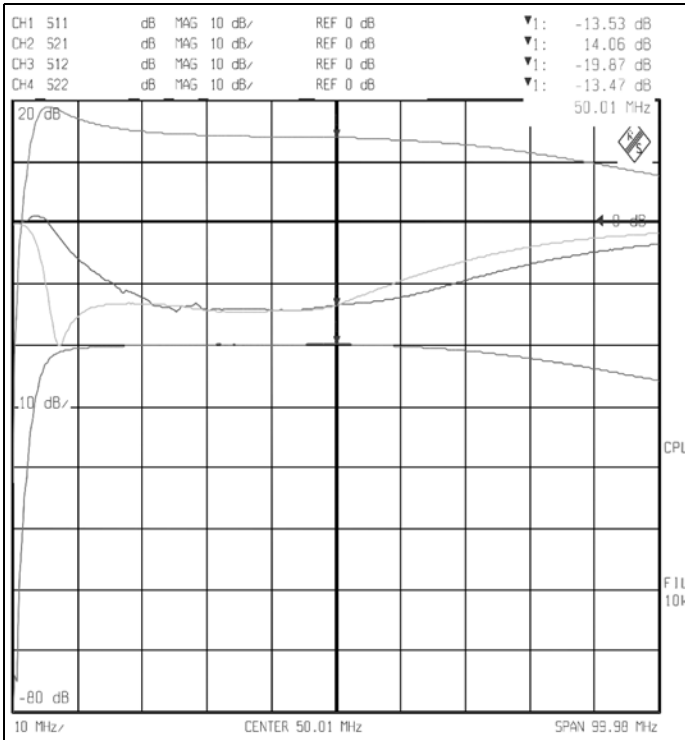
plated through holes around the earth pin of the ICs to transfer the heat to a 10mm wide and 20mm high aluminium cooling block fastened with M2.5 screws to the lower surface of the printed circuit board.

With a coax relay at the input of the driver the signal from the LNAs is fed to the mixer on receive. The relay is activated on transmit using the PTT signal and the signal to the driver is mixed up by 28MHz to 50MHz. A further relay at the output stage feeds the signal to the antenna.

## 8.

### Construction and alignment

The main double sided PCB is made



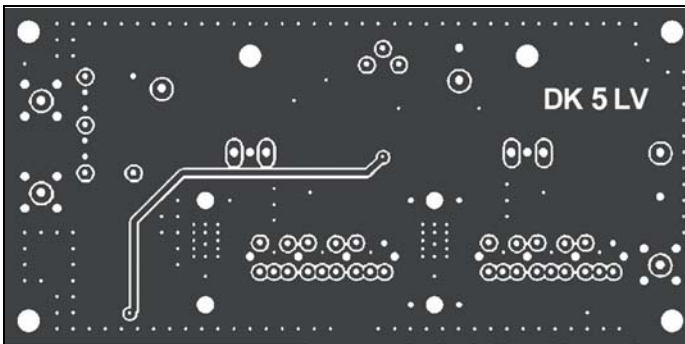
**Fig 10: Results of measurements on the AH101 amplifier.**

from 1mm thick FR4 material with plated through holes. The size of the board was chosen to fit into a standard 55mm x 111mm tin plate housing.

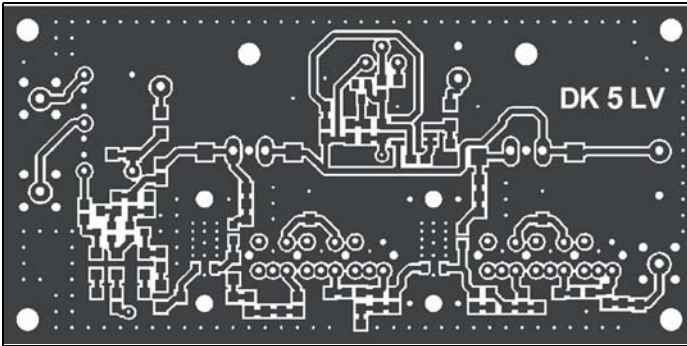
Fig 11 and 12 show the layout of the printed circuit board of the transverter for the SMD side and the component side. The layouts for the SMD side and component side are shown in Fig 13 and 14. Photographs of the prototype are shown

in Fig 15 and 16. It can be seen that some changes were necessary for the final version that are shown on the layouts.

Construction is best done in stages. First the pin diode PCB with IC3 and the voltage regulator IC2 is constructed and tested. Next the two band pass filters can be constructed and adjusted. The author used an 8753B network analyser to adjust the transfer function (S21) and the input



**Fig 11: PCB layout for the SMD side.**



**Fig 12: PCB layout for the component side.**

reflection (S11).

It is possible to adjust the filters for lowest attenuation in the pass band. As a reference the filter characteristics of the prototype are shown in Fig 17 for the frequency range 20kHz to 100MHz and in Fig 18 for the frequency range 50MHz  $\pm$ 5MHz).

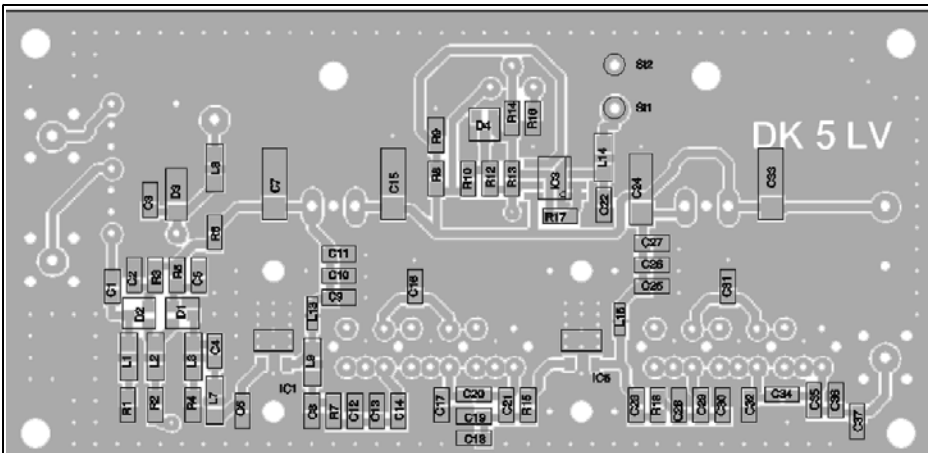
Finally the two amplifier stages are constructed. It should be noted that the earth surface under the ICs is in good thermal contact with the PCB because this is the way heat is dissipated. The ICs becomes perceptibly warm despite the aluminium heat sinks.

which was discovered by measurements via the 1 $\mu$ H drain choke. The failure happened because the earth surface was insufficiently soldered at the edges. Possibly the earth surface should be coated with SMD flux before ICs are fitted. Then fit the ICs by soldering the drain and gate connections and finally heat the earth surface by heating up the under side of the PCB with the soldering iron until the solder melts. An alternative would be to feed solder through the plated through holes form the underside of the PCB.

The AH3 dissipates 750mW (5V x 150mA) and the AH101 dissipates 1.8W (9V x 200mA).

After a long test the AH3 amplifier failed

**Continued on Page 35**



**Fig 13: Component layout for the SMD side of the PCB.**





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1999	Y	Y	Y	P
1998	Y	Y	Y	Y
1997	Y	Y	Y	Y
1996	Y	Y	Y	Y
1995	Y	Y	Y	Y
1994	Y	Y	Y	Y
1993	Y	P	P	P
1992	Y	Y	Y	Y
1991	P	P	P	P
1990	P	P	L	P
1989	P	P	P	P
1988	P	P	P	L
1987	P	P	P	P
1986	P	Y	Y	Y
1985	Y	Y	Y	Y
1984	Y	Y	Y	Y
1983	P	P	Y	Y
1982	P	P	P	P
1981	P	L	P	P
1980	P	P	P	P
1979	L	P	P	P
1978	P	P	P	P
1977	P	P	P	P
1976	P	P	P	P
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1974	L	P	P	P
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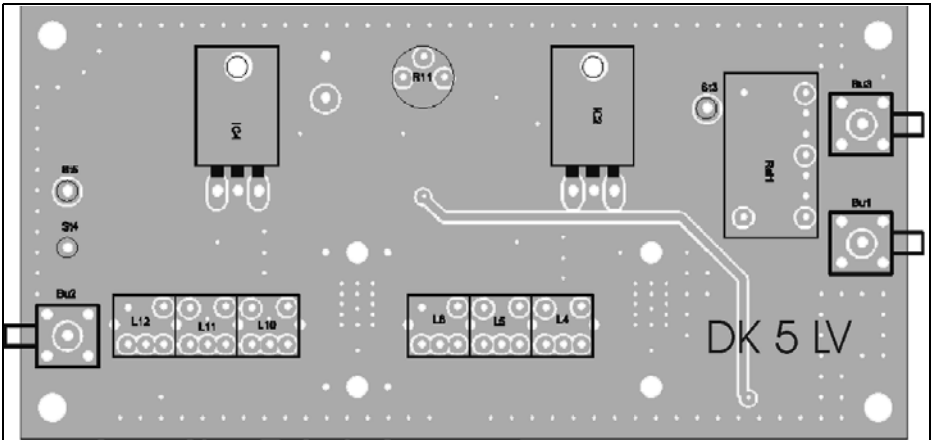
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**Fig 14: Component layout for the component side of the PCB.**

It is particularly important that there is good thermal contact of the AH101 to the earth surface of the printed circuit board.

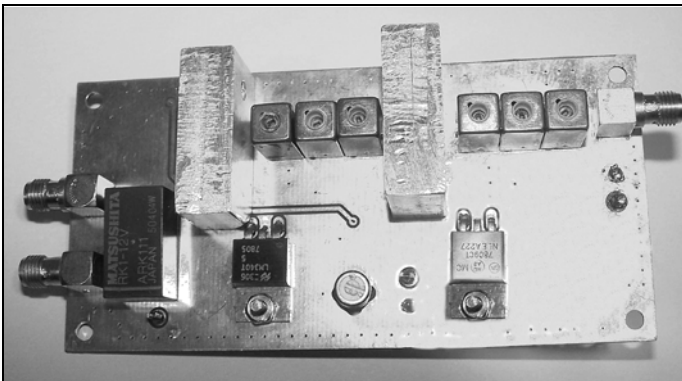
## 9.

