

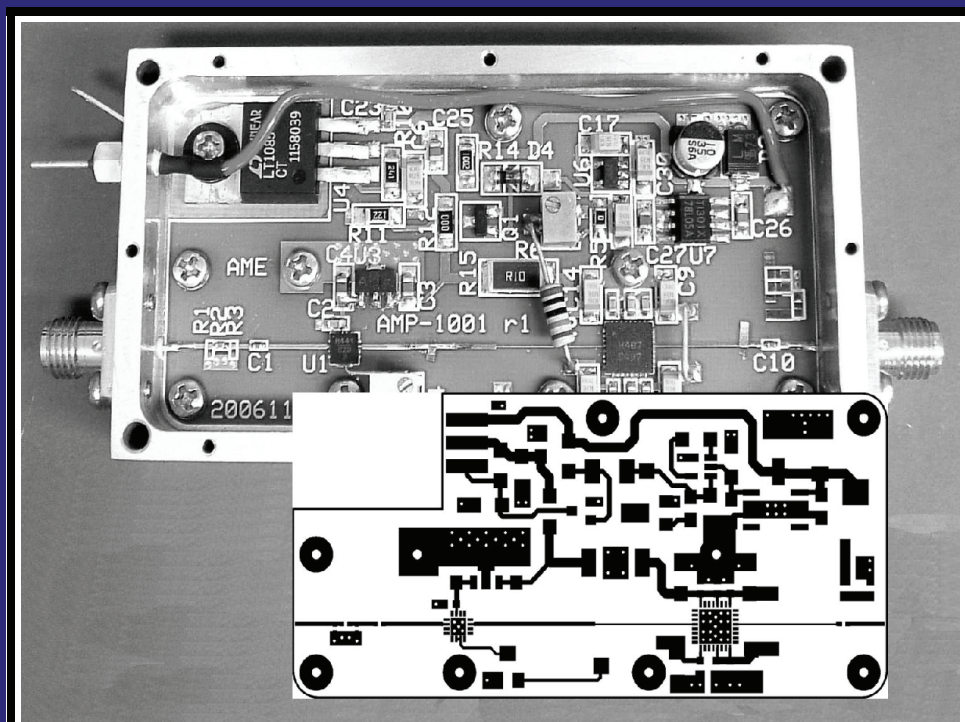


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**1 watt power amplifier for 9 to 11GHz**

*Alexander Meier, DG6RBP*

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# Contents

Alexander Meier DG6RBP	1 watt power amplifier for 9 to 11GHz	194 - 200
Wolfgang Schneider DJ8ES	2-tone audio generator for SSB transmitter measurements	201 - 206
DF4KS	Coax cable pulse generator used to measure amplifier performance	207 - 214
Gunthard Kraus	Practical project: Crystal filter recycling	215 - 233
André Jamet F9HX	Using DDS Aliasing to extend its frequency range	234 - 238
Bo Hansen OZ2M	Sequencer	239 - 241
Andy Barter G8ATD	Introduction to the JavaScript notebook	242 - 249
Gunthard Kraus DG8GB	Internet Treasure Trove	253 - 254

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



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Alexander Meier, DG6RBP

# 1 Watt power amplifier for 9 to 11GHz

Many 10GHz amateur band amplifiers were described years ago, all of these were narrow band. These amplifiers must be adjusted on the required 3cm transmit frequency. Wide band, integrated amplifiers have now been developed that can deliver high power. This article describes a wide band amplifier for the range from 9 to 11GHz with 30dB gain ( $\pm 1$ dB) and a power output of 1W. The amplifier is suitable for both the 3cm amateur band and as a laboratory-measuring amplifier because of its high gain and wide bandwidth.

---

## 1.

### Introduction

---

Wide band amplifiers are substantially more complicated and difficult to design than narrow band amplifiers that are only operated a specified frequency. However they are more universal and can be used for most diverse applications.

Matching circuits for broadband amplifiers are often tapered striplines or a gradual impedance transformation made from several quarter wave transformers around the frequency response to reduce different forms of negative feedback, like resistive shunt feedback, e.g. reduce the

gain. The transformation formulae can be computed with Tschebyscheff polynomials.

The use of an integrated amplifier that is already matched to 50 $\Omega$  at the input and output is substantially simpler. The manufacturer Hittite [1] has such an amplifier, the HMC487, for the frequency range from 9 to 12GHz. It is only 5mm<sup>2</sup> so cooling requires special attention, it must be soldered onto a printed circuit board and a large housings used. The gain is about 20dB and power output over is 1W. If the amplifier is made a two stage amplifier using an HMC441 as a driver, it makes a broadband amplifier from at least 9 to 11GHz with 30dB gain!

#### 1.1. Initial consideration

The performance curves for 25°C in the data sheet should not be used to estimate the achievable gain and power output. This is because the amplifiers convert a considerable amount of power into heat, which is the price paid for the wide band performance. The HMC441 driver gives about 14dB gain and a power output of approximately 17dBm (50mW) at 10GHz, sufficient to drive the HMC487. The latter gives about 16dB gain and 31dBm (something over 1W) power output at 10GHz at a somewhat higher temperature. The gain data is for the 1dB compression point.

For an application that requires high



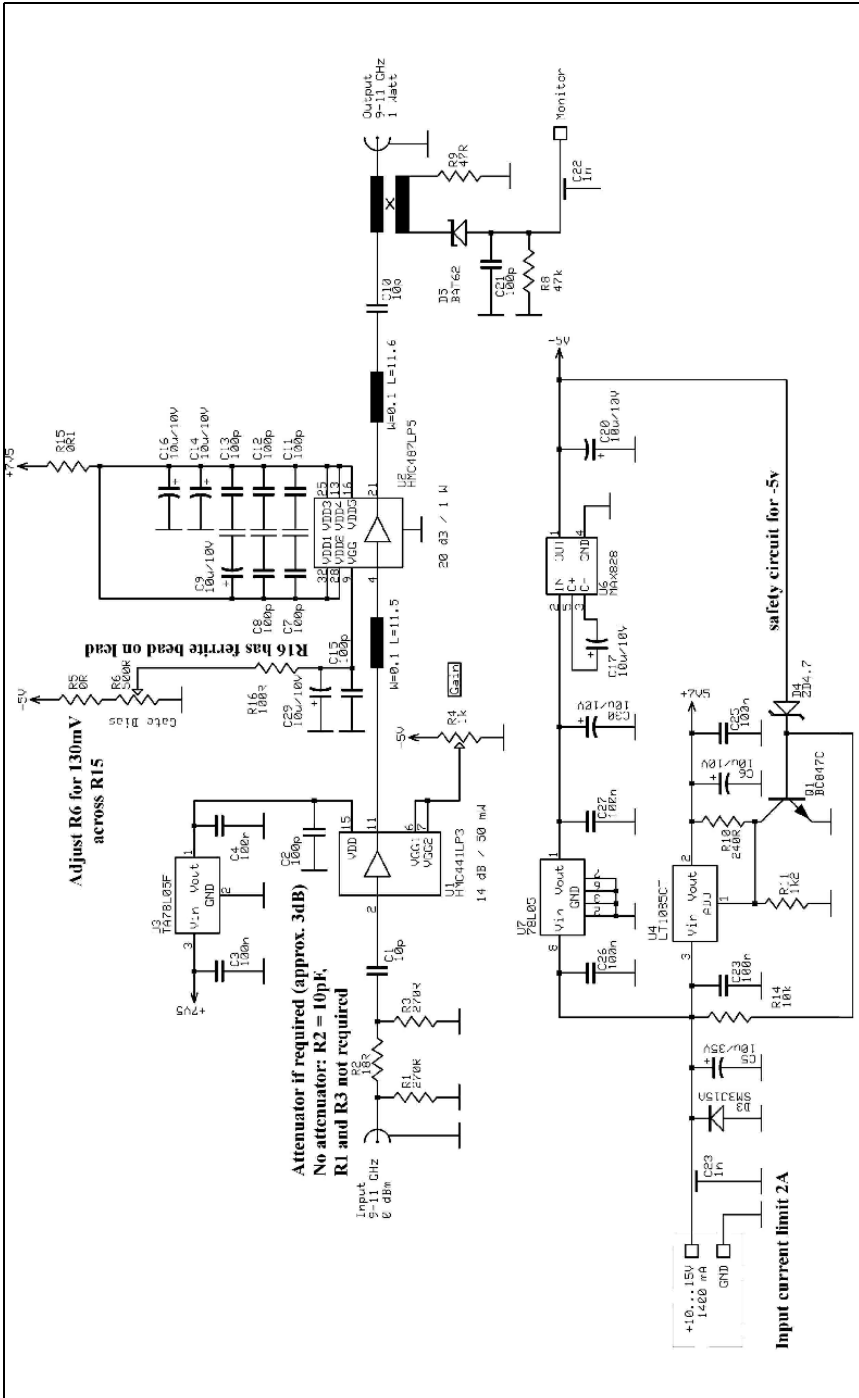
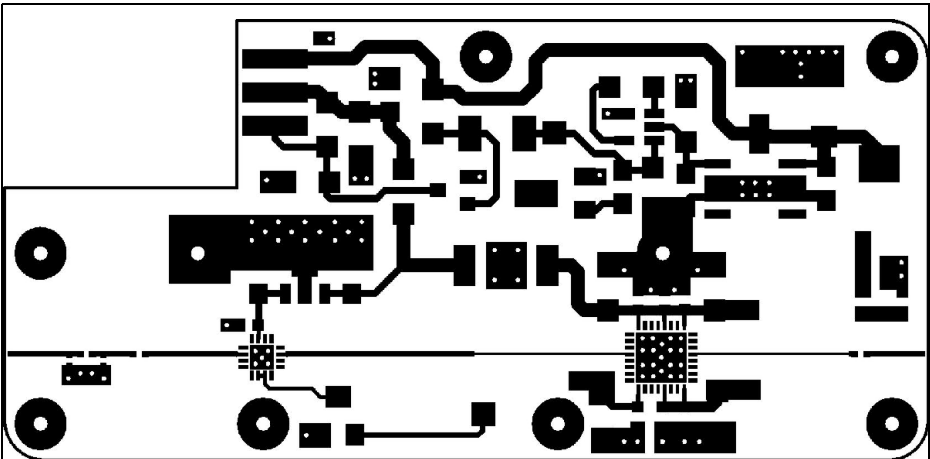


Fig 1: Circuit diagram of the two stage 1W, 9 to 11GHz amplifier.



**Fig 2: PCB layout for the 1W, 9 to 11GHz amplifier.**

linearity (e.g. SSB or D-ATV) one should remain somewhat under that for safety's sake, for FM (e.g. FM-ATV) one can go up to the saturation point, which should be between 1 and 2W. If the losses of the printed circuit board, the coupling capacitors and the transitions from the printed circuit board to the connectors are considered, the amplifier should supply an overall gain of approximately 30dB and 1W power output.

## 2.

### Circuit Description

The circuit diagram of the amplifier is shown Fig 1. There is an attenuator at the RF input made from R1 – R3, this is used when the application requires, if the attenuator is not required, simply replace R2 with a 10pF coupling capacitor and omit R1 and R3. The first amplifier stage uses an HMC441 (U1), it can operate in the frequency range between 6.5 and 13.5GHz. In the frequency range selected here (9 to 11GHz) the return loss at the input is about 15dB; which is perfectly

sufficient. The IC has its own 5v supply from U3. The trimmer R4 can be used to reduce the gain of U1 by up to approximately 5dB to be able to adjust the power output for use in certain systems. Under normal conditions this is not necessary, the trimmer was however fitted since sufficient space was available.

The second amplifier stage follows using an HMC487 (U2). The IC is matched for 50Ω, a simulation on the PC showed that this can be improved with a small impedance transformation. At 10GHz S22 is improved from the original 5dB to approximately 10dB. Striplines at the input and output are used for this transformation.

The amplifiers already contain coupling capacitors. The coupling capacitors C1 and C10 were nevertheless planned since they have a substantially higher voltage rating and so they make the amplifier somewhat more durable. The integrated capacitors can easily become defective with inappropriate use, which leads to DC voltage at the input or output! The disadvantage of the coupling capacitor at the output is that it costs about 0.2dB attenuation. The PCB also has a directional coupler to give a monitor output.

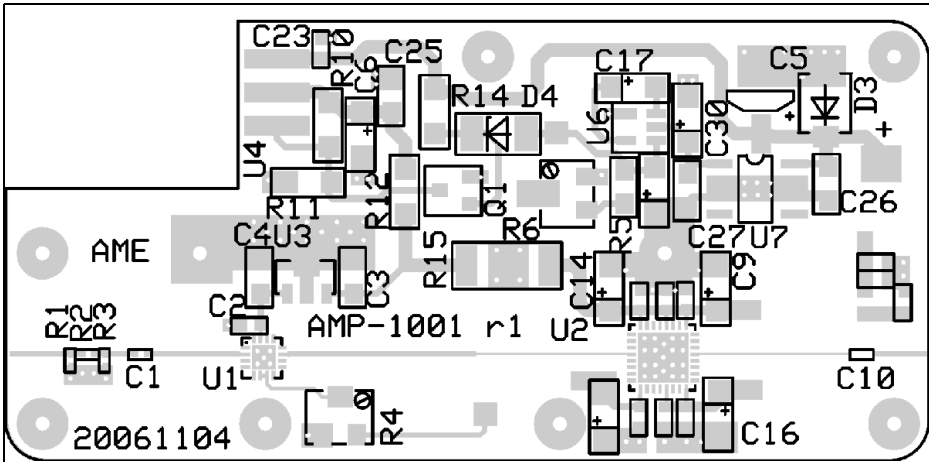


Fig 3: Component layout for the 1W, 9 to 11GHz amplifier.

The output stage has a small resistor of  $0.1\Omega$  (R15) to measure the supply current ( $100\text{mV/A}$ ). The bias supply to the gate of U2 is fed via a  $100\Omega$  resistor (R16) with a ferrite bead on the lead, this prevents oscillation at the gate. The trimmer R6 adjusts the operating point of U2 by adjusting the negative voltage on the gate.

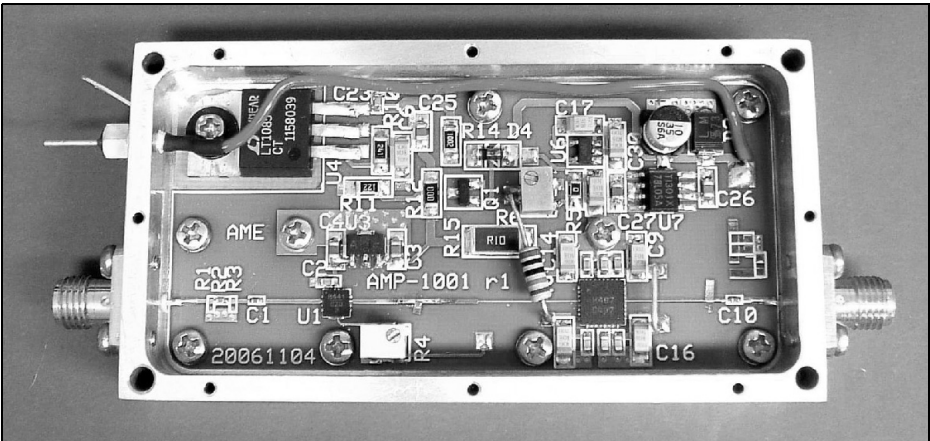
The external supply voltage (+10 to 15v) is fed via the feedthrough capacitor C23. An SMB J15A diode (D3) protects the amplifier against voltage peaks (transient ones). The supply for the output stage (U2) comes from the LT1085CT adjustable voltage regulator (U4). R10 and R11 set the supply voltage to 7.5v. The transistor Q1 protects the output stage if there is no negative gate bias, the supply for U4 is reduced to 1.25v. The voltage regulator U7 in conjunction with the charge pump U6 produces the negative gate bias. The MAX828 from Maxim [3] supplies the higher output currents needed for the gate of the output stage better than the familiar 7660.

### 3.

### Construction and alignment

The amplifier was developed on the proven microwave substrate RO-4003C from ROGERS [2]. The printed circuit board thickness 0.203mm was selected. This thin material makes very narrow  $50\Omega$  track widths possible and thus a better transition to the pads of the amplifier ICs. The thin material also has a thermal advantage because the high heat dissipation only has to be conducted through a thin board. The PCB layout is shown in Fig 2 and the component layout in Fig 3.

The PCB must be plated through under the output IC in order to dissipate the approximately 9W of heat energy. The IC (U2) is soldered to the PCB using flux and a hot air gun from the earth side of the PCB. The plated through holes should run freely with solder because this is the path to dissipate the heat, see Fig 4. Excess solder on the earth side of the PCB should be removed so that it does not prevent the PCB from laying flat.



**Fig 4: Picture of the completed PCB for the 1W, 9 to 11GHz amplifier a milled aluminium housing.**

The driver U1 is soldered in a similar manner with hot air. With this IC the plated through holes must be not completely filled, since only a small heat dissipation needs to be removed.

The entire PCB is soldered to a 2mm thick brass plate. For this, use an alloy with a lower melting point (Sn42Bi58) and a heating plate. The soldering procedure is carried out at 150°C. The brass plate gives faster heat distribution and thus better heat dissipation from the milled aluminium housing. Alternatively a milled brass housing can be used saving the transition to the brass plate just solder the PCB directly to the housing. However this is somewhat more expensive to manufacture, also the surface treatment of a brass housing e.g. gold, is substantially more expensive than a suitable plating for aluminium (e.g. chrome 6).

All further components, except R15 and R16, can now be fitted on the PCB. The supply voltage to the output stage U2 is made by a small cable link (see Fig. 4). Likewise a cable link must be used for the -5v supply of R4. C10 must be a high quality microwave capacitor [5], otherwise the desired gain and power output will not be achieved.

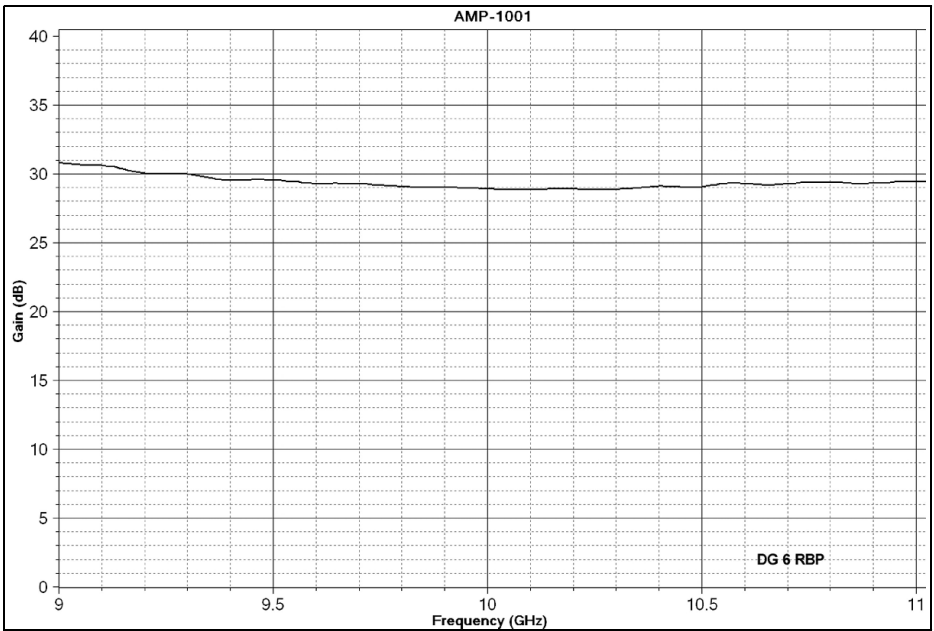
For the alignment procedure use a supply with a current limit of 2A to protect the amplifier in the event of an error. Connect the supply voltage and check all voltage levels (7.5v, 5v, approximately -5v). Set a voltage of -1.8v at the connection to R16 using the trimmer R6. Switch off and fit R15 and R16. Switch the amplifier on and set the quiescent current of 1.3A (130mVs across R15) using R6. The amplification is checked using a network analyser. Small tags soldered onto the strip lines can be used to adjust the frequency response. Thus the amplifier is now ready for use.

The heatsink for the output stage should be generously rated (approximately 1W/K).

