

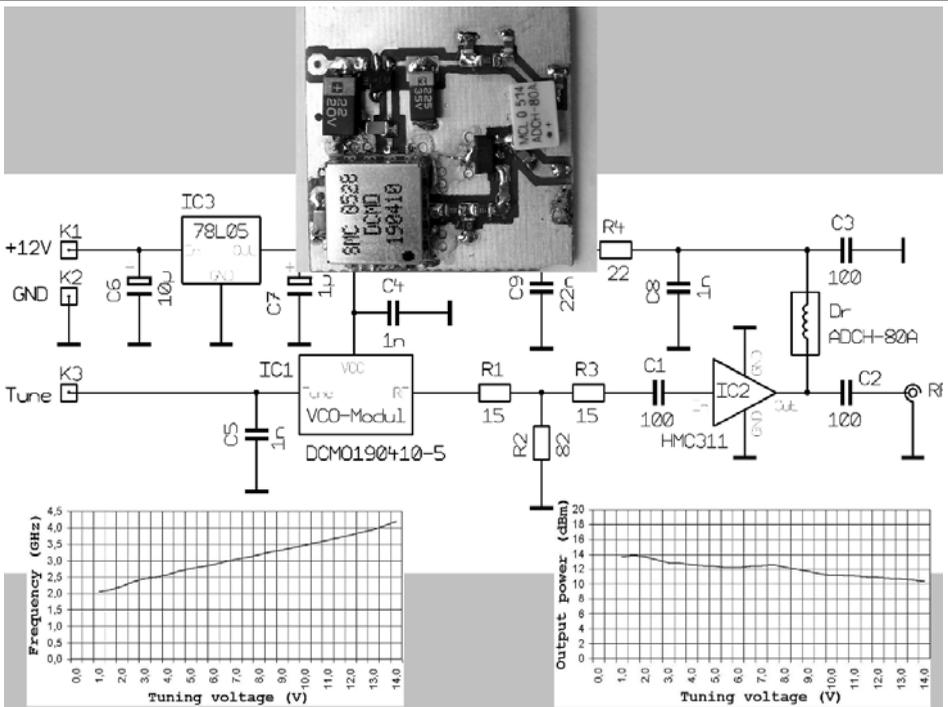


*A Publication for the  
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UHF and Microwaves*

# VHF COMMUNICATIONS

Volume No.38 . Winter . 2006-Q4 . £5.25



**VCOs as a replacement for YIG oscillators  
in the 2 to 4GHz range**

*Wolfgang Schneider, DJ8ES*

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*I hope that you have enjoyed the articles in the magazine this year. Don't forget to renew your subscription for 2007. The overall effect of the UK postal changes this year has meant that the magazine price will stay the same for 2007 but the additional price for air mail postage has risen in line with the increase in the postal rates. I now have a Paypal account (vhfcomms@aol.com) so you can use it to subscribe for 2007, if you use that method please make sure that your full postal address and callsign are part of the transaction.*

*I will be working on an update of The RSGB VHF/UHF Handbook over the winter, if you have any suggestions or interesting articles that would be suitable for the new book, please contact me. The new Handbook is planned to be available in Autumn 2007.*

*Merry Christmas and a Happy New Year. 73s - Andy*



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Wolfgang Schneider, DJ8ES

# VCOs as a replacement for YIG oscillators in the 2 to 4GHz range

YIG oscillators are popular because of their large frequency range. However the complex voltage and/or current supply is a large disadvantage. Also some types of YIG oscillators are in short supply. The solution can be to create a replacement for the YIG using modern, super wide band, oscillators for the GHz range. In the following article an appropriate circuit for the range 2 to 4GHz is described.

## 1.

### Introduction

YTOs (YIG Tuned Oscillators) are used as oscillator, particularly in wide band applications like Wobblers or Spectrum Analysers. Beside their wide band operation their linear frequency response to tuning voltage, or current, is important. The comparatively complex voltage / current supply of these oscillators is unfavourable.

Modern tuneable oscillators supplied by the industry offer wide band operation in the GHz range. Such a VCO component for the frequency range 2 to 4 GHz is the DCMO190410-5 from SYNERGY. With a simple amplifier module, not only the

wide frequency range of the YIG oscillator is obtained but the power output of nearly +16dBm (40mW), like the YIG oscillator, is achieved.

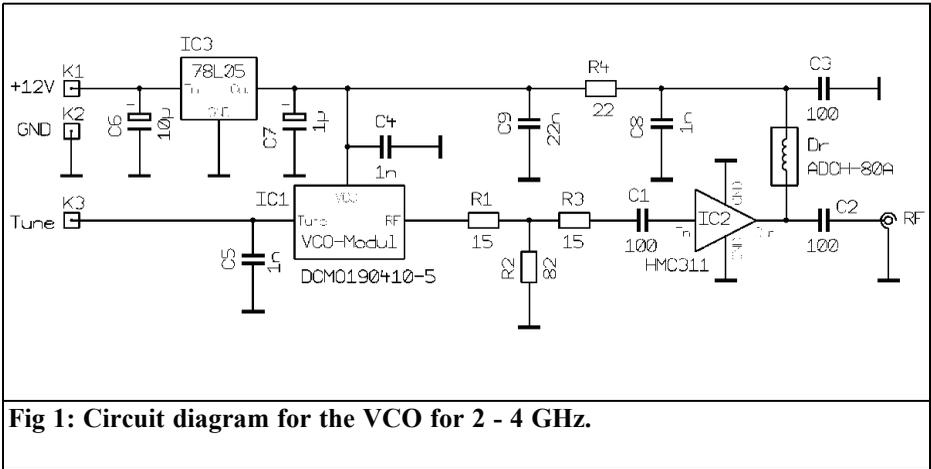
## 2.

### Description of circuit

The circuit for the oscillator module is very simple and shown in Fig 1. The voltage controlled oscillator DCMO 190410-5 (IC1) from SYNERGY is adjustable over a frequency range of 1.9GHz up to 4.1GHz. The oscillator is tuned with a control voltage between 0V and +15V, the transfer characteristic is shown in Fig 2.

The 3dB T section attenuator provides for a good adjustment over the entire frequency range. At the same time it adjusts the output level to the following amplifier stage.

The HMC311 integrated broadband amplifier supplies an output of approximately +13dBm (20mW). This amplifier has a more or less constant amplification over the relatively large frequency range up to 6GHz. The power output of the



**Fig 1: Circuit diagram for the VCO for 2 - 4 GHz.**

oscillator module over the entire tuning range is shown in Fig 3.

The complete circuit is operated from a 5V supply, the current required is approximately 100mA. A 78L05 (IC3) is used to stabilise the supply.

### 3. Construction

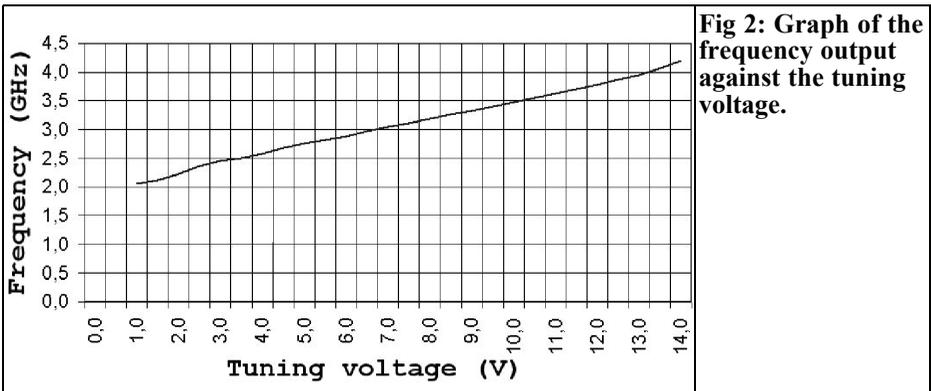
The circuit of the 2 to 4GHz oscillator is built on a 34mm x 34mm double sided PCB made from 0.51mm thick ROGERS

RO-4003C material.

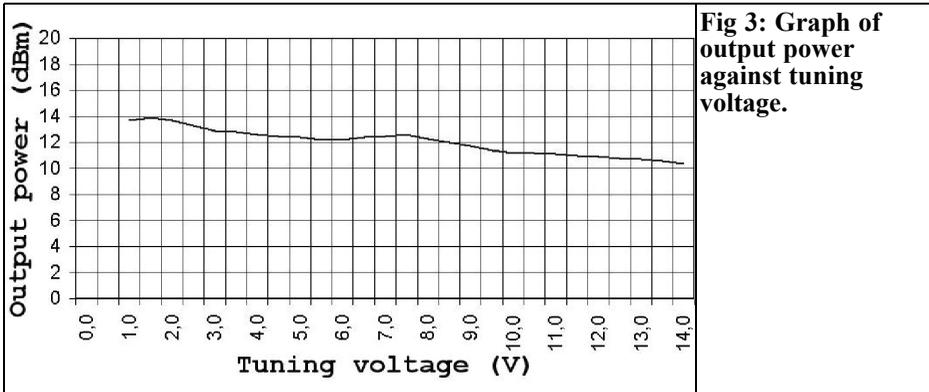
First the housing is prepared for the installation of the SMA socket and the two 1nF feed through capacitors.

Next the VCO module (IC1) is soldered on the track side of the PCB. All ground connections are to be soldered cleanly! Now the PCB is soldered into the standard tinplate housing (37mm x 37mm x 30mm).

The remaining components are fitted as required.



**Fig 2: Graph of the frequency output against the tuning voltage.**



**Fig 3: Graph of output power against tuning voltage.**

4.

R2 82Ω  
R4 22Ω

**Parts List**

**SMD capacitors 0805:**

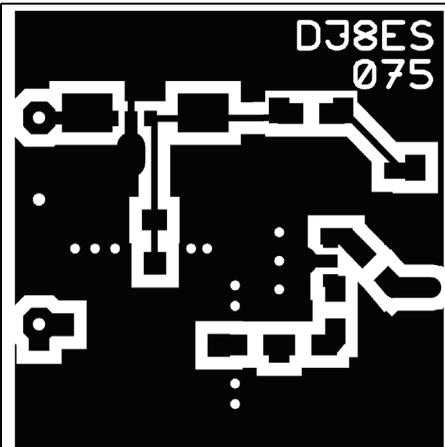
IC1 DCMO190410-5 VCO  
IC2 HMC311 amplifier IC  
IC3 78L05 voltage regulator  
Dr ADCH-80A rf choke  
C6 10μF/25V, tantalum  
C7 1μF/25V tantalum

C1, C2, C3 100pF  
C4, C5, C8 1nF  
C9 22nF

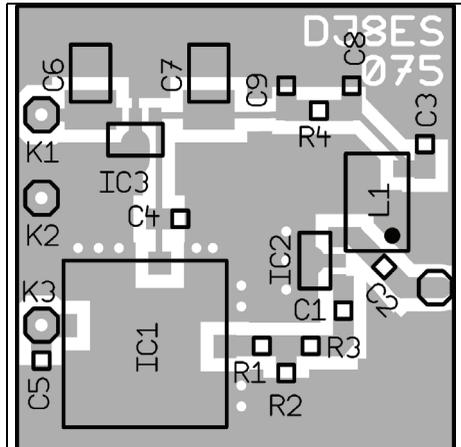
1 Tinplate housing 37 x 37 x 30mm  
1 SMA socket  
2 1nF solderable feed through capacitor  
1 solder point  
1 PCB DJ8ES-075, 34 x 34mm  
Rogers RO-4003C, 0.51mm

**SMD resistors 0805:**

R1, R3 15Ω



**Fig 4: PCB layout for VCO.**



**Fig 5: Component layout for VCO.**

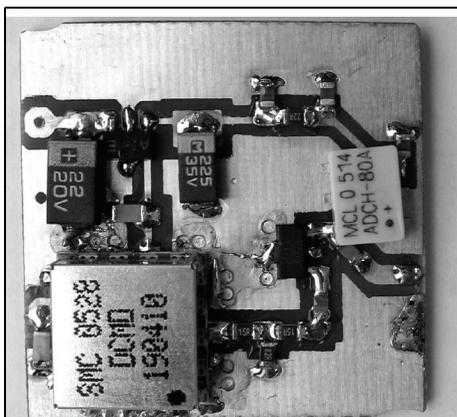


Fig 6 : Picture of the complete VCO.

## 5.

### References

[1] Wolfgang Schneider, DJ8ES: Frequency generator (Wobbler) to 4GHz, VHF Communications magazine 1/2002

[2] Synergy Microwave Corporation: Data sheet DCMO190410-5, [www.synergymwave.com](http://www.synergymwave.com)



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Bernd Kaa, DG4RBF

# Microwave milliwattmeter to 18GHz, WM 5000 PRO (-55dBm to +20dBm)

A good milliwattmeter is an important measuring instruments for an active radio amateur. To build such a measuring instrument yourself is a challenge, particularly if the following conditions are to be fulfilled: universal, high reading accuracy, comfortable controls, logarithmic display in dBm and simultaneous display in mW or volts, precise and reproducible measured values. In the following article a high quality milliwattmeter for home construction is described, it compares well with commercial devices. The use of modern components and a microcontroller makes hardware and construction costs relatively small.

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## 1.

### Specification

---

- Display: Logarithmic in dBm and simultaneous display in mW,  $V_{eff}$ ,  $V_{ss}$ ,  $V_{mf}$  or  $dB\mu V$
- Bar graph display:
- Simultaneous rough and fine display
- Centre display for simple alignment on maximum

- Switchable 6 stage amplifier
- Data for up to three detectors accessible
- Correction data for the frequency response the detector (21 storage locations per detector)
- Data for three selectable external attenuators (resolution: 0.01dB)

---

## 2.

### Technical data

---

- Frequency range: 1MHz to 18GHz (depending upon detector used)
- Range: -55dBm to +17dBm (+20dBm)
- Accuracy: approximately 0.3dB (depending upon detector)
- Resolution: 0.01dB over -40dBm and 0.1dB over -50dBm
- Dynamic range: 75dB



**Fig 1: Picture of the completed microwave milliwattmeter.**

### 3.

## Design

Industrial power measuring instruments usually use diode or thermal detector heads. A diode detector was chosen for this project because they have a large dynamic range of approximately 70dBs and a basic sensitivity of at least -50dB. There are many good diode detectors available for up to 18GHz from well-known manufacturers at reasonable prices. To build a good milliwattmeter an industrially manufactured detector is essential. These detectors usually have a negative output voltage; that is why the meter presented here has a negative diode input. All diode detectors have a problem: their output voltage is not linear with power, it follows the diode characteristics. If a micro controller is used, this problem can be solved perfectly.

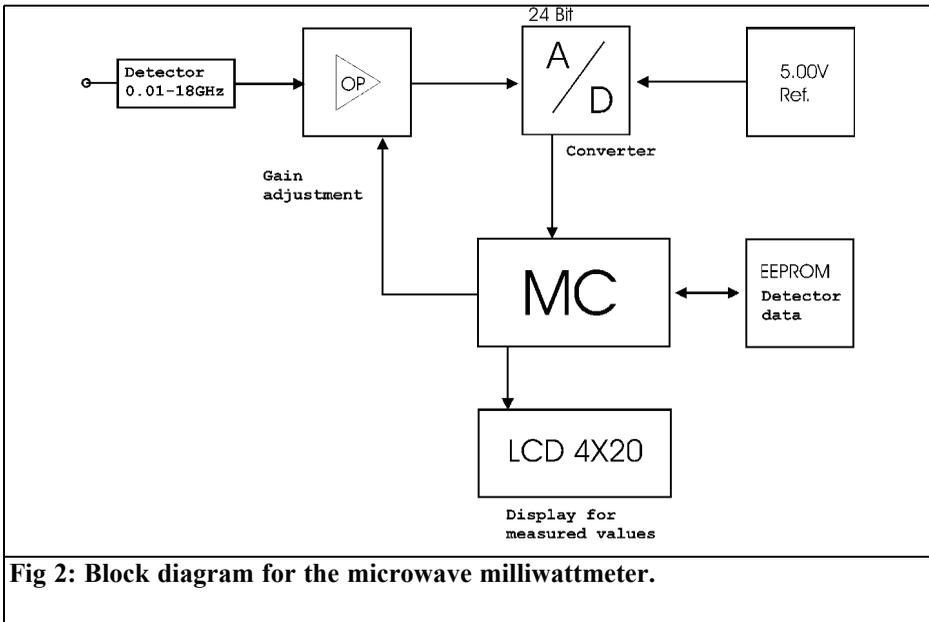
### 4.

## Description

The block diagram (fig. 2) shows the main sections of the milliwattmeter.

The large Dynamic range of approximately 75dBs is converted by a 24 bit A/D converter from Linear Technology (LTC2400), it uses delta-sigma technology to provide single cycle settling time. The converter is in an 8 pin SMD package and uses very few external components. It incorporates a 50Hz digital notch filter giving 110dB suppression of 50Hz noise and the A/D conversion contributes 0.3 ppm noise. This is a great advantage with the small input voltages involved and small measuring levels (-55 to -40dBm). Hence the measurements can be made with good resolution.

The resolution at lower levels is in-



**Fig 2: Block diagram for the microwave milliwattmeter.**

creased by using a preamplifier with switched gain. The minimum sensitivity is  $-55\text{dBm}$  and the maximum value is between  $+13\text{dBm}$  and  $+20\text{dBm}$  depending upon the detector used. The main display of the measured value is in  $\text{dBm}$  because this is most useful, particularly at low levels. The resolution is  $0.01\text{dB}$  in the range of  $+20\text{dBm}$  to  $-40\text{dBm}$ . From  $-50\text{dBm}$  the resolution is  $0.1\text{dB}$  measured in  $1\text{dB}$  steps. An additional display is possible in a second unit selected from:  $\text{mW}$ ,  $V_{\text{eff}}$ ,  $V_{\text{ss}}$ ,  $V_{\text{mf}}$  and  $\text{dB}\mu\text{V}$ . This can be selected at any time using the "Mod" key.

A measuring rate of approximately 6 measurements per second is used. In the lower measurement range of  $-55\text{dBm}$  to  $-50\text{dBm}$  an average of three sequential measurements is taken and displayed. The large  $4 \times 20$  display shows the two measured values plus a two line bar graph display. This allows level changes to be read in a similar way to using an instrument with pointer meter. Actually this is better than a pointer instrument, since there are always two bars visible, a

rough and a fine display. The bars are programmed so that the first bar consists of 75 single lines representing a change of  $1\text{dB}$  and the second bar shows  $1\text{dB}$  by a single line. This gives a resolution of  $0.0133\text{dB}$  per bar ( $1\text{dB}/75$ ).

There is also a centre display that is very helpful for alignment work, it shows the actual value as well as the adjustment of maximum or minimum values. The sensitivity of the centre display is adjustable in 6 steps, so that rough changes as well as very small changes ( $< 0.01\text{dB}$ ) can be shown.

Using the calibration function, the characteristics of up to three detectors can be taken and stored in the EEPROM. Additionally 21 frequency correction values (in the range 1 to  $20\text{GHz}$ ) can be stored for each detector and automatically included into the measurement. Normally these correction values are meant for frequencies from 1 to  $20\text{GHz}$  but they can be used just as well for other frequencies or other correction tasks. So for example  $\text{GHz}$  enthusiasts could use their



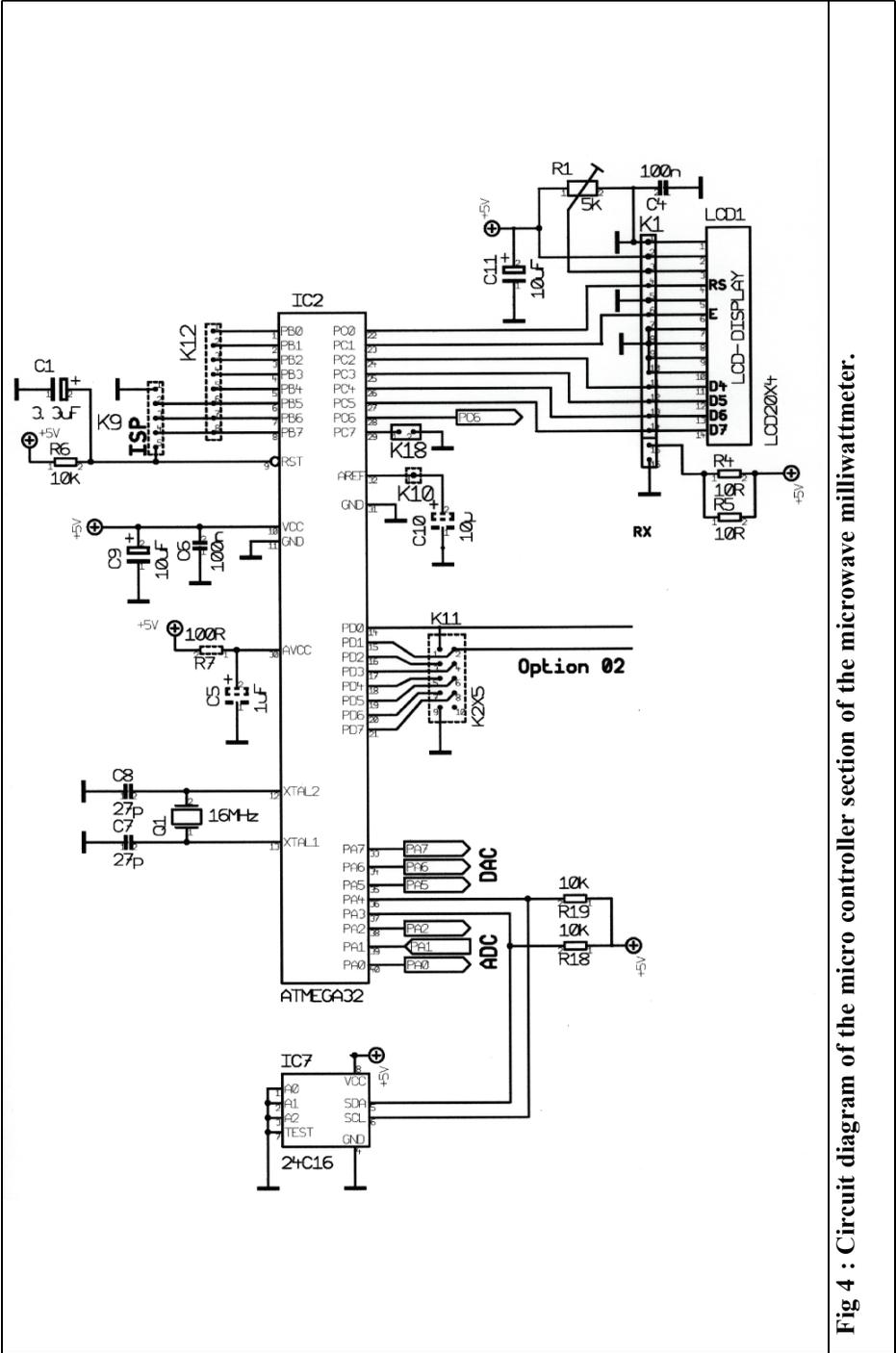
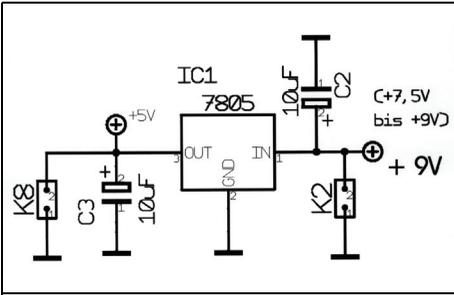


Fig 4 : Circuit diagram of the micro controller section of the microwave milliwattmeter.