### Measurement results

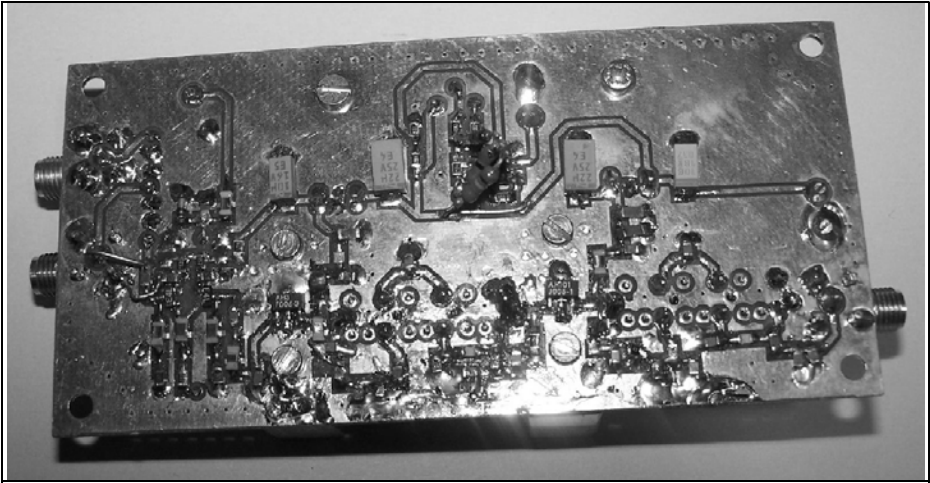
The frequency response of the complete driver is shown in Fig 19. The input adjustment is sufficient in the usable frequency range. It gets worse up to approximately  $-10\text{dB}$  for attenuation val-

ues for the pin diode around  $10\text{dB}$ . The overall gain amounts to  $16\text{dB}$ , which is about  $1\text{dB}$  more than the level plan. At  $44\text{MHz}$  an attenuation of  $60\text{dB}$  is obtained. At  $60\text{MHz}$  it is around  $80\text{dB}$  (the mix products from the  $50\text{MHz}$  converter on this frequency are attenuated by only  $29\text{dB}$  according to Table 1).

The intermodulation behaviour with a  $-7.6\text{dBm}$  two tone signal is shown in Fig 20. A signal-to-intermodulation ratio of third order products of  $-54\text{dB}$  is obtained. The signal-to-intermodulation ratios as function of the input power in the range  $-10\text{dBm/Tone}$  to  $0\text{dBm/Tone}$  is shown in Fig 21. The  $-8\text{dBm/Tone}$



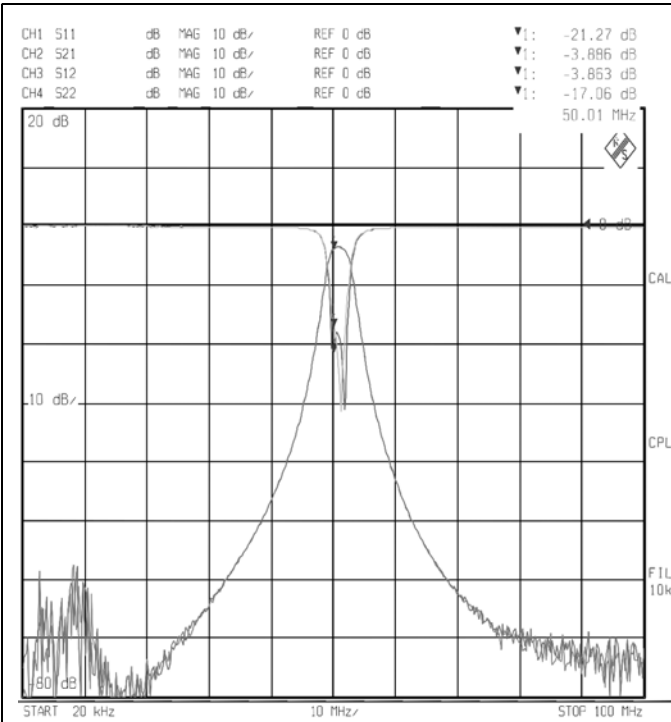
**Fig 15: Picture of the component side of the completed PCB.**



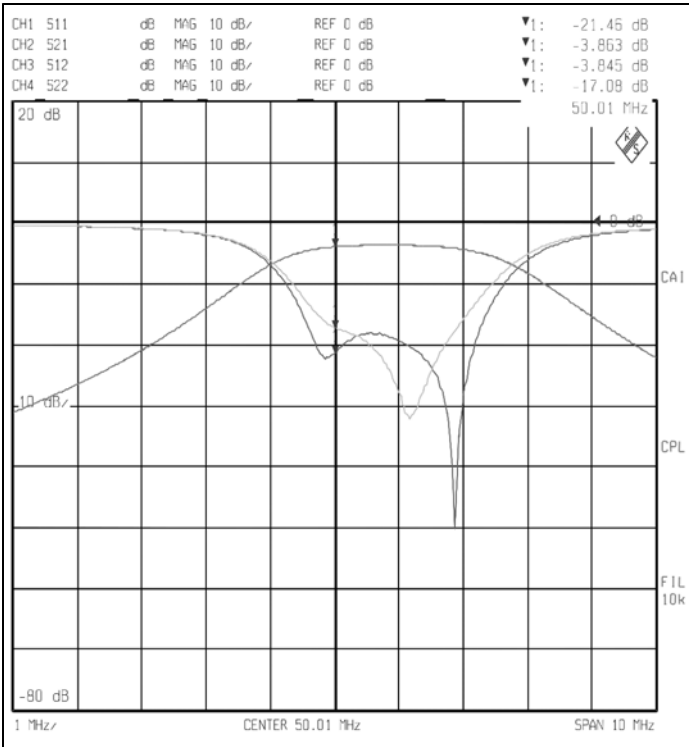
**Fig 16: Picture of the SMD side of the completed PCB.**

signal-to-intermodulation ratio for the driver is 60dB which means that the original specification is fulfilled. With

two input signals of 0dBm at the input of the transverter, the output obtained is 16dBm/Tone = 22dBm PEP. Considering



**Fig 17: Filter characteristics of the prototype for the frequency range 20kHz to 100MHz.**



**Fig 18: Filter characteristics of the prototype for the frequency range 50MHz ± 5MHz.**

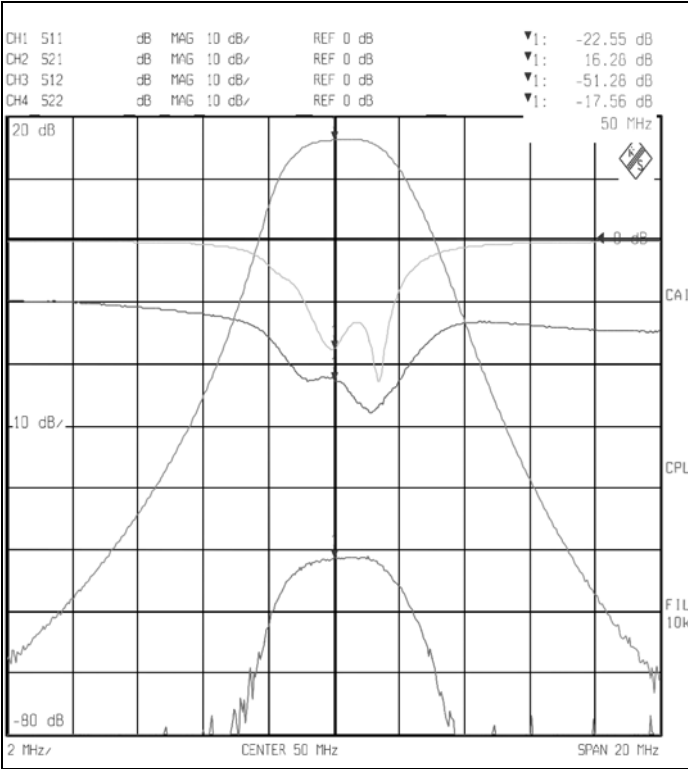
the attenuation of the band pass filter, the second amplifier supplies a power output of 25dBm PEP and therefore is operated 1dB below the compression point. The degradation of the signal-to-intermodulation ratio is explained because to data sheet shows for the output intercept point with two output signals of over +5dBm are measured with a frequency spacing of 10MHz.

**10.**

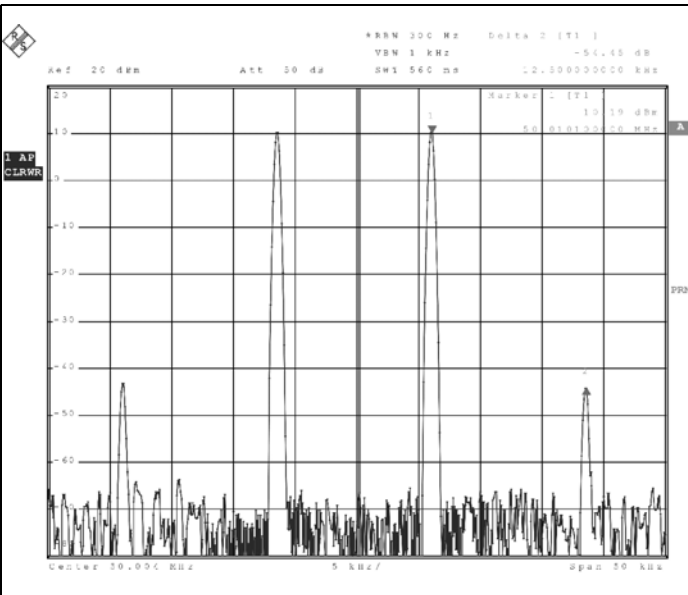
**Parts List**

- IC1 AH3 SOT 89, Watkin Johnson
- IC2 7805 TO220, National Semiconductors
- IC3 LM358 SO8 TI, National

- IC4 Semiconductors 7809 TO220, National
- IC5 Semiconductors AH101 SOT 89 Watkin Johnson
- D1, D2 HSMP 4810 Agilent
- D3 LL4148
- D4 BAS 16 SOT 23 Infineon
- R1, R4 560Ω SMD 0805
- R2 330Ω SMD 0805
- R3, R5 1k5Ω SMD 0805
- R6 680Ω SMD 0805
- R7, R15, 0Ω SMD 0805
- R18
- R8, 9, 10kΩ SMD 0805
- R12,
- R14, 16, 10kΩ SMD 0805
- R17
- R10 22kΩ SMD 0805
- R11 1KΩ 75 R Vishay
- R13 100kΩ SMD 0805
- C1, 2, 4, 10nF SMD 0805
- C5

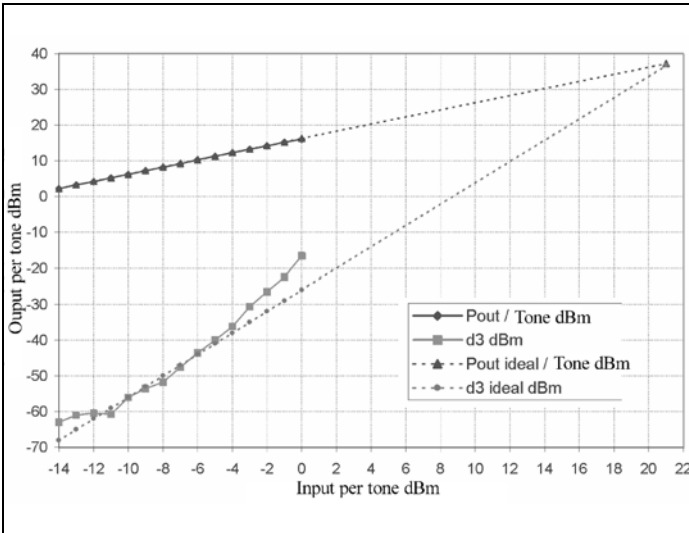


**Fig 19: The frequency response of the complete transverter.**



**Fig 20: The intermodulation products of the transverter with an input of -7.6dBm/tone.**





**Fig 21: The signal-to-intermodulation ratios as function of the input power in the range -10dBm/Tone to 0dBm/Tone.**

- C10, 22, 10nF SMD 0805
- C25
- C3, 11, 100nF SMD 0805
- C12
- C6 12pF SMD 0805
- C7, 15, 10µF/16V Kemet
- C24, 33
- C8 56pF SMD 0805
- C9, 19, 100pF SMD 0805
- C36
- C26-C28 100pF SMD 0805
- C14, 21 22pF SMD 0805
- C29, 35
- C16, C31 330pF SMD 0805
- C17, C32 18pF SMD 0805
- C20, C34 1pF SMD 0805
- C23 1000pF SMD 0805
- C30, C37 10pF SMD 0805
- L1 - 3, 1µH SMD 1206 Stettner
- L8, 14
- L4 - 6, 605nH, 5036 Neosid
- L10, 12
- L7 150nH SMD 1206 Stettner
- L9 120nH SMD 1206 Stettner
- L13, L15 1µH B82498-B1102 Epcos
- Bu1 - 3 SMA right angle socket
- Rel1 RK 12, Matsushita
- St1 - St5 1.3mm solder point