#### 4.

#### Parts list

D3	SMBJ15A
D4	ZD 4.7
D5	BAT-62 o.a.
Q1	BC847C
U1	HMC441LP3 [1]
U2	HMC487LP5 [1]



**Fig 5: The frequency response of the 1W, 9 to 11GHz amplifier. The gain only fluctuates from 30dB by 1dB.**

U3	TA78L05F	C3, C4, C23, 100nF, 0805
U4	LT1085CT	C25-C27
U7	78L05	C6, C9, C14, 10µF/10v tantalum SMD A
U6	MAX828EUK [3]	C16, C17, C20, C29, C30
R1, R2, R3	attenuator if needed, SMD 0402	C5
R4	1kΩ trimmer, Bourns 3214W	C22, C23
R5	0Ω 0805	1 x
R6	500Ω trimmers, Bourns 3214W	1 x
R16	100Ω fitted with ferrite bead	1 x
R15	0.1Ω, 2512	2 x
R9	47Ω 0603	
R8	47kΩ of 0603	
R14	10kΩ of 1206	
R11	1.2kΩ of 1206	
R10	240Ω 1206	
R12	0Ω 1206	
C1, C19	10pF, 0402, ATC [5]	
C2, C7-8,	100pF, 0603	
C11-13, C15,		
C21		
		10µF/35v SMD Electrolytic
		1nF Feedthrough capacitor, M3
		PCB from RO-4003C
		brass plate 2mm, milled
		aluminium milling housing
		SMA sockets with Micros trip tags

**5.**

**Measured values**

Fig. 5 shows the frequency response of the amplifier. The gain of 30dB was reached and varies only around ±1dB!



The amplification can be increased by alignment with a tag to a maximum in the 10GHz region to approximately 34dB. Then there are larger fluctuations over the entire frequency range (9 to 11GHz). Power output (1dB compression) over the entire frequency range is somewhat more than 1W (typically 1.2W - 1.3W). The saturation output is about 1.5W.

The second harmonic at 20GHz could not be measured on the spectrum analyser. The efficiency as PAE (power added efficiency) is 9%. Thus the desired values for gain and power output required were achieved.

[2] Rogers corporation, 100 S. Roosevelt Ave, Chandler, AZ85226

[3] Maxim Integrated Products, Inc., 120 San Gabriel drive, Sunnyvale CA94086

[4] Hittite Application note "Thermal management For Surface Mounted Devices", Hittite Microwave corporation

[5] American Technical Ceramics, One the north Lane, Huntington station, NY 11746

6.

Literature

[1] Hittite Microwave corporation, 20 alpha Road, Chelmsford, MA01824

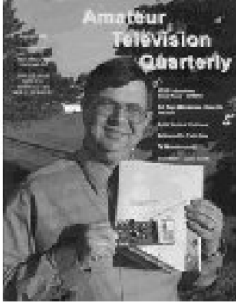
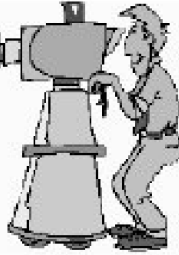
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Wolfgang Schneider, DJ8ES

# 2-tone audio generator for SSB transmitter measurements

The quality of SSB transmitters is usually determined by a 2-tone measurement. Relevant signal generators produce the two necessary tones and/or frequencies either at AF or at a high frequency. The latter was described in detailed in [2]. The following article describes the second option, a simple 2-tone audio generator.

For an initial functional test of an SSB transmitter a microphone attached to the input is sufficient. In order to make more exact and comparable measurements of range and intermodulation behaviour etc., a 2-tone generator is needed.

## 1.

### Description of the circuit

The concept of the audio frequency generator presented here (Fig 1) comes from American literature [1]. It produces two different audio frequencies, one at 750Hz, the other at approximately 1600Hz. These are not harmonically related.

Each signal is produced with a Wien bridge circuit (IC1 and IC3). Such generators always produce harmonics, in each case a low pass filter (IC2, IC4) is

placed on the output.

Two matched pairs of diodes provide the same output levels for each audio frequency. The diodes are selected from a quantity of diodes using a voltmeter to measure the apron and select ones within 10mV.

The single tones are summed in IC5 (LF356). The following amplifier IC6 (LF356) buffers the oscillators from the unit under test to prevent any reaction. The precision trimmer R31 (10kΩ) can be adjusted as necessary for a maximum output level of 50mVpp.

## 2.

### Construction

#### 2.1. Parts list

IC1 - IC6	LF356, operation amplifier
D1 - D4	1N4148, diode selected (see text)
9 x	Electrolytic capacitors 100µF/25v, RM 2.5 mm
3 x	2 pole socket strip also and jumpers
1 x	solder pin 1 mm
1 x	solder lugs 3 mm
1 x	PCB DJ8ES-075

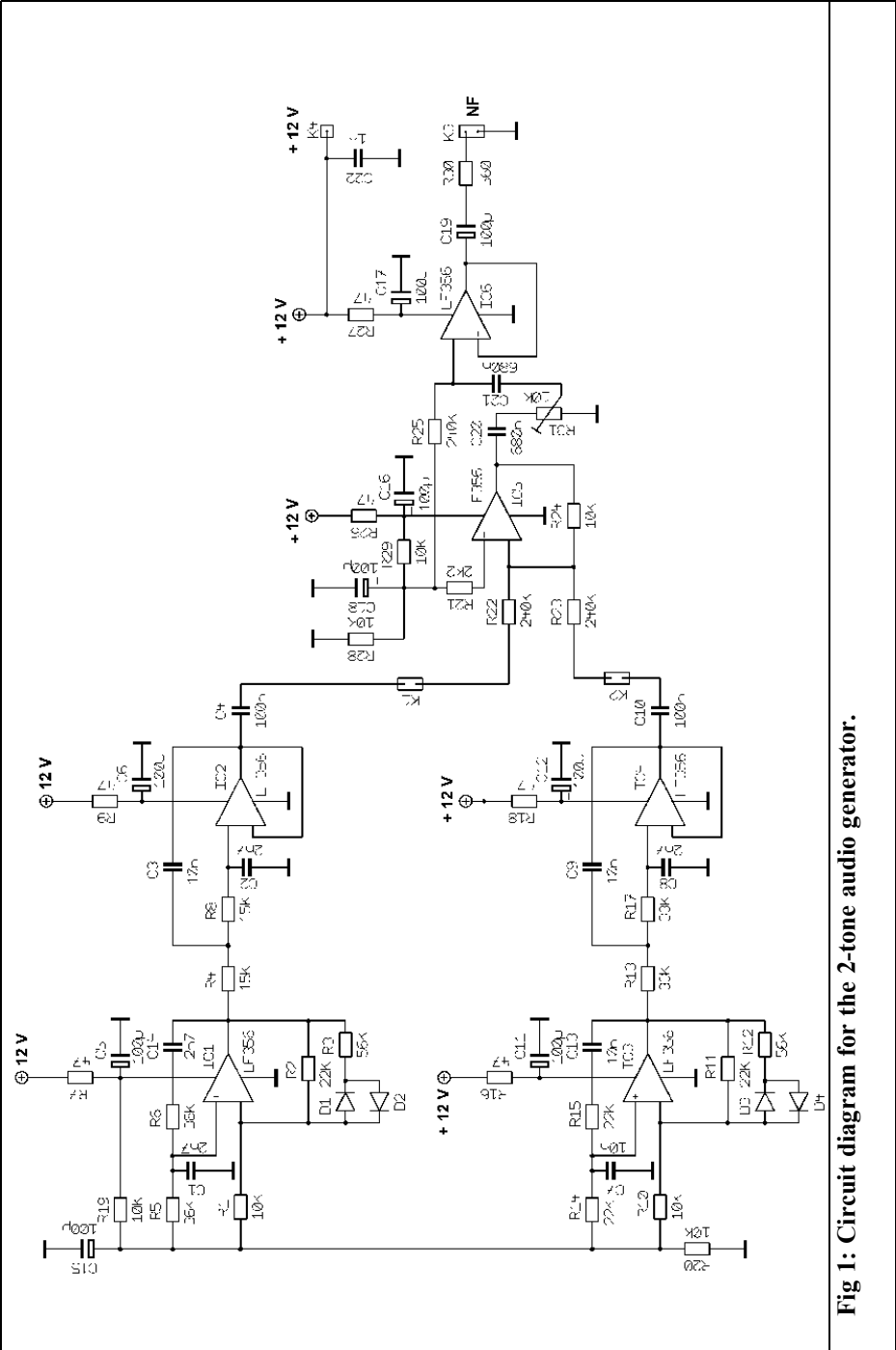
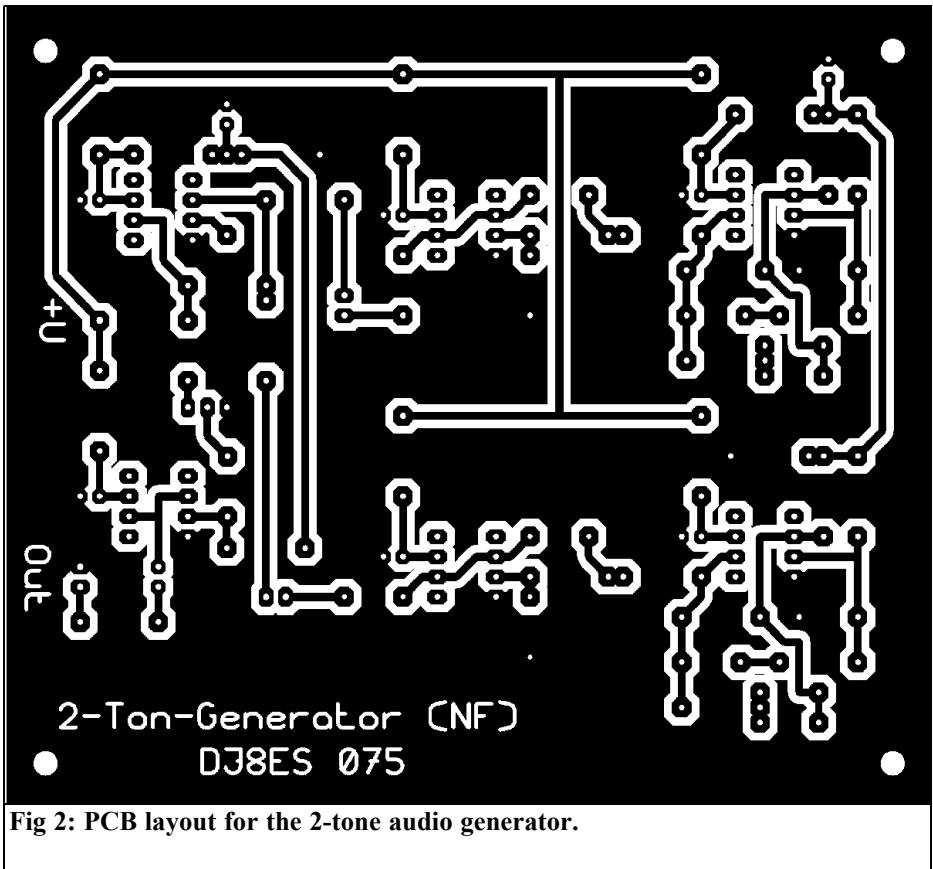


Fig 1: Circuit diagram for the 2-tone audio generator.





**Fig 2: PCB layout for the 2-tone audio generator.**

Capacitors MKH, RM 7.5 mm

1 x	1nF
4 x	2.7nF
4 x	10nF
2 x	100nF
2 x	680nF

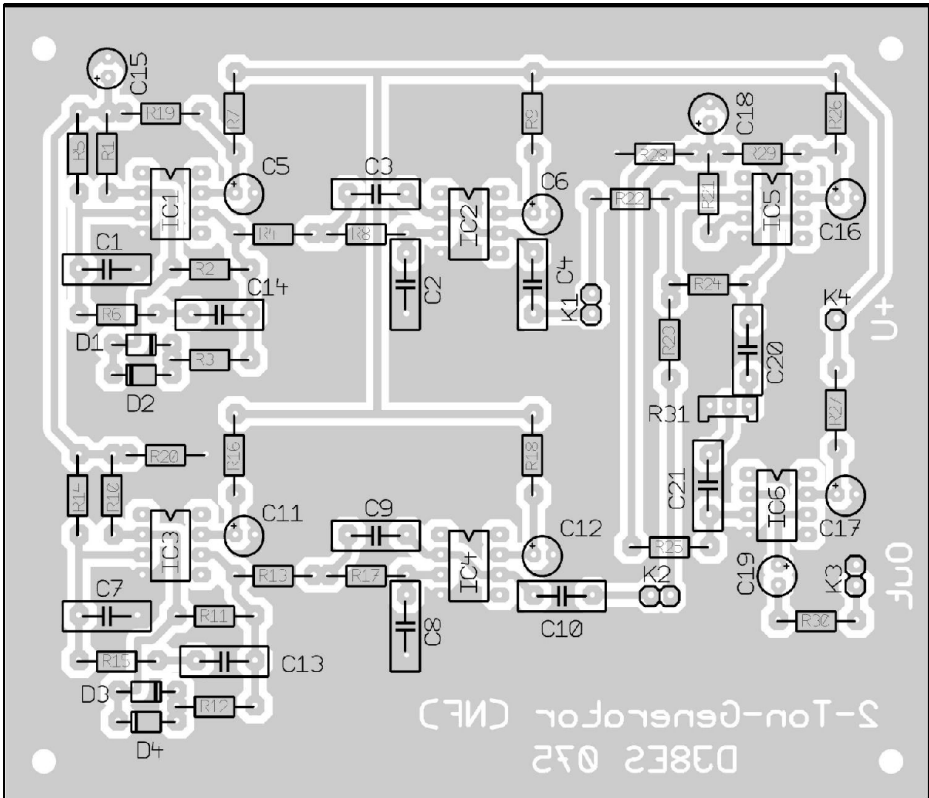
Resistors 1/4W, RM 10mm

6 x	47 $\Omega$
1 x	560 $\Omega$
1 x	2.2k $\Omega$
8 x	10k $\Omega$
2 x	15k $\Omega$
4 x	22k $\Omega$
2 x	33k $\Omega$
2 x	36k $\Omega$
2 x	56k $\Omega$
3 x	240k $\Omega$

### 3.

#### Alignment

After assembly of the 2-tone audio generator (DJ8ES-075) is complete, carry out a visual inspection before it is switched on for the first time. The power input required is 12v at 25mA. Jumpers can connect the two individual tones. Although the output signal looks very clean on an oscilloscope (Fig 4), a more precise measurement of the harmonics is available. These become visible using a spectrographic analyser (Fig 5). The harmonics are suppressed by at least 50dB. Mixing products are not measurable.



**Fig 3: Component layout for the 2-tone audio generator.**

The maximum output voltage of 50mVpp, is adjustable using the precision trimmer R31 (10k $\Omega$ ).

#### 4.

### Practical 2-tone measurement of an SSB transmitter

The 2-tone generator is connected to the microphone input of the SSB of transmitter to be tested. Use the precision trimmer R31 (10k $\Omega$ ) to adjust the necessary levels. The transmitter can be tested stage by stage.

Some interesting software to display a diagram of the frequency spectrum on a PC or a laptop is available as freeware to the download from the Internet: Spectran V2 - FFT software for signal analysis on a PC using a sound card (<http://www.weaksignals.com/>). Of course there are many other possibilities. A search on the Internet will result in numerous hits about FFT software.

Since the sound display on the PC/laptop is only in the AF frequency range, in our case up to a maximum of 25kHz, the frequency band that can be measured (e.g. 10.7MHz) must be mixed to AF (called the baseband). Such a baseband converter, including the display on a PC or laptop was described in detail in [3].

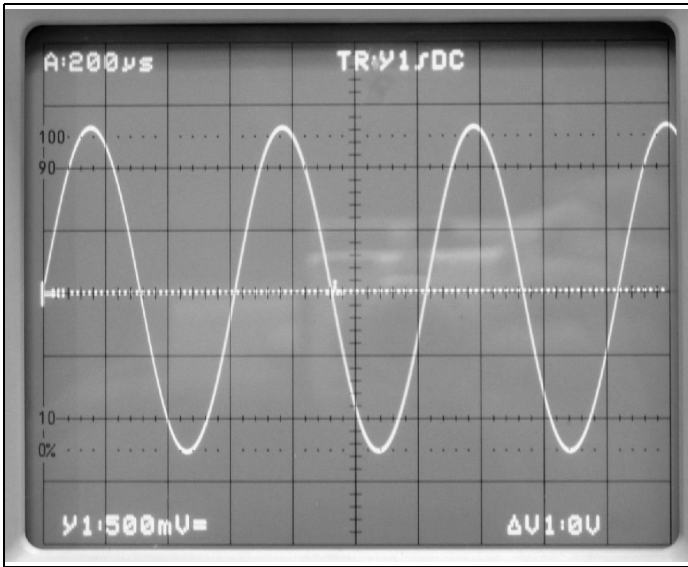


Fig 4: Clean sine wave output on the oscilloscope.

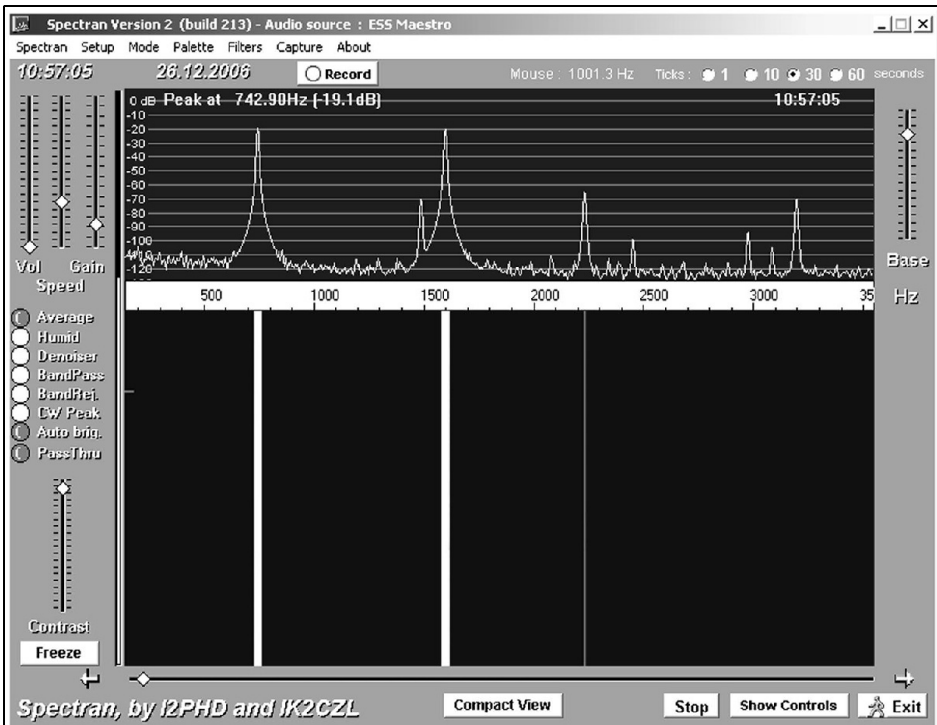
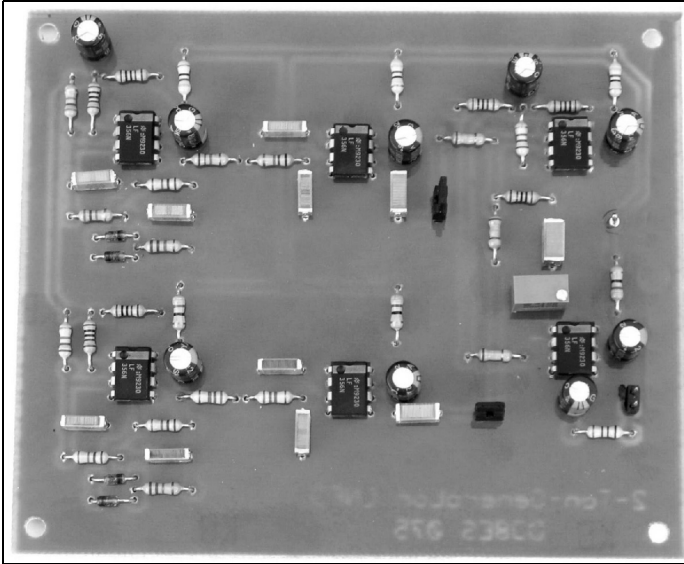


Fig 5: A spectrum display of the summed output signal showing that all mix products are less than 50dB.



**Fig 6: The finished 2-tone audio oscillator.**

**5.**

**Rerences**

[1] Hayward, Campbell and Larkin: Experimental Methods in RF Design ARRL

[2] Wolfgang Schneider, DJ8ES: 2-Tone Generator for 145MHz, VHF Communications Magazine 4/2002, pp 216 - 227

[3] Wolfgang Schneider, DJ8ES: Base-band converter for spectrographic analysis with FFT, Funkamateur 9/2005, pp 913 - 915



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DF4KS

# Coax cable pulse generator used to measure amplifier performance

How can the transmission characteristics of a broadband amplifier be assessed using a simple technique?