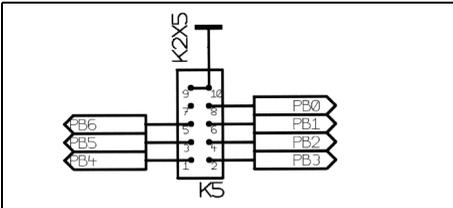
**Fig 5: Power supply for microwave milliwattmeter.**

“special detector for 76GHz” with the necessary correction values to give a useful display.

Three external attenuator values, programmable in 0.01dB steps, can be selected, these are then included into the measurement. Thus it is simple to extend the measuring range upward without having to convert the measured values. For example if the attenuation before the detector is set to 20dB then the measurement range will be +17dBm to +37dBm (5 Watts).

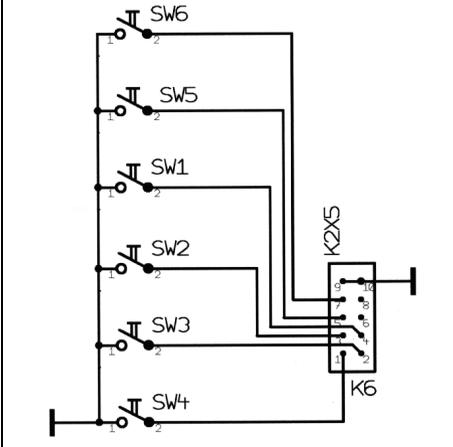
**5. Circuit description**

The circuit diagram is shown in Figs 3 - 6, it was designed to operate with negative output diode detectors which are the



**To micro-controller**

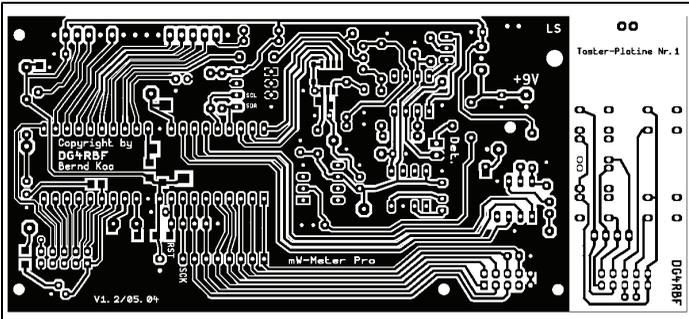
**Switch PCB**



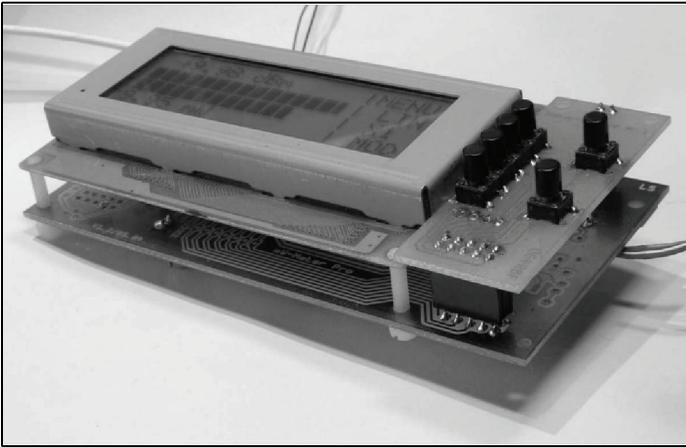
**Fig 6: Front panel switches for microwave milliwattmeter.**

most common commercial detectors.

The input feeds an ultra precision amplifier (OP177) which is ideally suited to this measuring task with it’s very low offset voltage of only 10μV and the very



**Fig 7 PCB layout for microwave milliwattmeter.**



**Fig 8: Picture of the assembled microwave milliwattmeter PCBs.**

small offset drift of  $0.1\mu\text{V}/^\circ\text{C}$ . It converts the negative output voltage of the detector into a positive voltage using a small amplification of approximately 2 which is fed to the A/D converter. The maximum input voltage to the LTC2400 A/D converter is 3.2V.

In order to increase the resolution in the lower measuring range, the amplification is changed by a factor of 10 using a 2N7000 FET. The change over is made by the micro controller at -20dBm. The five resistors, R9 to R13 that set the amplification of the amplifier should be precision resistors (0.1%). The balance potentiometer is used to set the zero point of the amplifier. The negative supply voltage for the amplifier is generated using an ICL7660 charge pump. A protection diode at the output of the amplifier prevents more than  $-0.3\text{V}$  being applied to the input of the A/D converter. The signal is fed to the A/D converter via an RC low pass filter. The A/D converter does not need any external components. Communication with the micro controller uses three lines (CS/, Dout, Clk). Since this A/D converter can measure extremely precise small voltages it is necessary to keep the reference voltage as constant as possible. An LT1021 CCN8-5 five volt regulator is used which has a temperature coefficient of typically  $3\text{ppm}/^\circ\text{C}$ . A 10 turn trimmer is used to

set the reference voltage to exactly 3.20V. The converter has a resolution of 24 bits, the maximum value is 16777215 (24 bits) this gives a resolution of  $0.0001907\text{mV}$  with a 3.2V reference.

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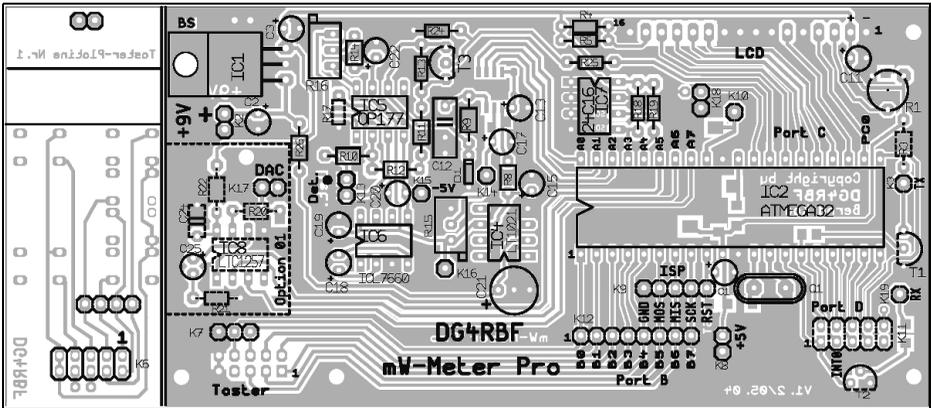
## 6.

### Micro controller and display

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To keep the cost of the unit as low as possible the micro controller and all of the components for the meter are mounted on a single printed circuit board (Fig 7). An ATmega 32 micro controller is used, this is a flash controller from ATMEL in a 40 pin DIP housing. It offers the necessary storage space for the extensive program and the complex arithmetic problems, it does not require any external EPROM or RAM. The display is attached directly to the main PCB (Fig 8), so no additional wiring is needed which improves the reliability. The trimmer (R1) sets the contrast of the display.

A plug connector beside the LCD on the main PCB is used to connect the small PCB for the controls. The different functions of the keys are indicated at the appropriate position in the display. This



**Fig 9: Component layout for the microwave milliwattmeter showing the component side of the PCB.**

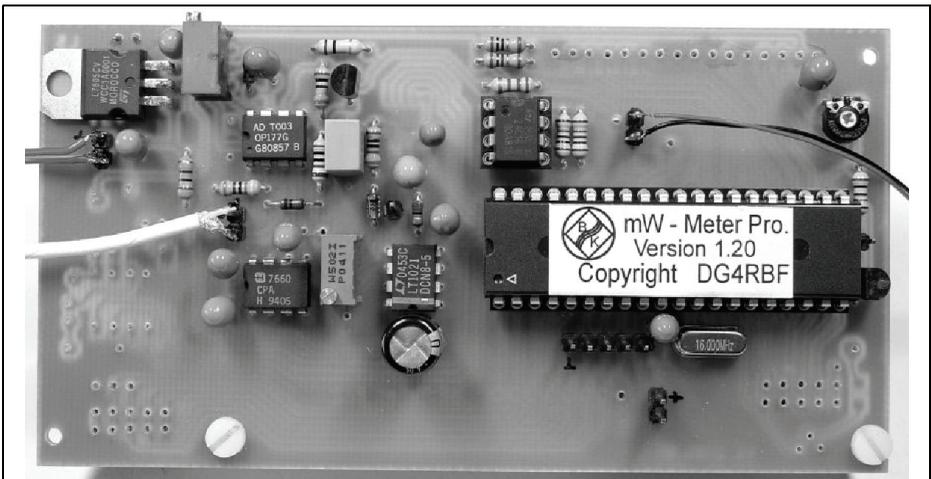
makes operation simple because the appropriate task of the controls is always shown.

The finished milliwattmeter is a complete unit that can be installed with the LCD and keys directly on the back of the front plate of a housing. An external EEPROM, which is addressed with the I<sup>2</sup>C-Bus, holds the calibration data of the detectors. It will retain the data without a battery, even if the unit is switched off.

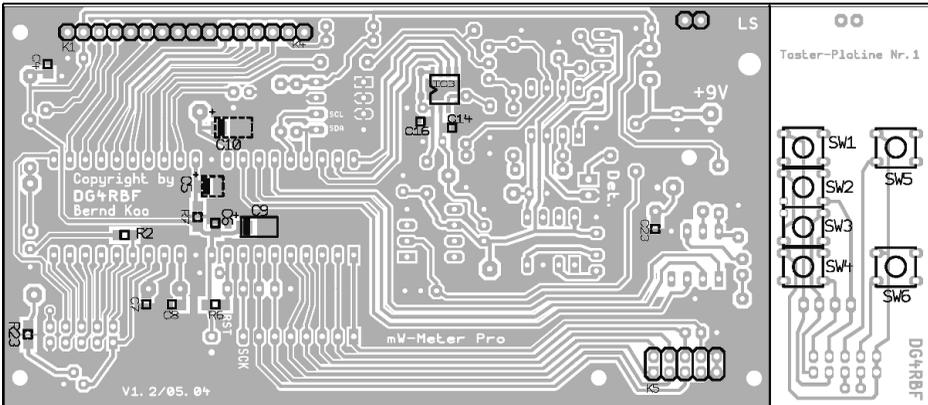
7.

**Construction**

The display has a 16 pin connector, a 16 pin socket strip is soldered to the track side of the main PCB so that the display can be plugged onto the main PCB. Since the socket strip is on the solder side of the single sided PCB, it must be soldered with approximately 1.5mm clearance for



**Fig 10: Picture of the components mounted on the component side of the main PCB.**



**Fig 11: Component layout for the microwave milliwattmeter showing the track side of the PCB.**

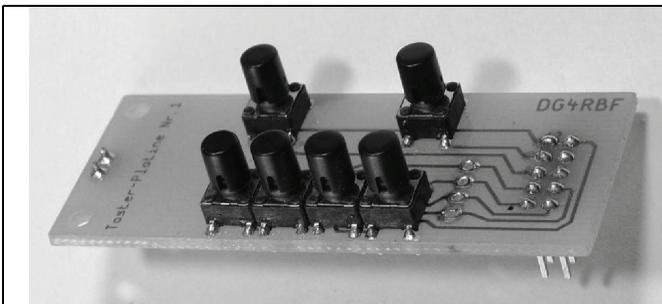
the other solder joints on the PCB. The micro controller and the EEPROM should be fitted in precision IC sockets.

A thin piece of coax cable should be used to connect the diode detector to the input of the milliwattmeter. A small plug top power supply can be used to power to unit. Battery operation is simple because only a single supply voltage is needed.

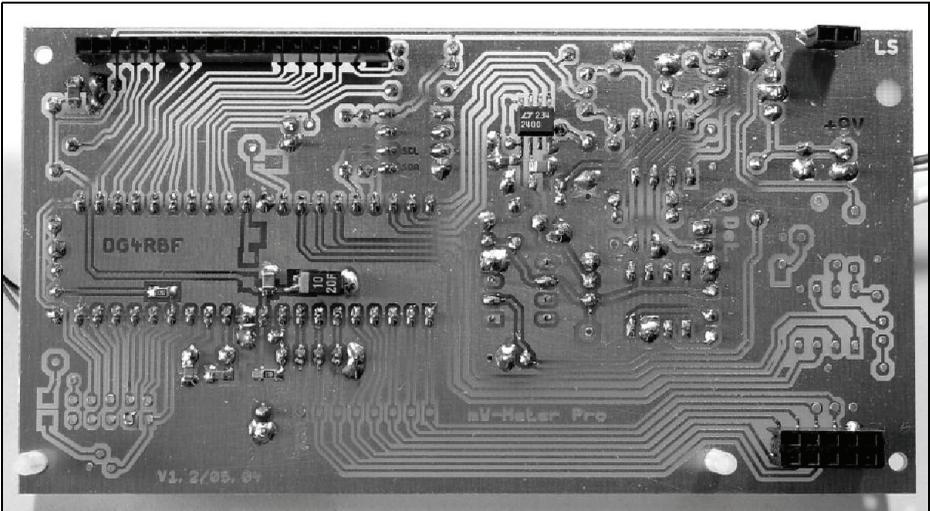
The case used for the prototype was made by BOPLA and purchase from CONRAD (#523348). Photo coated front panels are available for this case so professional looking panels can be made easily. The layout of the front panel is available as pdf file. If demand is sufficient, finished milled and engraved front panels will be available.

## 8. Alignment

The alignment is very simple and is limited to setting up of the reference voltage for the transducer and the zero offset adjustment of IC3. The diode detector must be attached for these adjustments. Use a digital voltmeter to adjust the reference voltage to exactly 3.20V measured at test point K14. To null the input offset, switch the equipment on and leave it for 10 to 15 minutes to “warm up”. After this set the zero value with trimmer R16 to approximately 0 which is indicated on the display by a value of <-55dBm (Adjustment tolerance approximately ±50). The zero value will vary by approximately 2 to 10 counts.



**Fig 12: Picture of completed front panel switch PCB.**



**Fig 13: Picture of the components mounted on the track side of the main PCB.**

**9.**

**Using the EEPROM to store detector data**

---

An empty 24C16 EEPROM can be used with the calibration function to store the characteristics of up to 3 detectors. A pre-programmed EEPROM is available from DG4RBF that contains the characteristics of two detectors: HP 33330B and GR-50. There is still space in this EEPROM

for another self-calibrated detector.

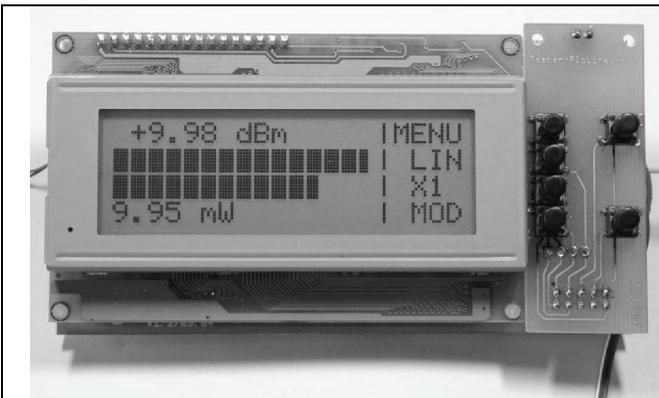
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**10.**

**Alarm function for maximum power**

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If the maximum permissible power level is exceeded (+13/+17 or +20dBm), then



**Fig 14: Picture of the completed PCB with the display in operation.**

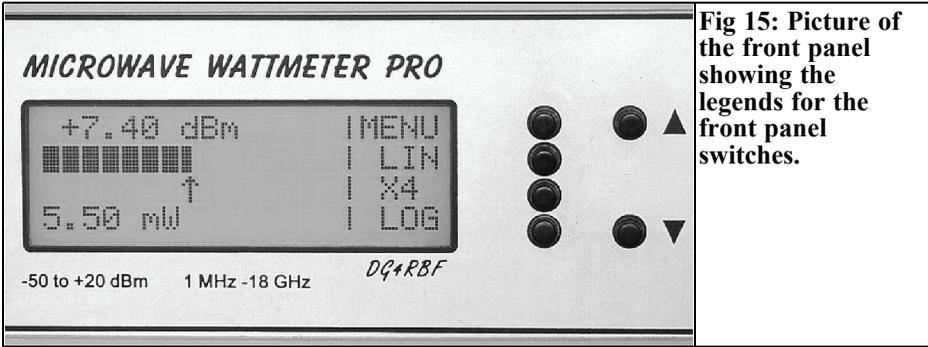


Fig 15: Picture of the front panel showing the legends for the front panel switches.

“max. power” alert is shown on the display. If a piezo sounder (transducer with no internal electronics) is connected to K18 this will sound when maximum power is reached.

This puts an arrow in the centre of the display independent of the actual power being measured. Changes in power caused by alignment of the circuit under test are immediately shown by the movement of the arrow.

## 11.

### Software and operation

*Zero point alignment (ZERO)* in addition to the hardware zero point adjustment using trimmer R16, this gives another null by software. This is used before a precise measurement by simply pressing the [Zero] key.

*Amplification factor (X 1-6)* with the third key [X1] can be used to change the amplification factor for this centre display from 1 to 6. Each time this is changed the [LIN] key must be pressed again in order to update the display with the new factor.

*Linear centre display (LIN)* with levels > -40dBm the legend key [LIN] is displayed beside the second key. Pressing the key put the display in linear centre mode.

*Mode (MOD)* the [MOD] function key is used to select the alternative power indication on the fourth line of the display. Pressing this key selects one of the following 6 display units in turn: mW, Veff, Vss, Vemf, dBµV, OFF (no display). The selected display unit is stored in the EEPROM so that the same unit will be displayed next time the unit is switched on.

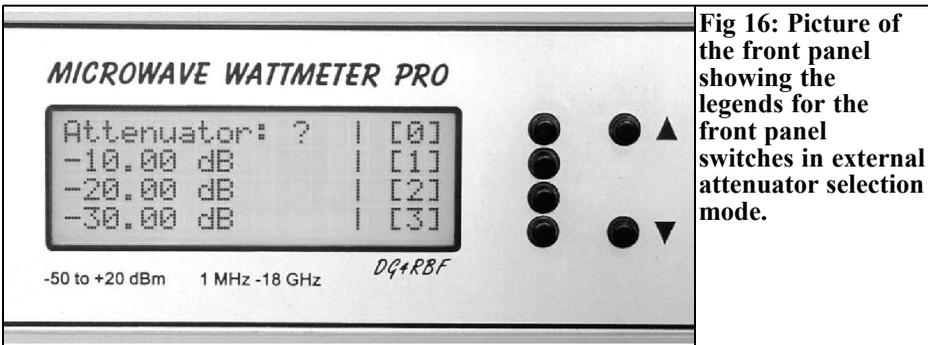
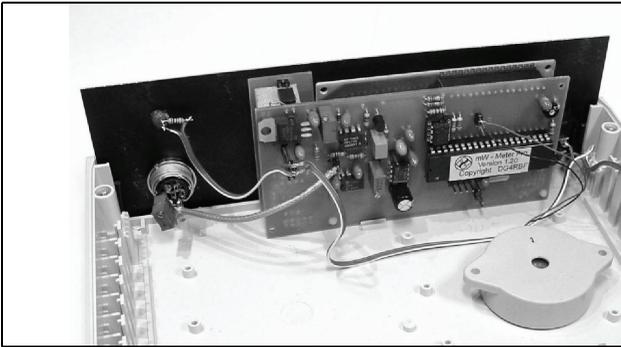


Fig 16: Picture of the front panel showing the legends for the front panel switches in external attenuator selection mode.



**Fig 17: Picture of the PCB fitted into the cas showing the piezo sounder.**

*Frequency correction (/) arrow keys* with the two arrow keys are used to select the 21 frequency correction values. The selected correction is then used to adjust the displayed value. During operation of the arrow keys the correction values are briefly shown on the lowest line of the display.