## 11.

### References

- 
- [1] The Charm of the 6m Band (1), special operating techniques, Martin Steyer DK7ZB, Funkamateur 3/2000, pp 299 - 301
  - [2] M57735 data sheet, Mitsubishi Semiconductors data book 1989, pp 2-82 to 2-87
  - [3] A wideband 30W push pull amplifier with two BLF244 MOS transistor (VDS = 28V); 25 - 110MHz application note NCO 8701, Philips Semiconductors, [www.Semiconductors.Philips.com](http://www.Semiconductors.Philips.com)
  - [4] Performance of a 30W push pull amplifier for the frequency range 25 - 110MHz with 2 BLF244 MOS transistors, Application note NCO 8702, Philips Semiconductors, [www.Semiconductors.Philips.com](http://www.Semiconductors.Philips.com)
  - [5] RD16HFF1 data sheet, Mitsubishi, [www.mitsubishichips.com](http://www.mitsubishichips.com)



[6] A Low cost Surface Mount pin diode Attenuator, Application note 1048, Agilent Technologies

[7] Data sheet AH3, Watkin Johnson, October 2003

[8] Data sheet AH101, Watkin Johnson, June 2003

[9] AH1 Thermal Resistance analysis, AH2, AH3, FH1, & FH101 Included by Similarity, Application note, Watkin Johnson

[10] SMT Inductors SIMID Series B82498-B data sheet, Epcos 04/00

[11] A modern 50/28 MHz, Henning Christof Weddig DK5LV, VHF Communications Magazine 2/2004 pp 95 - 115 and 4/2004 pp 23 - 248

[12] Design of a low-noise and large signal 50MHz transverter, Henning Christof Weddig, Conference booklet of the Munich amateur radio conference 2004, pp 37 - 54

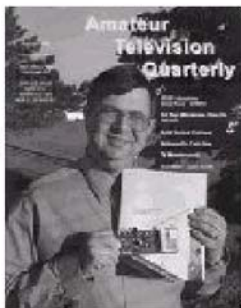
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*Gunthard Kraus, DG8GB*

# Practical project: Durable and reproducible antennas for the 2.45GHz WLAN band, Part 1

**It has always been the desire every radio amateur to increase the range of his or her station as simply and inexpensively as possible. Nothing changes, therefore in this article the same applies to the computer age WLAN range.**

## 1.

### Preface

Connection of computers into a network and to The Internet using a wireless "Access Point" usually uses around 2.45GHz. Radio amateurs use this techniques to make contact and try to increase the range within the legal limits. An alternative is to use directional antennas that concentrate the radiated power in a certain direction. They can be used for both the transmit and receive side with interesting possibilities:

- Higher receiver antenna signal over the same distance and thus better transmission quality (e.g. expressed as SINAD improvement) or a greater range for the same received signal.
- No more all round radiation thus reducing unwanted propagation

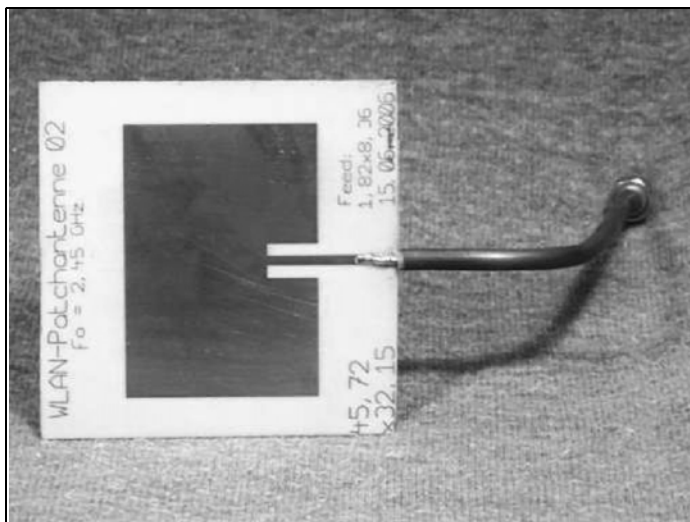
and/or directions of the signal.

- Mechanisms for practical bug proof point-to-point connections.

At 2.45GHz the wavelength is small (122.45mm in air) and the half wavelength, important for many antenna designs is six centimetres. Therefore quite useful antennas can be designed and the patch antenna was chosen for this design.

This antenna is:

- Extremely simply and durable, because it consists of just a double sided printed circuit board.
- It can be designed in such a way that it is matched to 50Ω therefore it does not need an additional matching network.
- It shields against backward radiation and therefore offers a higher gain and an excellent front to back ratio which is half the radiation pattern of a dipole.
- It has a maximum gain of 6.5dBi which doubles the receive signal in the forward direction compared to an isotropic antenna. Compared to a dipole or vertical antenna there is about 4.5dBi of gain.



**Fig 1: A rectangular patch antenna with a  $50\Omega$  feed.**

If the transmitter uses a similar antenna the overall gain of the connection is approximately 9dB over the original antennas and somewhat more than the “three-way” signal. (WLAN range measurements are often quoted for a three-way connection i.e. three connected PCs – ed.). If this thought process is extended for the far field, the received signal is greater in a greater distance therefore the “three-way” range can be increased.

Naturally there are some points to be considered:

The patch antenna is narrow band with a radiation pattern in form of half an eight for a single patch and is ideally operated on a single frequency. Therefore the performance at the band edges must be considered. During the design thicker material and larger width patches can be used to give a wider band. In practice this will probably reduce the three-way range.

The requirements of the printed circuit board are very high. Apart from absolute mechanical stability and small losses at the operating frequency, small temperature and humidity dependence of all electrical parameters are demanded. Ad-

ditionally the manufacturing accuracy must be kept within 0.01mm. On the positive side, if everything is under control, the finished antenna is very durable and can withstand falling to the ground without damage. The antenna can be reproduced easily, even for large production volumes.

Some articles have already appeared about the function and design of patch antennas for different frequency ranges, e.g. [1], [2] or [3]. Therefore the introduction is to be kept very short and aimed directly at the practical project.

Not only will the design procedure and the necessary steps up to production stage be considered but an important criteria is not to use expensive full versions of the EM Simulator programs. This discourages anyone from using pirate copies of these applications. Therefore one tries to get along with programs that can be obtained free of charge from the Internet, these have things missing that are replaced with an imaginative approach.



**Fig 2: On the left is a rectangular patch antenna with an SMA connector for the feed. On the right is a circular patch antenna with an SMA connector for the feed.**

2.

## Defaults and product specifications

Three different forms of single patch antennas are to be designed, built and tested.

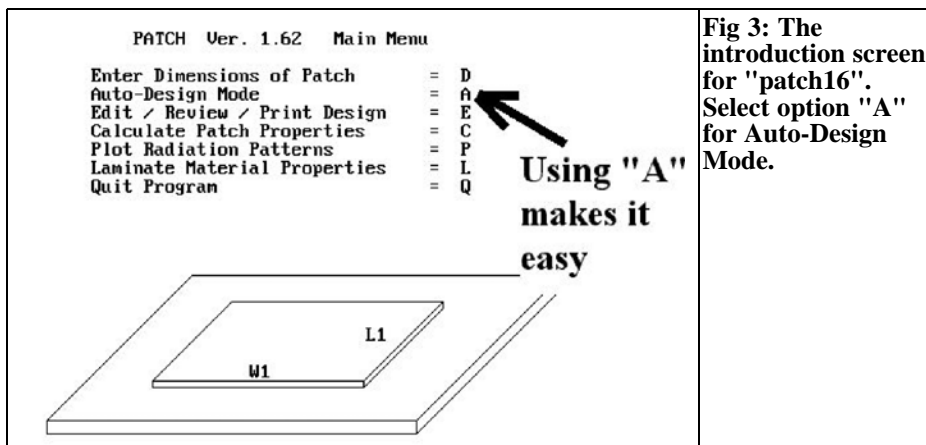
The size of the printed circuit boards is to be about 80mm x 60mm, fed by an SMA socket.

The resonant frequency is 2450MHz, with a bandwidth of 60MHz required to incorporate the whole frequency range from 2420 to 2480MHz.

The PCB material used is Rogers R04003 with its excellent mechanical stability and workability. It has a thickness of 1.52mm (= 60 MIL), 35µm copper coating, a dielectric constant of 3.38 as well as a dissipation factor of 0.001 at 2GHz.

A more detailed look at the finished products:

- A rectangular patch with a 50Ω “Insert feed” is shown in Fig 1. The start point of the feeder on the face of the patch is chosen to give an input impedance of 50Ω. A semirigid cable with an SMA plug is used as a temporary connection, it is better to use a surface mounted SMA plug soldered to the lower surface of the PCB with the centre conductor soldered to the stripline.
- Fig 2 shows the two other antenna designs. These are a rectangular and a round Patch. The feed is made within the patch face by an SMA socket, which is soldered on the lower surface. Both versions are handy and durable but the round version might prove better in hard everyday life because of the missing corners and edges. This is at the cost of a narrower bandwidth. In addition it must be connected to the transmitter and receiver with more care in order to align the polarisation correctly. Even for specialist this is a bit unusual.



**Fig 3: The introduction screen for "patch16". Select option "A" for Auto-Design Mode.**

3.

## Development of the rectangular patch antenna

To start, the simple rectangular antenna is chosen, fed from the lower surface. There is a free design program available on The Internet called "patch16". It will be interesting to test how it compares with a genuine high speed EM simulator and to manufacture the first PCB after the simulation. This will give an idea of the differences that can be expected.

### 3.1. Design from the program "patch16"

This program is loaded, installed and started from the Web onto your computer, it runs perfectly under Windows XP. When greeted with the message "Do you need Instructions" answer "no". The following steps are shown in Fig 3, choose "Auto design mode -A" and use the following values for the questions:

- Resonant frequency: 2.45GHz
- Range: 60MHz
- Dielectric constant: 3.38

- Dissipation factor: 0.001  
(Loss tangent)

After these values have been entered, the screen shown in Fig 3 is displayed, but there is a malicious trap to consider (Fig 4). The program has chosen a PCB thickness of 88 MIL (because of the square patch format), in order to give the desired range, however a thickness of 60 MIL was required. As long the square format is chosen, the program cannot change the calculated PCB thickness therefore another way must be found.

Consider the following:

- A larger range for a patch antenna means a thicker PCB or a bigger patch width with the same patch length (additionally a high dielectric constant helps).
- So the patch width must be increased (with a constant PCB thickness of 60 MIL) until the required range of 60MHz is reached. If necessary the patch length must be changed in order to give again the desired resonant frequency of 2450MHz. That can be done as follows:

Note the suggested values for the length (1.24 inch), the width (also 1.24 inch) as well as the feed point (0.45 inch) and press the letter "D" on the keyboard.