---

1.

## Introduction

---

I normally divide people who know about electronics into two groups: The first group is usually the younger group, who look at a piece of equipment, look at the back and ask, "Where is the interface"? The second group is the older group who remember using valves and when things are broken tend to complain. I don't consider myself as part of either group but both groups do seem to follow in their own footsteps and some techniques disappear from common use. Because I have modest means I like to use simple techniques and like to remember things that I learned in my education. I am sure to make a mistake but I hope this article will help the intellectual well being of everyone [1].

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2.

## Portrayal of a developing technology

---

The subject of technology can be approached in two ways: The first way is a practical way, using Shockley's equation as a creed, the second way is measuring the characteristic curves of the electronic devices. In addition there are some techniques for special cases, for instance, the linear characteristics for a given operating point can be expressed as a Taylor Series. A classical work on this technique is [2], this is a near classical work known to many German engineers but edition 11 contained many spelling mistakes so it set a revolutionary standard for spelling. Another way has far more theoretical beginnings and deductive methods. The starting point is the quantum mechanics model; many semiconductor properties can be explained using the shell model of atoms [3], where the outside shell of germanium and silicon with its four electrons, plays a crucial role. This fulfils the theory in detail and is more laborious to realise but gives a much better explanation of the effects. This technique requires a starting point set by expert knowledge followed by further analysis. In my experience this knowledge requires two or three years of academic



experience to gain.

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### 3.

## Trip into physics

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The operation of the pulse generator can be described by looking at the information described in the following paragraphs.

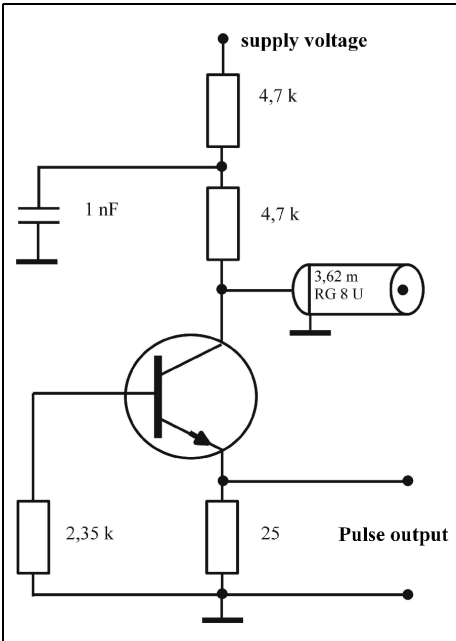
A main characteristic of an un-doped semiconductor, which differentiates insulators from conductors, is the large dependence of its resistivity on temperature. This characteristic alone is not sufficient for a perfect demarcation. A further criterion is a negative temperature coefficient of resistance, which is different to metals. Rising temperature breaks more electron connections and thus pairs of electrons and holes are produced, which contributes to the electrical conductivity.

This is somewhat different for doped semiconductor. The doping atoms possess a smaller ionization energy, which gives the same temperature behaviour but the conductivity is higher at low temperatures than for the un-doped semiconductor. The temperature range where not all doping atoms are ionized is called the impurity reserve. At mid temperatures, limited to  $-50^{\circ}\text{C}$  to  $+150^{\circ}\text{C}$ , almost all doping atoms are ionised. The conductivity is fairly constant but with rising temperature, increasing lattice vibrations and falling mobility of the charge carriers causes it to decrease. This reducing mobility of charge carriers is responsible for the positive temperature coefficient of metals; therefore semiconductors have metallic behaviour. This temperature range is called the range of the impurity exhaustion and is the range planned for use in electronic components. At higher temperatures intrinsic conduction occurs, as for the un-doped semiconductor, electron-hole pairs are produced. This

intrinsic conduction is of causal importance for the pulse generator being described. Intrinsic conduction (pin diodes have a third un-doped layer between the n doped layer and the p doped layer called the intrinsic or I layer) is not enough to detach whole atoms from the crystal lattice, which would be tantamount to melting. As long as the energy flowing into the crystal is limited by some means, the procedure of the thermal electron hole generation is reversible. If an electron arrives into the proximity of a defective connection (hole), it can take the place of the missing electron to release energy. Thus both the electron and the hole that it replaced are no longer available - electron and hole are recombined. Charge carrier production is therefore electron - hole pair generation and recombination represent a thermodynamic equilibrium. This is the influence of the temperature on the conductivity of homogeneous semiconductors.

How does this affect a p-n junction or inhomogeneous semiconductor? Operating in reverse bias a temperature dependent cutoff current flows. If the reverse bias voltage is multiplied by this current it results in a certain energy dissipation. Assuming that thermal resistances of the experimental assembly are constant, it can be equated to a certain energy dissipation of a certain junction temperature. A small reverse voltage increase causes a higher temperature, thus the cutoff current increases. Either the new operating point is stable or the characteristic has a negative upward gradient. Additional details are given in [3], chapters 7.9.1 and [4], chapter 2.9. This effect is called thermal break-through and affects the whole p-n junction. The pertinent characteristic is called quasi-static, i.e. determined very slowly in comparison to the thermal time constants.

As well as the thermal break-through just described there is a further break-through which is based on temperature effects, which is generally called "secondary break-through" or "secondary break-



**Fig 1: Circuit diagram of a coax cable pulse generator. It even runs with the old fashioned 2N3055.**

down”. Note: Avalanche or avalanche breakdown is used in Zener diodes from approximately 5 to 10v. There are no structures in a diode or transistor, where the doping is evenly distributed giving a basic path resistance of zero. In extreme load situations this can lead to local overheating due to position dependent current density where a positive feedback takes effect. A higher temperature leads to a higher charge carrier concentration (intrinsic conduction), which leads to increased conductivity giving further increase of the energy dissipation in the places where the temperature rises. This is a pinch effect producing “hot spots”. The following analogy explains this inhomogeneous current density: in an efficient power supply “n” diodes are fitted to “n” heatsinks and are connected in parallel. Each diode has its individual forward bias voltage therefore one diode will exhibit the smallest forward bias. This diode will draw the largest portion

to total current, so it becomes warmer, whereupon the forward bias continues to fall....

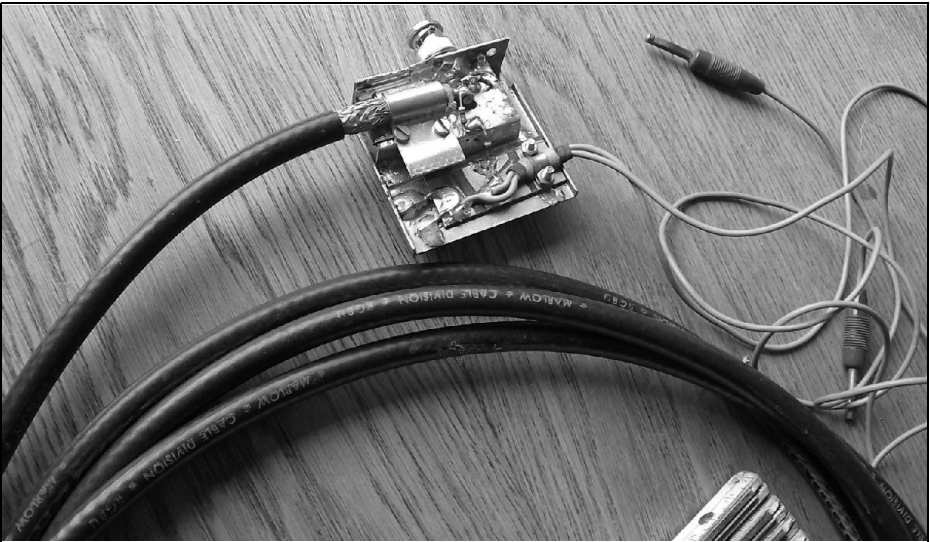
From the users point of view a thermal breakdown appears the same no matter which type it is. The explanation is complicated, in particular for secondary break-through, which is why most literature does not deal with it in great detail. Frequently the existence of secondary break-through is only mentioned as “the destruction of components is very probable and caution should be used”. More detail is given in [5], [6] and [7] with a technique for non-destructive measurement shown in [8]. This phenomenon affected the familiar 2N3055 transistor, now more robust transistors are used so there is less need to understand the problem. Input transistors in microphone amplifiers are less endangered than power transistors because of their application, however high frequency small power types are just as endangered because of their internal construction described in [4] chapters 2.8. Secondary thermal break-through can however occur with nearly all bipolar transistors.

#### 4.

### Production of fast pulses

There is at least one reason to look at the two forms of the thermal break-through: secondary break-through happens several powers of ten faster than the first type of break-through. A more exact view indicates the secondary break-through happens extremely fast, somewhere within the range of several picoseconds. This is of practical interest; such a fast event must have a use and if the energy is limited it can be repeated.

This is not new: In [9] a pulse generator is described that uses a 2pF capacitor as an energy reservoir in series with a 1MΩ resistor this produces approximately 90v



**Fig 2: Picture of the completed coax cable pulse generator.**

across a transistor which goes straight into secondary break-through giving a  $50\Omega$  resistance to discharge the capacitor. The rise and fall times are below 250 picoseconds! But it gets better: In [10] under the heading “Line controlled pulse generator” a circuit produces short rectangular pulses. As a reservoir it uses an n-structured LC arrangement, which mimics the behaviour of a line. As a switch it uses a thyatron, which had similar applications to a thyristor, with details shown in [11], [12] and [13]. Reference [14] has a chapter on Charged line + switches describing how pulses can be produced with rise times under a nanosecond using a mercury wetted reed-relay.

**Table 1: Typical data.**

Supply voltage	65v
Transistor type	2N2369a
Output breakdown voltage	18v
Pulse amplitude	12v
Pulse duration	37nS
Frequency	135kHz
Cable length	3.62m
Cable type	RG8U

It is not necessary to consider an aesthetic or compact design to make a circuit for study purposes. Several metres of coaxial cable can be used as a reservoir; electronically it is the ideal solution. A transistor can be used for the switch as described in [9] and shown in Fig 1. The circuit, if one can call it that, is constructed around a coaxial connector. The emitter resistor is  $25\Omega$ , made from four SMD resistors in order to reduce the parasitic inductances, the base resistor is two  $4.7k\Omega$  resistor in parallel. The coaxial cable is soldered directly onto the transistor case and to the collector. A broad clip offers strain relief and provides a good bonding.

The operating voltage depends on the transistor used. The parameters for secondary break-through are only roughly described in some data books for power transistors (SOA, Safe Operating Area) and I have never seen this for small power transistors. A stable operating point can be found by increasing the operating voltage from a low value. Further information is shown in Table 1 with a typical layout shown in Figs 2 – 4.

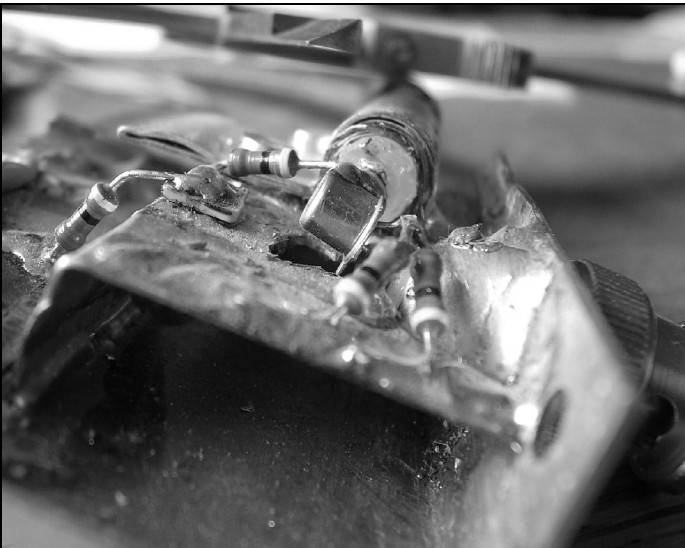




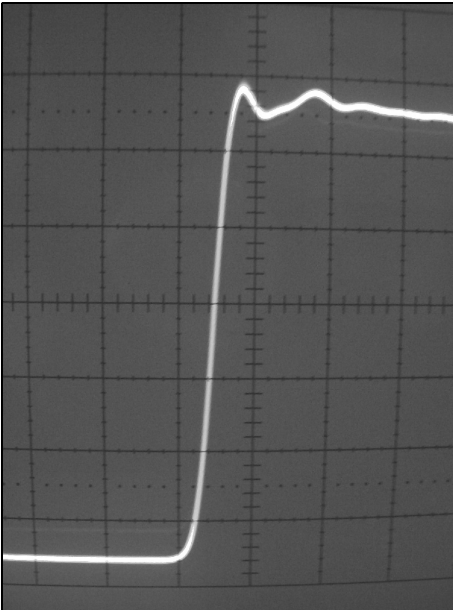
**Fig 3: Detailed view of the transistor in the coax cable pulse generator.**

Fig 5 shows a screen photograph of the pulse generator connected directly to the input of a 150MHz oscilloscope (Tektronix 2335). Obviously only the rise time of the oscilloscope is being measured at approximately 2nS. Unfortunately I could not determine the real rise time of the generator but the overshoot is within the specified limits of the oscilloscope.

Looking at the pulse and cable length: When the transistor breaks-through, a positive pulse edge is displayed on the oscilloscope and a pulse front moves toward the open end of the line. It is reflected there because  $r = +1$  and reaches the beginning of the line after approximately 37nS appearing on the oscilloscope as a negative pulse edge. Using the velocity factor for the cable of



**Fig 4: Detailed view of the coax cable pulse generator with the BNC connector on the right.**



**Fig 5: The positive pulse edge, x scale 5ns/division, y scale 2v/division.**

0.66 the calculated pulse width is 36.6nS, which agrees well with the measured value.

There is a mismatch at the beginning of the cable, this is displayed as a lower amplitude blip in Fig 6. This can be avoided by the choice of emitter resistor, this affects the pulse amplitude of the positive going edge, which is important for my application but not the quality.

The question is then what type of cable can be used? My first attempts used thin cable (RG174, RG58). The pulse clearly deviated from a flat pulse, which I attribute to the higher cable losses.

Without having the original documentation I know that companies like Tektronix and Hewlett Packard used such circuits in the 1960's for fast pulse generators or to produce scanning signals for diode gates in sampling oscilloscopes.

## 5.

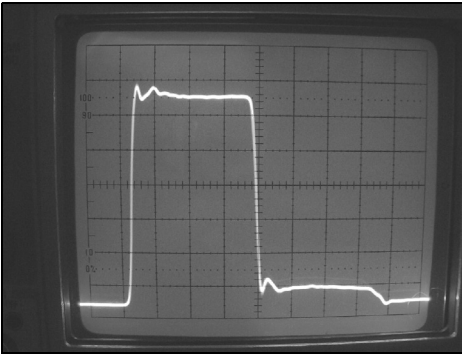
### Application

Finally the application shown in the title is described.

To determine the frequency response of a linear four-pole network e.g. a filter, an attenuator, an amplifier or a transmission line, measuring the complex quotient of the output voltage and input voltage ( $U_{out}/U_{in}$ ) at small steps across the frequency range enables the response to be plotted. This quotient can be expressed as an analytic expression, which is called the transfer function  $A(j\omega)$ . This apparently trivial measuring task requires a considerable amount of equipment: signal generator, voltmeter and phase meter. To use an oscilloscope for measurements up to 50MHz the  $-3\text{dB}$  response of the oscilloscope means that a 100MHz oscilloscope is required.

The measurements are made with time dependent values but the results are given as frequency values with no mention of time. But there is a fundamental relationship between frequency and time. This can be shown by the fact that one can represent the output signal of a signal generator either on an oscilloscope (amplitude as a function of time), or on a spectrum analyser (amplitude as a function of frequency).

This is described mathematically by means of integral transforms, in particular the Fourier transform [15] (Jean Baptiste Joseph de Fourier, 1768 - 1830) or the Laplace transform [16] (Pierre Simon Marquis de Laplace, 1749 - 1827). It does not matter whether the connection between time and frequency is regarded as a current, a voltage or as the transfer function of a four-pole network. If the latter is used its transfer function is  $A(j\omega)$ , then using the inverse Fourier transform gives the pulse response  $h(t)$ . To measure the pulse response requires a short sharp needle pulse for a reference.



**Fig 6: The whole pulse and echo pulse. x scale 10nS/division, y scale 2v/division.**

The presentation of a frequency range as a spectral display and the time domain are equivalent! That means that the characteristics of a transmission system represented in the familiar form can be achieved using mathematics as described in “Methode des scharfen Hinsehens (the method of sharp looking)”. Usually an oscilloscope is available rather than a spectrum analyser; therefore preference is nearly always given to measurement in the time domain. In addition a suitable test signal must be selected.

The sharp pulse described above is normally useless because the high amplitude drives the transmission system into the nonlinear range. The second best choice is the step function: At an arbitrarily time the voltage jumps from 0v to a chosen value “U”. In practice a square wave generator with a large time interval and sufficiently small rise time can be used. The first condition means that all transients produced by the voltage step must have faded away before the next step appears. The second condition means that the rise time of the input must be much smaller than that rise time that an ideal step function would produce at the output. Both conditions are perfectly fulfilled by the coax cable pulse generator described in this article. The amplitude is sufficiently high for most practical measuring tasks; if smaller voltages are re-

quired, then a suitable attenuator or stepped attenuator can be used at the input of the item under test.

The output signal of the test object should be viewed on an oscilloscope. Without needing to understand the mathematics in the detail, and without ambitious quantitative statements, a number of qualitative statements can be made [17]: The steeper the rise time, the larger the frequency range; Overshoot with a constant period indicates amplitude distortions; Overshoot with the oscillation period getting smaller indicates phase distortion.

If broadband amplifiers are being aligned the objective is to achieve the most beautiful step response, i.e. a steep rise time and if possible little or no overshoot. This can achieve the same optimal response as with an expensive measuring set-up!

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## 6.

### Conclusion

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At the end of this article the attentive reader will determine: Everything described was already available. I have produced some answers with the assistance of Pascal [18] (Blaise Pascal, 1623 - 1662). “You must not say that I have not said anything new, the arrangement is new. Everyone plays, if they are playing with the same ball, one of them plays with it better”.

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## 7.

### References

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*Most of these references are for German publications, the author thought that it would be better to refer any readers to the original article that appeared in*



*issue 2/2007 of UKW Berichte. I have decided to keep the list of references, with my poor German translations of the titles, because this is an interesting article and some readers may want to follow up the ideas described in the article. In fact the pulse technique described in this article is also referred to by Gunthard Kraus, DG8GB in his article in this magazine on crystal filter recycling . - Andy*

[1] K.P. Liessmann, Theorie der Unbildung (Theory of illiterateness), Paul Zsolnay publishing house, Vienna 2006

[2] U. Tietze, Ch. Schenk, Halbleiter-Schaltungstechnik (Semiconductor circuit techniques), Springer publishing house 1999

[3] Halbleiter-Elektronik (Semiconductor electronics), edited by W. Heywang and R. Müller, Springer publishing house 1979 from it: Volume 1, R. Müller, Grundlagen der Halbleiterelektronik (Fundamentals of semiconductor electronics)

[4] ditto 1991, volume 2, R. Müller, Bauelemente der Halbleiterelektronik (Devices of semiconductor electronics)

[5] Halbleiterschaltungen der Leistungselektronik (Semiconductor circuits for power-electronics), RCA 1971, publisher: Alfred Neye Enatechnik, Quickborn Hamburg

[6] Understanding Power Transistor Breakdown Parameter, AN1628/D, ON Semiconductor

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1991

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[11] K. Küpfmüller, Theoretische Elektrotechnik (theoretical electro-technology), Springer publishing house 1973

[12] L. Ratheiser, Röhrenhandbuch (tube manual), Franzis publishing house, Munich, technical publishing house ERB, Vienna, 1956, Reprint 1995, Franzis publishing house

[13] H. Meinke, F.W. Gundlach, Taschenbuch der Hochfrequenztechnik (Paperback of high frequency engineering), Springer publishing house 1956

[14] Fast pulses Techniques, E-H Research Laboratories, Inc. Oakland, California

[15] H.D. Lücke, Signalübertragung (Signal transmission), Springer publishing house 1979

[16] W. Ameling, Laplace Transformation, Vieweg publishing house 1984

[17] K. Küpfmüller, Die Systemtheorie der elektrischen Nachrichtentechnik (The system theory of the electrical communications technology), Hirzel publishing house 1974

[18] B. Pascal, Pensées, Über die Religion und über einige andere Gegenstände (Over the religion and over some other articles, collection world literature), Lambert Schneider, Heidelberg



Gunthard Kraus, DG8GB

# Practice project: Crystal filter recycling

Who knows about the shapeless metal enclosures of crystal filters (Fig. 1)? For many decades these professional high impedance components were seen on the printed circuit boards of receiver IF modules or SSB exciters in transmitters or transceivers. Embedded between transistor stages or the first integrated circuits they were an indispensable, highly effective and expensive component was the brand name for high quality equipment. Today they hide themselves in nostalgic devices or worse they are ignored in the printed circuit board junk crates at radio flea markets. The formerly silvered housings are spotted and unattractive. Nobody thinks that they are sealed in hermetically sealed enclosures; in melted glass envelopes (as with valves) the vacuum preserves the excellent electrical characteris-

tics in new condition for decades. These crystal filters today cost practically nothing and do not even interest even traders.