Menu 1 (MENU) pressing the [MENU] key enters Menu 1 the external attenuator menu

Attenuator > [SET EDIT NEXT BACK]

*Attenuator select* [SET] with this function three attenuators can be selected (1-3). Select (0) if no attenuator is to be included in the measurement display.

*Attenuator edit* [EDIT] this is used to set the dB value of the attenuator to a hundredth of a dB. The input the keys [ + ], [ - ], [ NO ] and [ SET ] are available.

Menu 2 from menu 1 pressing the [NEXT] enters menu 2 the detector select menu

Detector > [DET CAL FRQ BACK]

*Detector select* [DET] with this function one of three detectors can be selected. The detectors are designated 0-2. The characteristic that is stored in the EEPROM corresponding to the detector is then used.

*Calibrate detector* [CAL] this is used to calibrate detector 0, 1 or 2 by momentarily pressing [CAL]. Note: the existing data for the detectors are not write protected. Pay attention to which detector you selected for calibrating. You must confirm calibration of the selected detector by pressing [YES]. Then select the maximum permissible power for the detector: 13, 17 or 20dBm. Then the semi-automatic calibration program begins. Here the power is given that you apply to the detector and confirm with [SET]. The program starts at -55dBm and requires input at 1dB increments up to the maximum power indicated. Note that at -20dBm the amplification change-over is made and the program continues with -21dBm. The calibration value -20dBm is requested twice, once without and once with amplification. In order to do this calibration use a signal generator with an output adjustable from -55dBm to +20dBm (+17 or +13dBm depending upon detector) in 1db steps or a constant signal source with an output of +20dBm (+17 or +13dBm depending upon detector) and an attenuator switched in 1dB steps. Select the frequency between 10MHz and 100MHz, where detector and attenuator accuracy will be good and the affects of connecting plugs is still small. The calibration can be extended upwards at any time (+17, +20dBm), without having to begin again at -55dBm. Likewise it is possible to set or jump over individual values.



**Fig 18: Power detector for up to approximately 3 - 5GHz**

The values in the EEPROM are changed only if the key [SET] is pressed.

*Frequency correction [FRQ])* this menu option is used to enter correction values for the selected detector. 21 storage locations are used which are meant for 0 to 20GHz. For each storage location a correction value can be stored in dB (0.01dB steps). For input the keys [+], [-], [NO] and [SET] are available. During adjustment with the plus and Minus keys, the corrected power is displayed, so that the precise measurement of a microwave signal generator can be made. A separate memory is available for the 3 possible detectors (3 X 21 storage locations).

*RESET* if incorrect data is found in the memory of the micro controller a RESET function is displayed. A reset can be performed by pressing the [RESET] key



**Fig 20: Power detector type 33330B from HP for up to 18GHz and is the preferred detector for the milliwatt-meter**



**Fig 19: Power detector from Pacific Measurements for 1MHz to 18.5GHz**

(top left) when the unit is switched on. The calibration data for the detectors is not deleted.

## 12.

### Parts List

R1	5K $\Omega$
R4, R5	10 $\Omega$
R6, R11, R18, R19	10K $\Omega$
R8	22 $\Omega$
R9	1K $\Omega$
R10	47K $\Omega$
R12, R16, (R17)	100K $\Omega$
R13	1M $\Omega$
R14	47K $\Omega$
R15	5K $\Omega$
R25, R26	100 $\Omega$
R25	2.2K $\Omega$



**Fig 21: Power detector from NARDA for 10MHz to 18GHz.**



C1	3.3μF	IC5	OP177
C2, C3, C9,		IC6	ICL7660
C11, C13, C15,		IC7	24C16 (see text)
C18, - C20, C22	10μF		
C4, C6, C14, C21	100nF	LCD1	LCD 20x4
C7, C8	27pF	Q1	16MHz
C12	1μF	SW1 - SW6	Push button switch
C16	47nF		Conrad 705349 +
C17	22μF		705115
		K1	K1x16
D1	BAT42	K2, K4, K8, K13	
T3	2N7000	K18	K1x2
IC1	7805	K5, K6	K2x5
IC2	ATMEGA32	K7	K1x3
	(incl. DG4RBF	K14 - K16	K1x1
	software)	Case	BOPLA
IC3	LTC2400		
IC4	LT1021 DCN8-5		

## Amateur Television Quarterly

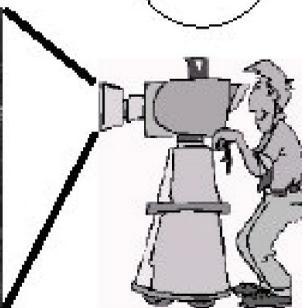
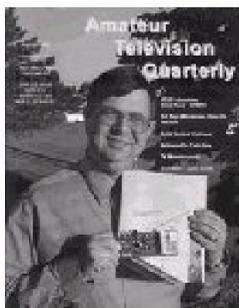
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Carl G. Lodström, KQ6AX and SM6MOM

# The Noble Art of measuring small resistances (and voltages)

Parts of this were published in 1986, in an article about coaxial relays [1], but I think it is time to air it again. Partially because the editor at the time proposed a change, without asking, erroneously suggesting the deletion of the separate voltage and current leads through the DIN connector. He must not have known anything about Kelvin connections, which was the whole point of the design. Another reason for a new article is the excellent Instrumentation Amplifier (IA), AD524, even 20 years later may not be fully appreciated by the readers of VHF Communications Magazine. I have used in at least three projects [3], [4] and [5], published in VHF Communications Magazine. With a few additions to the circuit in the earlier article, the amplifier itself becomes a really versatile instrument and it alone is reason grab the soldering iron!

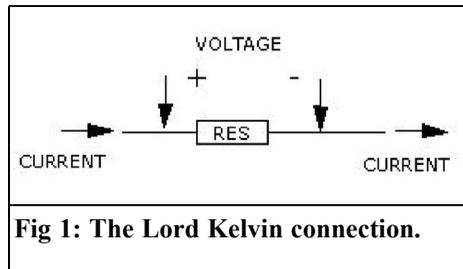
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## 1.

### Much wisdom from Lord Kelvin already 150 years ago!

---

To measure small resistances the connections to it must of course have even smaller resistances. When we are talking



about resistors on the order of fractions of an ohm, to begin with, this may not be possible. Since resistance is the voltage drop (V) caused by a known current (I), the problem can be reduced to running a known current through any resistor, and measuring the resulting voltage drop, Fig 1. As we all know  $R=V/I$ . To be able to do this, the resistor must have separate terminals for current, positioned outside the voltage clamps. Shunts, very low resistances, and precision resistors are built just like this, see Fig 2. This is the Kelvin connection. We can see the heavier bolts for current and the smaller screws for the voltage terminals. The larger, cylindrical, resistor is a laboratory standard (1k $\Omega$ ) by Hewlett Packard. There it is supposed to be 1000.000... ohm between the innermost (voltage) clamps and of course a little more between the outermost pair, the current clamps.

With a 4 pin probe, where the two



**Fig 2: Precision resistors and shunts.**

outermost supplies a current and the two innermost goes to the amplifier inputs (measuring the voltage drop), the resistance in almost any subject can be measured, even if very low. So now it is not so important with very good connections. Try measuring a  $1\text{m}\Omega$  resistor the regular way, even with quite good connections!

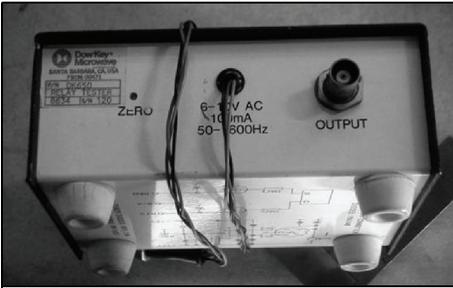
**2.**

**Coax relay contact measurement**

At the time I wrote [1] I was working for a manufacturer of coaxial relays. Some customers reported intermittent performance when used with small signals, such



**Fig 3: Front panel of the original relay tester.**



**Fig 4: Rear panel of the original relay tester.**

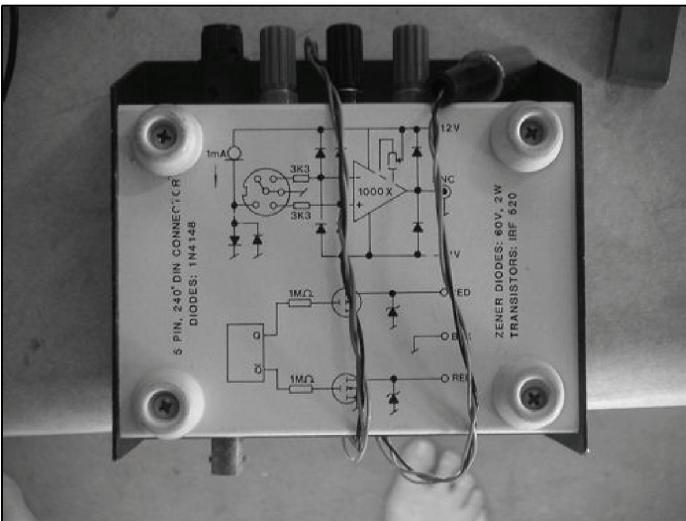
as receiver input signals. I figured out that it was caused by plastic dust, shaved off from the push rods, after many cycles of operation. Then I found that there were also cosmetics from the girls assembling the relays. Powder and silicone grease got onto the contact blades! The relays were “tested” by switching 6V / 0.3A dial lamps. If the lamp shone there was contact! Of course, it always shone. I realized that the contacts needed to be tested with a very weak current/voltage, one that could not burn through the problem. After some experimentation I came up with the device presented here, Figs 3, 4 and 5.

With this apparatus the effects of con-

tamination were clear to see. A  $\mu\text{V}$  signal to a receiver input will not burn through and “clean” dirty contact surfaces! The necessary degree of cleanliness was previously unknown (at least to this company) so the shipping quotas became seriously disturbed. The girls had to wear small rubber caps on their fingertips, but it did not help much as they touched their faces anyway... The Manufacturing manager was more interested in shipping the required volume every month than in the quality of the goods. I became a problem and got laid off. I do not know if they keep using the meter and if they got their contamination problems under control.

The readers of VHF Communications Magazine are more serious about what they do so this may be helpful reading, not only in regards to coaxial relays.

I will not get into any detail about this schematic diagram, as we will build ours a little different, not so many protection diodes for one, and it will be described well enough. But I think it is a good practice to have something like Fig 5 so the customers get an idea of what they are working with and what goes on inside the box. Fig 6 and 7 show typical test leads.



**Fig 5: Bottom panel of the original relay tester.**



**Fig 6:**The Kelvin connected probe for the relay tester.

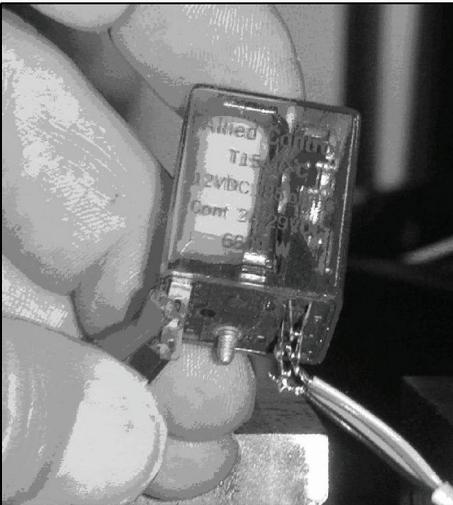


**Fig 7:** A Kelvin connected pair of test leads from a prime manufacturer, ESI.

With the instrument connected to an oscilloscope, we can cycle the relay from an oscillator. In the tester I included a very simple one, using two 4538 CMOS timers, one package, connected in a ring, and two MOSFET transistors that can handle just about any relay coil, but we will not repeat that here. The power supply was external. Instead I suggest using the gate pulse from an oscilloscope as the FET driver. I consider this a better method. On a dual time base scope, using the B gate pulse and setting the scope, with a free running sweep, to “A illumi-

nated by B” allows for a total control of how, when and where on the screen the relay is excited.

With this set up running it is interesting to study the influence such parameters as supply voltage, gravity, external magnetic fields, damping diodes on the coil and temperature will have on pull in and drop out times as well as ultimately on the contact resistances. Note that we can see both sides of a SPDT (Single Pole Double Throw) relay! The outer connectors can be joined as one. Both sides of the relay are measured while the scope makes one sweep.



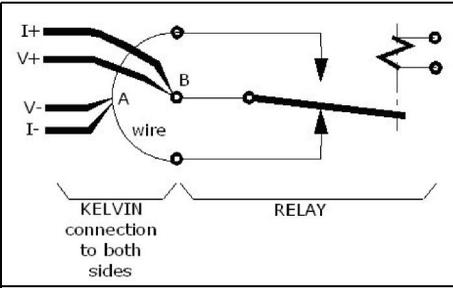
**Fig 8:** A relay under test.

### 3.

#### Measuring resistance.

In Fig. 8 we can see how the four leads (2 current and 2 voltage) are soldered directly to the contact pins of a small relay. I used a regular 4 conductor, unshielded, telephone cable for this.

The two wires from the current source, and the two voltage wires to the IA inputs, are connected a shown in Fig 9.



**Fig 9: Connection of both sides of a relay at the same time.**

Half the wire connecting either side of the relay is now included in the measured resistances. What is measured is the resistance between A and B in the picture. Here it does not matter if I or V are on the inside or outside each other.

At the beginning and end of the trace in Fig 10 we see the closure of one side. (The trace coming up and going horizontally.) Both closures are as nice and solid as one can wish for. The trace comes up and levels off. In the middle is a closure of the other side of the same relay and contact group. It has a horrible bounce before settling! A contact like this will soon be burnt up if one is hot-switching it. Hot switching is to switch with power

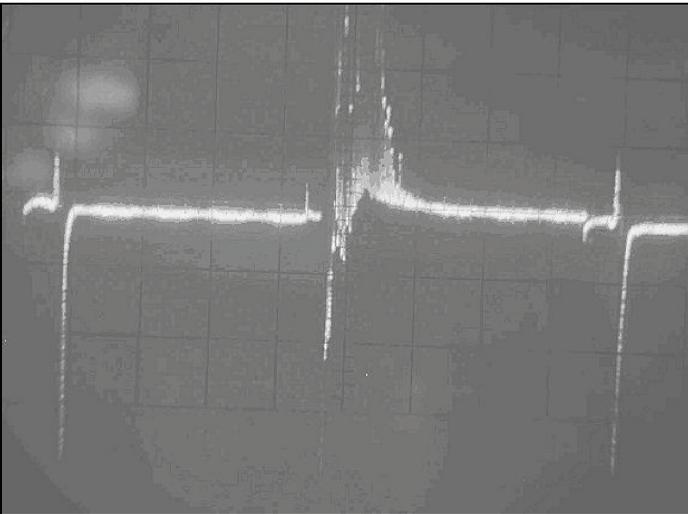
on. It may be as little as 10W at GHz frequencies.

By simply connecting the four wires to both sides of one SPDT switch we can see it all in one picture! In this case we can also see that the bouncing side has a slightly lower resistance (higher on the screen) than the good side. And we talk a total resistance of a few mΩ.

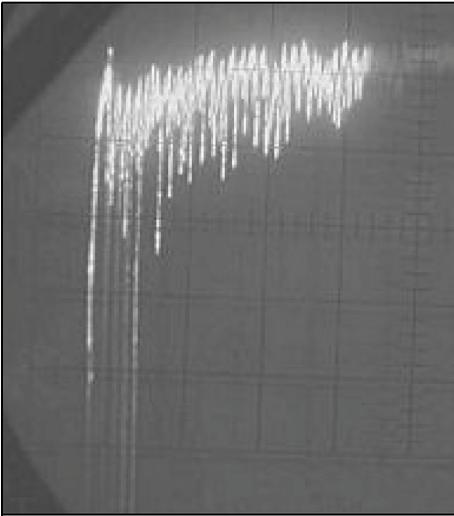
Temporarily removing one current pin, or turning down the current to a very low value, allows one to establish a base line.

The display disappears outside the screen for a moment as the relay presents an open circuit (voltage limitation  $0.25V \times 1000 = 250V!$ ) while switching over to the other side, and it re-appears when that side closes. Things to monitor are the resistance, when the trace has settled, and the stability of these lines. If they are “uneasy”, this indicates dirt in the switching cavity, or on the contacts.

Fig. 11 shows a detail of the bounce in this relay. This image was at times still, even though the relay cycled. Of course, it cannot be dirt that gives an identical erratic connection at every cycle! Here it is the contact blade that vibrates, like one



**Fig 10: Monitoring both sides of a switched group in a relay that is cycling at several periods per second.**



**Fig 11: Detail of the bounce in the relay.**

end of a tuning fork. We can even see the frequency, almost 10 cycles per cm screen. Most likely, the contact pressure was not sufficient. This time it could be due to the generator I used to drive the relay with that was marginal, but I doubt it. The issue here is not to check or repair relays anyway, but to write and illustrate this article! So I am glad that the contact was ringing! If I had to do something

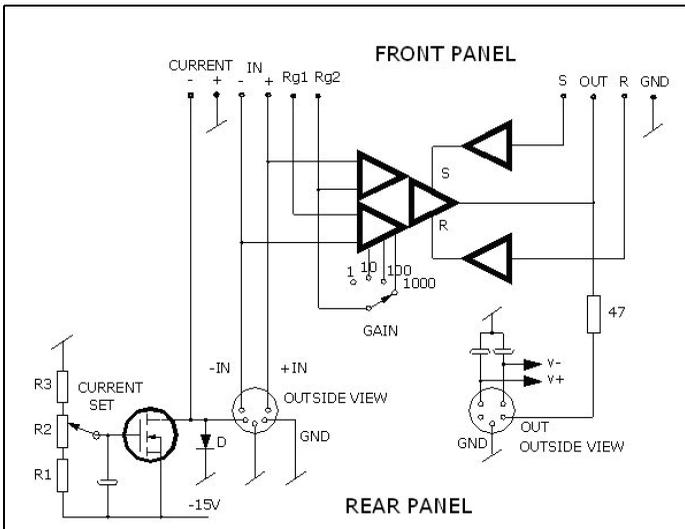
about it I would make sure the voltage on the coil was proper 12V. Next I would measure the contact pressure, and compare it to that of the other side, which switched very well.

#### 4.

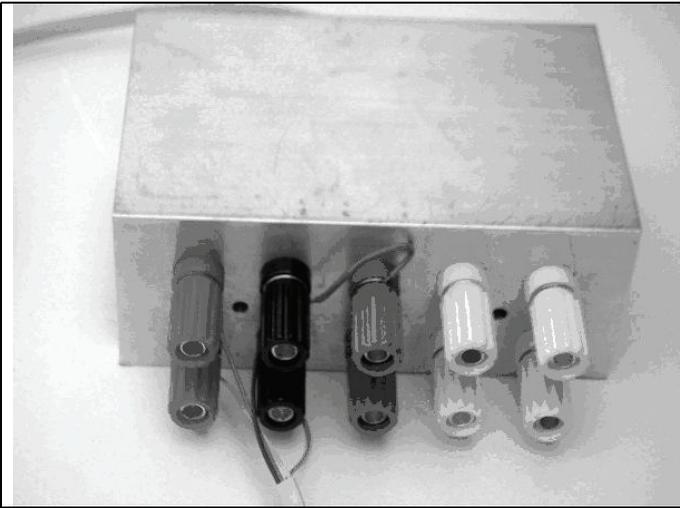
### Building up the instrument

I elected to use an “aluminium channel”, a U shaped extruded channel, for an enclosure. This one was a bit thick walled, but I can drill and cut the necessary holes in it. Otherwise, a sheet metal box may be a better choice. Fig 12 shows the circuit diagram of the unit that I built when writing this article. I wired the amplifier directly to the terminals and the DIN connectors. No circuit board, no socket. If one has to change the amplifier it was not a good idea. If one does not have to change it, the less stuff, the better the function! Maybe a ‘Veroboard’, with copper stripes and punched holes on a 0.1” grid, is a good compromise.

With the amplifier at 1000X, and 1mA of



**Fig 12: The circuit diagram of the IA made for this article.**



**Fig 13: The front panel of the resistance meter.**

constant current, the output is  $1V/\Omega$ . Most scopes can show  $5mV/cm$ , so we can easily see resistances in the single  $m\Omega$  range. Of course, if you use  $3mA$  instead, you will get  $3mV/cm$ . Such small resistances are found in a short piece of hookup wire, so we *must* use the Kelvin connection! Now we can easily have several feet of unshielded cable, and yet we are getting a clean signal! I like the DIN audio connectors, and have elected to use a 5 pin DIN connector of the  $180^\circ$  type for this task. I use the 5 pin  $240^\circ$  type for power supply, and the two can of course not be mixed up. One of the two leftover pins there I connect one to the amplifier output, and the power supply has BNC connectors to these pins so it can be hooked up to a scope. But this is not necessary.

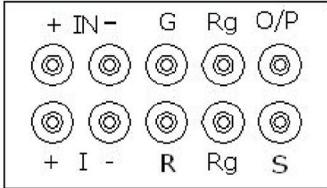
The MOSFET as a variable constant current source is very simple and practical. The drawback is that it is difficult to set a particular current with accuracy. It may be a good idea to make a calibrating plug out of a DIN connector with a precision resistor, permanently connected (using 4 leads!). The resistor can be of any, low, value, such as  $0.1\Omega$ . As long as you have verified its resistance with some accuracy it is OK. Then you just plug it in and adjust the current source

for whatever scale factor you desire on the display. Keep this calibrator with the tester!