**This is not wanted**

Feedpoint = 0.45 in  
THK = 0.008 in

PATCH Ver. 1.62 Main Menu

- Enter Dimensions of Patch = D
- Auto-Design Mode = A
- Edit / Review / Print Design = E
- Calculate Patch Properties = C
- Plot Radiation Patterns = P
- Laminate Material Properties = L
- Quit Program = Q

**Use option "D" to go further**

**Fig 4: The Auto-Design mode gives a PCB that is too thick. Use option "D" to change this.**

Enter these values including the data of the PCB material R04003. Once this is done the screen will be like Fig 4. Pressing "C" will calculate the patch parameters including the transformation line if required.

the small menu. The patch length, patch width and feed point can be changed, always return to the main menu and press "C" to calculate the results. Finally the results shown in Fig 5 can be achieved. This does not take as long as it sounds.

Now increase the patch width to get the required bandwidth of 60MHz. After optimisation the final parameters are shown in Fig 5 with the required resonant frequency and bandwidth. The final state after successful optimisation is to be seen in Fig 5, now tuned for both the resonant frequency and the range. In addition the radiation resistance for the patch edges has been computed. Looking at the main screen shown in Fig 6 the PCB thickness is now 60 MIL. The secret for doing this is revealed.

All important data is summarised in following table:

- Resonant frequency: 2.450GHz
- Bandwidth: 60MHz approximately
- Gain: 6.5dBi
- Radiation resistance: 116.88Ω  
At patch edge
- Patch length: 1.2595 inches = 31.99mm
- Patch width: 2.2 inches = 55.88mm

If "E" is pressed all of the design parameters are available to be edited as shown in Fig 7 and can be changed using

The Resonant Frequency is 2.450 GHz  
Qo is 31.3

The Edge Radiation Resistance is 116.88 ohms  
Zc of Quarter-wave transformer is 76.2 ohms  
Approx. width of the Quarter-wave transformer is 0.866 inches  
Length of Quarter-wave transformer is 0.756 inches at the Resonant Freq.

Input Resistance at probe location is 50.75 ohms

The 2:1 USWR Bandwidth is 2.4%  
Upper Frequency Limit = 2.479 GHz  
Lower Frequency Limit = 2.428 GHz

Press 'ENTER' to continue:

**OK!**

**OK!**

**Fig 5: The results of the "C" calculate option shows the results for the patch antenna.**



<p><b>Now it is 60 MIL</b></p> <p>Feedpoint = 0.34 in TH = 0.868 in</p>	<p>PATCH Ver. 1.62 Main Menu</p> <pre> Enter Dimensions of Patch      = D Auto-Design Mode              = A Edit / Review / Print Design  = E Calculate Patch Properties     = C Plot Radiation Patterns       = P Laminate Material Properties  = L Quit Program                   = Q         </pre> <p>DC = 3.38 Loss Tangent = .001</p>	<p><b>Fig 6: The PCB thickness is now correct.</b></p>
---	---	--

- Distance of the feed: 0.34inches  
distance from the front = 8.64mm  
patch edge
- Wavelength in air 122.45mm  
at 2450MHz

This basic information is enough to be able to analyse the design with a good EM simulator. Currently the best free of charge simulator is "Sonnet Lite".

### 3.2. Preparation for the antenna design with Sonnet Lite

The software can be downloaded and installed from the Sonnet homepage ([www.sonnetusa.com](http://www.sonnetusa.com)). When the software is started only six buttons appear on the task bar. Unfortunately the Farfield viewer is missing which is a shame. To overcome this select the Admin menu and choose the "Register Sonnet Lite" option as shown in Fig 8. Follow the

prompts and increase the available memory from 1 to 16Mb which will increase the number of available tasks significantly - and all free of charge!

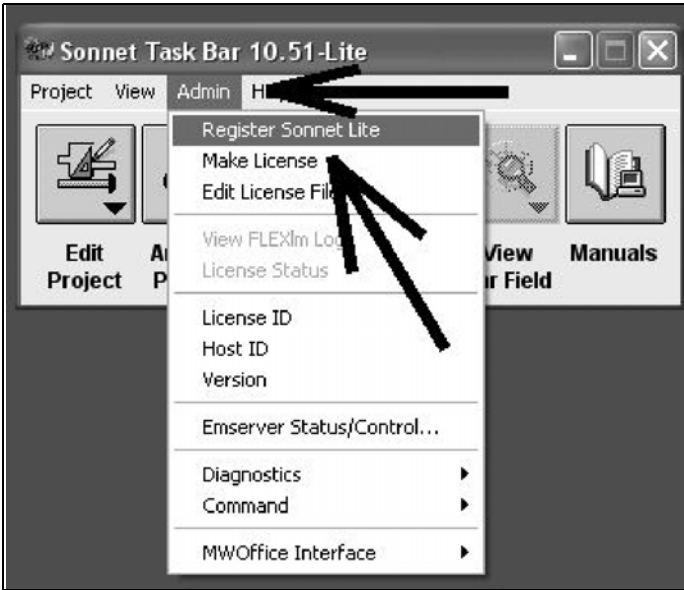
Now the simulation can be started by selecting "Edit Project" then "New Geometry". After the editor is loaded, first store the project in a suitable place with a suitable name. Then carefully carry out the following steps:

**Step 1:** In the "Circuit" menu, choose the "units" option. Ensure that millimetres and Gigahertz are selected for changes to physical values as shown in Fig 9.

**Step 2:** Again under "Circuit" select the "Dielectric Layers" option shown in Fig 10. The PCB for the antenna is on the lower line and above it is "Layer 0" which is used by Sonnet as an air layer with a thickness of half a wavelength

<p><b>These are the design parameters:</b></p> <p>Length (L) = 1.2595 inches  Width (W) = 2.2 inches  Height (H) = 5.999999E-02 inches  Dielectric Constant (D) = 3.38  Loss Tangent (T) = .001  Feedpoint Distance (F) = .34 inches</p> <p><b>Do you wish to edit any value? (Y/N):</b></p>	<p><b>Fig 7: The "E" (Edit) option goes to this menu where many parameters can be changed.</b></p>
--	--

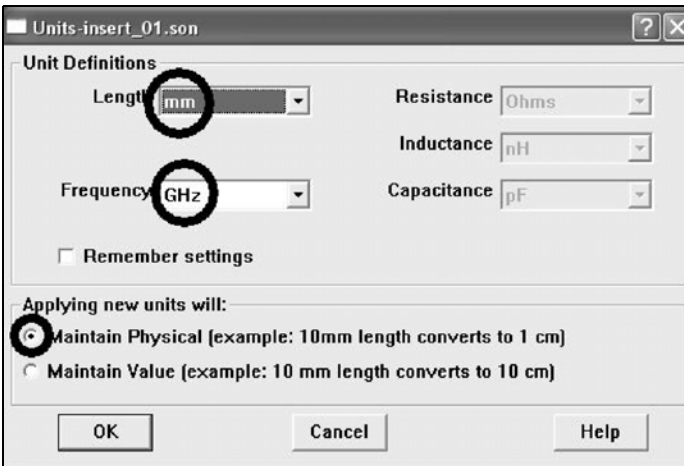




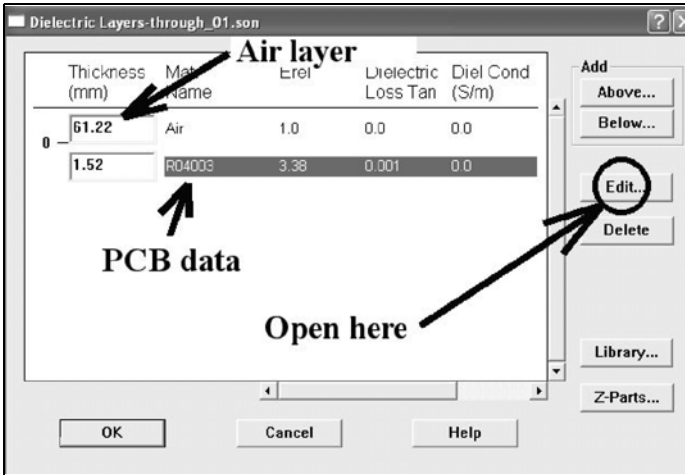
**Fig 8:** Use this menu to register Sonnet Lite and increase the available memory from 1 to 16Mb.

(61.22mm). First click on the lower line to highlight the line. Use “Edit” to change the data for the RO4003 material as shown in Fig 11. Then select the top line and edit the air layer with the following data (Thickness = 61.22mm, Erel = 1.0, Dielectric loss tangent = 0.0, Dielectric Conductivity = 0.0). The screen should then look like Fig 10. The magnetic characteristics (Mrel = 1.0, Magnetic loss tan = 0.0) take over!

**Step 3:** Again under “Circuit “ select “Metal Type”, at the start there is only one entry for “lossless”. Therefore click on “ADD” to open the metal editor. Click on “SELECT metal from LIBRARY” and then “global LIBRARY” which displays a metal selection menu. Choose “copper” to get the screen shown in Fig 12 and ensure that the thickness is adjusted to 0.035mm, then click “OK” to get to the menu shown in Fig 13. Change



**Fig 9:** Use this menu to change the dimensions to mm and frequency to GHz.



**Fig 10:** Set the correct values for the PCB material (R04003) and the air layer.

the “Metal for New Polygons” to copper and confirm with “OK”. All of the basic adjustments are now made and the antenna design can begin.

point is on the centre line and 8.64mm from the lower patch edge.

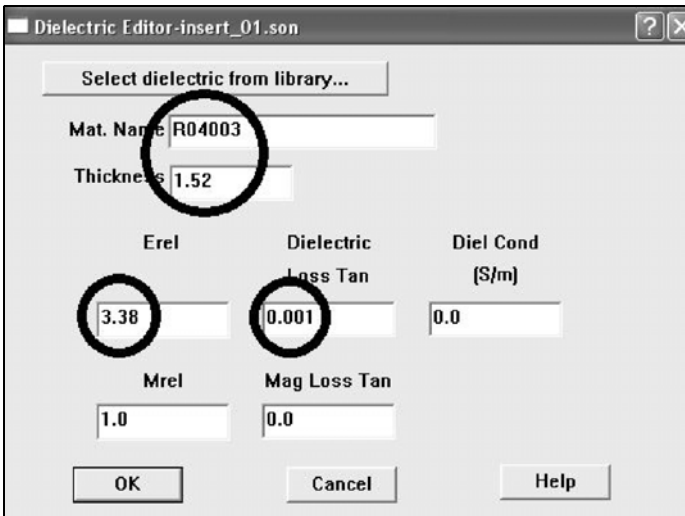
Now the most important work starts:

### 3.3. Antenna design with Sonnet Lite

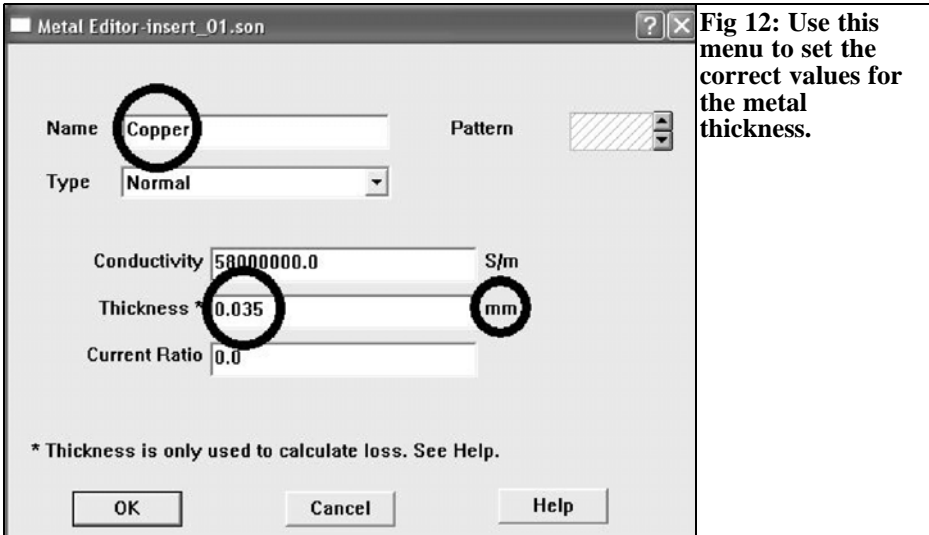
The antenna design must be broken down in small elements for the simulation, this is limited to the point where the memory limit of 16Mb is reached.