In the following article the uses of crystal filters today for useful solutions is considered.

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## 1.

### Technical data of filters

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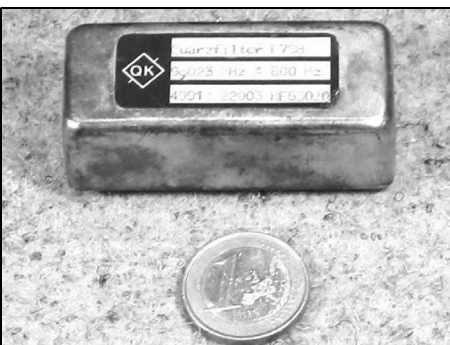
The questions that come to mind are:

- Where can we get data from the original data sheets, that have technical data which looks different, to use in modern circuits
- What work is necessary, in order to make a 50Ω component for modern applications?

Subscriber who have taken VHF Communications Magazine for many years will have no problems: more than 20 years ago there were advertisements from a familiar crystal filter manufacturer on the back of each issue (Fig. 2), this is the best source of information.

From this one can infer:

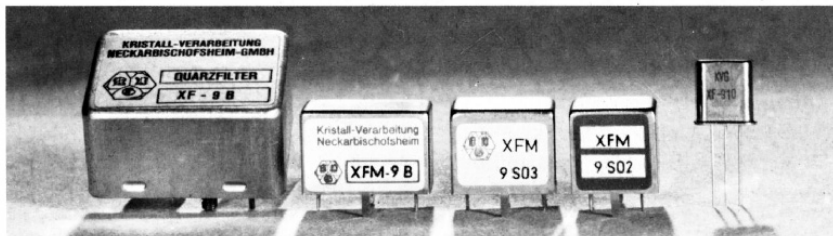
Crystal filters in the frequency range from some MHz to approximately 40MHz were used 20 years ago with



**Fig 1: This is the kind of crystal filter available.**



**OUR GREATEST now with reduced dimensions !**



Case:                    1                                    15                                    14                                    13                                    17

DISCRETE CRYSTAL FILTER	Application	MONOLITHIC EQUIVALENT					
		with impedance transformation			without impedance transformation		
		Type	Termination	Case	Type	Termination	Case
<b>XF-9A</b>	SSB	<b>XFM-9A</b>	500 Ω    30 pF	15	<b>XFM-9S02</b>	1.8 kΩ    3 pF	13
<b>XF-9B</b>	SSB	<b>XFM-9B</b>	500 Ω    30 pF	15	<b>XFM-9S03</b>	1.8 kΩ    3 pF	14
<b>XF-9C</b>	AM	<b>XFM-9C</b>	500 Ω    30 pF	15	<b>XFM-9S04</b>	2.7 kΩ    2 pF	14
<b>XF-9D</b>	AM	<b>XFM-9D</b>	500 Ω    30 pF	15	<b>XFM-9S01</b>	3.3 kΩ    2 pF	14
<b>XF-9E</b>	FM	<b>XFM-9E</b>	1.2 kΩ    30 pF	15	<b>XFM-9S05</b>	8.2 kΩ    0 pF	14
<b>XF-9B01</b>	LSB	<b>XFM-9B01</b>	500 Ω    30 pF	15	<b>XFM-9S06</b>	1.8 kΩ    3 pF	14
<b>XF-9B02</b>	USB	<b>XFM-9B02</b>	500 Ω    30 pF	15	<b>XFM-9S07</b>	1.8 kΩ    3 pF	14
<b>XF-9B10*</b>	SSB	—	—	—	<b>XFM-9S08</b>	1.8 kΩ    3 pF	15

\* New: 10-Pole SSB-filter, shape factor 60 dB : 6 dB 1.5

Dual (monolithic twopole)                    **XF-910**; Bandwidth 15 kHz, R<sub>T</sub> = 6 kΩ, Case 17

Matched dual pair (four pole)                **XF-920**; Bandwidth 15 kHz, R<sub>T</sub> = 6 kΩ, Case 2 x 17

**DISCRIMINATOR DUALS** (see VHF COMMUNICATIONS 1/1979, page 45)

for NBFM                **XF-909**                Peak separation 28 kHz

for FSK/RTTY        **XF-919**                Peak separation 2 kHz

**CW-Filters** – still in discrete technology:

Type	6 dB Bandwidth	Crystals	Shape-Factor	Termination	Case
<b>XF-9M</b>	500 Hz	4	60 dB : 6 dB 4.4	500 Ω    30 pF	2
<b>XF-9NB</b>	500 Hz	8	60 dB : 6 dB 2.2	500 Ω    30 pF	1
<b>XF-9P*</b>	250 Hz	8	60 dB : 6 dB 2.2	500 Ω    30 pF	1

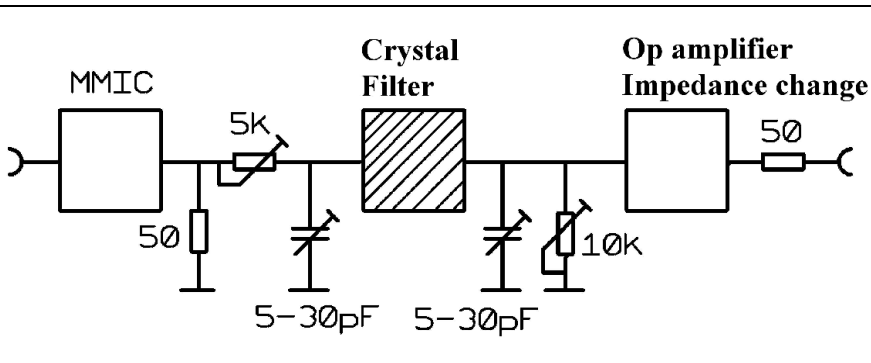
\* New !

**KRISTALLVERARBEITUNG NECKARBISCHOFSHAIM GMBH**

D-6924 Neckarbischofsheim · Postfach 61 · Tel. 072 63 / 6301



**Fig 2:** If there is no documentation available you can help yourself using this information that appeared in adverts on the back page of VHF Communications magazine in the 1980s



**Fig 3: This shows how historical and modern devices can be combined.**

internal and terminating resistors, which were somewhere between 500Ω and 3kΩ. In addition the input and output of the filter had a parallel capacitor (up to 30pF) connected to ground.

The exact resistance and capacitance was only found on the data sheet and must be adjusted in the circuits developed today in order to achieve the published transmission curve.

## 2.

### Circuit concept

Highest priority must be given to “the N-port-principle”. Simplified this means a component (Twoport) with 50Ω input and output resistance, which can be measured correctly with a network analyser and can be described using S-parameters. Therefore the following concept was devised (Fig. 3):

The circuit is divided into two building blocks on two separated small PCBs. Each one is 30mm x 30mm and consists of 1.52mm thick FR4 Material (Epoxy).

An input stage with a modern broadband MMIC (e.g. MAR6) gives the required input impedance of 50Ω in the frequency range between 3 and 100MHz. The inter-

nal resistance of the output stage is 50Ω, which is why this value of resistor is fitted to the PCB as termination. This gives a basic gain of approximately 20dB. The crystal filter would like to see a source resistance of at least 500Ω, so it can be connected to the 50Ω terminating resistor of the MAR6 by inserting an SMD 5kΩ trimmer without disturbing the output matching too much. A small SMD trimming capacitor with a range of adjustment from 5 to 30pF is fitted directly to the input of the filter.

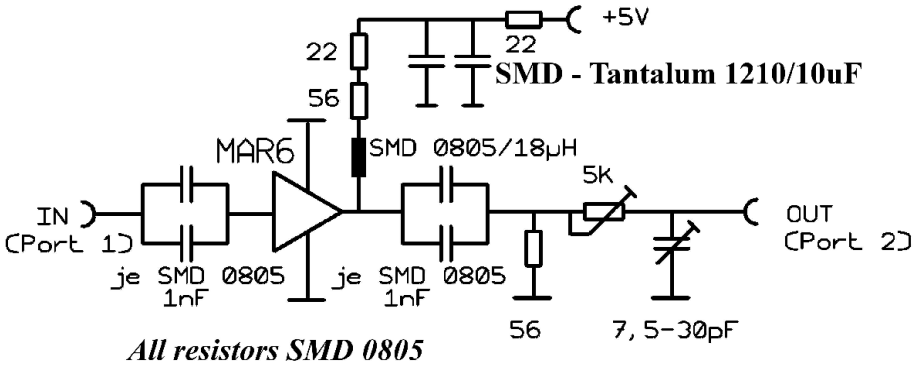
On the second PCB there is a parallel trimming capacitor followed by a trimmer potentiometers to correctly match the filter. The output signal is supplied to a wide band OP amplifier type OPA355. It is connected as a non-inverting amplifier with  $G = 1$  and has a high input impedance, the resistor of 50Ω at the output satisfies the N port.

The overall gain of the assembly is approximately 20dB – 6dB (filter adjustment) – 6dB (Adjustment at the output) = 8dB plus the basic attenuation of the filter.

#### 2.1. Input stage circuit

The input circuit developed is shown Fig. 4; these are the design considerations:

The power requirements of the broadband MMICs (MAR6) are 15mA at 3.5v.

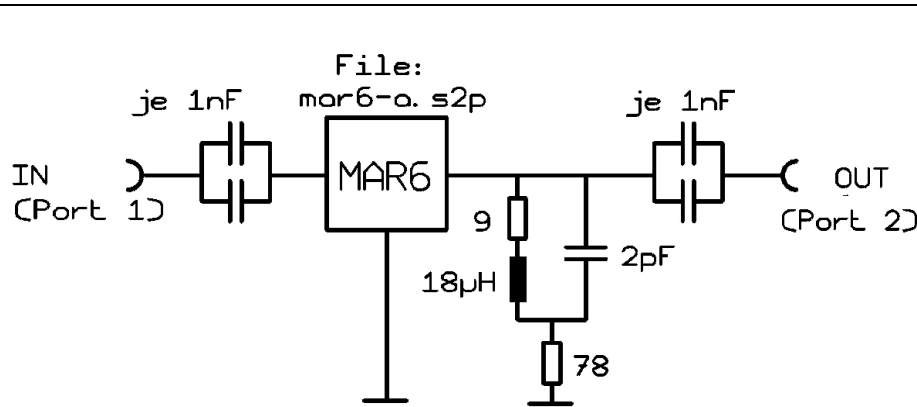


**Fig 4: The circuit of the input stage. It is important to read the information in the datasheet about the operating voltage.**

A choke must be inserted between the output pin of the IC and the supply. This choke that feeds the IC must have a reactance that is at least 3 x the terminating resistance of 50Ω otherwise the gain will be reduced too much. The reactance of an 8µH choke is 150Ω at 3MHz. This value was doubled and the next standard value of 18µH was selected to keep the reduction in gain at 3MHz small. A current limiting resistor is always advisable in series with the choke. With a 5v supply, 1.5v must be dropped to feed the

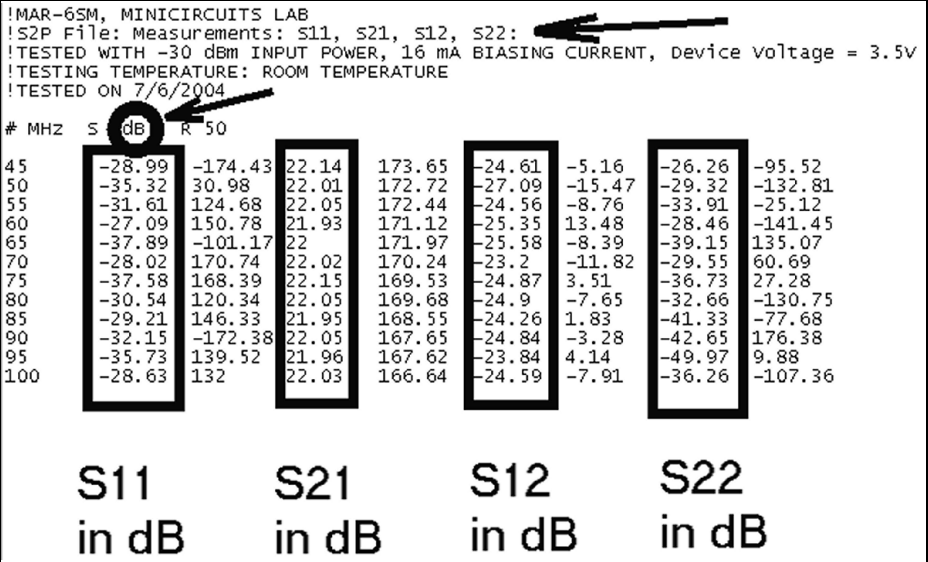
IC with the recommended 3.5v, this requires a resistor of 100Ω with 15mA supply current. From this, 22Ω is connected to the supply with two tantalum decoupling capacitors, the remaining 78Ω is made from 56Ω + 22Ω.

The coupling capacitors for DC isolation at the input and output consist of two parallel SMD capacitors in order to halve the total self-inductance; an important habit for microwave circuit developers! The 50Ω termination, trimming resistor



**Fig 5: The simulation circuit used for the PUFF.**





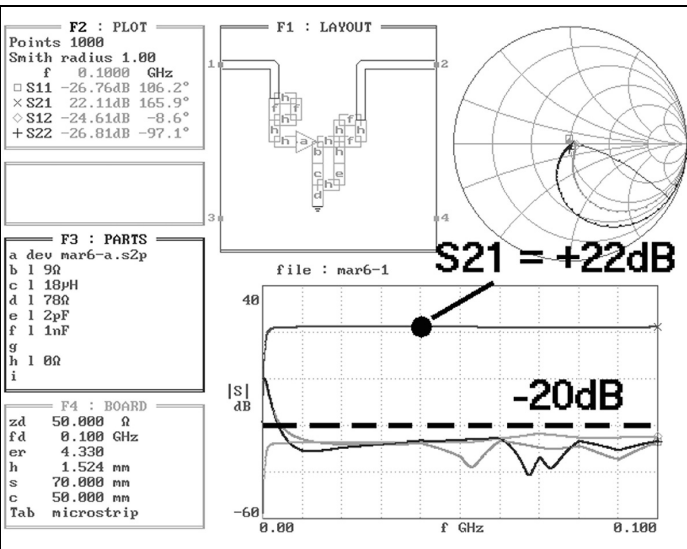
**Fig 6: The S-parameter file has magnitudes in dB that are no good for use with PUFF.**

and capacitor have already been discussed.

### 2.2. Simulation of the input stage

The simulation diagram shown in Fig. 5 was developed for the circuit. According to the data sheet a 9Ω series resistor and

a 2pf parallel capacitor are added to the 18μH choke to simulate the self resonant frequency of approximately 25MHz. Next the “microwave pocket calculator” (PUFF) is used. The S-parameter file for a MAR6 is downloaded from the Internet and copied it into the PUFF file. The first



**Fig 7: This shows that the input stage is ideal.**

**Table 1: List of S-parameters to be used with PUFF.**

```

!MAR-6SM, MINICIRCUITS LAB
!S2P File: Measurements: S11, S21, S12, S22:
!TESTED WITH -30dBm INPUT POWER, 16mA BIASING CURRENT,
device Volyage = 3.5V
!TESTING TEMPERATURE: ROOM TEMPERATURE
!TESTED ON 7/6/2004

!Translated to linear values by Gunthard Kraus
!File-Name: mar6-a.s2p
# ghz s ma r 50
0      0.04      0      12.79  180      0.044      0      0.005      -90
0.01   0.04      5      12.79  178      0.044      0      0.01      -90
0.02   0.04      10     12.79  176      0.044      0      0.02      -90
0.03   0.04      15     12.79  174      0.044      0      0.025     -90
0.04   0.04      20     12.79  173      0.044     -10     0.03      -90
0.05   0.017    30.98  12.96  172.72  0.044    -15.47  0.034    -132.81
0.06   0.044    150.78 12.48  171.12  0.054    13.48  0.0377   -141.45
0.07   0.04     170.74 12.61  170.24  0.069    -11.82  0.033     60.69
0.08   0.0297   120.34 12.59  169.68  0.0568   -7.65  0.0232   -130.75
0.09   0.0247  -172.38 12.66  167.65  0.057    -3.28  0.007    176.38
0.1    0.037     132     12.79  166.64  0.059    -7.91  0.015   -107.36

```

simulation gave completely unreasonable results, the error was quickly traced: in the S-parameter file from Minicircuits, all the magnitudes are in dB, these are not good for use with PUFF (Fig. 6)! It can be seen that all S21 values are in the order of +22dB and all other parameters are at least -20dB, this does not help the PUFF simulation.

The dB values are number crunched with a pocket calculator into factors for the 45 – 100MHz range and a new S-parameter file produced. The values for frequencies below 45MHz were roughly estimated from experience. The new file for the PUFF simulation is shown in Table 1, it is saved using the name “mar6-a.S2P” in the PUFF file. The new simulation result for the circuit in the range from 0 to 100MHz is shown in Fig 7.

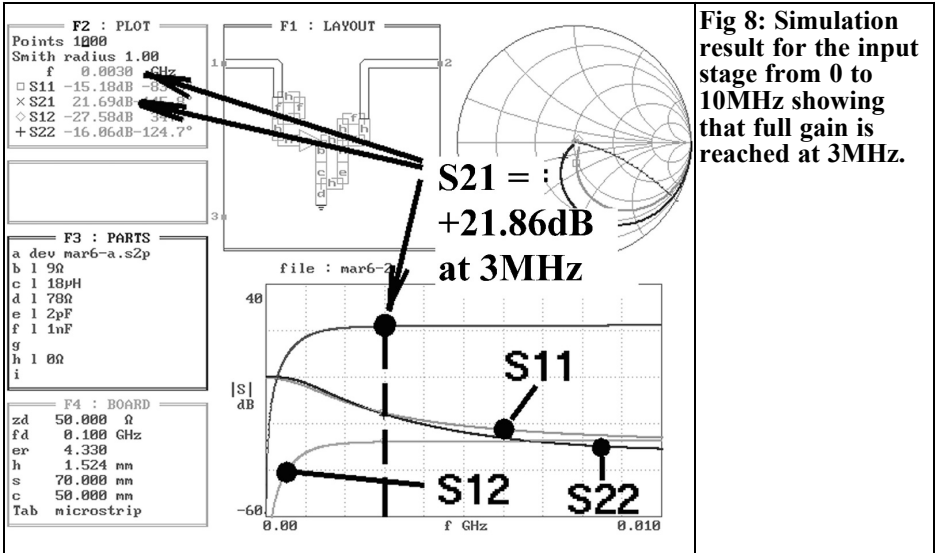
Fig 8 shows an interesting view with a smaller sweep range from 0 to 10MHz, this shows that by 3MHz the full gain of +22dB is reached, in addition the other

parameters are smaller than -20dB.

### 2.3. Measurements of the finished PCB

The circuit was built on a 30mm x 30mm PCB made from 1.5mm thick FR4 Material using SMD and microstrip techniques (Fig 9). The lower surface of the PCB is the ground surface and the earth areas on the top surface are connected to the lower surface by silvered hollow rivets. The output line has a characteristic impedance of 50Ω. Fig 10 shows the first PCB, the things to note are:

- The terminating resistor on the PCB was omitted to measure the S-parameters because the network analyser provides it. Similarly the two trimmers (trimmer resistor and capacitor) used when the filter is being matched, are missing in this stage.
- The IC output is temporarily connected to the output port with a zero Ω 0805 SMD resistor.