When wiring the  $180^\circ$  connector, it would be logical to put the current source on the two outermost pins, ground on the middle one and the amplifier inputs on the two remaining, the ones used for the right side channels in audio stereo systems. We will abandon logic this time! If we hook up the amplifier inputs to the outermost pins we can use it for direct connection of for example microphones and other audio equipment with 3 pin  $180^\circ$  DIN connectors! Ground on the middle is still a good idea, and the input of the unused microphone pin can be connected to ground on the front panel. The current source and return are now on the “right channel” pins!

The front panel of the unit is shown in Fig 13, the connections are shown in Fig 14:

- Top left two terminals are the + and - IA inputs. Underneath are the + and - current output and return. These can be strapped vertically when making resistance measurements.



**Fig 14: Front panel connections.**

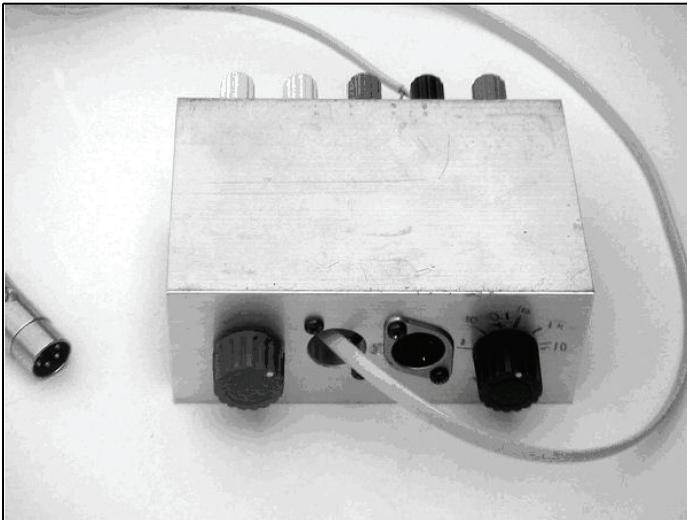
- Third is Ground, underneath is Reference. They can also be strapped vertically.
- The fourth in the upper row is Rg1, and below it is Rg2. They are normally left open, but here is where an impedance is connected that will determine the gain, with the gain switch at 1. At these Rg pins we can also monitor what is on each input!
- Rightmost, upper terminal is Output and below it is Sense. They would normally be strapped together, just like Reference and Ground.
- The two holes between the terminals are the trimmers for input offset (to the left) and output offset.

To adjust the input offset: connect a short, or a resistance like what you will use the amplifier with, between the input terminals, set the gain to 1000 and adjust for a zero output. The zero may vary slightly for different input loads. The inputs cannot be left floating. The cure is to run a 1MΩ or so from one input, or both, to ground!

To adjust the output offset: leave the same short in place, set the gain to 1 and adjust for 0 output. Repeat if necessary. You should now have a beautifully zeroed, general purpose, amp!

The rear panel of the unit is shown in Fig 15:

- On the left is a knob for a rotary switch. This selects the gain of 1, 10, 100 or 1000 by connecting either nothing, pin 13, pin 12 or pin 11 to pin 3 (Rg2).
- To the right of this knob there will be a 5 pin, 240° DIN connector. It is on order from Newark [8], the only distributor I found here in the US who seems to understand this connector format and have an intention to carry them all. They probably have



**Fig 15: The rear panel of the resistance meter.**



all the rest of the components too for this project.

- Next is the 180° 5 pin DIN connector. It has the IA inputs, the constant current output and return, as well as Ground. This is straight in parallel with the front panel. Sometimes it is just handy to have it ready for a DIN connector; sometimes one may want to build up some network on the front panel terminals!
- The knob on the right end is the control for the constant current. Mine give 1 $\mu$ A at 9 o'clock, then 100 $\mu$ A at about 12 o'clock, 10mA at about 3 o'clock. 50mA is about all I get out of it, but that is plenty. The 1 $\mu$ A is probably less than ever needed, but I leave it in place.

My MOSFET, from the top layer in the Junk Box, happen to be IRF 120. 100V and 9A, it is an N channel, so I connected it to the negative supply. A P channel one can be connected the same way to the positive supply. The gate needed to be varied from just over 2V to 3.5V (compared to Source, the -15V rail) and the linear potentiometer was 47k $\Omega$ . 1.5V/47k $\Omega$  in other words. Multiply this by 2V and get R1 = 56k $\Omega$  and R3 = 330k $\Omega$ . I have not yet tried the capacitor from gate to ground, but since this is referenced to the 15V line and not to ground, maybe it is needed.

So our instrument will have a source of (relatively) constant current. The article about a true Rheostat [7] showed the way. Any MOS FET will be quite a constant current sink, or source, given a constant gate voltage. The gate voltage is proportional to the logarithm of the current, which is beneficial for us! We can set a small current, as well as a larger one, with the same pot. It is easy to derive some fairly stable 5V for the gate voltage pot, and its padding resistors from the 15V supply. If not: employ a separate little 5V regulator! It would be nice with as much constant current as 1A,

but it will require quite a bit of power supply, and transistor dissipation, if it is derived from a 15V source, but it is possible.

I realised that the voltage on the test leads themselves, not only across the contacts measured, had to be limited! Even if the test current is limited to 1mA, it may not do much good if the leads are charged up to 15V at open circuit. When suddenly shorted, by the closing contact, the leads discharge, and a much stronger current than 1mA flow until they are discharged. This current I guess would be 15V divided by  $Z_o$  of the cable (possibly 100 $\Omega$ ) => 150mA.

So I limited the voltage with a forward diode across the current source pins. Had I had one handy at that time I would have chosen a Germanium diode; 0.25V is more than enough for measuring voltage drops in the  $\mu$ V range, but a Si diode, giving some 0.56V at this low current, will work! In the present design we can chose any diode we want and I have stocked up on Ge diodes since then. Beware that if you use a Ge diode, the drop of some 0.25V, is close to the 0.2V burden common for digital mA meters! So use a higher range than the most sensitive for a true reading on the meter!

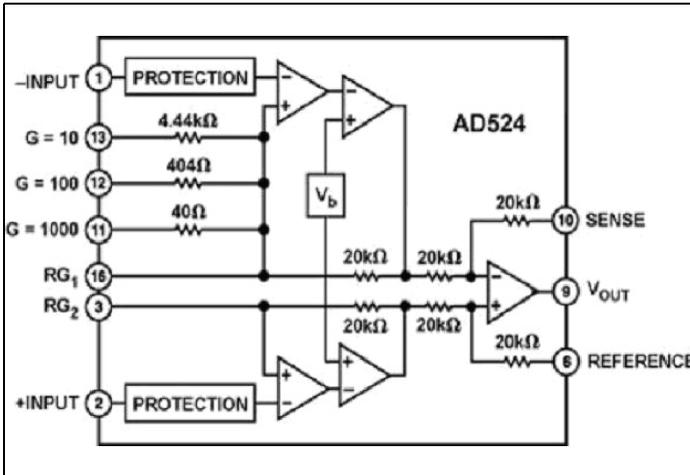
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## 5.

### What is an Instrumentation Amplifier (IA)?

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The "IA" is an arrangement of three operational amplifiers and some resistors, see Fig 16. In a way, it is an "ideal" amplifier, as it has very high input resistances (on both inputs), very good suppression of common mode voltages and one can control the gain of differential inputs by selecting the value of one



**Fig 16: The AD524 Instrumentation Amplifier (IA) block diagram.**

resistor,  $R_g$ , connected between pins 16 and 3, and none of the fixed gains selected. Gain =  $(40k\Omega/R_g) + 1$ .

The brave at heart sometimes roll their own IAs, but there are 6 resistors (the 20k $\Omega$  ones) to be matched, beyond anything practically possible, to even get close to the performance of the AD524. If your time is worth \$1/hour, and if you get all the parts for free, you will go upside down on the project anyway! So spend some \$10 and buy one. Naturally, this IC is too complicated to describe here. Download the data sheet!

### 5.1 The Instrumentation Amplifier as an Instrument in itself

There are several stellar properties in an IA. One of them is called Common Mode Rejection Ratio, describing how well a common mode (as opposed to a differential mode) input signal is rejected. In this amplifier, at a gain setting of 1000X, it is better than 120dB! Think for a moment! With a common mode signal of 10V, the output reacts as for a 10 $\mu$ V differential signal, or less! Slightly different guaranteed value for the different grades and gains, but to my experience, they are usually better than promised. Rather than pay \$30 for best grade, buy 3 of the lowest for the same amount, and test

them!

Another great property is that both inputs have very high resistances, on the order of 1G $\Omega$  or more!

Whatever the input voltage times the gain produces, the output is not necessarily referenced to ground! In an IA it is referenced to the Reference pin voltage! If it is grounded, the output voltage is referenced to ground. If it is connected to a 6V AC source, the output voltage will be referenced to it, and with a very good accuracy!

The Sense pin is confirming that the output voltage is what it is supposed to be. If the output is driving a voltage divider and, say 1/3, is going back to S, then the O/P pin will have 3 times the voltage we would expect, keeping the O/P Sense pin at what the output "should be". Additional gain in the output stage of the IA makes this possible. This can of course be useful, but I would consider it a "spare gain" to be used only if you have to, and limit it to 10 or maybe 100X. These R and S functions are useful when feeding long, lossy, lines.

The ability to reference the Ref and Sense to remote points allows us to



manipulate the output voltage, for example so we can get an output current as opposed to an output voltage! See Fig. 40 in the data sheet for an example! For normal use, the S is connected to the output and R to ground. I have placed them accordingly on the front panel to make it easier to strap them.

Notice that both the Ref and the Sense pins have 20kΩ resistances to drive! The actual voltage on the pins is what counts, so if you are driving them from a source that has almost *any* impedance but 0Ω, you will introduce an error! Considering the general purpose nature of this project, we want to be able to drive the R & S pins by the neighbour's cat walking by, almost! So for these pins: put in a good dual OP as voltage followers, one half for each pin! If you need better accuracy than a mV or so you may have to choose OP amps with offset trim. I suggest that any OP of good quality will be good enough. In the schematic diagram I have just indicated these, but I take it that you know how to make a follower with a gain of +1 and that it too needs +/- 15V.

The gain pins. There are three fixed Rg provided on chip. The selection of either provides 10 – 100 – 1000 times voltage gain, with good accuracy, and 1 X is had by not selecting any of them. Furthermore: Rg, connected between pins 3 and 16, can set the gain to any value. The formula is  $|G| = (40k\Omega/Rg) + 1$  with a ±20% tolerance, and it does not have to be just a resistor. A parallel resonant LC circuit here will display high impedance at resonance => low gain at resonance => a notch filter. The Rg, or parallel LC resonator, can be connected in series with a capacitor, following the gain formula for higher frequencies, yet moderating the gain for lower fq where the C reactance is high. As the DC resistance is very low, leading to "infinite" DC gain, the circuit may need a larger series capacitor.

A series LC circuit displaying low impedance at resonance will display low impedance => high gain => band pass filter.

## 5.2 Other applications for the IA.

Many transducers are differential in nature. Any Wheatstone bridge will benefit from an IA as a null amplifier. For example load cells with strain gauges. The load cell is often built as a "half bridge" with two strain gauges (a thick film resistor, on a piece of tape, that increases its resistance when stretched) and two fixed precision resistors. They are connected as a Wheatstone bridge. If you are trying to build this yourself, the notion "Precision Resistor" will take on a new meaning! This holds true even if you use 1% metal film resistors... In your search for better resistors, the name Vishay [9] will surface. Your problems are over, it is one of the few companies who can make a Precision Resistor. I will remember that they bought up Micro Measurements, who makes some of the best strain gauges, so now it is all under one roof.

One can also use four strain gauges, in a full bridge, with twice the output voltage for a given load. It is also the most stable design since all resistors can be strain gauges of the same type.

Using a ±15V DC excitation of the bridge, 1k gauges and the AD524, will result in an incredible sensitivity! 350Ω gauges are more common, running on ±5V (they cannot handle 15V). Already without LP filtering the output signal from noise, you can see how a good piece of aluminium bends from under its own weight when held horizontally! The AD524 has a GBW of about 25MHz, so 25kHz in the 1000X configuration. Thus there are no problems feeding the bridge with AC, for example 1kHz, and use synchronous detection (use a FET for Rg and modulate it!) Depending on your filtering, the BW of the system drops and



the already low noise drops. Of course, LP filtering of a DC output signal to similar BW will give similar sensitivity.

Another use is to find equipotentials on metal structures, complicated or not so complicated; they can be 2 or 3 dimensional. For example: on a Hi-Fi amplifier chassis or a PCB. Ground the IA somewhere on the part to be measured, apply an excitation current to the object and use the input leads to find points with the same potential ( $\Rightarrow$  no output voltage)! It may be easier to use an AC voltage for excitation and have a scope or headphones on the IA output.

You can also use it for to track down traces on a PCB that have excess current draw. Make a "fork" with two sharp pins about a cm apart and connect them to the IA inputs. Already single mA currents can be detected in a trace a few mm wide!

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## 6.

### References:

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[1] 1990 in the ARRL UHF/Microwave Experimenters Manual pp 7-53 & 7-54, Carl G. Lodström SM6MOM.

[2] The Analog Devices page from where a PDF data sheet can be downloaded

[http://www.analog.com/en/prod/0,,759\\_782\\_AD524,00.html](http://www.analog.com/en/prod/0,,759_782_AD524,00.html)

Articles in which I have used, and sometimes praised and described, the AD 524:

[3] Measurement Arrangements for Complex Impedances, Carl Lodström, VHF Communications Magazine, 1991/2, pp 93 - 101

[4] An RF Power Meter with Linear Scale, Carl Lodström, VHF Communications Magazine, 1994/4, pp 201 - 204

[5] A Simple Noise Figure Meter, Carl Lodström, VHF Communications Magazine, 2002/2, pp 105 - 110

[6] The Noble Art of measuring optical power, Carl Lodström, VHF Communications Magazine, 2006/1, pp 35 - 46

[7] The Noble Art of the Rheostat, VHF Communications Magazine, 2003/4, pp 242 - 249

[8] <http://www.newark.com/> in general and [http://www.newark.com/NewarkWebCommerce/newark/en\\_US/endecaSearch/searchPage2.jsp;jsessionid=5GQDIMRXXEA0QCXDUY2SFGIK2OTCHIV1?N=1001963++&=0&N=0&=true&No=0&comSearch=true](http://www.newark.com/NewarkWebCommerce/newark/en_US/endecaSearch/searchPage2.jsp;jsessionid=5GQDIMRXXEA0QCXDUY2SFGIK2OTCHIV1?N=1001963++&=0&N=0&=true&No=0&comSearch=true) in particular!

[9] <http://www.vishay.com/company/brands/foil-resistors/>

<http://www.vishay.com/company/brands/transducers/>

<http://www.vishay.com/company/brands/micromasurements>



John Fielding, ZS5JF

# Transistor Frequency Multipliers

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## 1.

### Introduction

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In a previous article we looked at the use of diode multipliers. These all suffer from one serious disadvantage, the conversion loss when multiplying to high orders can be quite high. This makes them only useful for low orders of multiplication of perhaps 3 as a practical maximum.

Using an active device such as a transistor can overcome the inherent loss of the simple diode multiplier, because the transistor can have appreciable gain when used as a multiplier, which eases the design of multiplier chains where high multiplication is required. If the transistor is operated in grounded emitter configuration with the input RF drive applied to the base the impedance is medium and a greater RF voltage swing can be obtained from relatively low drive power.

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## 2.

### Principle of Operation

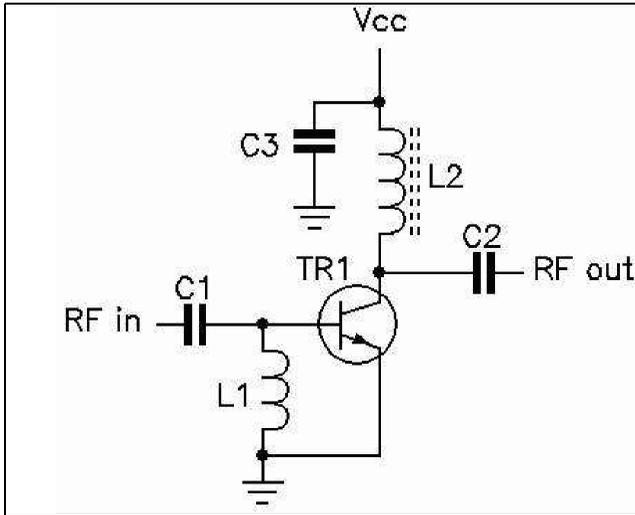
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If a transistor is used as an amplifier for a

signal requiring linear amplification then it is limited to the output power it can produce because of the maximum voltage swing on the base or collector. If the input voltage swing exceeds the base-emitter forward voltage the junction starts to behave like a rectifier diode and distortion occurs. We can simplistically think of a multiplier transistor as being two separate items. The first is the diode junction contained within the base-emitter and the second is a very linear voltage amplifier between the base and collector. By overdriving the base-emitter junction we generate harmonics of the input frequency and these are then amplified in a highly linear manner by the collector-emitter portion of the transistor.

Unfortunately it isn't as simple as this, but the simplistic idea forms the basis of all transistor frequency multipliers. We could place a multiplier diode across the base-emitter and couple into the transistor base via a capacitor. By generating harmonics in the external diode these are then amplified by the linear biased transistor. Some commercial two-way radio multiplier chains used this technique. Unfortunately the efficiency this achieves is worse than the diode on its own if we include the DC power being drawn by the transistors before and after the diode multiplier.

By utilising the inherent transistor base-emitter diode junction we have a better



**Fig 1: Basic multiplier circuit.**

method of generating high levels of harmonic power. Of course we still have to filter out all the unwanted harmonics at the collector and select the desired one with as little loss as possible. However, by altering the conduction angle of the transistor we have a method of enhancing certain harmonics and hence reducing the unwanted ones.

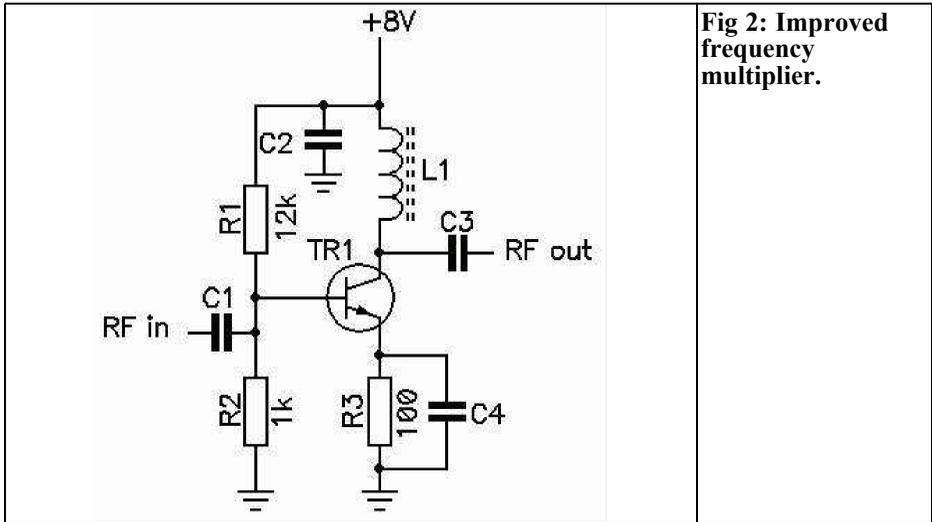
The most basic circuit is shown in Fig 1. Although this will work it does have a number of problems, which we will consider later.

The drive signal to be multiplied is applied to the base of TR1 via C1. The inductor L1 across the base to ground holds the base at DC ground and looks like a high impedance to the input frequency. The emitter is also directly grounded, so for the transistor to conduct the RF voltage across L1 needs to exceed the base-emitter junction voltage of about 0.6V. This means that when the RF voltage is below the threshold nothing happens, but as soon as the base voltage exceeds about +0.6V on the positive peaks of the RF waveform the transistor will start to turn on. If the RF drive is raised even more the transistor will be conducting for a greater portion of the RF envelope and so more collector cur-

rent will flow. The transistor is being operated as a Class C amplifier and the waveform at the collector is similar to a square wave, which is rich in harmonic content. All that is required is a suitable band pass filter to select the wanted harmonic and reject the remainder.

On the face of it this seems to be perfect. But one thing to be aware of is that the multiplier is very critical to the RF drive level, too little and it will snap off and too much and it will saturate and the gain will fall dramatically. Also the threshold is very dependent on the base-emitter forward voltage. If the ambient temperature varies over a wide range the base-emitter voltage does the same. At very low temperatures the voltage required to make the base-emitter junction conduct will be greater than when the ambient temperature is high. So although it works fine in the shack it may not work at all from a contest site on top of a mountain in winter!

The danger with this simple circuit is that the base-emitter junction can be driven very negative on the opposite half cycle and small signal bipolar junction transistors have a limited reverse base-emitter break down voltage, as little as 2V can destroy the junction.



**Fig 2: Improved frequency multiplier.**

Another failing with this simple circuit is that the impedance presented to the driving stage varies wildly as the drive level varies. Below the threshold voltage the multiplier appears close to an open circuit, at high drive levels the multiplier appears as low impedance, just like the diode multipliers.