You should be familiar with the representation of the antenna outline by coordinates related to the lower left hand corner as shown in Fig 14 because this is the convention that the Sonnet editor uses. The length (31.99mm) and width (55.88mm) are taken from 3.1. The feed

Experience must be used to decide on the best compromise, this indicates that the length and width of the smallest cell ( $\Delta x$  and  $\Delta y$ ) are approximately 1% to 2% of



**Fig 11:** Use the edit menu to set the parameters for the R04003 PCB material.



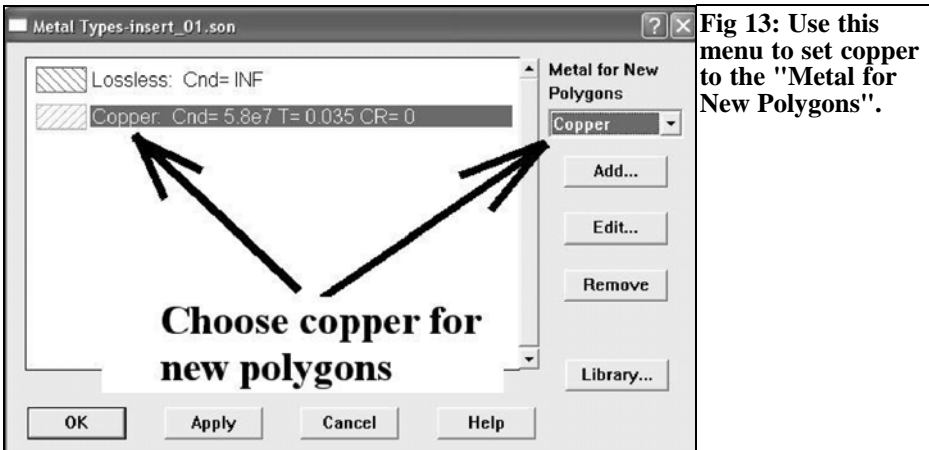
the wavelength. The pain threshold for rapidly rising simulation inaccuracies is up to approximately 5% and down to approximately 0.4%. In addition the manual suggests that the antenna edges should be at least one wavelength from the housing wall (Sonnet simulates everything in a rectangular metal box), in order to keep disturbing influences of the box wall small. But, the larger the better however the 16Mb memory limit sets the ultimate limit.

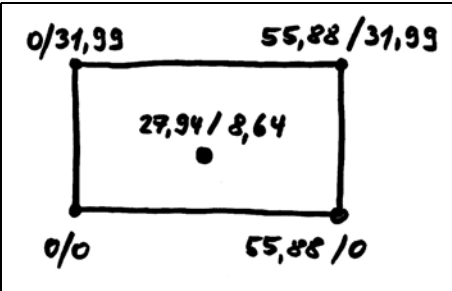
wavelength, the cover of the box forms the free space area where the radiation begins. That was taken into account in the Dielectric Layers setup.

The definition of  $\Delta y$  is for the patch length. It corresponds about the half a wavelength (50% of  $\lambda$ ) and therefore only needs the length to be divided by 50 to be 1% of the wavelength (in the PCB material, not in air!):

$$\Delta y = 31.99\text{mm}/50 = 0.6398\text{mm}$$

The air layer above the PCB is half a





**Fig 14: You should be familiar with the way that Sonnet Lite uses coordinates by making a sketch showing magnitude and coordinates.**

If a wavelength in air of 122.45mm is added to the patch length on both sides the calculation becomes:

$$122.45\text{mm} + 122.45\text{mm} + 31.99\text{mm} = 276.89\text{mm}$$

And then:

$$276.89 \text{ mm} / 0.6398 \text{ mm} = 433 \text{ cells in the y direction.}$$

The patch width can use the same cell size,  $\Delta x$ , as  $\Delta y$ . However further thought is needed because the next step in the antenna design is to add the feed. This is placed on the patch face to match to 50Ω. The width of this is determined using a suitable microstrip calculator or micro-

wave CAD program (PUFF, Ansoft designer SV, Appcad etc.) using the available PCB data. All programs agree a value of 3.48mm, since no housing with covers is required (and PUFF or Appcad are not able to include the option "cover height" in their simulation).

Therefore strip line width is used to set the cell size and get the best from the antenna simulation

Thus simply divide the width value of 3.48mm into four equal parts to obtain:

$$\Delta x = 3.48\text{mm} / 4 = 0.87\text{mm}$$

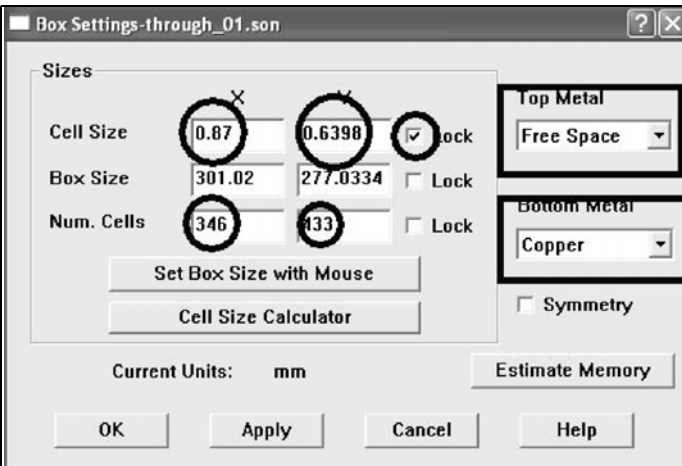
That is about 1.36% of the PCB wavelength and thus everything is in the green range for the Sonnet simulation. To determine the number of cells in the x direction:

The patch width is 55.88mm. On each side, add a wavelength of 122.45mm to get:

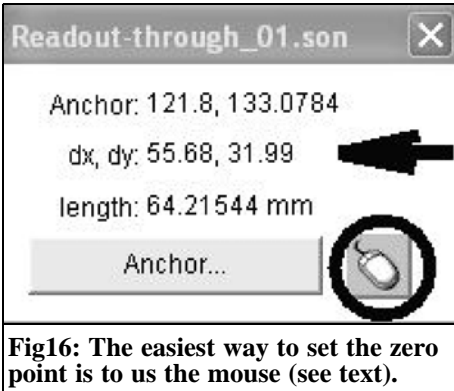
$$122.45\text{mm} + 122.45\text{mm} + 55.88\text{mm} = 300.7 \text{ mm}$$

That results in:

$$300.78\text{mm} / 0.87\text{mm} = 346 \text{ Cells in the x direction.}$$



**Fig 15: These are the "Box Settings" for Sonnet Lite to operate with. The x and y cell details are shown as well as the "Top Metal" set to "Free Space" and the "Bottom Metal" set to "Copper".**



**Fig16: The easiest way to set the zero point is to use the mouse (see text).**

The associated “box” parameters for the Sonnet circuit menu are shown in Fig 15.

- The line “Cell Size” is set to  $\Delta x$  and  $\Delta y$  with the “Lock” box checked
- The line “Num. Cells” is set to the correct number of cells for x and y
- “Top Metal” is set to “Free Space”
- “Bottom Metal” is set to “Copper”

If the cell size  $\Delta x = 0.87\text{mm}$  is used there is a problem because the patch width of  $55.88\text{mm}$  is not accurately calculated. This is not so tragic because a difference in the patch width up to  $1\text{mm}$  does not

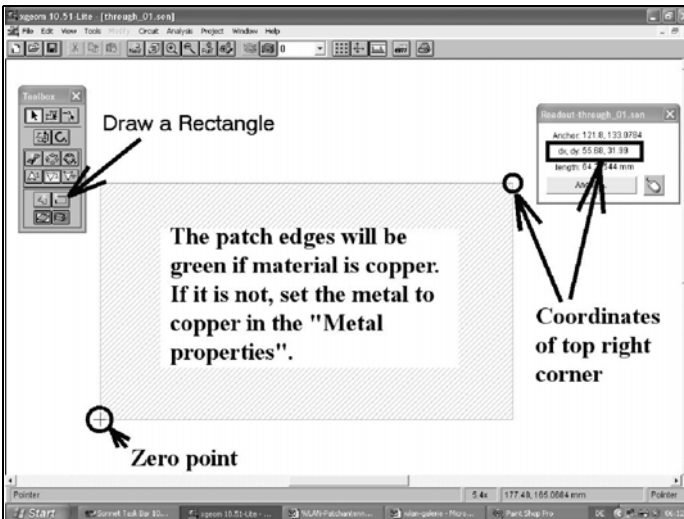
make much difference but a difference of  $0.1\text{mm}$  in the feeder width makes a big difference in the feeder impedance.

The simulation can be started by drawing the antenna but before that select “Measuring Tool” on the “View” menu and click the mouse symbol as shown in Fig 16. The cursor pointer changes to a small cross which is the zero point or lower left corner of the patch, it can, for example, be set on the centre of the display surface. The patch can now be drawn by clicking on “Draw Rectangle” and pulling with the mouse. The sizes of the two values  $\Delta x$  and  $\Delta y$  can be seen in “Measuring Tool”, these should be set to;

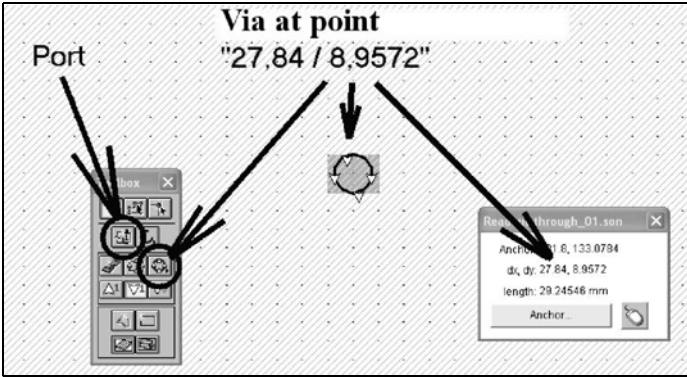
$$\Delta x = 55.68\text{mm} \text{ and } \Delta y = 31.99\text{mm}.$$

The display surface can be zoomed to make this easier. The patch area should be green, which means copper metallisation. If this was not set in “Metal Type” and “Material for New Polygons” it can be set by right clicking the mouse on the patch area and using “Metal Properties” as shown in Fig 17.