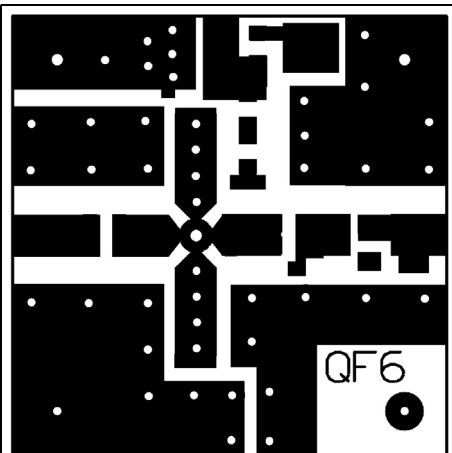
- The input and output of the circuits are connected to SMA sockets fitted to small copper angle strips. This makes a reflection free transition up to 10GHz from the round pins of the SMA sockets to the microstrip lines on the PCB.

The power supply used was an old 200W PC power supply with four output voltages (+5v, -5n, +12v, -12v) found in the

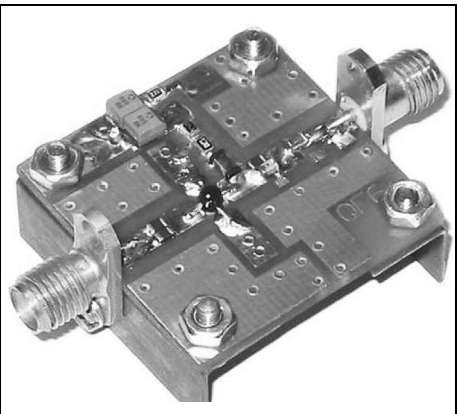
cellar. The measuring equipment was an HP8407A Analyser and an HP8601 sweep oscillator with a range from 10kHz to 110MHz. The result were:

- S21 (and hence gain) was constant at approximately 21dB in the range of 3MHz to 110MHz.
- The gain only begins to fall below 2MHz.

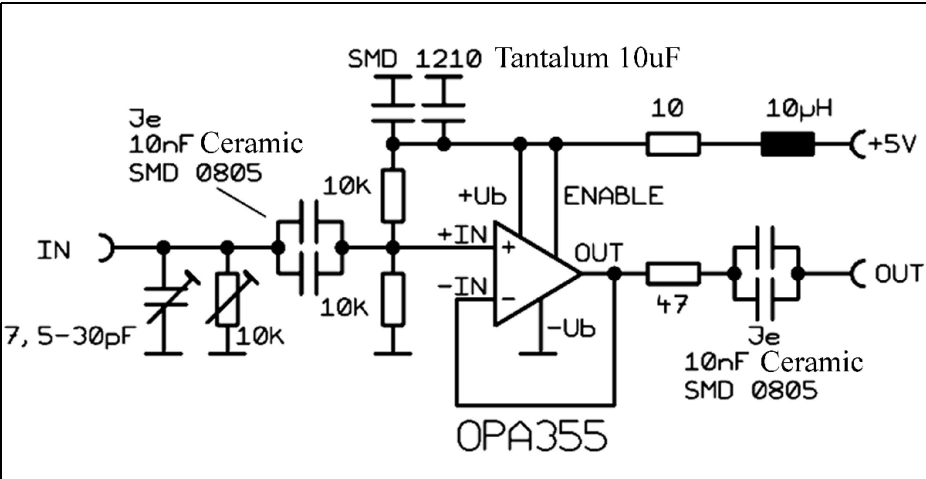
These simulation results are acceptable



**Fig 9: The PCB layout for the input stage.**



**Fig 10: A picture of the completed input stage showing how 50Ω connectors are fitted so that it can be tested using a network analyser.**



**Fig 11: Circuit diagram of the output stage which is more complex. Unfortunately there are no high impedance MMICs for this application.**

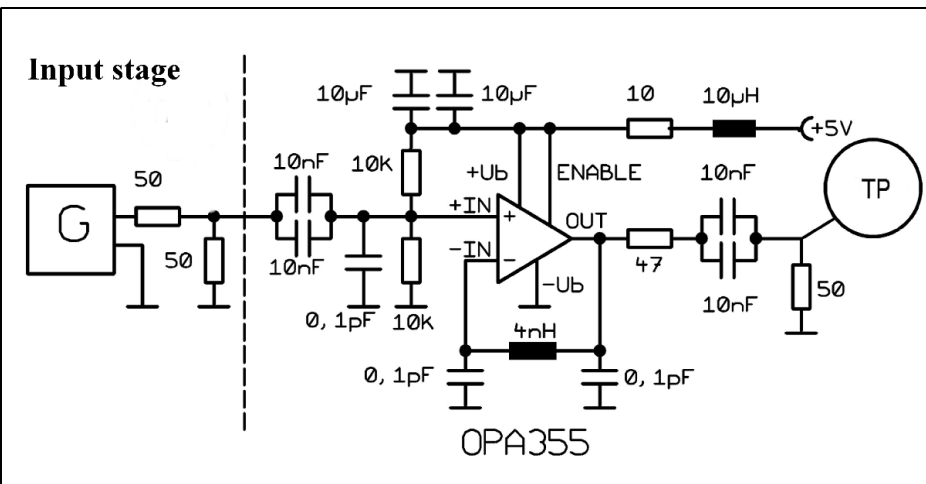
so work on the second circuit can be started.

### 2.4. Output stage circuit

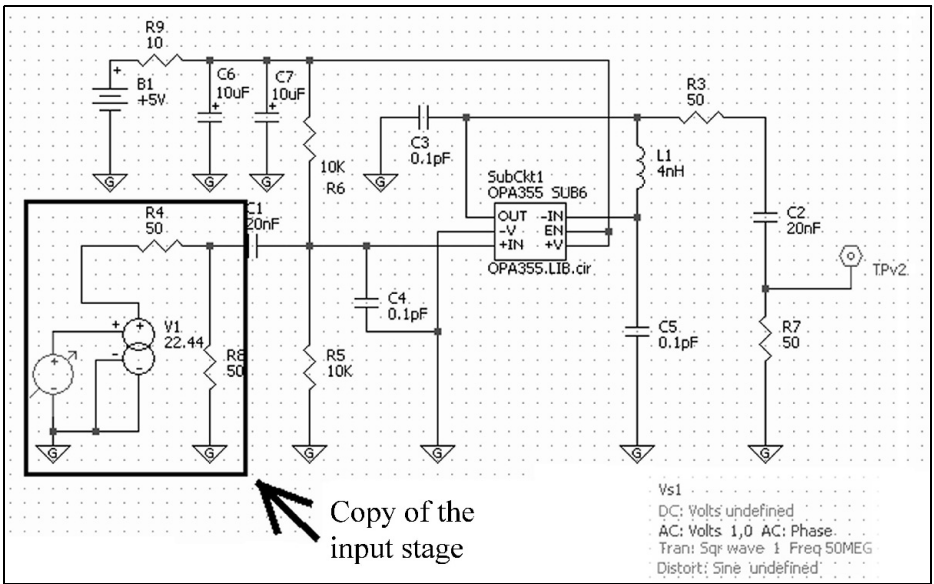
Because a high input impedance is required no 50Ω system microwave MMIC can be used. One of the modern fast operation amplifiers is necessary. The OPA355 was chosen with a gain up to 200MHz, which is enough for the in-

tended purpose for frequencies up to 50MHz.

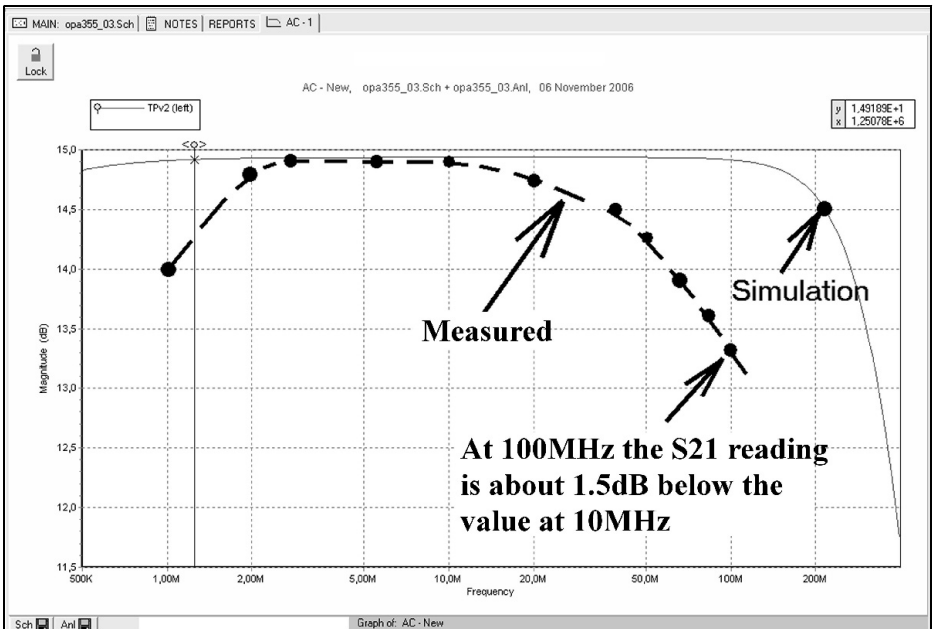
The circuit diagram of the output stage is shown in Fig. 11. It is a familiar non-inverting amplifier circuit with full negative feedback of the output signal to the inverting input (gain = 1). Since only a single supply voltage is used (the data sheet specifies a maximum of +5v), a



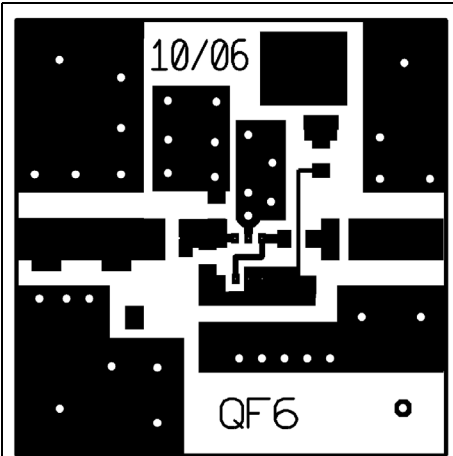
**Fig 12: The circuit diagram of the output stage with the additional components required for the PSPICE simulation.**



**Fig 13: The circuit representation of Fig 12 used for the 5SPICE simulation.**



**Fig 14: Up to 50MHz the measured results are exactly the same as the simulation results.**



**Fig 15: The PCB layout for the output stage.**

half supply voltage of +2.5v is generated using a voltage divider for the non-inverting input.

### 2.5. Simulation of the output stage

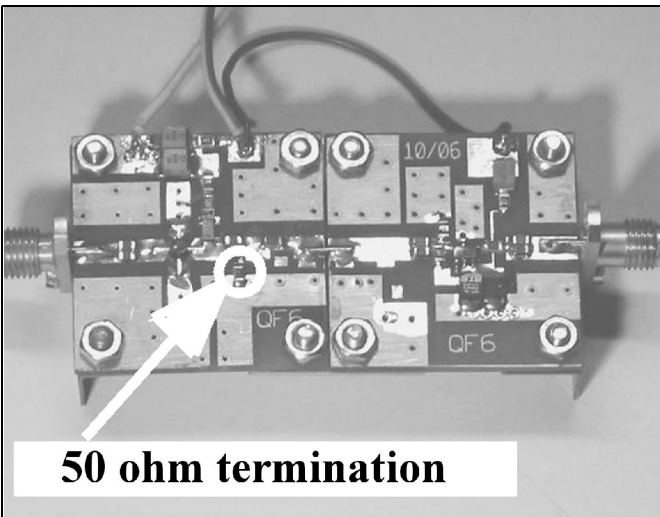
There are no S-parameter files for this type of IC but a SPICE model can be used, working with the free “5spice” software [1]. As well as all of the components in fig. 11 the other things that must be included in the simulation

are: the capacitance of the SMD IC connection pads of 0.1pF, which takes part in the large gain range (the amplifier has a gain-bandwidth-product of 200MHz) and the inductance of the connection from the output pin of the IC to the inverting input (approximately 4nH for the 4mm connection path). This produces the circuit shown in Fig 12 and Fig 13 shows how such a complex circuit fits into the limited workspace of the free 5spice test version.

Comparing Fig 12 and 13 shows how the input stage with its constant gain, measured as 21dB, was included into the simulation:

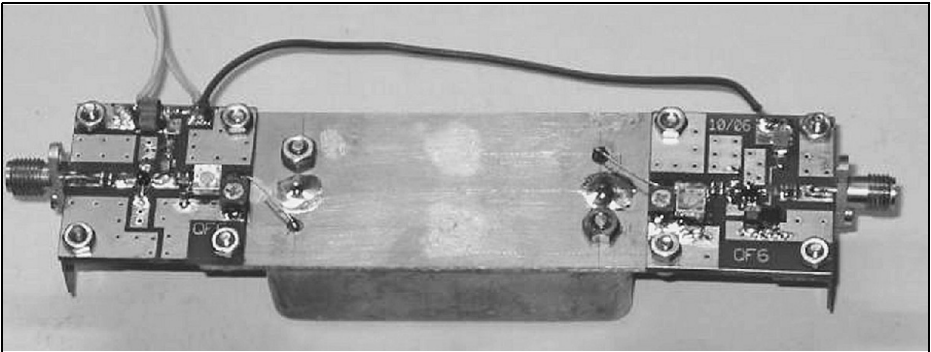
The generator from Fig 12 and the 21dB gain was replaced by a voltage controlled voltage supply with a transfer function of 22.44. As for the real input stage it has an internal resistance of 50Ω with a 50Ω resistor connected to ground.

If the output voltage is simulated in dB, then it is referred to as the unloaded source voltage. Because the gain of the input stage has been included the result is the same as what would be measured with a network analyser as “S21” so it is an exact simulation. Fig 14 can verify this statement; the simulation and measurement are directly compared. Over the



**Fig 16: Picture of the input and output stages connected together for measurements to be taken.**

**50 ohm termination**



**Fig 17: Picture of the completed assembly with input stage, crystal filter and output stage.**

intended frequency range of 3MHz to approximately 50MHz the results are acceptable.

Fig 15 shows the 30mm x 30mm FR4 PCB and Fig 16 shows how the two stages were connected using a suitable piece of PCB to make measurements. It clearly shows how the 50Ω terminating resistor, already mentioned, is fitted.

### 3.

## The hour of truth

Finally the complete test arrangement is shown in Fig 17. The trimming resistor and capacitor were fitted to the output stage and the bridge on the output of the input stage was removed. A suitably fashioned piece of PCB was used to connect the three building blocks (input stage, crystal filter and output stage). Interestingly enough, the alignment and measurement with the network analyser only took a few minutes. Both trimming resistors are adjusted to easily find the positions where the ripple becomes minimal but still visible. Mutual adjustment of the trimming capacitors quickly has the correct transmission curve appearing on the screen.

The overall gain of the arrangement was

stated earlier in the article as:

$$+21\text{dB gain} -6\text{dB (filter matching)} -6\text{dB (output adjustment)} = +9\text{dB}$$

That only applies if the crystal filter has a pass band transmission loss of zero dB - a somewhat unrealistic assumption! The measurements resulted in a gain of approximately +5dB at the centre frequency of 8.94MHz, so the basic attenuation of the filter in this case was approximately 4dB.

Testing the filter over a wide range showed that over the entire range from 1MHz to 100MHz the stop band attenuation was better than the range of the measuring equipment (approximately 60dB). 60dB is very good for filters of this age and confirms the original idea.

Fortunately this time again all results agreed with a professional network analyser. The transmission curve shown in Fig 18 has minimum ripple and a maximum gain, measured accurately, +4.47dB! The 3dB range determined from this curve is 15kHz and this is exactly what was on the identification plate of the filter.

The test over a wide frequency range is shown in Fig 19. It shows a beautiful response but it must be confessed that the data sheet for the filter says: "The attenuation in the stop band is at least 60dB" which is exactly what was measured.

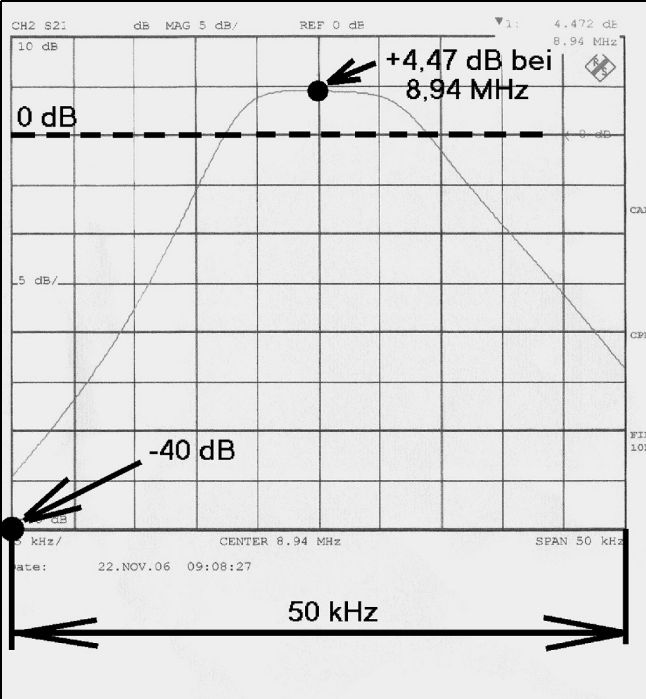


Fig 18: The transmission curve measured with a modern analyser.

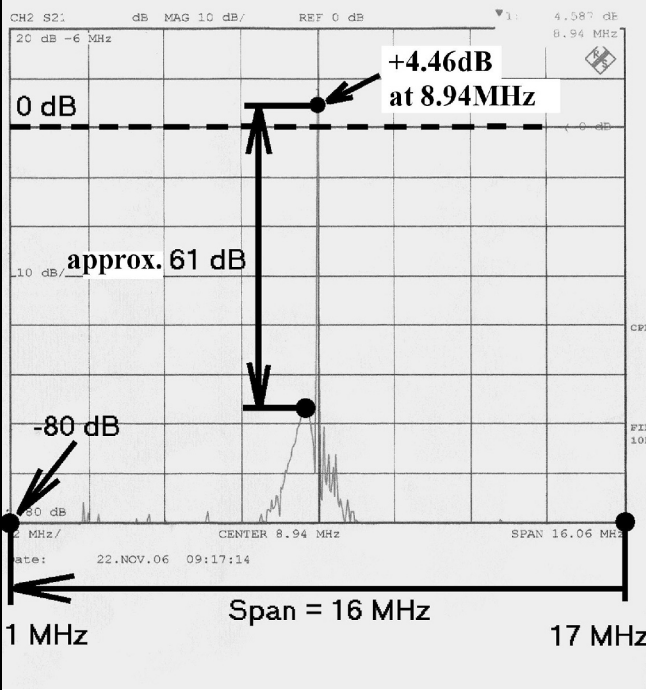
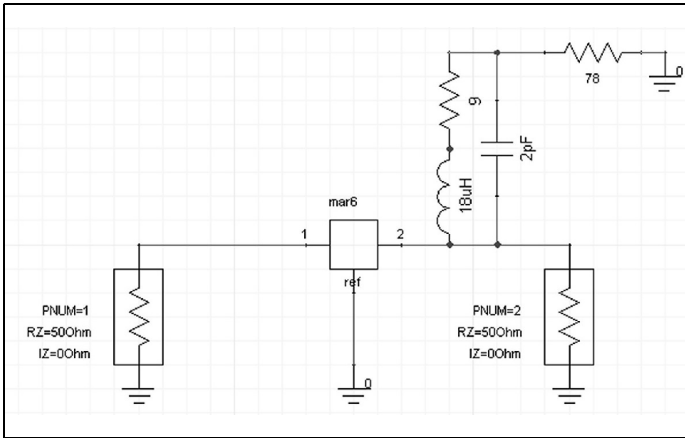


Fig 19: The transmission curve measured from 1 to 17MHz showing the response in the stop band.