### 3.

#### Improving the Basic Circuit

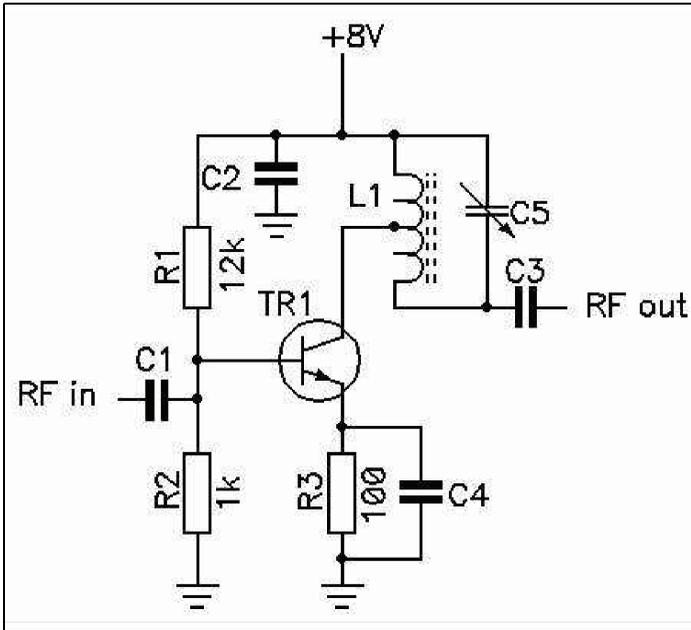
The first problem we will address is the temperature sensitivity. In a linear amplifier using the common emitter configuration we normally include a resistor in series with the emitter to ground to provide a small bias voltage. This helps to hold the collector current more stable with variations in temperature.

The second problem is the sensitivity to the RF drive. We can to some extent reduce this tendency by applying a small amount of forward bias to the base-emitter junction with a potential divider of two resistors. By selecting the resistor values such that the base-emitter voltage

is just below the conduction point the RF drive voltage swing is greatly reduced. This translates into less drive power being required and higher input impedance. Fig 2 shows the modifications needed.

The value of the emitter resistor is determined by the order of multiplication required. For high multiplication factors the resistor value is higher than for low order factors. This is to reduce the conduction angle to enhance the higher harmonics. In simple terms we need to move the conduction more towards Class C as the harmonic order increases.

The value of the base potential divider inter-reacts with the emitter resistor. Essentially what we need to achieve is the DC voltage on the base with no RF drive to be just below the conduction point or a very small collector current flowing, a few hundred  $\mu\text{A}$  typically. With the values shown in Fig 2 the base voltage with respect to ground is approx. 0.62V. But the collector current flowing in the emitter resistor raises the emitter voltage above ground. For a collector current of 20mA flowing when the base is driven the emitter will be approx. 0.2V above ground, making the effective base-emitter voltage about 0.42V. Hence the input



**Fig 3: Incorrect collector coupling.**

RF drive needs to be greater than about 0.3V peak to turn the transistor on.

The disadvantage of this circuit is the emitter decoupling capacitor; this needs to be a low reactance to not only the input frequency but also all harmonics. This requires quite a low inductance capacitor and a SMD chip type is normally the best choice. If leaded capacitors are used, for example ceramic disks, the emitter decoupling may not be good enough for the higher harmonics and some loss of gain will occur at the higher harmonic products.

#### 4. Coupling Out Of the Collector the Required Harmonic

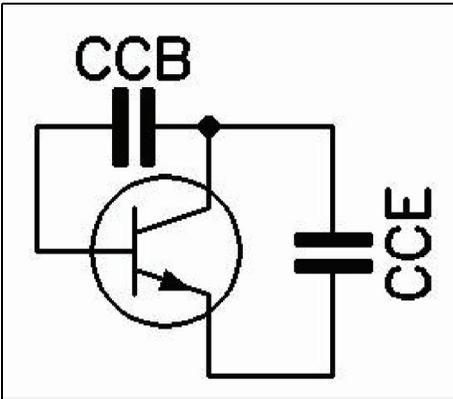
This is where most published designs get it wrong. Although the multiplier stage appears to work satisfactorily often rear-

ranging the circuit a little can make a distinct improvement.

In Fig 1 and 2 the collector is connected to a radio frequency choke with a wide bandwidth. The coupling capacitor serves to block the DC supply and also transfers the AC signal to the following band pass filter. In Fig 3 is shown the wrong way of coupling out the harmonic signal.

The collector of TR1 is tapped down the inductor L1. L1 is resonated to the required frequency by the variable capacitor C5 and this tuned circuit could be the first portion of a multi-pole band pass filter.

This tapped inductor method is commonly used for amplifiers such as intermediate frequency amplifiers where linearity is critical. The tapping of the inductor allows a good impedance match to the collector, the conjugate match. However, in a frequency multiplier this is exactly the opposite of what we require. To understand this logic we need to look more closely at how the bipolar junction

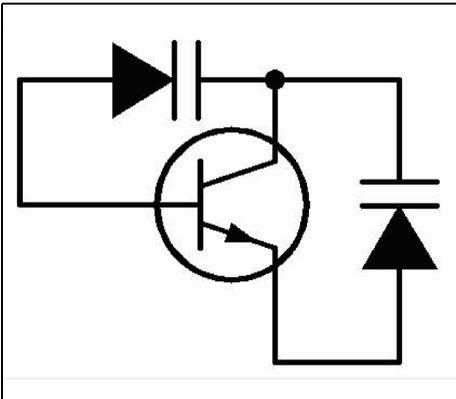


**Fig 4: Transistor parasitic capacity.**

transistor works and its internal structure.

Fig 4 shows the two major parasitic capacitances inherent in the bipolar junction transistor. These are the collector to base (CCB) and collector to emitter (CCE). There is one other which exists between the base and emitter but this does not affect the operation, only the input matching.

Normally in a small signal linear amplifier application the capacitance is relatively constant and can be tuned out. Because the RF voltages appearing across the junctions are small the capacitances are constant. But in a frequency multiplier with correct collector matching the RF voltages are not small, in fact we want to make them as large as possible, without exceeding the collector-emitter secondary breakdown voltage of the device. The reason is that with a large voltage swing the parasitic capacitance is not constant but behaves like a varactor diode. Motorola in an Engineering Bulletin [1] gave details of how the varactor diode characteristic can be exploited to make more efficient frequency multipliers. This used the MRF629 a 12V/2.5W UHF power transistor as a *parametric multiplier*. This exhibited a power gain of over 6dB when driven by a 100mW signal at 150MHz and developed as much as 700mW output at 450MHz

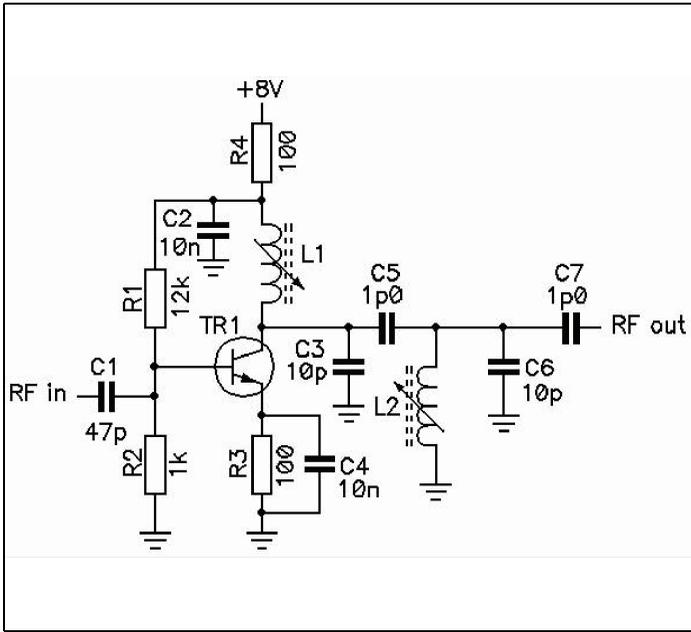


**Fig 5: Inherent varactor diodes in a BJT.**

when configured as a tripler. The writer used this technique when building the 1152MHz LO chain for a 2C39A power-mixer SSB transverter for 23cm and obtained nearly 1W output at 384MHz. This used 150mW input at 96MHz with the circuit configured as a 4x multiplier. The use of idler circuits at the input and output is the secret to efficient multiplication.

This technique is not just applicable to the MRF629 (which today is obsolete) but even small signal BJT RF transistor exhibit this effect when a sufficiently large RF voltage is developed at the collector. Hence, we need to match the collector to a much higher impedance than suggested by the formula for conjugate matching. This also explains why junction field effect transistors (JFETs) are poor frequency multipliers, as they do not exhibit the same varactor effect as BJTs. Fig 5 shows the inherent varactor diodes in BJT

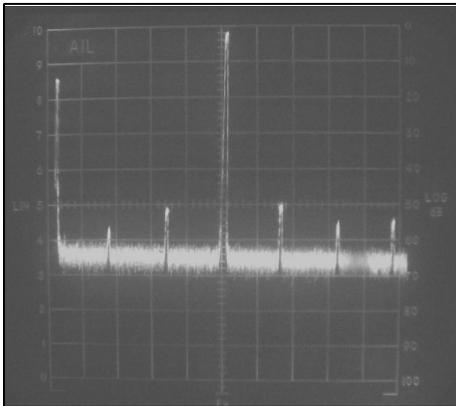
In Fig 6 is shown an example of the correct collector matching for a frequency multiplier. This uses a two-section top-coupled band pass filter to select the required harmonic. The Q of the tuning components needs to be sufficiently high to develop the required RF voltage swing.



**Gig 6: Typical frequency multiplier stage.**

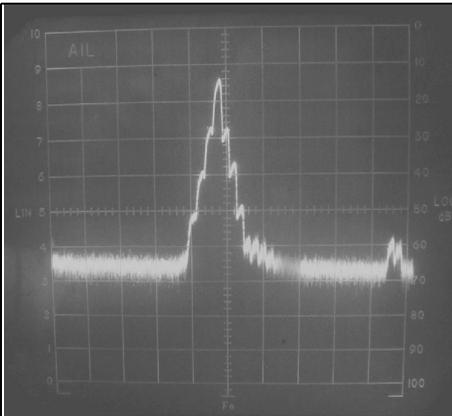
The input to the multiplier is via a capacitor with a medium impedance at the operating frequency. The drive source is a low impedance stage such as an emitter follower and with an output voltage swing of at least 1V p-p. The tuned circuit of L1 and C3 selects the required harmonic. This develops the necessary high impedance, and hence the higher voltage swing, at the collector of

TR1. The coupling into the next section of the filter is made via C5, a very low value capacitor to prevent loading of the collector. This capacitor is also the top coupling for the band pass filter so it needs to be a very low value, which suits the circuit configuration well. The second inductor L2 is resonated by C6, the coupling to the next stage is performed by C7 which is also a low value capacitor because we need to prevent loading on the band pass filter, which would alter the selectivity response if excessive loading occurs. The values of the inductors and capacitors need to be chosen to suit the frequency. If more selectivity is required it is often better to incorporate this into the following buffer amplifier before the next multiplier stage.



**Fig 7: Typical spectrum for a BJT multiplier.**

Fig 7 shows the typical spectrum at the output of the band pass filter. This circuit is the one shown in Fig 6 and uses an MPSH-10 transistor as a tripler. The input frequency is 50MHz at a level of 5mW in 50Ω and the wanted output at 150MHz measures approx. +13dBm (20mW). The worst spurious products



**Fig 8: Unstable multiplier spectrum.**

are about 50dB down on the 150MHz signal. This will require further filtering in another band pass filter before it is applied to the next multiplier or used by the equipment.

One thing not obvious is why a bipolar junction transistor is the preferred device. Some schematics show junction Fets as multipliers. This is a poor choice because the Fet does not contain the same type of varactor junction as the bipolar. Hence, the drive power required is much greater to get any significant harmonic generation. Fets make good mixers because they are used in a “additive mode” with a second carrier, the local oscillator. The local oscillator energy causes the drain-source current to swing up into the saturation region and the non-linearity is greatest under this condition. The Fet is operated as a saturated switch to “chop” the input waveform and cause additive and subtractive products, one of which is the wanted mixing product. If a junction Fet and a similar die size bipolar junction transistor are compared in a multiplier circuit the bipolar will deliver more harmonic output power and require less drive power. Hence, junction Fets are not very suitable for straight multiplier duty.

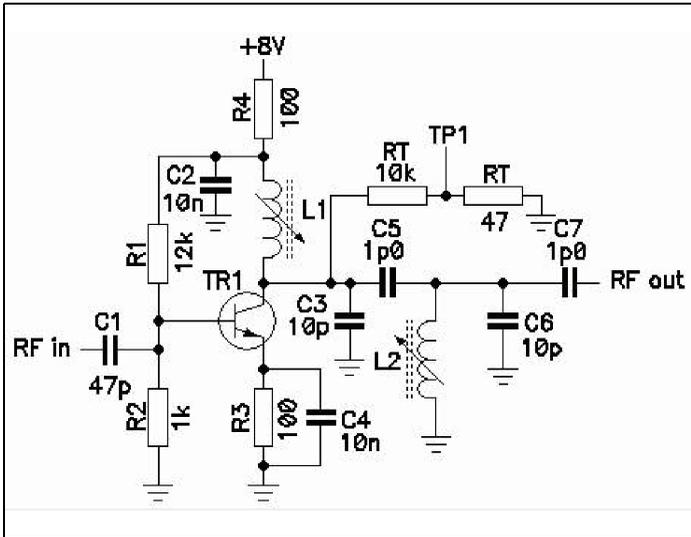
## 5.

### Multiplier Stability

In general, frequency multipliers using BJTs are inherently very stable because the input and output are tuned to different frequencies. Unlike a tuned amplifier where the input and output are the same, the only thing that can make a frequency multiplier unstable is either poor layout or the internal parametric components of the transistor. If the drive level is marginal the transistor can exhibit erratic operation. If the input drive varies over too great a range the multiplier can snap on or off as the level varies. Hence, it is necessary to have some reserve drive in hand. A good minimum is about 3dB above that required to ensure reliable multiplication.

When tuning up a multiplier stage the power output should vary smoothly as the input and output are aligned. If the multiplier snaps on and off as the tuning is altered it is usually a sign that the input drive level is marginal. A good check of this is to measure the base voltage of the BJT with full input drive. If the drive is adequate the base voltage should be a voltage slightly negative with respect to ground. A value of -0.25V to -0.75V is normally about right. When the input drive level is more than required the top potential divider resistor in the base bias chain can sometimes be omitted and the drive develops a negative base voltage across the base to ground resistor. In cases such as this it is prudent to connect a reverse biased diode across the base to ground to clamp the base reverse voltage to less than 0.7V to protect the base-emitter junction.

Fig 8 shows the spectrum for a multiplier stage where the input drive level has been deliberately reduced to show the effect that instability causes.



**Fig 9: Test point added to multiplier stage.**

**6.**

**Optimum Supply Voltage**

In all the years of designing frequency multipliers using BJTs it has been found that there is a certain range of supply voltage where the optimum performance occurs. It is not quite clear why this is so but is probably to do with the inherent parametric varactor diodes in the BJT. However, from many experiments the optimum supply voltage on the collector-emitter junction is about 7 to 9V, the so called “sweet-spot”. With voltages far from these limits the efficiency suffers, with too low a voltage the power output falls away and with too high a voltage the efficiency also falls off. Hence, when designing a multiplier chain using BJTs it is often best to use a supply voltage of 8 to 9V. This is convenient as a 3-terminal regulator can be supplied by the nominal 12V-vehicle supply.

If the drive level into the next stage is too high a simple cure is to starve the inter-stage buffer amplifier transistor or MMIC following the multiplier by in-

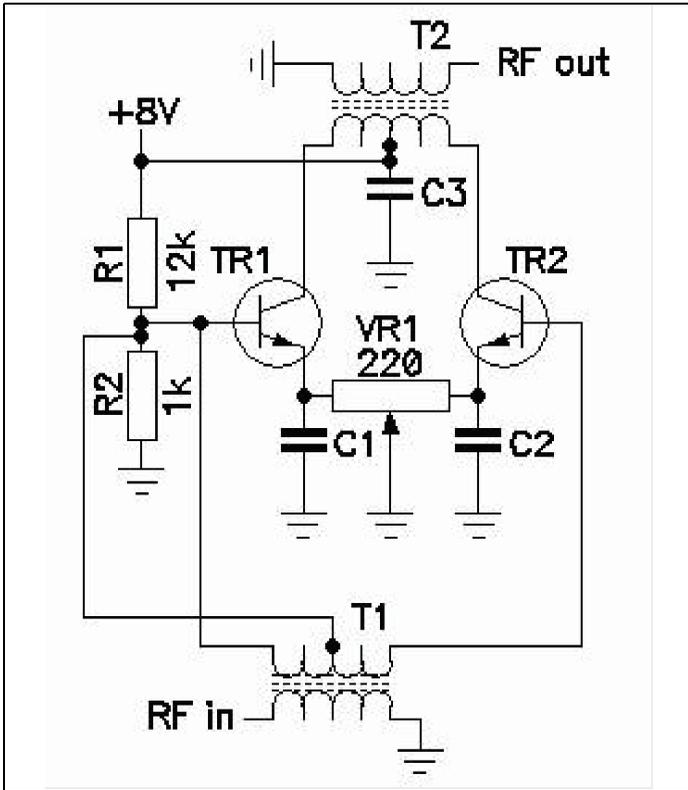
creasing the collector feed resistor value to limit the collector voltage.

**7.**

**Choice of Transistor Type**

As strange as it may seem we do not need to use a “super-transistor” for an efficient multiplier. Often the choice of a transistor is dictated by the frequency of operation and hence the fit as an amplifier needs to be quite a bit higher than the operating frequency. This however is for a transistor operated as a linear amplifier, when used as a highly non-linear multiplier we can often get by with less exotic devices.

Form experiments made on a wide selection of transistor types it has become apparent that even lowly devices such as the 2N2222 with an  $f_t$  of only 100MHz work perfectly well up to a few hundred MHz when operated as a doubler or tripler stage. For the frequencies higher than this a move to a popular type such as the MPSH-10 or the BFR92A or



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

BFR93A in the SMD SOT-23 package work well even as high as 2GHz, which the manufacturers data does not suggest. Of far more importance is the drive level and the filtering components used to select the wanted harmonic. If the filter has excessive loss then the advantage of using an exotic transistor is wasted.

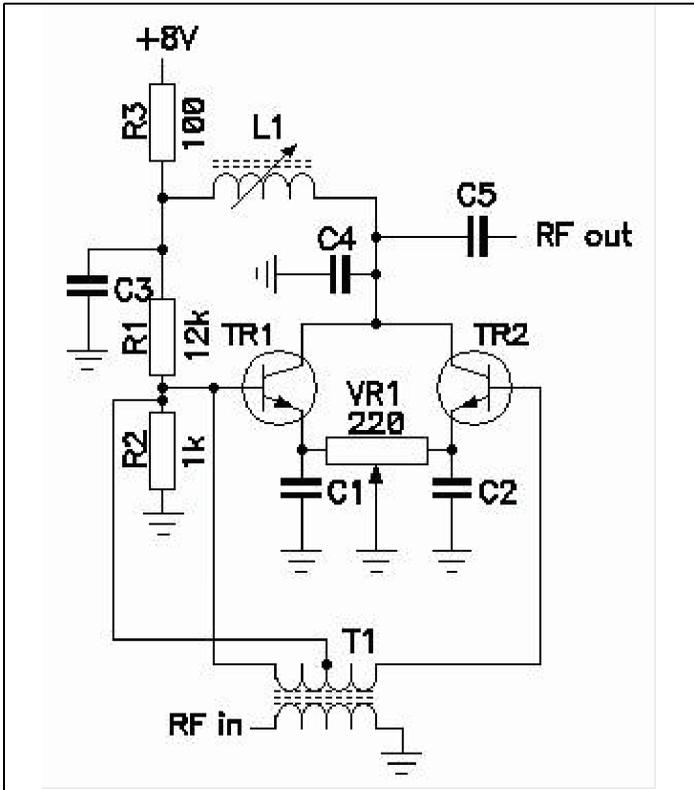
## 8.

### Test Points

When aligning a multiplier chain it is helpful to be able to break into the chain at critical places to attach test equipment to see what is going on. Some designers make all the multiplier stages to termi-

nate in  $50\Omega$  so that normal test equipment can be connected. This not only introduces extra loss but also adds unnecessary components to the circuit. We can safely assume that the base input impedance of a typical BJT when driven at the optimum level is about  $200\Omega$  to  $500\Omega$ . Therefore the drive power is not going to be too high; perhaps  $+10$  to  $+13\text{dBm}$  in  $50\Omega$  when transformed up will give sufficient base-emitter voltage swing.

Probing a high impedance portion of the circuit with a RF AC voltmeter often causes detuning because of the probe capacity. A simpler method is to incorporate some test points where a  $50\Omega$  instrument can be attached without breaking the chain. Often two resistors can achieve this and the  $50\Omega$  equipment can be attached as and when required without upsetting the system. This con-



**Fig 11: Push-Push tripler.**

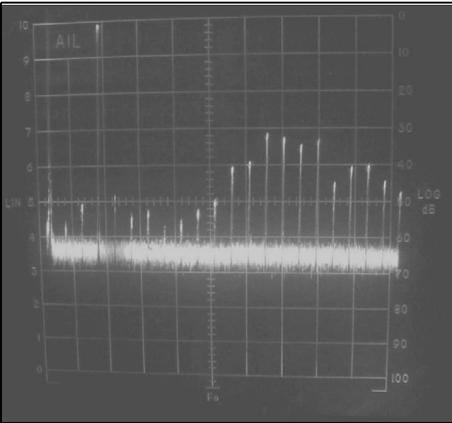
cept is illustrated in Fig 9. The two resistors are a high value one to tap into the high impedance point and a second termination resistor to suit the test equipment. A coaxial cable can be connected and taken to a spectrum analyser to see the spectrum. The amplitude will not be correct but the spectrum should still be the same.