Now set a “Via” (plated through hole inside diameter =  $1.27\text{mm}$ ) at the feed-point ( $27.84\text{mm}$ ,  $8.9572\text{mm}$ ). The “Via” should be copper and will show in the



**Fig 17: All of the information about the patch is shown here.**

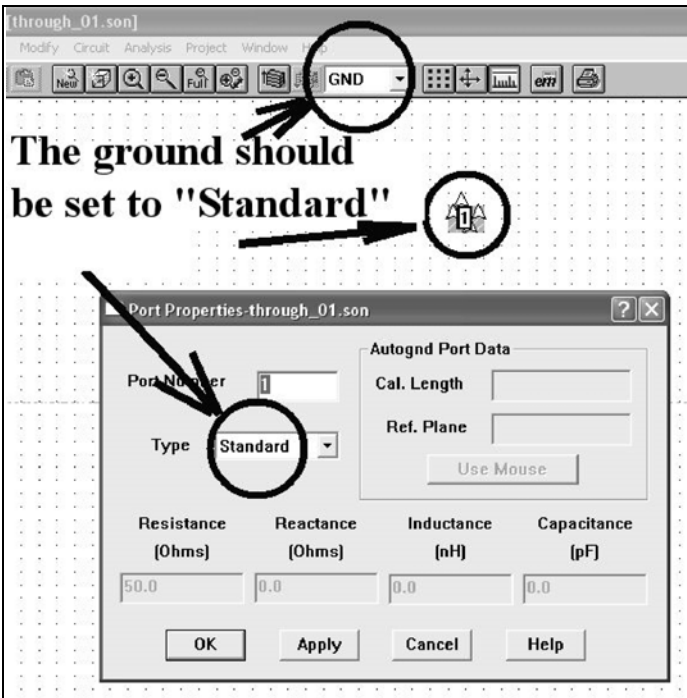


**Fig 18: Setting a Via, it must be set to Copper.**

same green colour as the patch area. Sometimes that is not the case so the Via must be marked by a right hand mouse-click and then the property in the appearing menu changed to "Copper" (Fig. 18). At last a Port must be set exactly at the position of the Via. Then switch Layer to "Ground", mark there the Port by a double-click and choose the property "Standard" in the Port Property Menu.

Now the drawing work is completed! (Fig.19).

Things are fast from now on. In the "Analysis" menu there is a Setup for the sweep. Fig. 20 shows the attributes necessary for "Adaptive Bandsweep" in the frequency range from 2.4 to 2.5 GHz. Fig. 21 shows the Simulation Results from clicking "Project", then "Analyze" after waiting for the computing time (if



**Fig 19: Switch the layer to GND and then the Port Type to "Standard".**

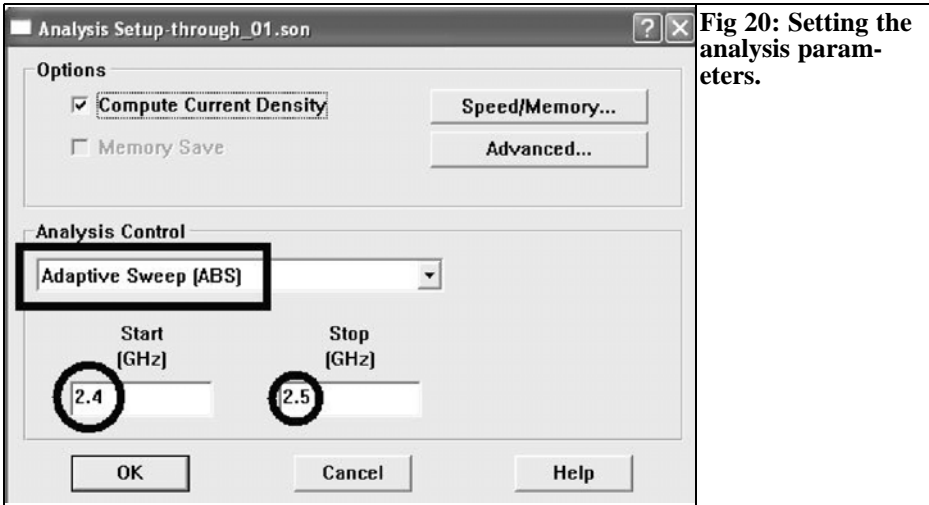


Fig 20: Setting the analysis parameters.

the project has been correctly stored). The key to show the results in a diagram can also be found in Fig. 21 and we see that the resonant frequency  $f_0$  2449 MHz

could not be better, but the S11 curve is more worrying. A Smithchart can be chosen from Fig 22 which is much more interesting and shown in Fig 23. S11 of

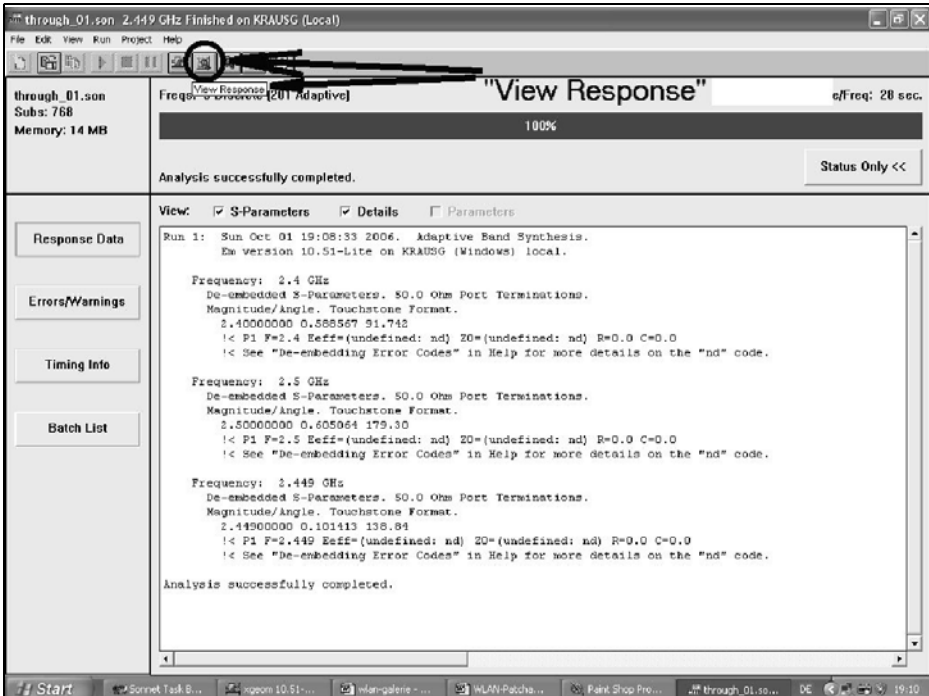


Fig 21: The results of an analysis.

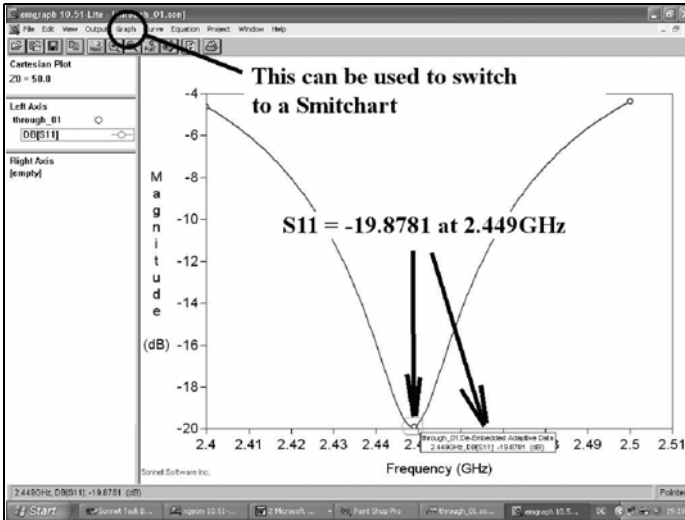


Fig 22: The results of the simulation of the design from "patch 16".

approximately  $-20\text{dB}$  is a reflection factor of approximately 0.1, BUT the feed point is too far from the lower patch edge because the loop in the  $S_{11}$  curve does not enclose the centre of the Smithchart and that means that the impedance value is under  $50\Omega$ ! The simulation is repeated after moving the Via to the next smaller distance of 8.3174mm using the mouse. This gives a good result with  $S_{11}=-30\text{dB}$ , near the correct matching point at  $50\Omega$  (Fig 24) with an acceptable increase of

the resonant frequency of 2505MHz. Unfortunately the feed point can only be moved in steps of  $y = 0.6398\text{mm}$  (because we have to stay within the 16Mb memory limit) and that is simply too rough for a finer approximation. So we have to be content with this simulation result, and a PCB with the external dimensions of 80mm x 60mm with the feed point as shown in Fig 25 is designed and produced.

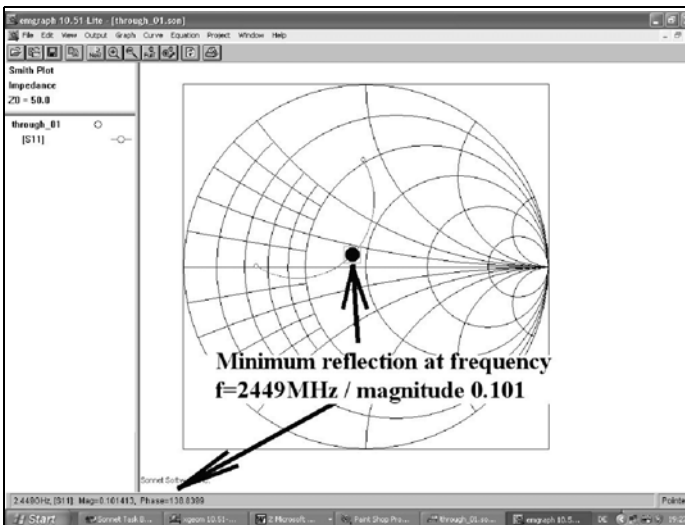
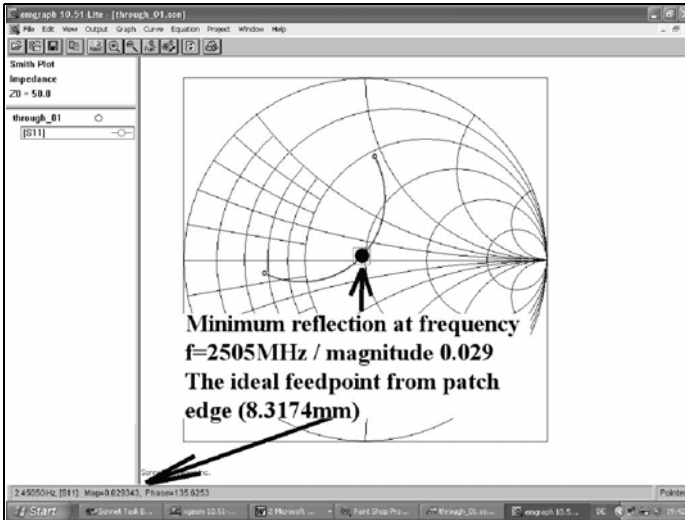


Fig 23: The results shown as a Smithchart.