**Fig 20: The input stage circuit diagram used for Ansoft Designer SV.**

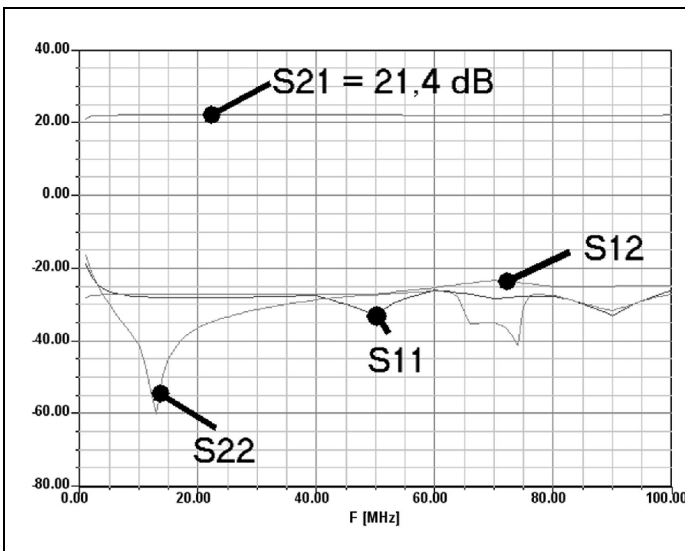
### 3.1. What is the tendency for self-oscillation?

Modern circuit simulation removes the question “will the circuit work?” Better still is the fact that the question about self-oscillation is unnecessary, since this problem can be answered equal well with simulation. In the following section the self-oscillation for this project is investigated in more detail.

#### 3.1.1. Stability of the input stage

Some more information is required,

which cannot be delivered by PUFF. So an ANSOFT Designer SV project for the MAR6 was started. The simulation circuit was taken directly from PUFF and is shown in Fig 20 as an ANSOFT diagram. The S-parameter simulation (Fig 21) does not show any surprises, because the results agree completely with those from PUFF (please turn back to Fig 7 for comparison!). Fig 22 shows the stability factor “k”, which must be larger than 1 in the frequency range required for absolute stability. That is the case and thus the output stage can be checked next.



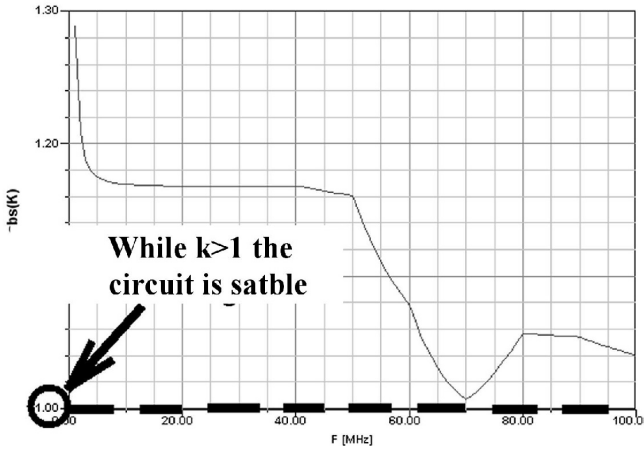
**Fig 21: The S-parameter simulation of the input stage, compares well with Fig 7.**



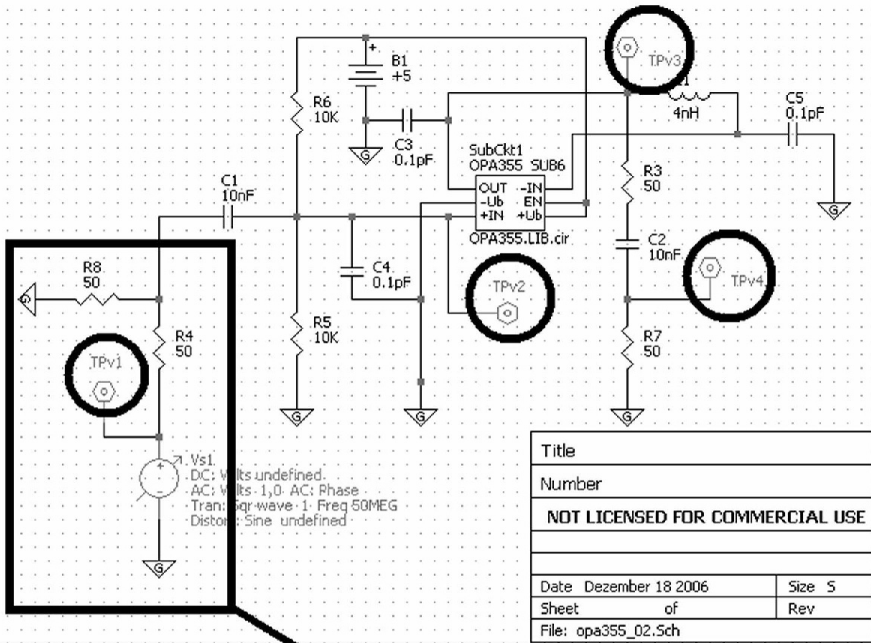
12 Nov 2006

Ansoft Corporation  
XY Plot 2  
Circuit1

09:23:52



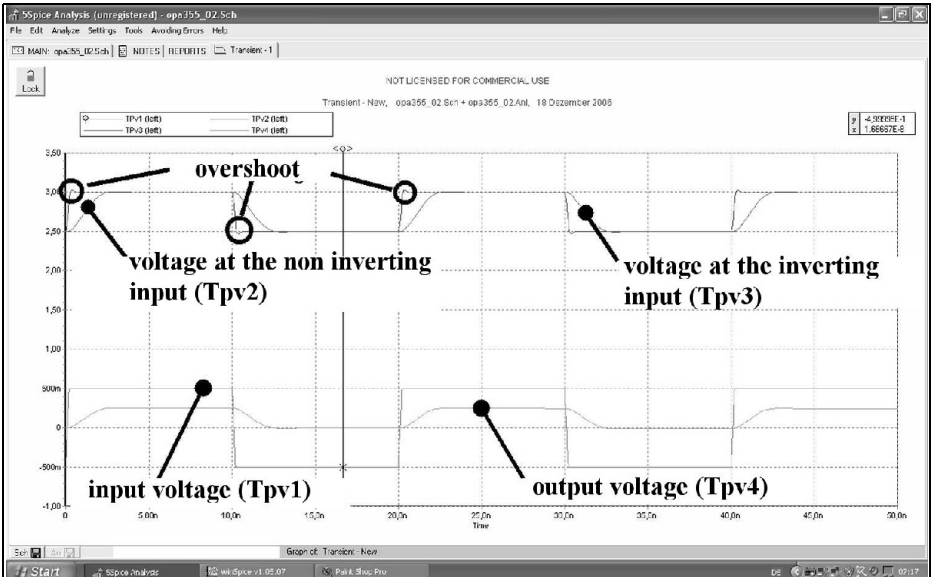
**Fig 22: The stability factor "k" for the input stage must be greater than 1 for stability.**



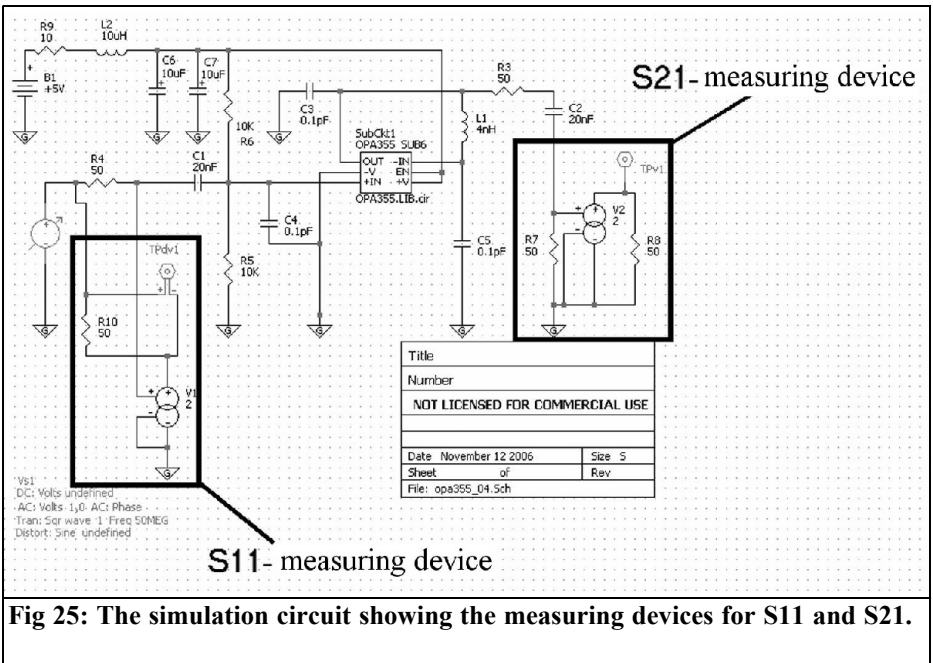
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NOT LICENSED FOR COMMERCIAL USE	
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Sheet: of	Rev
File: opa355_02.Sch	

**Replacement for the input circuit**

**Fig 23: The PSPICE circuit used to measure the stability of the output stage by feeding a 50MHz square wave into the input and measuring the response at 4 test points**



**Fig 24: The results of the simulation on the output stage showing something that must be investigated at the input of the non inverting input.**



**Fig 25: The simulation circuit showing the measuring devices for S11 and S21.**

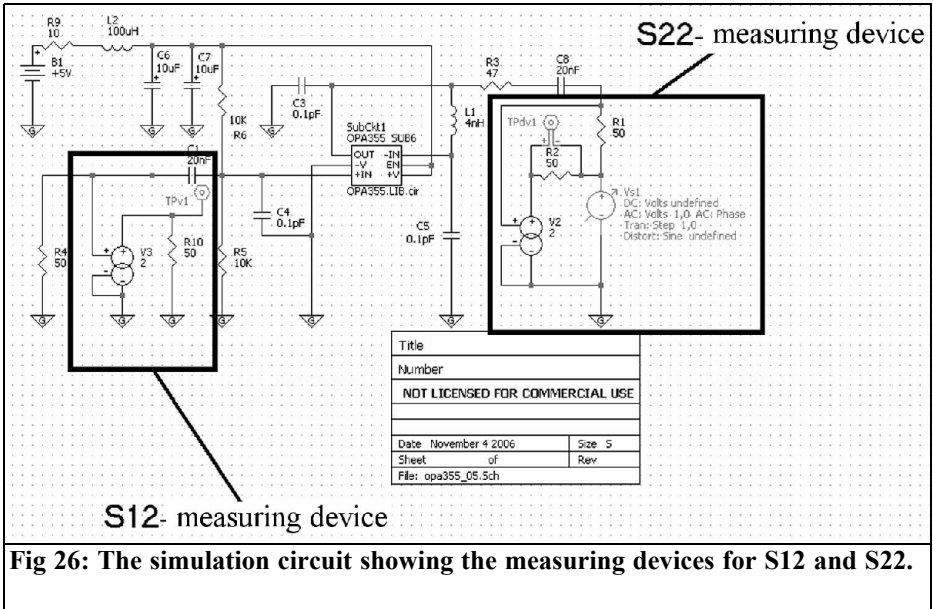


Fig 26: The simulation circuit showing the measuring devices for S12 and S22.

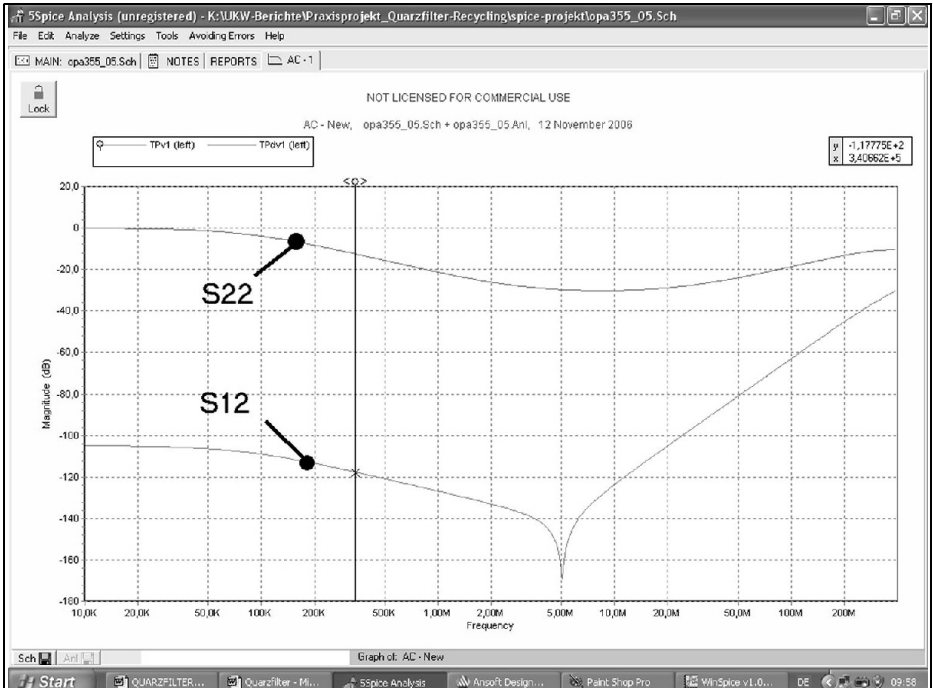
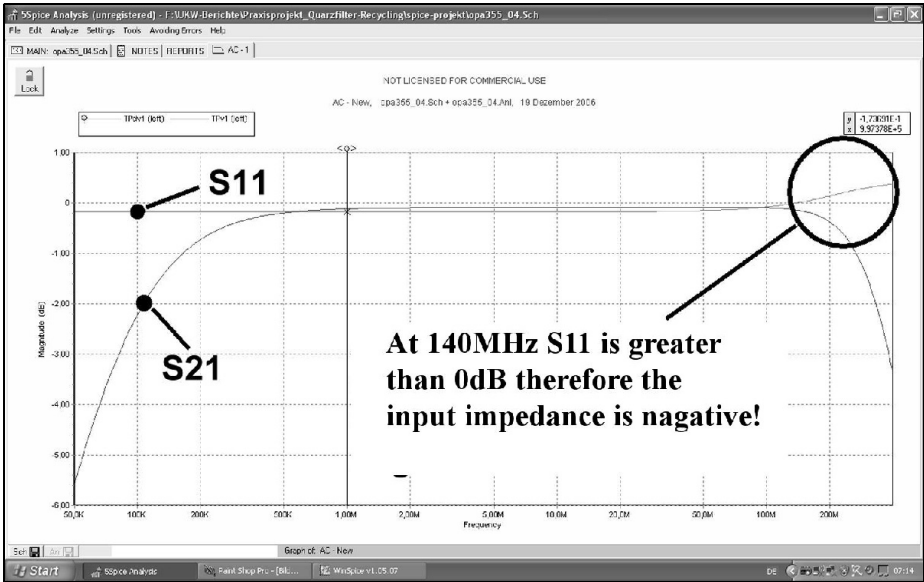


Fig 27: The response curves for S12 and S22 that don't show any problems.



**Fig 28: The response curves for S11 and S21 that show a problem beyond 100MHz.**

### 3.1.2. Stability of the output stage

This is more difficult because it has to be done with a PSPICE simulation, which works in the time domain. Before using the big guns, it is best to have a quick look over the fence of RF technology. There are some complex methods used in control engineering to determine stability of a system by measuring the frequency response using complex equipment. A faster technique is to use the pulse method by applying an input step and measuring the response of the system. A dangerous tendency to self-oscillation shows itself immediately by an overshoot at the output with clearly visible oscillation at the natural frequency of the system that slowly dies away.

To try this method simply feed the input with a symmetrical square wave at a frequency  $f = 50\text{MHz}$  and examine the output. The PSICE circuit used for this is shown in Fig 23 with 4 test points (generator output, non-inverting input of the amplifier, inverting input of the am-

plifier and output of the circuit). The control input is provided by a voltage source followed by a voltage divider made from two  $50\Omega$  resistors. The simulation result is shown in Fig 24.

It shows that the calculated voltage at the non-inverting input shows an overshoot which must be investigated further. That is complicated, but re-reading [2] in detail, shows how to derive S-parameters from the PSPICE simulation circuit. Fig. 25 shows the representation of S11 and S21 and Fig 26 shows the circuit diagrams for S22 and S12. The simulation results for S12 and S22 are shown in Fig 27, they don't show any problems but Fig 28 shows something unpleasant: Starting from 140MHz, S11 becomes greater than zero dB and results in negative input impedance in this range!

Once this has been found, the remedy is not difficult: a real resistance is slowly added in parallel with the input from 100MHz so that it compensates the negative input impedance. The simple solution is shown in Fig 29: a resistor of

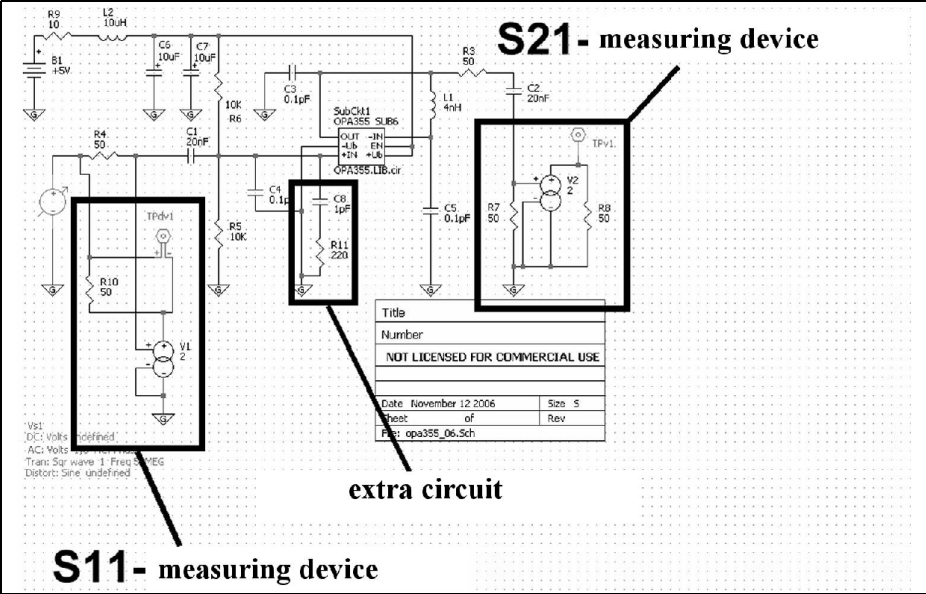


Fig 29: Fortunately the effect can be controlled easily with an extra circuit added to Fig 25.

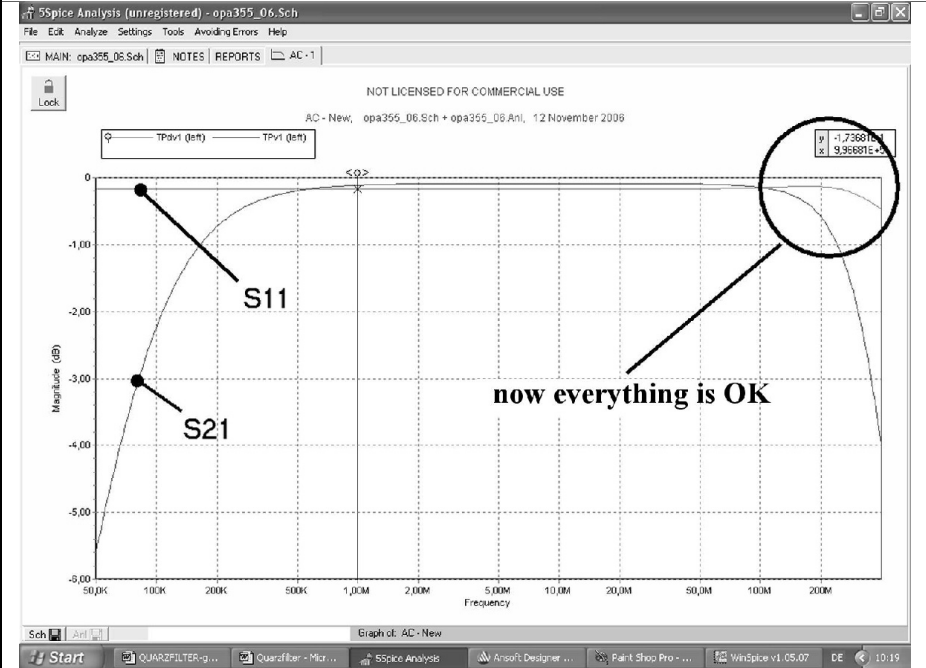


Fig 30: Now the S11 curve is OK.



approximately  $220\Omega$  in series with a 1pF capacitor is enough for correction, Fig 30 shows the resounding success of this measure.

It is inconvenient that PCB layout of the output stage must be re-done in order to accommodate both components as close as possible to the inverting input with the connecting wires as short as possible...

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#### 4.

### Conclusion

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Using modern components and modern CAD programs can give old crystal filters a new lease of life. The final result is something to be proud of. The problem with self-oscillation demanded some work from the grey cells. Having completed the project and added it to the collection of personal building blocks, thoughts turn to a suitable application.