## 9. Using MMIC Amplifiers as Multipliers

There have been several application notes and other articles published concerning the use of wide band microwave

monolithic integrated circuit amplifiers (MMICs) as frequency multipliers. The choice of devices is somewhat limited but several writers have reported good results. However, because of the internal biasing components of typical MMICs these devices do not really lend themselves to this application. Whereas it is true that any transistor can be over driven to force it into a non-linear operating portion this seems not so easy with the typical MMIC. As the typical MMIC is designed to interface with 50Ω terminations intuitively we can see that we cannot run the collector into the high impedance necessary to induce the varactor diode effect.

Where MMICs can be made to work is with an external shunt diode across the input, with DC blocking capacitors to prevent the bias being upset, to generate



**Fig 12: Wide band sweep of LO signal.**

the harmonic energy and then to use the MMIC as a conventional amplifier to raise the level. However, when you run the calculations of the total power required the efficiency still ends up being quite low and a cheaper BJT will probably out perform it at far less cost. MMICs are good as inter-stage buffer amplifiers but not so good as multipliers.

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## 10.

### Special Multipliers

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Where we need additional suppression of certain harmonics we can utilise some of the ideas found in diode multipliers. Fig 10 shows a push-pull doubler using two BJTs. The circuit suppresses the odd order harmonics by about 30dB and leaves just the even order harmonics. The variable resistor VR1 allows the circuit to be balanced by adjusting the emitter currents of TR1 and TR2. This provides the best null to the unwanted products. For optimum performance a dual-transistor in a single package offers the best results.

The wideband transformers T1 and T2

need to be wound for the best balance and this limits the higher frequency performance.

This circuit can also be modified to form a push-push tripler where the predominant harmonics are the odd order ones. This is shown in Fig 11.

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## 11.

### Multiplication Order

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In many cases we need to multiply by a fixed number and this is clearly defined by the starting and finishing frequencies. For example a receive-converter for 23cm may require a 24x-multiplier chain to arrive at the final LO injection frequency. A 24x-multiplier can be arranged in several different ways. We need to multiply by 24, which can be broken down into doublers, triplers or quadruplers (2x, 3x or 4x). Choosing which multiplier factor to use and where in the chain is often not simple. As a general rule it is often better to make the first multiplier stage with the greatest multiplication factor, so in this case we might choose the first multiplier stage to be a 4x. Against this must be weighed the difficulty of filtering out the harmonics close to the wanted multiplier product.

As an example, consider the 23cm receive-converter for an IF of 144MHz. The starting frequency is 24x less than the LO injection, which in this case is 1152MHz. This determines the crystal frequency as being 48MHz. This is a good choice as the 3x product is 144MHz and if a test port is provided on the first multiplier then a small amount of the 144MHz signal can be used to set the band edge on the 144MHz tunable IF. If the 144MHz receiver is accurately calibrated in frequency readout the receiver can be used to zero beat the crystal



oscillator exactly onto 144MHz. Hence, in this case the first multiplier needs to be a tripler.

However, consider the case where the filtering of the first multiplier is insufficient to reject all the 48MHz spaced products. If these are allowed to pass through the subsequent multipliers then we will end up with 48MHz products spaced either side of the final LO frequency. The 1152MHz product is not of concern but the image at 1008MHz (288MHz away) will cause degradation in the receiver performance.

Hence, it is better to first double the 48MHz to 96MHz, band pass filter the result as well as possible to reduce the 48MHz products to negligible levels, and then proceed from there. No integer multiplication factor using 96MHz can produce a signal at 1008MHz so this is a better plan.

Assuming we choose the first multiplier as a doubler, this leaves a total of 12x further multiplication to arrive at the final LO injection. Filtering at 1152MHz will require a fairly good filter and to ease the design of the filter we need to place the unwanted harmonic products as far away as possible. We could choose the final multiplier to be a doubler and hence the injection frequency into the final multiplier would then be 576MHz. If this choice is made then we need to assess if it is possible to arrive at 576MHz from 96MHz. The answer is we can by multiplying by 6, this would be a 2x and a 3x multiplier in either order.

If we chose the more traditional route of multiplying by 3 from 384MHz to arrive at 1152MHz then we need a total multiplication of 4x from 96MHz; this could be 2x & 2x in two separate stages or a 4x in one stage. Again we need to assess the difficulty of making a band pass filter at 384MHz that can reject the 96MHz spaced products with sufficient attenuation. The safest method would be the dual 2x approach as the filtering of a 192MHz spaced product with a 384MHz

band pass filter is likely to be easier. This also simplifies the filter design and it is likely to have less insertion loss than the narrower filter previously required. In some cases commercial helical filters can be obtained aligned for the required frequency which allows a more compact design. Companies such as Toko and Temwell list suitable filters in their standard product range that occupy very little board area and these have adequate attenuation to the unwanted products.

Unfortunately not all amateurs have access to a spectrum analyser to ascertain if the multiplier chain filtering is adequate. Many constructors have little more than a RF diode probe and perhaps an absorption wave-meter or milli-watt meter to align the multiplier chain. Although it may appear to work satisfactorily it is only when it is examined on a spectrum analyser that the truth emerges, sometimes causing a big surprise! Unwanted products at the final LO of a receive-converter only cause the operator a problem, but if the same dirty LO is also used to mix up in a transmit converter then we are potentially causing spectral pollution to other users. This is not only irresponsible behaviour but probably in contravention of our license conditions. These spurious signals may interfere with sensitive receivers used by other services, perhaps radio astronomy, civil or military communication services and should be avoided for obvious reasons!

One factor which is misunderstood by most amateurs is how clean the LO signal needs to be. If the signal is being used to drive a mixer to down-convert or up-convert then the harmonic content of the final signal is not important, in fact given the correct phase relationship a bit of 2nd and 3rd harmonic content can be beneficial to the following mixer. The following mixer will regenerate the harmonics of the input signal so there is no need to suppress the 2nd and higher harmonics of the LO signal. What is of far greater importance is to eliminate as far as possible the lower multiplier prod-



ucts especially the ones closest to the wanted LO frequency.

To illustrate the problem take a look at Fig 12. This is the tripler circuit shown in Fig 7 but with the spectrum analyser sweep increased to cover up to 1GHz. Although the products close to the wanted 150MHz are acceptable the far out products around 700MHz are very bad. They are barely 30dB down and are due to a poor layout where the harmonic energy was able to hop over the filter and get picked up in the output amplifier stage. A contributory factor is the topology of the band pass filter; it is a top-coupled design and hence tends to act like a high pass filter to the higher harmonics. If the spectrum analyser sweep had not been opened up we would be blissfully unaware of how bad the spectrum was. With the writer's spectrum analyser covering from 10kHz to 12.4GHz it is possible to see significant levels of harmonic energy all the way up to nearly 3GHz. This is with a transistor "officially spec'ed" only to an  $f_t$  of 650MHz.

If significant levels of spurious signals at the LO exist then these are potential interference generators for frequencies far removed from the wanted channel frequency. Most microwave receiver front ends do not have very high selectivity because of the nature of LNA design; few LNAs have anything like the required rejection characteristics because this would destroy the low noise figure required. If the unwanted signal is of sufficient amplitude to get past the front end filtering and into the mixer then these will appear as "birdies" or "sprogs" when the receiver is tuned across the band. If you are really unlucky then you may find that your weak signal DX channel is completely obliterated by a strong signal far from the wanted frequency. Running some simple mixing spurious calculations can highlight potential problem areas.

The choice of the 1st IF and hence the

LO frequency can have a dramatic effect on the image and other spurious response characteristics. If the 23cm band is used as an example then the choice of 28MHz as the 1st IF is not a good choice. For this we need a LO at 1268MHz and the image will then be at  $1268-28\text{MHz} = 1240\text{MHz}$  which is right in the middle of the aeronautical DME and radio altimeter band worldwide! Typical radio altimeters are downward looking radar systems fitted to most aircraft and these radiate up to 10kW of pulsed power. DME systems are used for distance measuring and the aircraft radar interrogates a land-based transponder, this then sends a reply pulse delayed by a known amount. By measuring the time delay the aircraft can calculate the exact distance to the ground transponder. A typical aircraft DME transmitter is about 300W and the ground station anything up to 10kW.

The writer recently had to do a significant system revision for a radio telescope receiver for the Hydrogen Line at 1420MHz that is under development. It was originally proposed to use a 1280MHz 1st LO and the 1st IF was at 140MHz. Only recently it became apparent that a significant amount of L-band microwave signals on or near the image frequency of 1140MHz from aircraft DME transponders existed with high signal levels caused severe interference and this forced a change in 1st LO and 1st IF. The 1420MHz receiver has only 40dB of image rejection because of the nature of the LNA and front-end filter design to obtain the required sensitivity. The minimum discernible signal of this receiver with the LNA noise figure of 0.18dB is less than  $-146\text{dBm}$ , which is about 20dB better than a good 23cm receive converter. In this case it was found that changing from a "Low-Side" injection scheme to a "High-Side" injection placed the image frequency at about 1700MHz which is also a protected portion of spectrum reserved for SETI research and the 1st IF then occurred at 152MHz which is another protected piece of spec-



trum for Radio Astronomy. The 1st LO then occurred at approx. 1572MHz.

## 12.

### Special Local Oscillator Techniques

During the upgrade to the writer's 23cm transverter an unusual approach was used. The original design used a 5th overtone crystal at 96MHz, but this suffered from excessive warm-up drift and other problems. Because the desire was to make as little change as possible it was decided to make a new 96MHz oscillator module that then fed into the existing modules. This new module used a 12MHz TCXO, which was multiplied up to 96MHz, a factor of 8x. The first multiplier used a 2x and the second a 4x. After filtering and amplification the 96MHz signal was fed to the original oscillator multiplier chain. A test port on the second multiplier provided a small amount of 144MHz harmonic energy to calibrate the 144MHz transceiver.

Because the first multiplier output spectrum contained products spaced 12MHz and 24MHz either side of the wanted frequency the filtering required was quite severe. If these products are not removed at the source then it practically impossible to eliminate them further down the multiplier chain. The aim was to attenuate these close-in products by at least 70dB, in the end a figure of over 80dB was achieved by careful attention to the early band pass filters. The 96MHz final output was clean; the worst spurious product was -86dBc, which was considered sufficient. Filtering, shielding and supply line bypassing achieved a compact and clean module suitable to drive the existing multiplier chain. One of the other problems with the 96MHz crystal oscillator was its susceptibility to strong

RF fields when transmitting 250W on 23cm. This caused FM when using SSB or a chirp when CW was used. With adequate shielding and supply line bypassing these problems disappeared with the new oscillator module.

It must be appreciated that the filtering inductors need to be well shielded, the best option is pre-wound canned coils or helical filters such as those made by Toko. If air wound inductors are used the coils radiate spurious energy and this can hop over shielding plates and effectively bypass the filter. The RF voltage at the collector of a multiplier stage can be many volts, and these make good "mini-transmitters" unless the RF field is contained in shielded compartments. Grounding, layout and shielding are all vital parts of any oscillator chain design and construction.

## 13.

### References

[1] EB-70A "Frequency multiplication simplified by internal faraday shields in MRF629", Motorola Semiconductors

### HEROS technology Ltd

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- WinPIC2006 USB full speed programmer.
- CPU microcontroller modules.
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Gunthard Kraus, DG8GB

# An interesting program: Ansoft designer SV 2.2

Part 2, continuation from issue 3/2006

## 4.

### Analysis of an amplifier

Following the successful filter design in the first part of this article, the next step is to analyse a low noise 23cm preamplifier. The description will be in small steps so that the process is easily understood and can be applied to your own projects.

#### 4.1. Analysis of the available S parameters

An active circuit is described by its four S parameters S11, S21, S12 and S22, plus its noise figure "NF". Additionally

the stability factor "k" can be used to test if self oscillating will happen. Then the start point is to decide if the amplifier will use a transistor or HEMT, if The Internet is searched for "LNA design" there are many application notes from other people to help to make the choice. They should be studied to find out the right device for your purpose. Having made that choice the S parameters of the chosen device must be found, again these are available on The Internet.

#### Step 1:

Having decided on the device to use, enter the device number in a search engine to find the data sheet. This will show the RF parameters, the S parameters and some suppliers for the device.

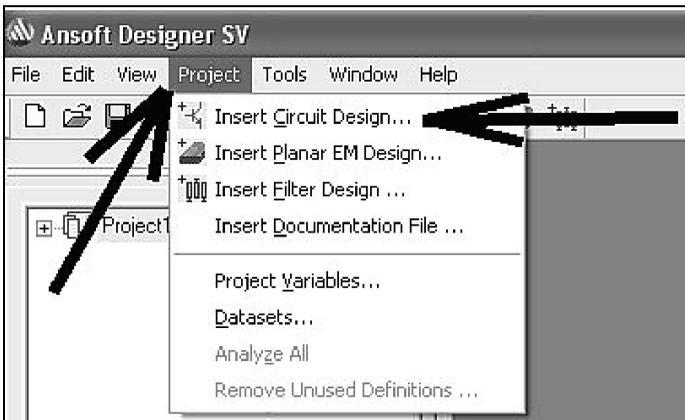


Fig 24: The project menu showing the "Insert Circuit Design" selection.

**Table 1: S-Parameter file and noise figure parameters for the ATF-33143.**

```

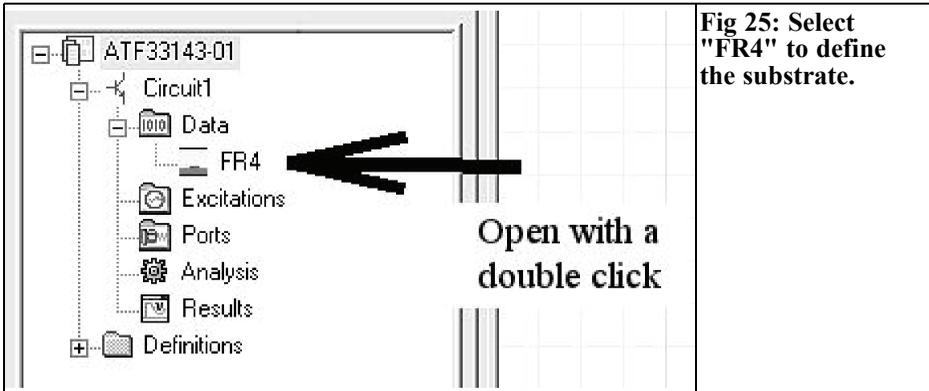
!      ATF-33143 S PARAMETERS
!      Vd=4V Id=80mA LAST UPDATED 10-15-01
# ghz s ma r 50
0.5  0.86   -76.9  14.93  132.1    0.037   55.4    0.26  -126.6
0.9  0.77   -115.9  10.77  109.1    0.051   43.9    0.34  -155.5
1    0.76   -123.2  10     104.8    0.053   42.1    0.35  -160.5
1.5  0.72   -151.7  7.18   87.4     0.063   36      0.39  -180
2    0.72   -171.1  5.51   74.3     0.071   32.1    0.43  166.6
2.5  0.72   170.1   4.45   62.6     0.077   28.1    0.45  158.7
3    0.73   157.4   3.7     52.9     0.084   24.6    0.47  151.2
4    0.74   135.9   2.79   35.4     0.099   16.4    0.49  138.7
5    0.75   116.6   2.28   17.7     0.114   5.7     0.5   124.7
6    0.77   97.6    1.91   -0.7     0.126   -6.9    0.52  108.3
7    0.79   80      1.6    -18.6    0.133   -20.6   0.54  91
8    0.82   64.5    1.35   -34.4    0.136   -32     0.57  75.3
9    0.84   50.5    1.18   -48.6    0.141   -42.1   0.61  63.1
10   0.86   36.4    1.06   -63.7    0.15    -54     0.64  51.5
11   0.88   21.6    0.94   -79.8    0.151   -67.3   0.67  38
12   0.9    7.4     0.81   -95.5    0.148   -80.2   0.71  22
13   0.91   -4.9    0.68   -110     0.138   -92     0.74  6.4
14   0.91   -15.5   0.59   -122     0.132   -100.8  0.76  -5.6
15   0.92   -27.4   0.53   -134.5   0.131   -110.8  0.79  -15.5
16   0.93   -40.5   0.48   -147.4   0.131   -121.5  0.81  -25.4
17   0.94   -52.3   0.43   -161.2   0.128   -134    0.82  -37.6
18   0.93   -61.3   0.37   -171.9   0.119   -142.9  0.84  -49.5
0.5  0.3     0.4    28.2    0.08
0.9  0.35    0.31  44.1    0.07
1    0.36    0.3   47.4    0.07
1.5  0.4     0.23  79.1    0.05
2    0.46    0.22  117     0.05
2.5  0.52    0.26  157.7   0.04
3    0.58    0.29  171.1   0.04
4    0.69    0.39  -157.2  0.044
5    0.8     0.46  -132.4  0.07
6    0.9     0.52  -109.4  0.13
7    1.02    0.57  -88.8   0.25
8    1.12    0.63  -70.5   0.42
9    1.21    0.66  -54.1   0.63
10   1.32    0.76  -40.4   0.83

```

There is a low priced HEMT FET on the market manufactured by Agilent (Hewlett Packard) which will be used for this article. The S parameter file and noise figure parameters are shown in Table 1. (I found the datasheet at: <http://www.datasheetarchive.com/search.php?q=ATF-33143> - Andy).

Note: In the first block of figures the S parameters are listed in the sequence; S11, S21, S12, S22 for the frequency range from 0.5 to 18GHz. Each indi-

vidual parameter is shown in the format "Magnitude + Angle". The second block of figures is the noise parameters. This shows the minimum possible noise figure in dB for the frequencies shown followed by the associated values of Magnitude and Angle of the source – this is the same as S11 for a "One-Port". The last figure represents the relationship between the optimal source resistance and the system resistance of 50Ω. From these figures the minimum noise figure for 23cm (1300MHz) would theoretically be a

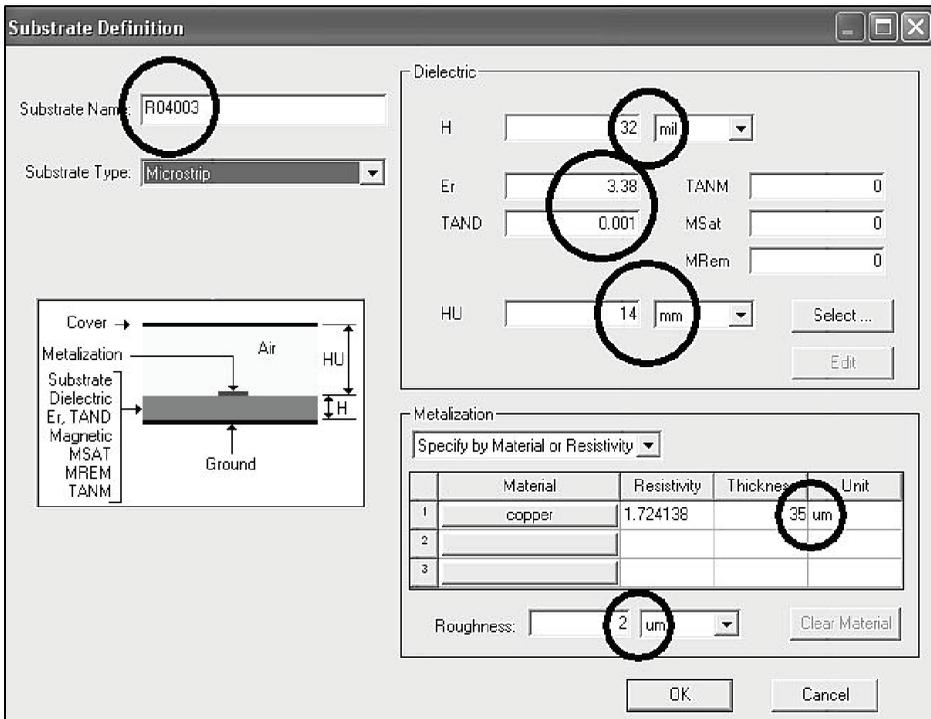


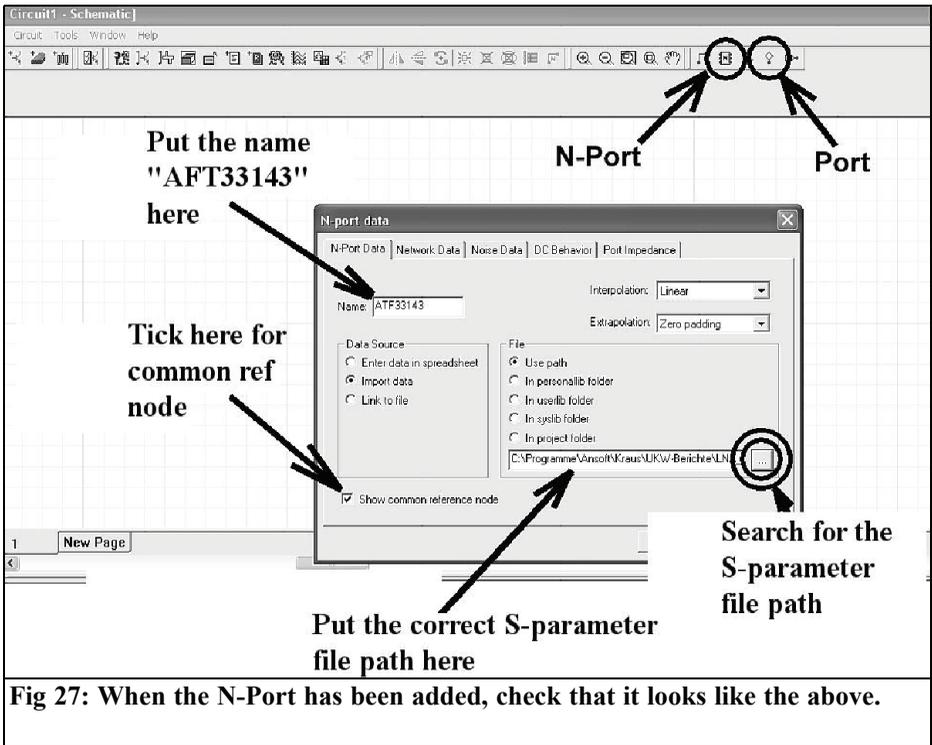
mouth watering value of 0.4dB.