**Fig 24: Smithchart of the results after moving the feed point.**

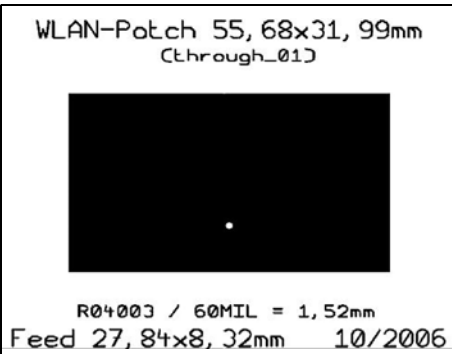
Because of many reasons e.g. PCB external dimensions may be smaller than assumed in the simulation, manufacturing inaccuracies and edge under etching, influence of the soldering the feed point within the patch area to the SMA connector, differences in the material data etc., the result of measurement using a network analysers will surely deviate from the simulation. Then a second PCB is necessary with some changes!

To be continued

**X.**

**References for Part 1:**

- 
- [1] Practical Project: A patch antenna for 5.8GH, Gunthard Kraus, VHF Communications Magazine, 1/2004, pp 20 - 29
  - [2] An interesting program, SonnetLite 9.51, Gunthard Kraus, VHF Communications Magazine, 3/2004, pp 156 - 178
  - [3] Modern Patch Antenna Design Part I and Part II, Gunthard Kraus, VHF Communications Magazine, 1/2001 and 2/2001, pp 49 – 63 and 66 - 86



**Fig 25: Finally the finished PCB drawn using TARGET3001. Remember to add enough information to uniquely identify the PCB.**



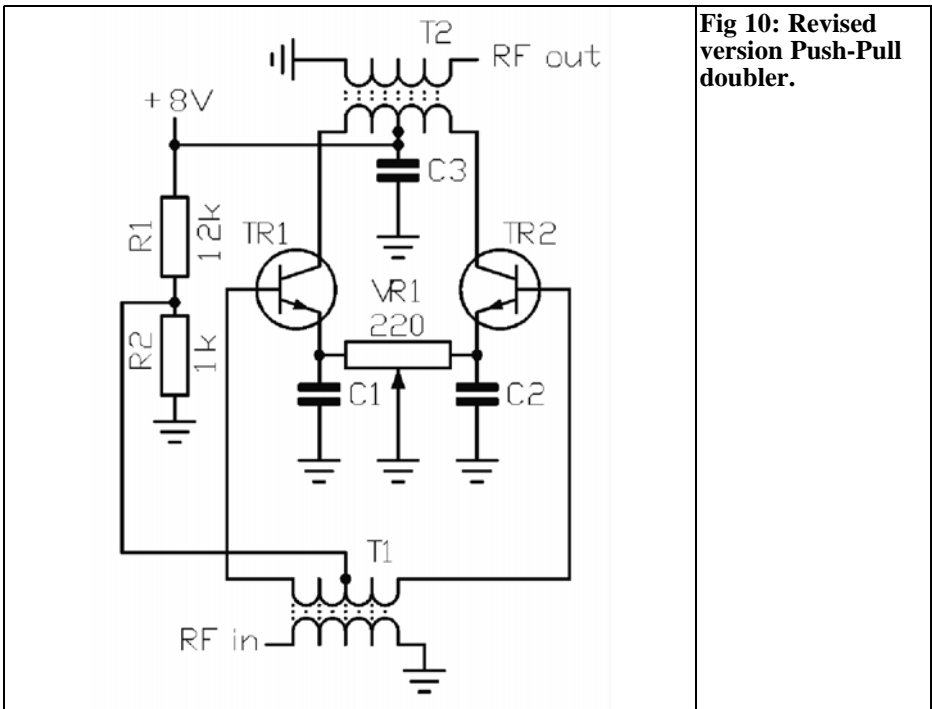
John Fielding, ZS5JF

# Corrections to transistor multiplier article published in issue 4/2006

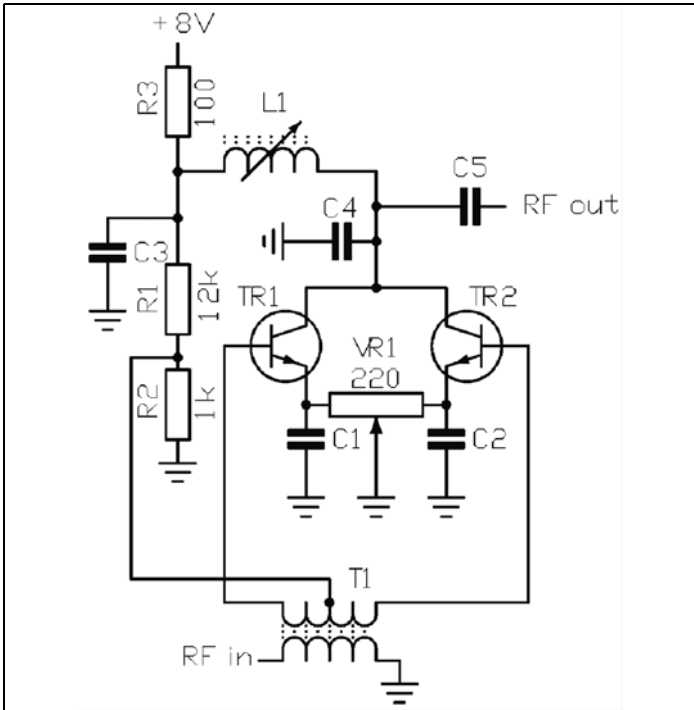
## 1. Error in Fig 10 and 11

Following the publication of the transistor multiplier article John noticed that there was an error in Fig 10 and 11.

Essentially he made a mistake with the base bias connection. He drew Fig 10 and then modified the file to generate 11 so the error was carried over. The base of TR1 should only connect to the secondary of T1 and not as originally drawn. The corrected Fig 10 and 11 are shown here.



**Fig 10: Revised version Push-Pull doubler.**



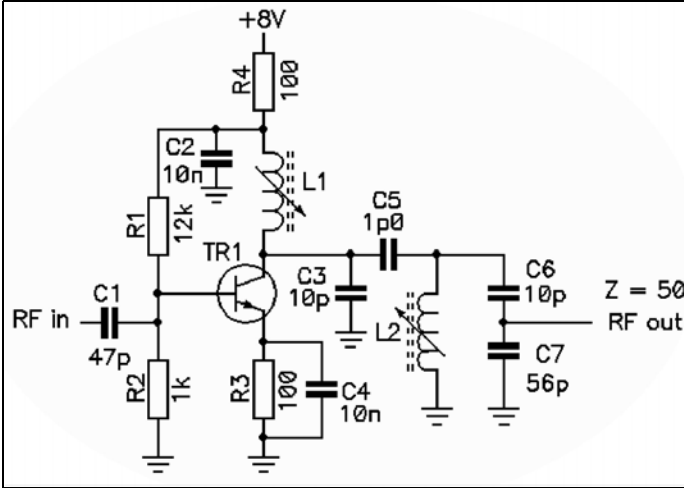
**Fig 11: Revised version of Push-Pull tripler.**

**2. Power output of tripler**

John has received an enquiry from an Italian reader who had built the 50 to 150MHz tripler shown in Fig 6 & 9. He was unable to obtain the same power output that John had achieved and was puzzled as to why.

John dug out the prototype and ran some tests only to find it gave the power claimed. It was then that John realised what the problem was. Fig 6 & 9 are the same circuit but Fig 9 has the extra test point resistors. These are not the problem. The output matching is where the loss of power occurs. The schematics show a 1pF output coupling capacitor (C7), this assumed the multiplier is feeding another transistor as either a buffer amplifier or a second multiplier. The

component values are dimensioned to feed into an impedance of about 500Ω. If trying to drive 50Ω there will be a large mismatch. To drive 50Ω requires a change to the output matching and the capacitor values. The modified version of Fig 6 is shown here. The values of C6 & C7 will depend on the value of L2 and the operating frequency. Normally C7 will require varying in value to give the correct impedance match, this depends on the value of L2 and the dynamic impedance occurring at resonance. The value of 56pF is a good starting point when the inductor is about 100nH with a Q of 100. If the inductor Q is lower then C7 reduces in value and if L2 has a higher Q then C7 increases in value. The matching is a standard "capacitor-tapped" arrangement.



**Fig 6a: Version of the typical frequency multiplier stage with 50Ω output matching.**

### 3.

#### Input matching

Some confusion arises about the input drive power measurement. John stated that +7dBm was sufficient to drive the multiplier. This of course is measured from a 50Ω drive source. The base of the multiplier looks like about 500Ω, so the input drive needs to be matched from the 50Ω driver. As long as the input base voltage swing is about 1V p-p or more it will work OK. To drive from a 50Ω source requires a step-up transformation to obtain the voltage swing. Either a 4:1 or 9:1 transformer or another BJT configured as an amplifier will work. In his prototype the 50MHz crystal oscillator output was about 1V p-p into a high impedance and this drives an emitter follower which then drives the multiplier input. Hence, the input drive is still about 1V p-p.

A signal generator when correctly terminated (e.g. driving a pure 50Ω load) delivers the indicated voltage, this is the potential difference (PD) mode. But if the signal generator is operated into an

open circuit it delivers an output voltage of twice the PD mode, the electro motive force mode (EMF).

If the multiplier base is driven from a 50Ω signal generator the drive voltage approaches that of an open circuit mode because the base looks like a relatively high impedance, about 500Ω. So evaluating the exact drive signal level is very difficult if a signal generator is used as the drive source. The reason this occurs is because signal generators are NOT conjugately matched devices but have an output resistance of 50Ω, usually furnished by an attenuator network in series with the output and the signal source.



Carl Lodström, KQ6AX and SM6MOM

# Instrument amplifier noise (additional information for article published in issue 4/2006)

A minor issue that maybe should be added to "The Nobel Art of measuring small resistances (and voltages)" article published in issue 4/2006. It was about the AD524 as an Instrument with many uses. I did not mention anything about the input noise. It is low, but it is there! The more resistance of course the stronger the noise.

The measured noise, 3.7mV for example, with 10K $\Omega$  to ground on each input is then 3.7 $\mu$ V on the input as I used the 1000X gain selection for each measurement.

I took some pictures from the scope when I had one resistor from each input to ground. In no case did I use any filtering. At this gain the BW is ~25KHz. If I had selected a lower gain the noise would have got weaker and the BW increased. All pictures are taken with 1ms/sweep speed.