The author hopes that the readers have fun with such projects.

#### 5.

### References

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[1] An interesting program: Circuit simulation using PSPICE, Gunthard Kraus, DG8GB, VHF Communications Magazine 4/2005, pp 223 - 238

[2] Determining S-parameters with PSPICE, Gunthard Kraus, DG8GB, VHF Communications Magazine 2/2006, pp 95 - 106



André Jamet F9HX

# Using DDS Aliasing to extend its frequency range

## 1.

### Introduction

The Nyquist theorem limits the normal DDS range up to the half of the clock frequency. In practice, it is advised not to exceed a third. To obtain frequencies around 100MHz we need at least a 200MHz clocked DDS. Above the Nyquist limit, the output is no longer a unique sine wave. It shows an infinity of frequencies like the discontinuous wave spectrum. They are called aliases or images.

Fig 1 shows the spectrum envelope of a 100MHz clocked DDS for a 45MHz nominal output. The aliasing generates frequencies that could be interesting as they are beyond half of the clock fre-

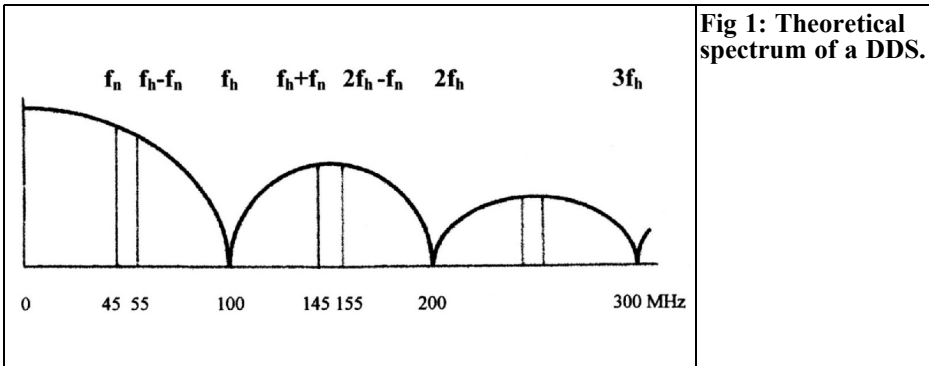
quency. Their levels decrease with the frequency but they remains usable. In the appendix, maths readers will find some references about the cardinal sinus function that governs the spectrum generation.

We can use a DDS far beyond half its clock frequency using aliases produced. Only careful filtering is needed to extract the desired frequency.

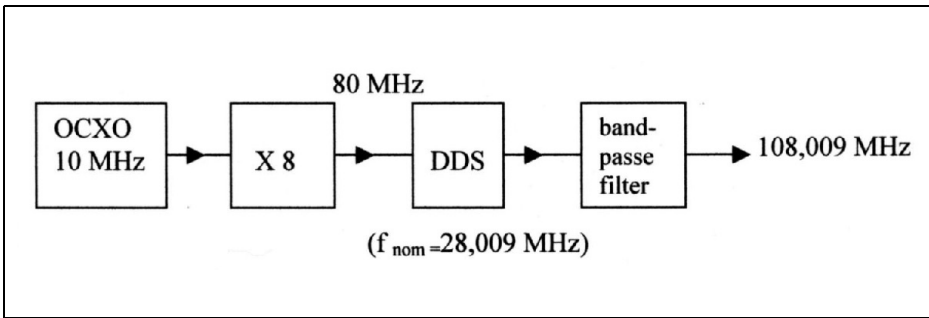
## 2.

### An example

We need a 108.009MHz signal as a source for a 10368.900MHz beacon. Let us take a 100MHz DDS but with an 80MHz clock. Set up the nominal fre-







**Fig 2: Block diagram of the VHF source.**

quency at 28.009MHz.

We obtain an alias at :  
 $f_c + f_n = 80 + 28.009 = 108.009\text{MHz}$ .

Another alias is at :  
 $2 f_c - f_n = (80 \times 2) - 28.009 = 131.991\text{MHz}$

Other frequencies are present as shown by Table 1 which gives some measurements (for a 28MHz nominal DDS frequency).

We have to filter the output in order to eliminate undesirable frequencies, mainly at 131.991MHz which is the closest to the desired frequency. A double pole bandpass filter like the NEOSID 22522/51138 or a crystal filter as described in [2] can do that. Fig 2 shows the block diagram of this VHF source.

To produce an 80MHz clock signal, we

can use the same method as for the synthesizer F5CAU/F9HX [2]. The precise and stable output of a 10MHz OCXO is squared by a 74HC4046 to get a CMOS signal that is harmonic rich. The desired harmonic is selected by a band-pass filter and then amplified to get the required level for a DDS clock. This method can also be used to produce a high stability 100MHz clock (Fig 3). We can also use a 100MHz clock and an 8MHz nominal frequency DDS to get a larger spacing between the two aliases:

$$f_c + f_n = 100 + 8 = 108\text{MHz}$$

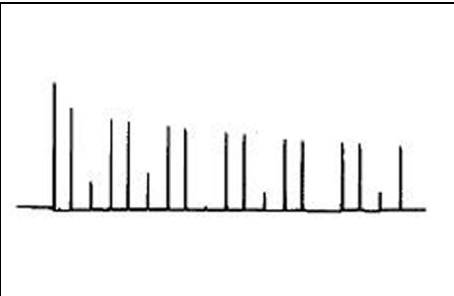
and

$$2 f_c - f_n = (100 \times 2) - 8 = 192\text{MHz}.$$

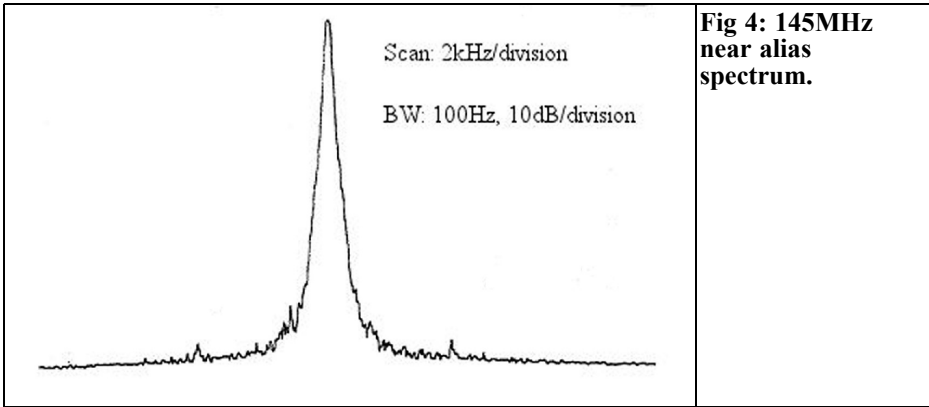
The drawback is a lower output that is only -32dBm at 108MHz. Easier removing of the “bad” alias or better output power: what is the choice!

**Table 1: Other frequencies present at the output of a 28MHz DDS**

Frequency MHz	Output dBm
28	-14
52	-20
80	-66
108	-28
132	-30
160	-52
188	-30
212	-30
268	-34



**Fig 3: The output spectrum of a 100MHz clocked DDS. Nominal frequency : 45MHz.**



**Fig 4: 145MHz near alias spectrum.**

**Table 2: Other frequencies present at the output of a 45MHz DDS**

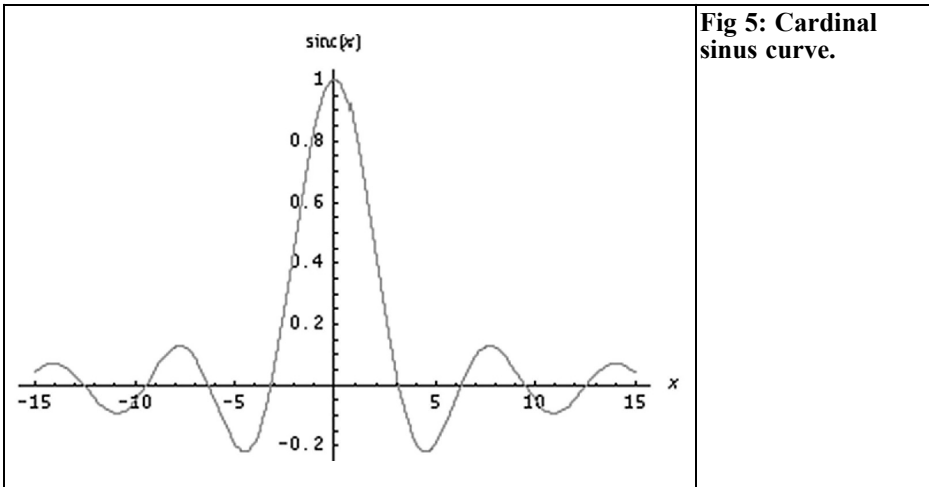
Frequency MHz	Output dBm
45	-10
55	-12
100	-60
145	-20
155	-20
200	-45
245	-24
255	-24
300	-60
345	-27
355	-28

**3.**

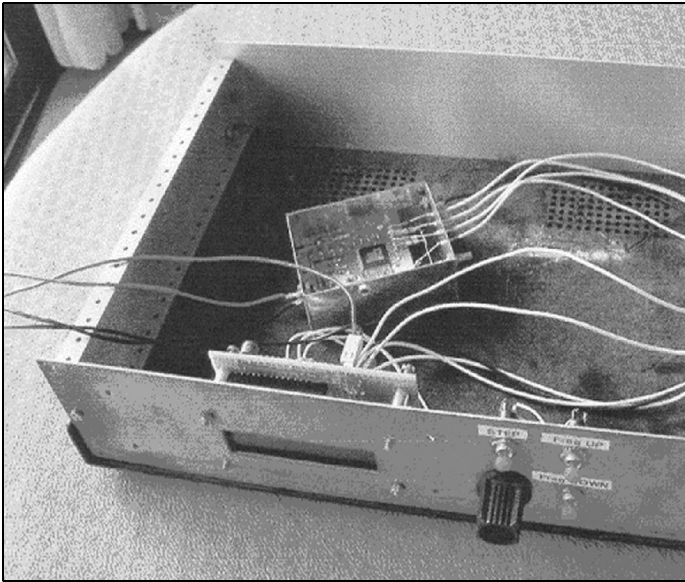
**Another example**

This method can be used to produce any VHF frequency. For example, with a 100MHz clocked DDS and a 44 to 46MHz nominal frequency, we can get a 144 to 146MHz output. See Table 2 and figures 4 and 5.

For a first approach, we can determinate the clock frequency required by dividing the desired frequency by 1.4. As shown in Fig 1, we stay in a favourable zone where the alias level is near the maximum. Too near 1.5 and the “bad” alias is



**Fig 5: Cardinal sinus curve.**



**Fig 6: Picture of the F5CAU 100MHz clocked DDS.**

close to the “good” one and it would be difficult to eliminate. To get higher frequencies, we have to divide by 2.4 to be in the favourable zone; nevertheless, the output will be quite low.

#### 4.

### Phase noise

Referring to [1], the alias phase noise is degraded compared to the nominal output, as the level is reduced. Further from the Nyquist limit more it is noticeable.

We also have to take into account the advantage of a direct a 144 to 146MHz signal generation instead of multiplying a 24.166MHz by six. That multiplication induces a phase noise increase equal to the squared frequency ratio:

$$10 \times \log(62) = 15.6 \text{ dB}$$

Figure 4 shows the near spectrum of a 145MHz alias.

#### 5.

### Realization

The DDS module, assembled by F5CAU comprises an AD9852 and a PIC to control the frequency, the LCD display and memories. The frequency steps are from 1Hz to 1MHz. A digital encoder controls the frequency steps.

I have made several experiments to get various frequencies. I obtained consistent results with the theory and spectrum analyser curves are convincing.

Unfortunately, I am not skilled enough to cope with the very small pins on the DDS (do not forget F9HX is licensed since 1946!). So, I only use the F5CAU module and I am not able to describe one built by myself. Sorry, but you can't do the impossible! I only hope to attract attention of readers fond of exotic devices.



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**6.****Conclusion**

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It is obvious that an up-to-date DDS working with a 500MHz clock is able to deliver a VHF signal without the need of aliases. Nevertheless, it is easier to produce a VHF frequency clock (80 and 100MHz as described above) instead UHF one. Moreover, this method opens a possibility to generate UHF signal with a common DDS instead of gigahertz DDS that are still state of art devices.

If any readers know about any other similar DDS use, apart [4,5], please let me know ( [agit@wanadoo.fr](mailto:agit@wanadoo.fr) ). I'll take it!

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**7.****Appendix**

---

The frequency envelope of a DDS is given by one function called cardinal sinus:

- $\sin c(x) = \frac{\sin x}{x}$  (not normalised form)

When the variable is null, the cardinal sinus is undetermined 0/0. So, by derivation, we can find the limit that is 1.

That function is shown in Fig 5.

In practice, we cannot have negative values for both frequency and magnitude. So, the actual curve is given by

- $|\sin t| = |\sin \pi t/\pi|$  with  $t > 0$

which correspond to Fig 1.

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**8.****References**

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[1] A Technical Tutorial on Digital Signal Synthesis, ANALOG DEVICES - <http://www.analog.com/en/DCcList/0,3090,770%255F%255F65,00.html>

- Section 2: Understanding the sampled output of a DDS Device
- Section 10: Using DDS Images as Primary Output Signals in VHF/UHF Applications

[2] Multiplication division and addition of a 10MHz source to get a synthesised VHF signal, F5CAU & F9HX, VHF Communications 2/2003, pp 119 - 124

[3] A Low Budget Vector Network Analyzer for AF to UHF, DG8SAQ QEX 3/4 and 7/8 2007

[4] Using DDS Aliases to Extend the Frequency Range, Sam Wetterlin, [www.wetterlin.org/sam/AD9952/MultiplierClockAliasing.pdf](http://www.wetterlin.org/sam/AD9952/MultiplierClockAliasing.pdf)



Bo Hansen, OZ2M

# Sequencer

1.

## Sequencer description

If you add a power amplifier to your station it is important to make sure that the changeover from transmit to receive does not feed your powerful signal into the receiver input as the aerial is switched and the amplifier signal decays. This is the job of the sequencer, it makes sure that things happen in a controlled sequence. The timing diagram Fig 1 shows the sequence of events. The PTT signal is the input that you control to

switch from receive to transmit, the sequencer then controls the timing of the aerial changeover relay, the power amplifier and the transmit/receive signal to the prime mover (transceiver). To ensure that the receiver input never receives any signal from the power amplifier the aerial changeover relay is switched first when going from receive to transmit and released last. Next the power amplifier is switched on and finally the transmit signal is applied.

The circuit diagram of the sequencer is shown in Fig 2, it uses a PIC to produce the timing signals, the source code for the PIC can be downloaded from Bo's web

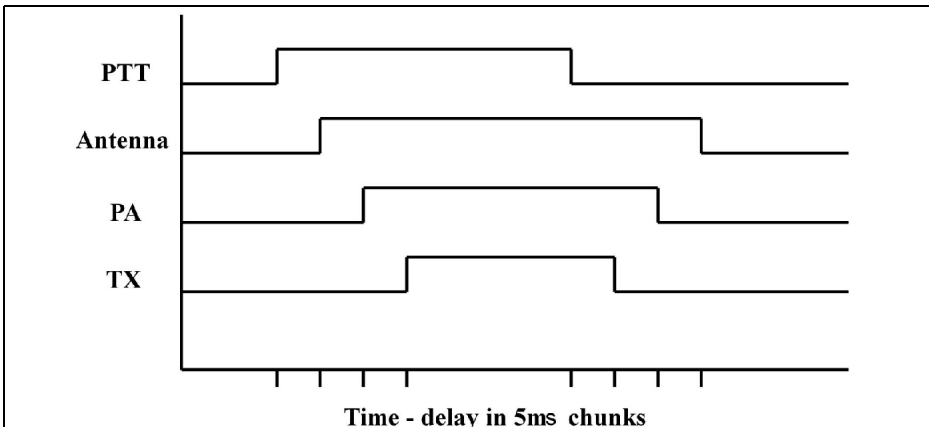
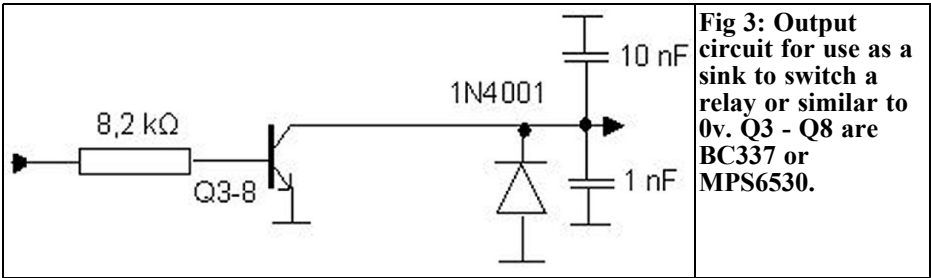


Fig 1: Timing diagram for the sequencer.





**Fig 3: Output circuit for use as a sink to switch a relay or similar to 0v. Q3 - Q8 are BC337 or MPS6530.**

page [1]. It has the following facilities:  
 Control three separate power amplifiers and aerial changeover relays for three different bands.  
 Different delays for SSB with a “K” signal at the end of SSB transmission.  
 All timing parameters changeable in software.

To use the sequencer, one of the three band inputs is set; the corresponding PA and ANT outputs will be activated when

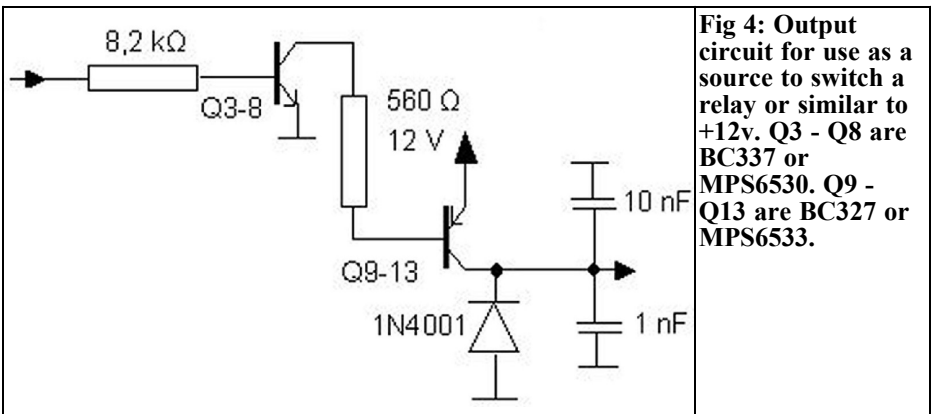
the timing starts. Table 1 shows the pin assignments on the PIC for the three sets of outputs. These pins are connected to either a source or sink circuit shown in Figs 3 and 4. The source circuit has a transistor that switches the output to +12V and the sink circuit has a transistor that switches the output to ground. There are three inputs to start the timing sequence (PTT, PC or voice keyer) and two to abort the sequence (Abort PC or Abort VK). If the SSB input is set, the “K” output will be sent to the Mic audio when switching from transmit to receive.

**Table 1: Pin assignments for the outputs of the sequencer.**

Pin	Name	Function
8	RB2	PA band 1
9	RB3	Antenna band 1
10	RB4	PA band 2
11	RB5	Antenna band 2
12	RB6	PA band 3
13	RB7	Antenna band 3

**2. References**

[1] Bo Hansen, OZ2M web site <http://www.rudius.net/oz2m> the pic source code is available from this site under “sequencer”



**Fig 4: Output circuit for use as a source to switch a relay or similar to +12v. Q3 - Q8 are BC337 or MPS6530. Q9 - Q13 are BC327 or MPS6533.**



Andy Barter, G8ATD

# Introduction to the JavaScript Notebook

1.