### Step 2:

Start a new project and go to "Insert Circuit Diagram" as shown in Fig 24. This automatically opens the PCB mate-

rial selection list, select any Microstrip material e.g. "MS FR4 (Er=4.4) 0.030 inch, 0.5oz copper" from the upper half of the list and click "open". Save the current project in a suitable new file with a suitable name (e.g. "ATF10136\_01"). The appropriate S parameter file for the





**Fig 27: When the N-Port has been added, check that it looks like the above.**

HEMT FET from step 1 should be saved in the same place.

The finished screen is shown in Fig 26, confirm this by pressing “OK”.

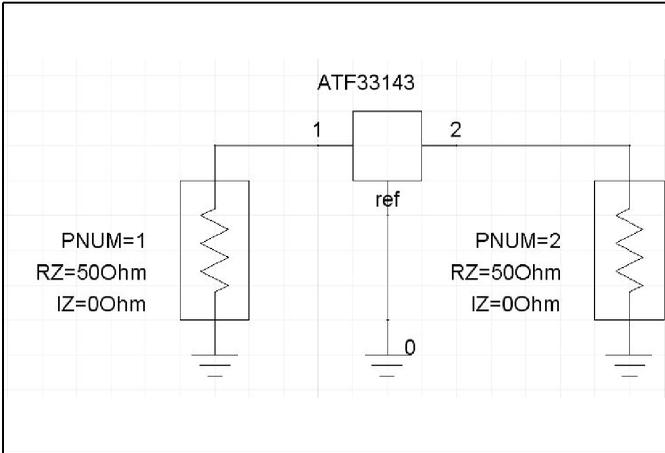
**Step 3:**

The whole menu structure is visible in “the Project manager” as shown in Fig 25. Now, using the mouse, double click on “FR4” under “Data”. This displays the substrate definition where the required changes can be made (To make the changes click on the “Edit” button):

- Metallization  
 Thickness = 35 um  
 Roughness = 2 um
- PCB material  
 PCB thickness H = 32mil  
 Dielectric constant  $\epsilon_r$  = 3.38  
 Loss factor TAND = 0.001  
 Cover distance HU = 14mm  
 Substrate name = R04003

**Step 4:**

Now the actual test circuit is set up. Start with two Microwave Ports by clicking on the Port Symbol in the upper border and setting two of them on the Editor screen. Afterwards double click on each Port Symbol and change it's properties to "Microwave Port". Then click on "N-port" (again in the upper border), give the name "AFT33143" and tick the common reference node. Open the correct path to the S-parameter file of our device and then check carefully everything by comparing your screen with Fig 27. When you close the menu then the Two-Port “hangs” on the cursor. This is now attached to the left and right port and to the Ground (you find the ground symbol between Port and N-Port symbol in the upper border). To finish double click on



**Fig 28: If all your details are correct, as shown here, you can continue with the simulation.**

the still anonymous N-Port to make it's name visible. In the Property Menu select "Property display" and press the "ADD" button followed of "OK" to see the diagram shown in Fig 28.

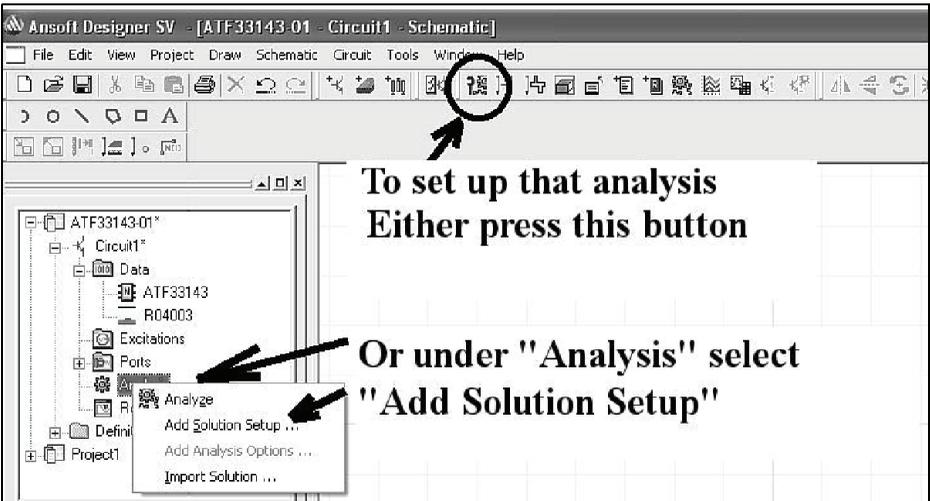
**Step 5:**

The sweep parameters are now set up using the menu as shown in Fig 29. Click "NEXT" for the entries "linear network analysis" and "frequency Domain". Click "ADD" in the next menu to program of a

linear sweeps from 0.5 to 10GHz in steps of 0.1GHz as shown in Fig 30. Click "ADD" to set the sweep values and then "OK" then press the "Analyze" button in the upper menu border to start the simulation.

**Step 6:**

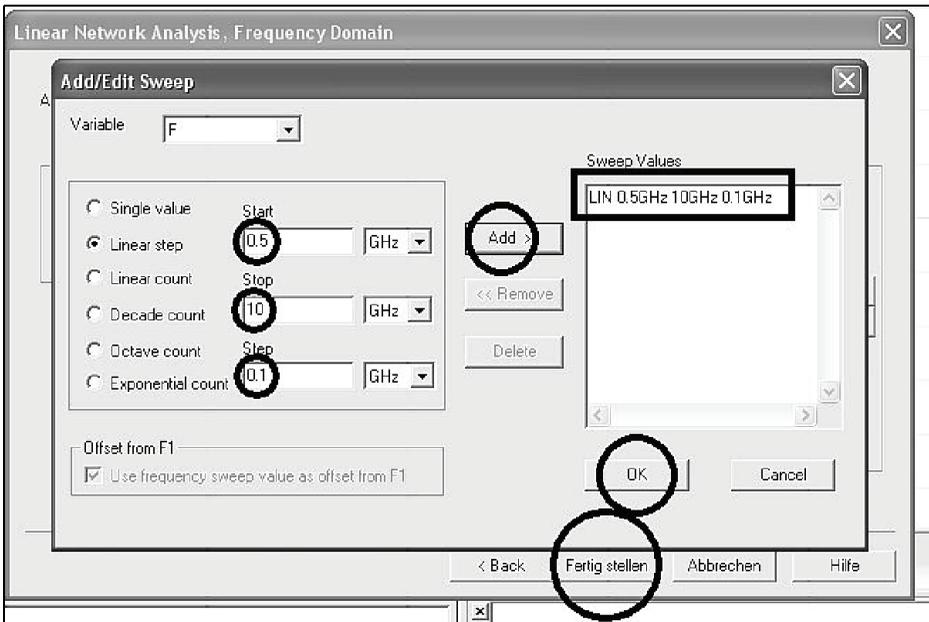
Initially the display is blank because you must specify what you want to see. Fig 31 shows how to start this process by opening the "Create report" menu, then



**To set up that analysis  
Either press this button**

**Or under "Analysis" select  
"Add Solution Setup"**

**Fig 29: Setting up the programmed sweep. This can be done in one of two ways.**

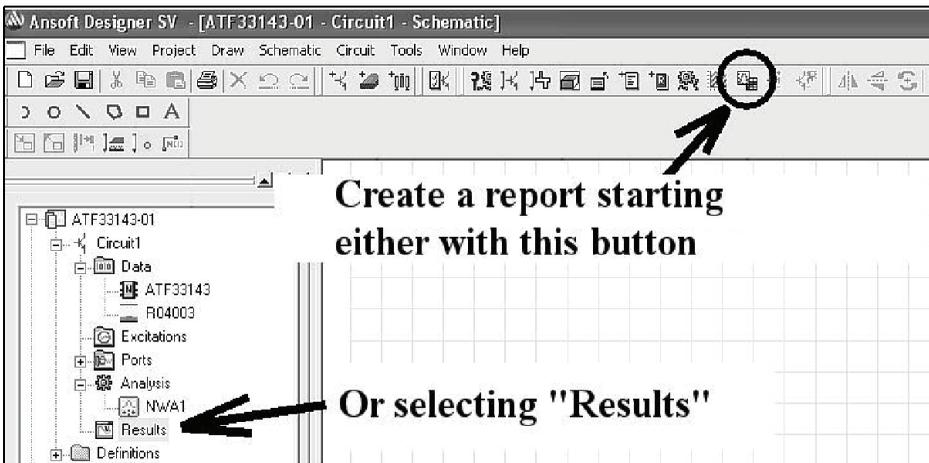


**Fig 30: The sweep parameters to set-up.**

selecting “Rectangular Plot”. Start with “S parameter” to put S11, S12, S21 and S22 into the task list. Pressing “Done” displays the results shown in Fig 32.

**Step 7:**

The noise figure and stability in the chosen frequency range can also be displayed. To do this repeat “Create report”



**Fig 31: Setting up the report, this can be done in one of two ways.**

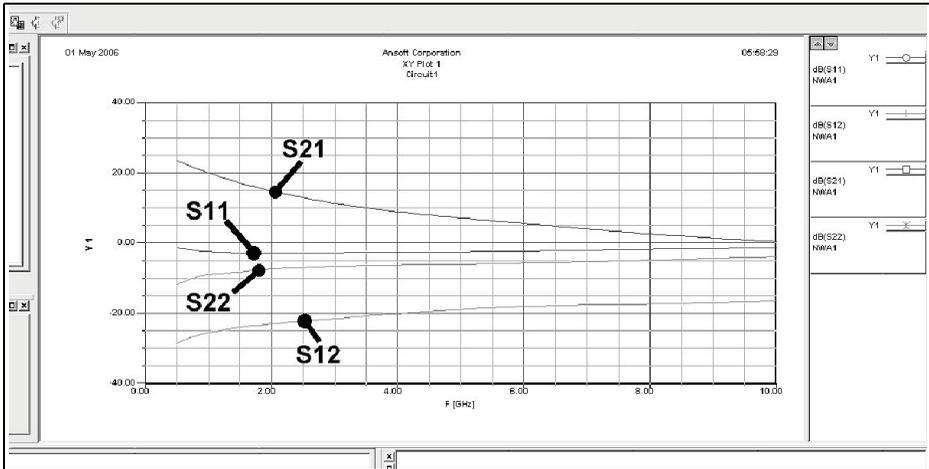


Fig 32: The report showing S11, S21, S12 and S22 from 0.5 to 10GHz.

from step 6, but select “Stability” and then “k” as an absolute value (function “abs”). Then select "Noise" from the offered column “FMIN” and “NF” in "dB" to show all in the same diagram. Double click on the numeric values for the axis and set the scaling for the Y axis as 0 to 2. The result should be as shown in Fig 33.

From these results the fundamental characteristics of the circuit can be examined for possible improvements. The noise figure is very good up to 4GHz remaining less than 1dB. However the stability factor k is below 1 and the circuit is unstable and will probably oscillate at some point.

#### 4.2. Noise Figure adjustment for 1300MHz

Some preliminary work must be carried out before work can start. The noise figure parameters from the data sheet show the source impedance needed for minimum noise figure. Now the best circuit to achieve this must be chosen.

In practice two different methods are normally used:

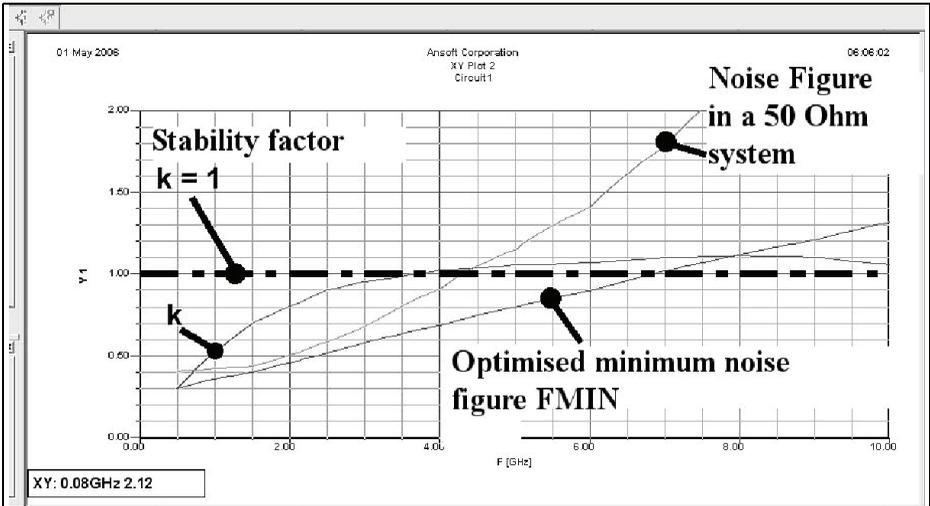
The first method uses the principle of power balancing. The rule for this is that for a complex load impedance, the internal impedance of the source is the conjugate of the load to achieve the highest power transfer from the source. In this case the impedance varies with frequency so it is a complex process to match source to the load. This is achieved using

**Table 2: Modified S-parameter file.**

```

!ATF-33143 source parameters for minimum noise
!Vd=4V Id=80mA
# ghz s ma r 50
0.5      0.4      28.2
0.9      0.31     44.1
1        0.3       47.4
1.5      0.23     79.1
2        0.22     117
2.5      0.26    157.7
3        0.29    171.1
4        0.39   -157.2
5        0.46   -132.4
6        0.52   -109.4
7        0.57   -88.8
8        0.63   -70.5
9        0.66   -54.1
10       0.76   -40.4

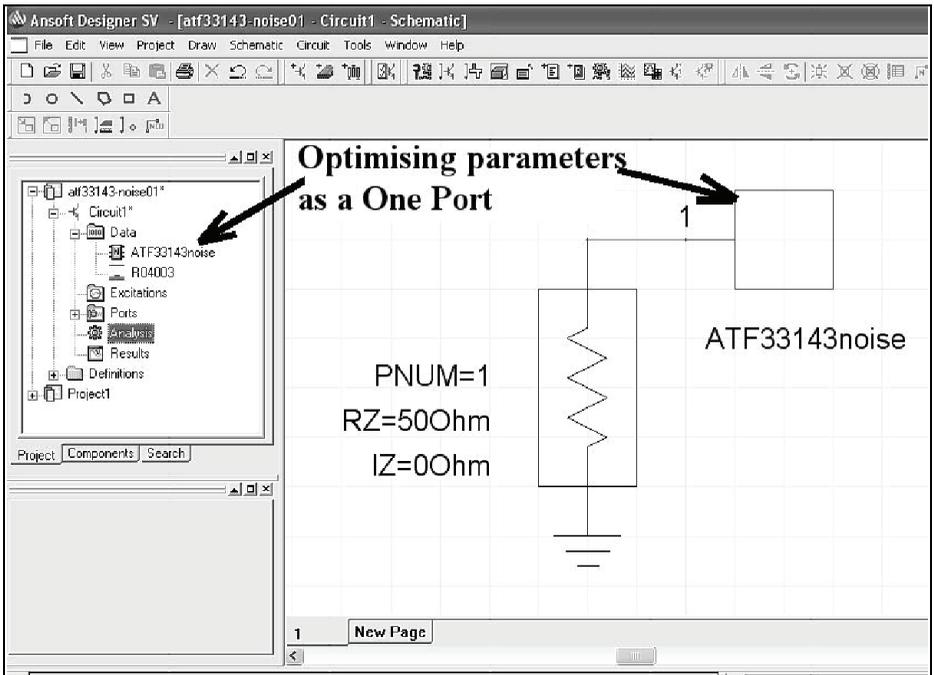
```



**Fig 33: The report showing the noise figure and stability factor.**

the S parameters from the ATF33143 data sheet to make a new S parameter file

where the imaginary part of the impedance is the opposite sign. The trick is to



**Fig 34: Circuit used to simulate for minimum noise.**

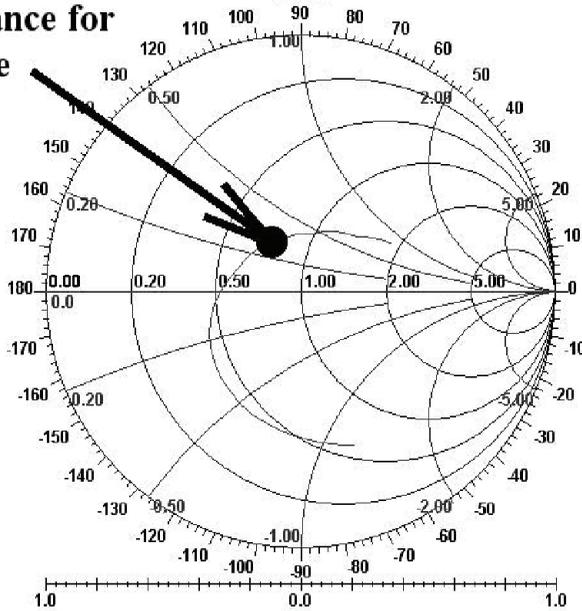


19 May 2006

21:09:46

Ansoft Corporation  
Smith Plot 1  
Circuit1

### Source impedance for minimum noise at 1300MHz



**Fig 35: The details of the matching circuit can be read off from this diagram.**

arrange that the series connection of matching circuit and HMET input must result in a “One-Ports” with a pure real input resistance of exactly 50Ω at the desired frequency of 1300MHz. In practice this means:

The matching network must be changed until the S11 curve at 1300MHz runs accurately along the centre of the Smith-chart.

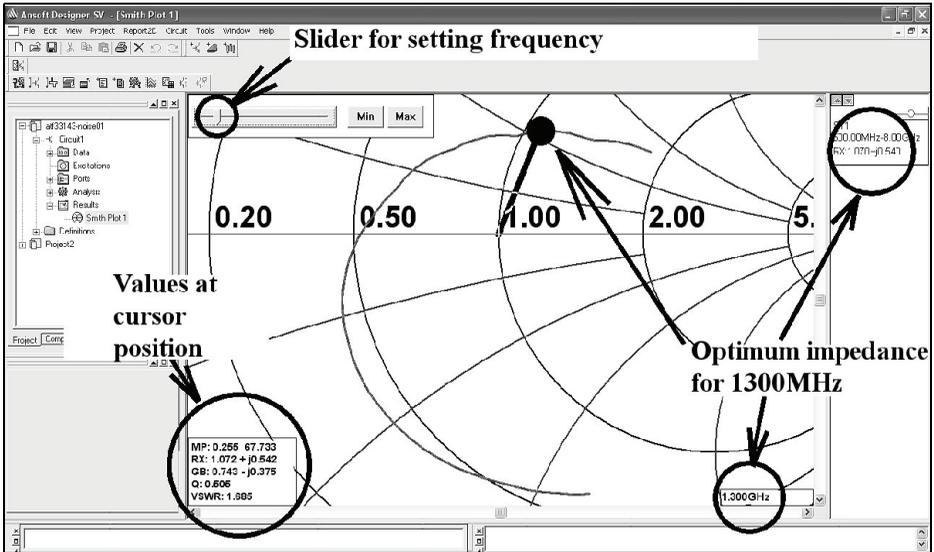
Only in this case is the input impedance 50Ω, which results in exactly the required power balancing to the 50Ω of the signal source. The author freely admits that he has used this method, but was never really inspired.

The second case is more pleasant. You assume that you are at the gate of the

HEMT and looking to the left (in the direction of the source). Now the necessary matching network is a “Two-Port” always terminated with the source resistance of 50Ω. So you can regard it as a “One-Port” and write a simple small S-parameter file which only contains the necessary S11 values.

This needs less effort: using a text editor, the S-parameter file of the AFT33143 is radically modified by deleting all of the data except the magnitude and angle values for the source impedance for minimum noise. The result is shown in Table 2 and is stored as “ATF33143noise.S1P”.

Now repeat the complete procedure from 4.1:



**Fig 36: Using the "Mark all Traces" option gives a diagram as shown.**

Open a new project, set this up for one "Microwave port" and attach the "One-Port file" "ATF33143noise.S1P" to this port. Make this circuit visible, as shown in Fig 34, before pressing the "Analyze" button.

To see the result of the analysis a report must be created. This time set up a report for S11 in the frequency range 0.5 to 10GHz as a Smithchart. This result is shown in Fig 35 and is the S11 curve required for minimum noise figure, however the indication of frequency is missing. It was not forgotten but it was not specified and the report only shows what is requested. There are several possibilities available to show the frequency:

**First possibility:**

Click on the diagram with the right mouse button and select "Zoom in" then mark a cut out of the S11 curve by pressing and dragging with the left mouse button. Click on this with the right mouse button and select "Mark all traces". Now the desired frequency can

be chosen by using the left or right cursor keys or using the slider control or using the mouse to move along the S11 curve until the frequency of 1300MHz is reached. The result is very beautifully shown in Fig 36.