The pictures are:

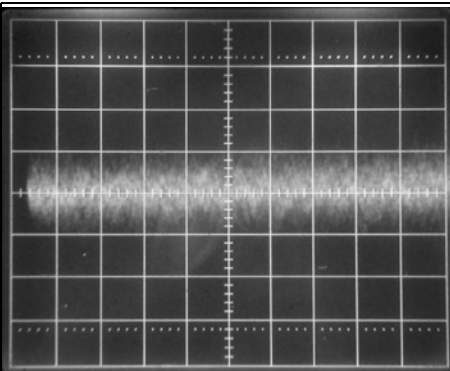
Rin (per input to ground)	Scope gain mV/div	Noise mV RMS
1M $\Omega$	100	36
100K $\Omega$	20	10.2
10K $\Omega$	5	3.7
1K $\Omega$	5	1.75
1K $\Omega$	2	1.75

If the inputs are left open it may seem that even lower noise can be achieved, but it is the amplifier that bottoms out from it's own input bias current ( $\pm 50$ nA

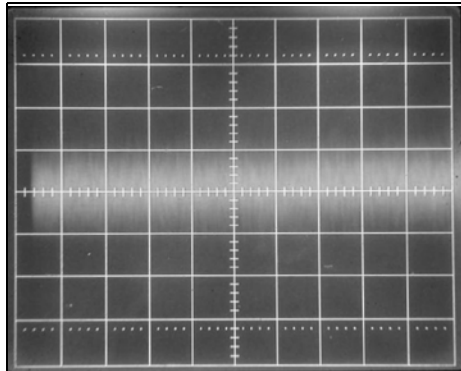
or less) so at least 100M $\Omega$  are needed to keep the inputs in the Common Mode Range.

This noise is by far weaker than the signal from an old dynamic microphone (2.5K $\Omega$ , built in transformer) for a reel-to-reel tape recorder. Even in the middle of the night I cannot cover the microphone and get it so quiet that I can see much of the amplifier noise on the microphone signal! There is at least a 20 - 30dB difference.

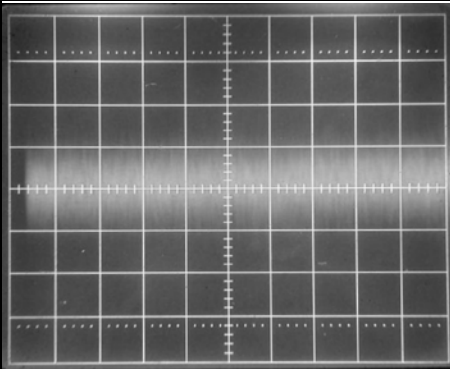
It was the DC resistance of the transformer secondary that I measured, it is an input transformer for a valve amplifier. I measured the impedance at 1KHz and it is 25K $\Omega$ . Using the AD524 instrument amplifier is "as good" (if not better) than a valve amplifier. The input resistance is over 1G $\Omega$ , so the transformer operates into an open circuit. I have looked at the signal one night with the microphone wrapped in a pair of jeans (does not help much) and at a time when the fridge was not running. There still was some 12mV output from the amplifier, meaning 12 $\mu$ V input, and the oscilloscope shows barely a mm wide band of noise from the amplifier on a cm or more peak to peak signal from noise I cannot hear! So the input noise on this amplifier is really something else! I have found, at higher sound levels, that the signal (from this particular old microphone, with transformer) is roughly 100 $\mu$ V at 40dB sound pressure. This corresponds to 1 $\mu$ V at 0dB, which is about where the amplifier



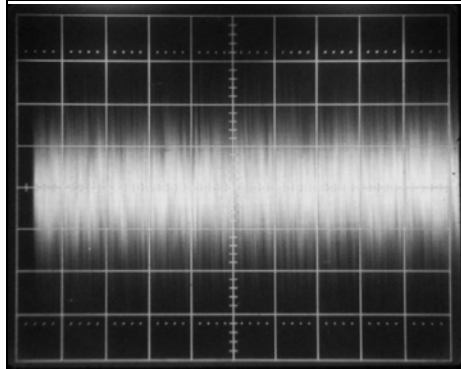
**Fig 1: Noise with 1MΩ input resistors**



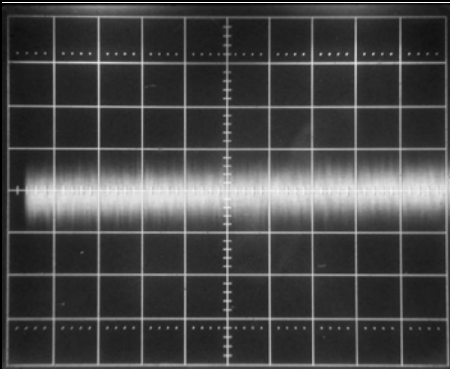
**Fig 2: Noise with 100KΩ input resistors**



**Fig 3: Noise with 10KΩ input resistors**



**Fig 4: Noise with 1KΩ input resistors**



**Fig 5: Noise with 1KΩ input resistors**

input noise is. So one could build a sound meter that can work down to about 0dB with a microphone like this and the AD524. That is pretty awesome. One will not be able to see any such low levels as one may have to be in an abandoned mine or such. Even an analogue meter would make more noise when it deflects! With the microphone ~2m distant I get good readings when I breath very quietly through my nose!

Ah, well. I figured afterwards that the noise of an amplifier is a pretty important parameter for an amplifier that is going to be used around the house for a lot of measurements.



# EADX 6m contest

**OBJECTIVE:** To increase the activity and experimentation on this magic band, field day operations as well as CW. In Spain there are still some areas where operation on this band is not allowed or highly restricted. So a list of all the stations taking part in this contest will be sent to the Spanish telecommunications ministry in order to avoid such restrictions in the excluded areas.

We would appreciate that club stations taking part in this contest send us a letter with their logs signed by the association explaining that they agree with our demand. The letter will be attached to the stations list explained above. Please show clear in the summary sheet if you do not want to be included in this list.

**ENTRANTS:** Licensed radio amateurs and SWL from all over the world.

## CATEGORIES:

- Single operator
- Multioperator or club stations
- Short Wave Listeners (SWL)

Only one transmitter can be used at the same time. Multi operator stations can use two transmitters (one for each mode: SSB and CW) provided the antennas of both transmitters are separated by less than 50 meters. All stations must be operated from the same location during the contest period.

**CONTEST PERIOD:** From Saturday 10:00 UTC 9<sup>th</sup> June 2007 till Sunday 16:00 UTC 10<sup>th</sup> June 2007. Single operators must have a minimum resting period of 6 continuous hours.

**CONTACTS:** One station can only be worked once on each mode (SSB and CW). All duplicate QSOs must be clearly stated as DUPE in the logs with no points claimed. Contacts made through repeaters or EME are not allowed.

**MODES OF OPERATION:** Contacts can be performed on SSB and/or CW.

**EXCHANGE:** RST plus full QTH locator (i.e. 599 JN11BH). Contact time must be logged but you do not need to exchange it during the QSO.

**SWL:** They must receive and write down in the log the following data of both stations: Callsign, QTH locator, date and time. It's not allowed to log more than 5 QSOs with the same station.

**POINTS:** One (1) point per kilometre and Two (2) points per kilometre for QSOs with special stations (these stations will be identified during the QSO).

**POINTS for SWL:** One (1) point for each QSO in SSB or CW.

**MULTIPLIERS:** One (1) multiplier for each different main grid square of each WW Locator achieved during the contest (i.e. JN11, IN62, etc.). One (1) multiplier for the first contact made with a station in each DXCC and WAE entities.

**LOGS:** The following formats are accepted:

**Paper logs:** Normalised IARU contest sheets or similar and a summary sheet showing essential information and leaving enough space for contest manager or contest committee notes. Multi operator stations must be clearly indicated. Final claimed score must be clearly shown in the upper part of the summary sheet.

**Digital logs:** Only in Cabrillo format.

We would be very grateful if you would send your logs in Cabrillo format in order to make the correction task easier.

Logs must be postmarked not later than 15<sup>th</sup> August 2007 and you can send them to:

**Paper logs:** EADX6M CONTEST  
PO Box 68  
E-08960 Sant Just Desvern  
Barcelona  
SPAIN

**Digital logs:** eadx6mcontest@gmail.com

## VERIFICATION and DISQUALIFICATION:

The contest organisation will take the responsibility of log verification. Minor errors can be penalised with loss of points. A QSO won't count if an obvious error in QTH locator or a time error of more than 10 minutes is found. Claiming points for a dupe contact will be penalised with 10 times its value.

The following cases will be deemed cause for disqualification of the contestant:

- Any violation of contest rules or the IARU band plan.
- Violation of amateur radio regulations in the country of the contestant.
- Use of DX Cluster for self announcements or using it as a personal log.
- Operators of multi operator stations participating with his own callsign at the same time.
- Giving false data to other contestants or the organisation.
- Helping out certain contestants at the expense of the rest of contestants.

**PRIZES:** 1st, 2nd and 3rd place in each category. Special prize for:

- Achieving most SSB contacts
- Achieving most CW contacts
- First place of each EA district
- First place of each DXCC or WAE entity
- Longest distance contact

Contestants assume total acceptance of the above rules. All decisions and/or actions of the organisation are official and final.



Gunthard Kraus, DG8GB

# Internet Treasure Trove

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## MuRata

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A very well known Far East company famous for coils, capacitors, filter etc. There are not only the product specifications, but also the collection at S-parameter files to the Download. The documentations with the measuring conditions are worth reading with the S-parameter files for separate frequency ranges.

Addresses:

a) homepage:

<http://www.murata-europe.com/>

b) S-parameter collection:

<http://www.murata.co.jp/sparameter/index.html>

c) Measurement of the S-parameters:

<http://www.murata.co.jp/sparameter/measure.html>

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## University of Guelph (Canada)

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One finds interesting things again and again at universities - one must search very thoroughly with them. On this homepage there is a tremendous collec-

tion at electronic hobby circuits. Just as interesting however there are links to a data sheet collections, SMD code, etc.

Address:

[http://www.uoguelph.ca/~antoon/circ/circuits.htm\\_\\_](http://www.uoguelph.ca/~antoon/circ/circuits.htm__)

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## Eesoft - Agilent

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This simulator is the “Rolls Royce” of RF circuit simulation - both from the price, as well as the extent and the possibilities. Nevertheless the search is worthwhile because the non Rolls Royce content. Here there are introductions or Application Notes for the simulation of different RF and microwave circuits. Also most can be used with little trouble for other cars (Pardon: Simulation programs). A really interesting site.

Address:

<http://eesof.tm.agilent.com/applications/>

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## Helix Antenna Design and Construction detail

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Who is interested in this type of antenna or would like to experiment. Here is a beautiful empirical report on this topic.

Address:

<http://www.jcoppens.com/ant/helix/index.en.php>

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## Raltron

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Some on-line computations among other things converting SSB phase noise to jitter are to be found with RALTRON. Also the application for quartz crystal jitter and noise are worth reading.

Address:

<http://www.raltron.com/cust/tools/default.asp>

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## An experimental 10Mbit/sec Microwave Data Link

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This project takes some time to appreciate, besides the empirical reports and/or results of measurement of the completed project, there are the planning documents, the initial considerations and the PCB layouts etc.

Unbelievably accurately and the last detail inside documents; even the views for RF security and the protection of systems are there.

Address:

<http://www.qsl.net/ke5fx/uwave.html>

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## Notice

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

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



## The UK Six Metre Group

### [www.uksmg.com](http://www.uksmg.com)

With over 700 members world-wide, the UK Six Metre Group is the world's largest organisation devoted to 50MHz. The ambition of the group, through the medium of its 56-page quarterly newsletter 'Six News' and through its web site [www.uksmg.com](http://www.uksmg.com), is to provide the best information available on all aspects of the band: including DX news and reports, beacon news, propagation & technical articles, six-metre equipment reviews, DXpedition news and technical articles.

Why not join the UKSMG and give us a try? For more information contact the secretary: Dave Toombs, G8FXM, 1 Chalgrove, Halifax Way, Welwyn Garden City AL7 2QJ, UK or visit the website.



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