**Second possibility:**

After right clicking with the mouse on the diagram and zooming as above, select "Data marker" from the menu. The impedance curve immediately changes colour and a "Tag" can be adjusted to different frequencies with the mouse. When the Tag is placed on the curve it may not be at 1300MHz, pressing "Escape" steps back to Tag adjust mode so that it can be adjusted. This is shown in Fig 37.

**Note:**

- To exit from the first example, use a right mouse click and select "Exit Marker Mode".

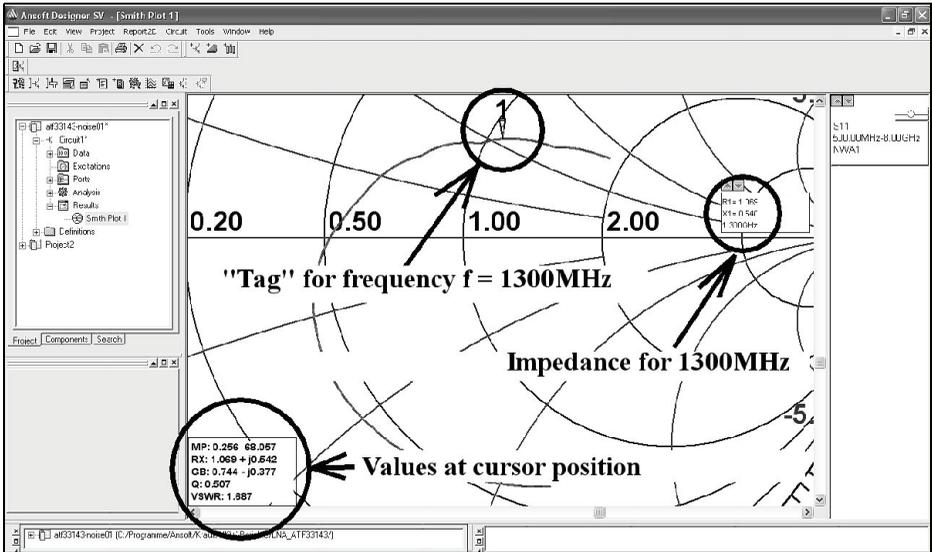


Fig 37: Using a "Tag" to mark the frequency of 1300MHz.

- To exit from the second example, use a right mouse click and select "Delete all Tags".
- To return to the original frame use a right mouse click and select "Fit all".
- A window in the lower left of the work area shows interesting values for the current position of the mouse pointer on the Smithchart curve, these are:
  - MP, the Measuring point, shown as Magnitude and Angle of S11
  - Associated standardised impedance as a series connection
  - The associated standardised conductance as a parallel connection
  - The quality factor Q
  - The voltage standing wave ratio

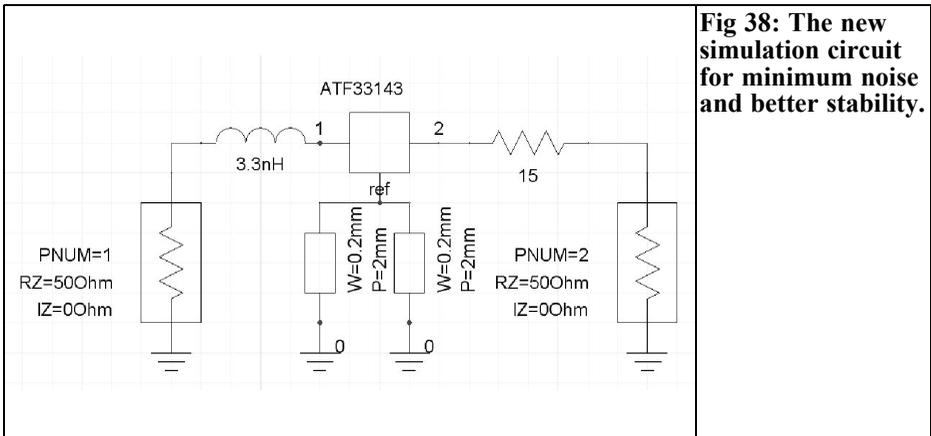


Fig 38: The new simulation circuit for minimum noise and better stability.

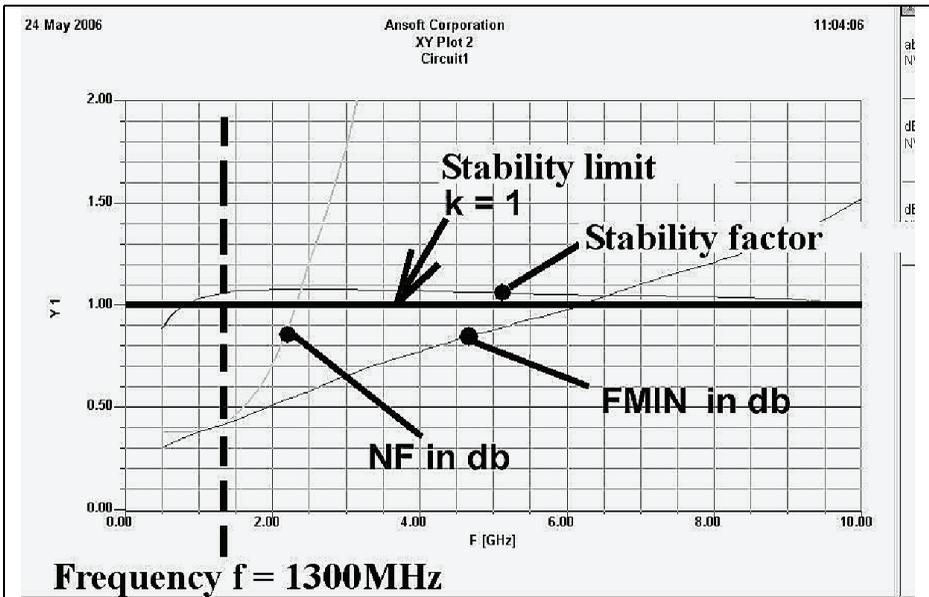


Fig 39: The result of the improvements shows the noise figure optimised for 1300MHz and stability factor > 1 down to 700MHz.

VSWR

The acid test....

The simulation is to be evaluated. The two possibilities above produce the same values at 1300MHz:

$$X_{\text{normalised}} = 1.069 + j0.54$$

multiplying by  $50\Omega$ , results in:

$$X = 53.4\Omega + j27\Omega$$

The practical conversion is very simple:

The real part of  $53.45\Omega$  required is left as the source resistance  $Z = 50\Omega$  but the imaginary part of  $j27\Omega$  at 1300MHz is converted to 3.3nH using a pocket calculator. This must be inserted between the source and the gate connection of the HEMT. (The author cannot help remarking that this conversion is much simpler when using PUFF, clicking on the equals sign for S11 would give the value 3.3nH).

4.3. Noise figure adjustment with stability improvement

There are two things that can be used to increase the stability that work satisfactorily:

- A small inductance is inserted between the source and ground. Since all microwave transistors have the source connecting on two leads it is simple to simulate two small coils in parallel. In practice it is easy to use two short pieces of Microstrip, these are part of the PCB manufacture and cost nothing. Two pieces 0.2mm wide and 2mm long have been used for this project.
- A small resistance inserted in the drain connection. A 15Ω resistor has been used for this project.

The complete simulation circuit is shown in Fig 38. This can be set up by saving the project shown in Fig 28 under a new name and then adding the new compo-

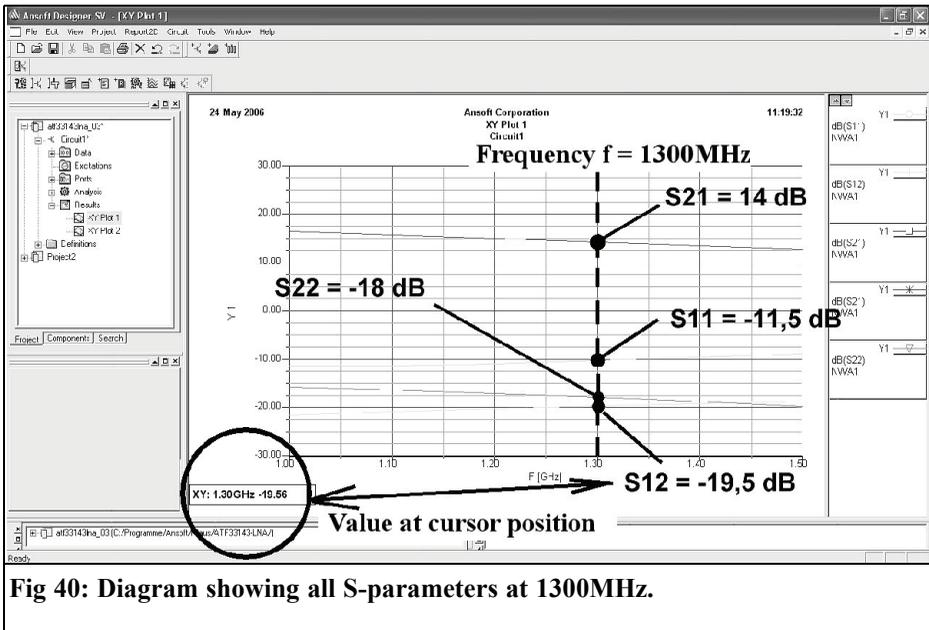


Fig 40: Diagram showing all S-parameters at 1300MHz.

nents. A Sweep from 500MHz to 10GHz shown in Fig 39 shows the stability factor, minimum possible noise figure “FMIN” and the actual noise figure “F” in dB. The noise figure is now optimised at 1300MHz and the stability factor is greater than 1 from 700MHz to 10GHz significantly reducing the possibility of oscillation.

For a better evaluation, S11 to S22 were simulated and the result zoomed in on the range between 1 and 1.5GHz. The exact values for 1300MHz are noted on the diagram shown in Fig 40 and are completely acceptable for a first draft.

How would the development continue? First the stable range would be extended, but from experience this will be paid for by a degradation of the noise figure (approximately 0.1 ... 0.2dB). The practical parts of the PCB are missing e.g. the SMD pads, feeder lines to neighbouring circuits, the steps between SMD pads and feeder lines and through plated holes to the lower ground surface etc....

Fortunately all of these are available without restrictions in the student version of ANSOFT Designer, but the simulation circuit becomes then a little cryptic. But in most cases the software rests more strong than the normal designer.

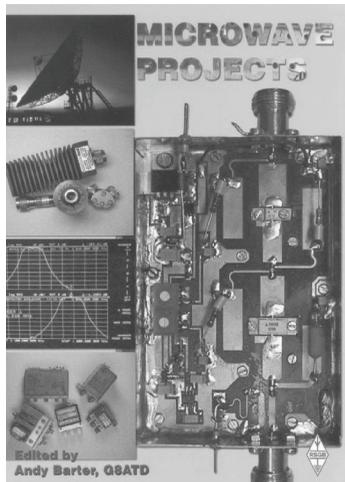
## 5. Summary

A completely new world is available with unbelievable possibilities to become acquainted with RF and microwave circuits on the computer before they are actually made. However the learning process is never ending and it is necessary to exhaust all possibilities.

But: all of these tools are completely free so it is necessary to say a hearty thank-you to ANSOFT.



## Microwave Projects Books edited by Andy Barter, G8ATD



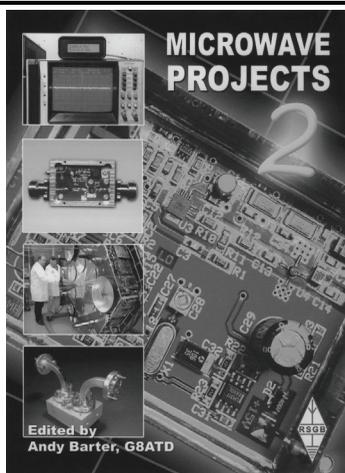
Microwave Projects is aimed at those who are interested in building equipment for the amateur radio microwave bands.

Packed full of ideas from around the world this book covers the subject with a variety of projects. The book has many contributors who have a wealth of experience in this area and they have produced many projects, design ideas, complete designs and modifications of commercial equipment, for the book.

This title provides much useful information as to what can be achieved effectively and economically. Aimed at both the relative novice and the "old hand" the book also covers useful theory of designing microwave circuit and test equipment for the projects. The book includes chapters covering:

- Signal Sources
- Power Amplifiers
- Design
- Transverters
- Test Equipment

Microwave projects is a must have book for all those who are already on the microwave bands.



Following the huge success of Microwave Projects Andy Barter, G8ATD brings you more innovative projects from around the world. If you are interested in building equipment for the amateur radio microwave bands, the designs in this book are sure to please you. Projects have been selected from international authors and all of the projects use modern techniques and up to date components. Details of how to obtain ready-made boards are included with most projects. The book includes chapters covering:

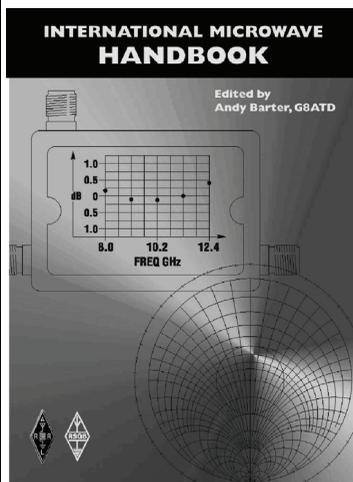
- Transverters and transmitters
- Receive amplifiers
- Filters
- Miscellaneous, including rare articles on using YIG oscillators and microwave absorbers.
- Power amplifiers

If you are interested the microwave bands or just in amateur radio construction Microwave Projects 2 provides great ideas and projects to satisfy everyone.

**Both books are available from The RSGB for £14.99 - [www.rsgb.org/shop](http://www.rsgb.org/shop)**



## The International Microwave Handbook - £24.99

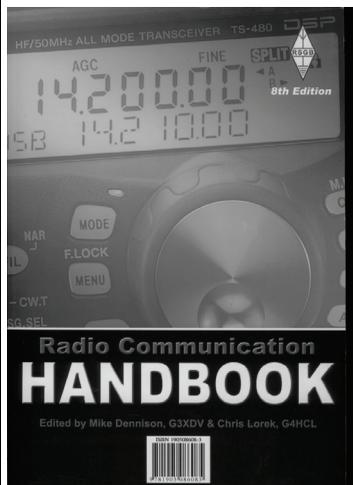


The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area, along with excellent designs from around the world to fire the imagination. Material has been drawn from many sources including the RSGB journal RadCom and the ARRL publications QST & QEX. Alongside this material a truly international range of sources have been used including items from Germany, Denmark, New Zealand, Slovenian and many more.

The earlier chapters of the book provide invaluable reference material required by all interested in this exciting area of experimentation. Techniques and devices are covered in depth, leading the reader to understand better the wide range of equipment and techniques now available to the microwave experimenter. This book contains a wide selection of designs using the latest technology that can reasonably be used by radio amateurs and ranges from ones that can be reproduced by most radio amateurs to those that require a high degree of skill to make.

With the explosion in consumer electronics using microwave frequencies the opportunity to experiment has never been greater and this book is simply the best guide to the area of microwave radio.

## The Radio Communications Handbook 8th Edition - £29.99



The 8th edition of the Radio Communication Handbook has been extensively revised with several chapters completely rewritten. This is the largest revision for many editions, and includes a great deal of new material. Once again, the RSGB has recruited experts on a wide variety of subjects to produce the most comprehensive guide to the practical side of amateur radio. Covering the entire spectrum from the basics through to advanced projects, and including many classic circuits, the Radio Communication Handbook makes an essential shack accessory.

The beginner will find much of value as the first few chapters explain the basics, leading to the chapters on specialist topics. For the experienced radio amateur there is a host of new ideas, including modern techniques such as microprocessors, surface mount components and computer aids to designing circuits and antennas.

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- 26 chapters and two appendices

If you only ever buy one book on amateur radio, this should be the one.

Both books are available from The RSGB - [www.rsgb.org/shop](http://www.rsgb.org/shop)



*Gunthard Kraus, DG8GB*

# Internet Treasure Trove

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## Microwave Antenna Book

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For anyone who wants to understand the field of microwave antennas, this has all of the basic information. It is a complete textbook with individual chapters downloadable as pdf files. Very well produced, very informative and kept up to date with many revisions. Recommended!

Address:

<http://www.qsl.net/n1bwt/contents.htm>

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## Prosig

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This site covers the topic “digital signal processing and digital signal processors”. This includes interesting and free of charge tutorials and Application Notes for applications in the audio range.

Address: <http://www.prosig.com/signal-processing/>

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## L.B. Cebik, W4RNL

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An interesting site for antenna enthusiasts. There is a lot of information on the site plus downloadable information, including a comprehensive history of antennas but the emphasis is on modern technology, in all variations.

Address:

<http://www.cebik.com/radio.html>

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## University of Idaho

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Universities specialised in certain areas, in Idaho there is “a centre for Advanced Electronics and Biomedical Research”. The “NASA symposium on VLSI Design” takes place at the university, afterwards interesting articles are made available as pdf files. There are things like “A 1.2V 2.4GHz Integrated Direct Down-conversion Receiver front-ends” or “Design of a CMOS Low Noise Amplifier (LNA) for 5.8 GHz” to read, plus more

Address:

<http://www.cambr.uidaho.edu/symposiums/>



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## Antenna software

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This homepage gives many useful links to antenna design and calculation software.

Address:  
<http://www.ac6v.com/antsoftware.htm>

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## Applied Computational Sciences

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Computer simulation of microwave circuits is indispensable for modern design work. This site has some beautiful and detailed examples on LNA design for minimum noise or improved adjustment. They are worth reading, even if you do not use the LINC2 design program used in these examples.

Address:  
<http://www.rfcircuitdesign.com/>

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## Down East Microwave

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Down East Microwave are a rare company, they not only publish a picture of the finished component and the technical data but also a photo of the PCB as well as a circuit diagram. Interesting for own work! The example is a low noise amplifier but the full catalogue has much of interest.

Example LNA article Address:  
<http://www.downeastmicrowave.com/PDF/915lna.PDF>

Full catalogue Address:  
<http://www.downeastmicrowave.com/Catalog.htm>

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## IST

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Institute for the Telecommunications Sciences, offers a large selection of information on a wide range of different topics. Worth reading!

Address:  
<http://www.its.bldrdoc.gov/pub/pubs.php>

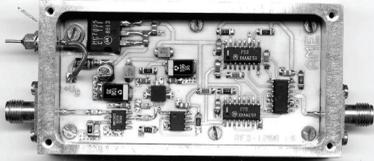
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We wish to point out that neither the compiler nor the publisher has any liability for the correctness of any details listed or for the contents of the sites referred to!



## Ready built units, Kits and PCBs available for recent projects



**12GHz Divide by 1000 prescaler by Alexander Meier, DG6RBP from issue 4/2003**

Ready built in diecast housing with SMA connectors     **£160.00**

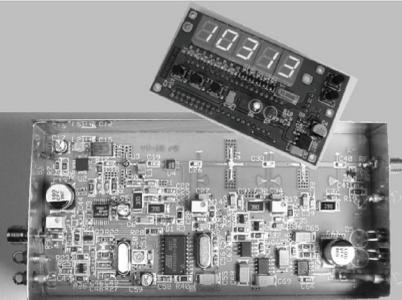
Also available an 18GHz version of the divide by 1000 prescaler     **£230.00**



**Microwave milliwattmeter by Bern Kaa, DG4RBF from issue 4/2006**

PCB     **£27.50**

Micro-controller with software     **£95.00**



**ATV transmitter for 10GHz by Alexander Meier, DG6RBP from issue 4/2005**

Ready built unit including latest improvements     **£216.30**

Ready built frequency input module     **£41.30**

Versions of this ATV transmitter are available for 13cm and 6cm



**FC4000 4GHz frequency counter by Bernd Kaa, DG4RBF from issue 1/2000**

Complete set of 4 PCBs     **£20.00**

EPROM     **£10.00**

4GHz divider module kit     **£42.00**

100MHz divider module kit     **£18.50**

Frequency counter module kit with TXCO     **£105.00**

Micro controller module kit with EPROM     **£85.00**

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K M Publications: Tel +44 (0)1582 581051, email [andy@vhfcomm.co.uk](mailto:andy@vhfcomm.co.uk)



# VHF COMMUNICATIONS

*A Publication for the Radio Amateur Worldwide*

*Especially Covering VHF, UHF and Microwaves*

Volume No.38

Winter

Edition 2006-Q4

**Publishers**

KM PUBLICATIONS,  
63 Ringwood Road, Luton,  
LU2 7BG, United Kingdom  
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Fax: +44 (0) 1582 581051

Email:  
andy@vhfcomm.co.uk

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Andy Barter G8ATD

**VHF  
COMMUNICATIONS**

The international edition of the German publication UKW-Berichte is a quarterly amateur radio magazine, especially catering for the VHF/UHF/SHF technology. It is owned and published in the United Kingdom in Spring, Summer, Autumn and Winter by KM PUBLICATIONS.

The 2006 subscription price is £20.85, or national equivalent. Individual copies are available at £5.25, or national equivalent each. Subscriptions should be addressed to the national representative shown in the next column. Orders for individual copies of the magazine, back issues, kits, binders, or any other enquiries should be addressed directly to the publishers.

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Translated using Translution translation software:  
www.translution.com

Printed in the United Kingdom by:  
Printwize, 9 Stepfield, Witham,  
Essex, CM8 3BN, UK.

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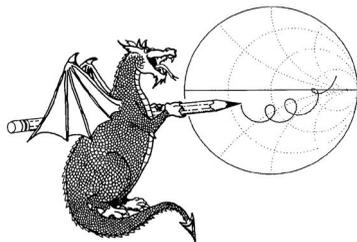
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