## Introduction

While I was doing research for the RSGB VHF/UHF handbook that I edited recently I came across the JavaScript Notebook by Martin E Meserve, K7MEM [1]. It is a useful collection of calculations for many amateur radio projects. The site is made interactive using JavaScript so that there is no need to download or install

any software just enter your answers into the relevant form and the answers are displayed. All of the calculations are easy to use with descriptive notes and are intuitive to use with good graphical representations of the results. Fig 1 shows the front page of the web site with the menu of calculations available on the left hand side. Hovering over the main menu items shows a sub menu of the detailed options for that type of calculation as shown in Fig 2

K7MEM Amateur Radio Home Page for Martin E. Meserve		
<b>JavaScript® Electronic Notebook</b> <ul style="list-style-type: none"> <li>• Resistors</li> <li>• Capacitors</li> <li>• Inductors</li> <li>• Power Supplies</li> <li>• Filters</li> <li>• Attenuators</li> <li>• Antennas</li> </ul> <p><small>Place your mouse pointer over any of the categories above, for an active menu of their contents.</small></p>	<b>Introduction</b> <p>Hello, and thank you for visiting my web site. The main focus of this site is what I call my JavaScript® <b>Electronic Notebook</b>, accessible from the menu on the left. It consists of a group of JavaScript® driven web pages that are intended for use as a quick reference for Ham Radio operators or just anyone interested in electronics. Use the menu bar on the left to access any one of them.</p> <p>The Notebook is similar to other applications on the net, like HamCalc by George Murphy, VE3ERP, or a windows version of HamCalc on John Owen's, G4VWL, web site. The only real difference is that my pages do not require any downloading and installation. All you need is a web browser and an internet connection.</p> <p>This web page is kind of an experiment. Initially the content may seem a little scant. But, I only started this site on June 5, 2004 and as time goes on, I will add much more. Use the menu at the top to navigate this site. I may expand into other areas over time, but this is it for now. If you have some ideas on expanding some pages, adding new topics, or even have a page of your own that you might like to include, let me know. I may adjust it so that it fits the patterns of my pages, but you will receive full credit.</p> <p>Because this site is experimental, don't expect everything to work right off. If there is a real problem let me know. At the top of each page is a link to a email page (E-Mail Web Page) so that you can get ahold of me. This helps me keep spam down, by not having to list my email address. This link will allow you to send email to me, even if you don't have email configured on your local machine.</p>	<ul style="list-style-type: none"> <li>My Amateur Radio Biography</li> <li>What's New On My Web Site</li> <li>Contact Me</li> <li>Web Site Map</li> <li>Developing In HTML and JavaScript</li> <li>Web Server Information</li> <li>Web Browser Compatability</li> </ul>

Fig 1: The home page of the JavaScript Notebook.





<p><b>JavaScript® Electronic Notebook</b></p> <ul style="list-style-type: none"> <li>Resistors</li> <li>Capacitors</li> <li>Inductors</li> <li>Power Supplies</li> <li>Filters</li> <li>Attenuators</li> <li>Antennas</li> </ul> <p>Place your pointer over categories about active menu contents.</p> <p>Click to view the Resource Credits</p>	<p style="text-align: center;"><b>Introduction</b></p> <p>Hello, and thank you for visiting my web site. The main focus of this site is what I call my JavaScript® <b>Electronic Notebook</b>, accessible from the menu on the left. It consists of a group of JavaScript® driven web pages that are intended for use as a quick reference for Ham Radio operators or just anyone interested in electronics. Use the menu bar on the left to access any one of them.</p> <p>The Notebook is similar to other applications on the net, like HamCalc by George Murphy, VE3ERP, or a windows version of HamCalc on John Owen's, G4WVL, web site. The only real difference is that my pages do not require any downloading and installation. All you need is a web browser and an internet connection.</p> <p>This web page is kind of an experiment. Initially the content may seem a little scant. But, I only started this site on June 5, 2004 and as time goes on, I will add much more. Use the menu at the top to navigate this site. I may expand into other areas over time, but this is it for now. If you have some ideas on expanding some pages, adding new topics, or even have a page of your own that you might like to include, let me know. I may adjust it so that it fits the patterns of my pages, but you will receive full credit.</p> <p>is experimental, don't expect everything to work right off. If there is a real problem let me know. h page is a link to a email page (E-Mail Web Page) so that you can get ahold of me. This helps a lot, by not having to list my email address. This link will allow you to send email to me, even if mail configured on your local machine.</p>	<p>My Amateur Radio Biography</p> <p>What's New On My Web Site</p> <p>Contact Me</p> <hr/> <p>Web Site Map</p> <p>Developing In HTML and JavaScript</p> <p>Web Server Information</p> <p>Web Browser Compatibility</p>
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**Fig 2: Selection on the antenna calculations and the sub menu.**

**2.**

**VHF/UHF Yagi antenna design**

There are two options for Yagi calculation, one is verbose stepping through each design choice in detail the other is a quick option that uses a simple question and answer technique, plus an option to calculate a suitable feed.

Fig 3 shows the page of questions to design a Yagi using the quick option. They allow the selection of the design option for:

- Frequency
- Gain or boom length
- Reflector and director spacing
- Boom type, diameter and correction
- Driven and parasitic element diameters

the results for an antenna with the following design parameters:

- Frequency: 1296MHz
- Forward gain: 15dBd
- Reflector spacing: 0.20 - DL6WU
- Director element spacing: DL6WU spacing
- Boom type: Metal boom
- Boom diameter: 10mm
- Boom correction: default
- Driven element diameter: 3mm
- Parasitic element diameter: 3mm

As the required entries are made a diagram of the antenna design appears in the bottom half of the window. Fig 4 shows



**Design Frequency**  
440 MHz

**Reflector Spacing**  
0.20 - DL6WU

**Boom Type (Mounting)**  
Metal Boom (Bonded)

**Select Forward Gain or Boom Length**  
12 dBd - Forward Gain

**Director Element Spacing**  
Use DL6WU Element Spacing

**Boom Diameter**  
1.25 inch

For 12.0 dBd Gain, the Estimated Boom Length is:  
 $2.284 \lambda = 61.3" = 1556 \text{ mm}$

**Boom Correction**  
0 6 7 6 8

**Driven Element Diameter**  
0.25 inch

**Parasitic Element Diameter(s)**  
0.25 inch

At this frequency, the Boom Diameter should be limited to:  
 $0.06 \lambda = 1.6" = 40.9 \text{ mm}$

It is recommended that the element diameters be limited to between 0.001 and 0.02 wavelengths. At 440.000 MHz, this is 0.027" (0-1/32") to 00.536" (0-17/32"), or 00.681 mm to 13.627 mm. The idea here is to keep the diameter small in relation to the element length. Thus, as frequency increases, the useful element diameter decreases.

**Fig 3: The questions for the quick design of a Yagi.**

Design Synopsis		Antenna Dimensions		
View Dimensions:	Metric	Cumulative Spacing (mm)	Element	Element Length (mm)
1296 MHz, 19 Elements, 15.049 dBd Estimated Gain		Zero	REFL	118.55
26.7 Degrees <i>Horizontal</i> Beam Width		46.26	D.E.	109.32
27.5 Degrees <i>Vertical</i> Beam Width		63.61	D1	104.06
10.00 mm Diameter, <i>Metalic</i> Boom with <i>Bonded</i> Elements. Boom Correction of 00.661 applied.		105.25	D2	102.67
Electrical Boom Length of 1344.1 mm. Allow for overhang when cutting boom to length.		154.99	D3	101.21
3.000 mm Driven Element Diameter.		212.82	D4	99.86
3.000 mm Parasitic Element Diameter.		277.59	D5	98.68
Suggested Stacking Distance for 2 Yagis:		346.98	D6	97.66
441.8 mm Horizontally		419.85	D7	96.79
429.7 mm Vertically		496.19	D8	96.02
00.7 mm Dimensional tolerance required for element lengths.		575.99	D9	95.35
		659.27	D10	94.74
		746.01	D11	94.19
		836.23	D12	93.69
		928.76	D13	93.23
		1113.81	D15	92.4
		1206.34	D16	92.03
		1298.87	D17	91.68

**Fig 4: Design page for a 15dBb 1296MHz Yagi.**



## Introduction

The VHF/UHF Yagi Antenna Design page does a good job of calculating the lengths of all the antenna elements but it kind of leaves you flat for actually feeding the antenna. Although it's not really mentioned, the long boom yagis, designed from that page, are intended to use a **Folded Dipole** antenna feed. That is what this page is all about.

The **Folded Dipole** has several useful characteristics. It's easy to build and adjust, its bandwidth is good for over an octave. For example, 50 Mhz to 100 Mhz or 120 Mhz to 240 Mhz. And, its characteristic impedance is a more or less a constant 300 ohms. This doesn't mean that the whole yagi will have that kind of bandwidth. The directors and reflectors respond best for the design frequency, giving you an overall bandwidth of about 7% of the design frequency. So, a antenna designed for 145 MHz should give you an effective bandwidth of about 10 MHz.

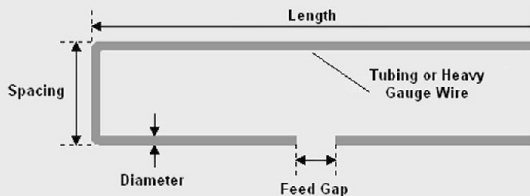
Below is a somewhat exaggerated drawing of the basic **Folded Dipole**. But, it's only intention is to show you the configuration you are working with.

The most important dimension is the **Length**. You want to adhere as closely as possible to the calculated length from the VHF/UHF Yagi Antenna Design page.

The second most important dimension would be the element **Diameter**. For these designs, the diameter of the element should be limited to between 0.001 and 0.02 wavelengths. As an example, this would mean that for 2 Meters you should keep the diameter between 0.061" (2.053 mm) and 1.617" (41.068 mm). For 70 CM the diameter should be limited to between 0.027" (0.681 mm) and 0.536" (13.627 mm).

The **Spacing** between the horizontal sections can vary rather widely. This nicely accomodates do-it-yourselfers, but also give you some latitude for varying boom diameters and mounting methods.

The **Feed Gap** is exactly what it sounds like. This is where you break the loop for attaching the feed. An exact dimension is not necessary, but try to keep the gap as small as practicable. A good rule of thumb would be to keep it no larger than one boom diameter.



**Fig 5: Description of the folded dipole.**

### 3.

## Feed design

Once the antenna is designed the VHF/UHF fed option can be used to see how to construct your antenna and design a suitable feed. Fig 5 shows how to construct the folded dipole feed and Fig 6 shows how to mount the elements onto the boom. Finally the feed impedance of about 300 ohm must be matched to the coax feeder. Fig 7 shows the section of the feed page used to calculate the length of the half wave phasing loop with an assembly diagram.

### 4.

## Inductor calculations

Of the other options on the main menu the most useful to the VHF constructor are the calculations of inductance for a coil, straight wire and flat strip. Figs 8 and 9 show the calculation page for a coil, Fig 10 show the straight wire calculation and Fig 11 shows the flat strip calculation.

### 5.

## Reference

[1] JavaScript Notebook by Martin E Meserve, K7MEM - <http://www.k7mem.com>



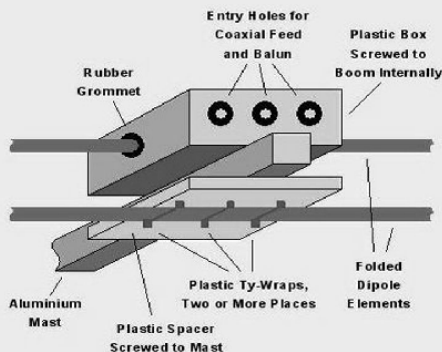
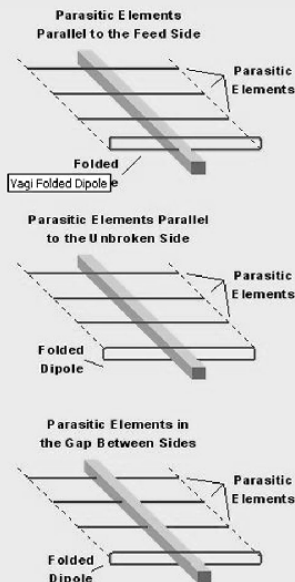
## Mounting

Here again, you have several choices. Some of this depends on your choices for the antenna boom. Initially you have to decide on the orientation of the folded dipole feed in relation to the parasitic elements. The picture on the right shows three different orientations. On **top**, the folded dipole feed point is parallel to the parasitic elements. In the **middle**, the unbroken section of the folded dipole is parallel to the parasitic elements. And then on the **bottom**, the parasitic elements are parallel to the aperture between the feed point section and the unbroken section.

Neither of the choices has any particular advantage over the other, so you can choose the mounting method that best fits your application. The boom running through the folded dipole doesn't seem to have any significant effect of the performance of the overall antenna.

In all of the methods, however, you should be careful not to make an electrical connection between the folded dipole and the antenna boom. The center of the unbroken side of the folded dipole should be at an electrical zero, but there is no guarantee that you have a perfect balance and this could cause unwanted RF currents to flow on the boom.

The most popular method for mounting the folded dipole is to secure a small plastic box above, or below, the boom. The box serves the dual purpose of providing an insulated place for making the feed connections and allows you to easily weather proof the connections. Again, this is probably best described in a picture, like the one below. No actual dimensions are given, because it's unknown to me exactly what you are using.



The plastic box can be secured to the boom with two sheet metal screws. If you place solder lugs under these same screws, the coaxial cable shields can be tied here as well, but more on that later. Properly sized rubber grommets in all of the entry holes will help prevent the weather from entering.

Don't drill any feedline entry holes until you decide on the kind of feedline to use. Otherwise you may have some extra holes that you will need to patch up.

The plastic spacer, on the bottom, serves to insulate the folded dipole element from the boom and to secure it in place. For UHF folded dipole feeds, all you should need to do is use a couple of UV stabilized ty-wraps to secure it in place. Larger and heavier elements would be better held in place by a small clamp of sorts. The drawing doesn't show it very well, but, take care to make sure the dipole element, held down by the plastic spacer, does not touch the screws that secure the spacer to the boom. There is nothing that says that the plastic box needs to go on top. If you want it on the bottom, just flip the drawing over.

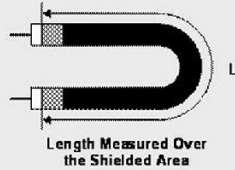
The drawing shows a square boom but this kind of arrangement can just as easily be done with a round boom. If you are using a non-metallic boom insulating everything is going to be much easier.

**Fig 6: Mounting the elements on the boom.**



Frequency (MHz)	Type	Velocity Factor	Impedance
1296	<input type="checkbox"/> RG-59	.66	75 Ohm
	<input type="checkbox"/> RG-59/U Foam	.79	75 Ohm
	<input type="checkbox"/> RG-11/U	.66	75 Ohm
	<input type="checkbox"/> RG-11/U Foam	.80	75 Ohm
	<input type="checkbox"/> RG-6	.75	75 Ohm
	<input type="checkbox"/> RG-58	.66	50 Ohm
	<input type="checkbox"/> RG-8	.66	50 Ohm
	<input type="checkbox"/> RG-8x	.78	50 Ohm
	<input type="checkbox"/> RG-8/U Foam	.80	50 Ohm
	<input checked="" type="checkbox"/> RG-213	.66	50 Ohm

Operating Frequency = 1296 MHz  
 Cable Type = RG-213  
 Velocity Factor = 0.66  
 1/2 Wavelength Phasing Section (L)  
 0' 3" (7.6 CM)



The diagram on the right illustrates the assembly of the 1/2 Wavelength Phasing Section and Folded Dipole Feed. The feedline connects to one side of the Folded Dipole Feed section along with one side of the 1/2 Wavelength Phasing Section. The 1/2 Wavelength Phasing Section exits the plastic box through a rubber grommet and comes back in through another. This end is then connected to the other side of the Folded Dipole Feed. The shields for all of the coax ends are tied together and soldered to a lug inside the box. If the boom is metal, and the screw holding the ground lug creates a physical connection, that's OK. If the boom is not metal, that's OK too.

You have a lot of space to get creative here. The drawings are only a suggested method. For example, if you had a box big enough, you could probably keep everything but the feedline in the box. But that is usually not the case, except for higher frequency antennas. Or, you could use coaxial connectors at the places where the coax enters the box.

In one of the drawings the 1/2 Wavelength Phasing Section is shown as a simple U shape, whereas, in the drawing above, it is shown looped at the bottom. Either way is acceptable. For lower frequencies, the 1/2 Wavelength Phasing Section may be quite long. Just roll it up into a 6 or 8 inch loop, secure it with electrical tape, and then lash it to the boom. Higher frequency antennas will have shorter 1/2 Wavelength Phasing Sections and may not need to be spooled up.

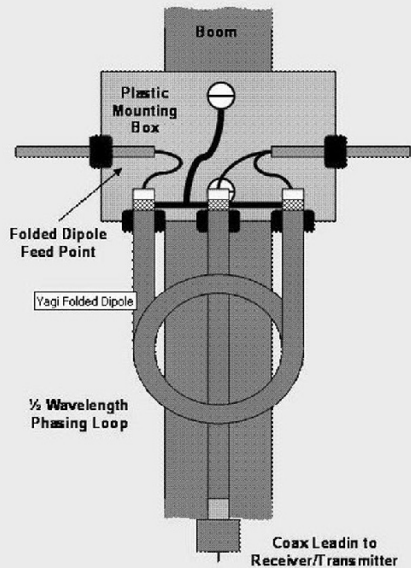


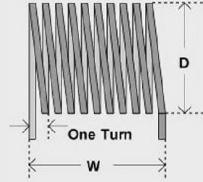
Fig 7: Calculating the half wave phasing loop.



As the title indicates, this page is for the design of a **single-layer, air-core, coil**. It is not equipped to handle multi-layer coils. The page is long and is split into several pieces. Much of it is explanation, but the intent is that, you start out with a required inductance and physical properties, then, through a series of calculations, finish with a practical coil design. These formulae provide close approximations of values for frequencies in the 1-30 MHz range that are sufficiently accurate for most Amateur Radio purposes. The calculations are useful in the VHF and UHF range but only as a basis for further calculation and experimentation.

Calculations propagate from one section to the next so that you can fine tune your requirements, and your design, as you go. Initially, start with the Program Description and then use the navigation menu for getting around.

A diagram of the coil and the basic equation used for the calculations are shown below.



$$L (\mu\text{H}) = \frac{D^2 + T^2}{18 \cdot D + 40 \cdot W}$$

Where:

- L = Inductance (uH)
- D = Coil Diameter (inches or mm)
- T = Number of Turns
- W = Coil Width (inches or mm)

**Fig 8: The coil design layout.**

**Initial Design Input Data**

Required Inductance (uH)

Wire Size (AWG, in, or mm)  **AWG#** ▼

Wire Type (\* See Below)  ▼

Turns per Inch (\* 25.4 mm)

Length/Diameter Ratio (\* 1.5:1 Recommended)  ▼

The **Turns per Inch** area, above, is used as a **Output** when Bare or Enamelled wire is selected, however, it is used as a **Input** when Insulated wire is selected. See the information in the section on Design Considerations for details on how to handle different turns spacing.

**Initial Design Output Data**

Number of Turns

Wire Size

Turns Spacing

Coil Form OD

Coil Length

L/D Ratio

Turns\_per\_Inch

[View Design](#)

[View Coil Taps](#)

**Fig 9: The coil inductance calculation parameters.**

Any length of wire has some inductance, however, at low frequencies, 1-30 MHz, the added effect of this inductance is usually minimum. At VHF and UHF frequencies the inductance of a wire may have a great effect on a circuit. Now, there is no diagram for this because it is simply a straight round wire in **free space**. By **free space** I mean, it is far enough from grounds or other wires so that they have minimum effect.

At VHF frequencies, and above, skin effect will reduce the inductance. The constant, 0.75 will approach infinity, however, skin effect will not reduce the inductance by more than a few percent.

Enter your data below. Internally, all input data is converted to inches for calculation purposes, however, for entry purposes, the dimensions can be mixed or matched.

$$L = K \cdot b \cdot \left[ \frac{2b}{a} \ln\left(\frac{2b}{a}\right) - 0.75 \right]$$

Where:

- L = Inductance in uH
- K = 0.00508 (dims in inches)  
0.0002 (dims in mm)
- a = wire radius (in or mm)
- b = wire length (in or mm)
- ln = Natural Logarithm  
= 2.303 \* Common Log(base 10)

Wire Size (a)  
(AWG, in, or mm)

**AWG#** ▼

Length of Wire (b)  
(in, or mm)

**inches** ▼

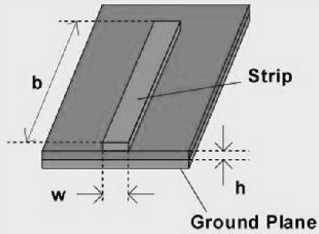
Inductance

**Fig 10: The straight wire inductance calculation page.**



The figure to the right shows the dimensional requirements for calculating the inductance of a *Flat Strip* over a *Ground Plane*. The *Flat Strip* is usually etched into a printed circuit board and the *Ground Plane* is usually a continuous layer of copper on the opposite side of a printed circuit board.

This program uses formulae published in the ARRL Handbook For Radio Amateurs. The equation listed below is not as messy as the one for a Straight Wire over a Ground Plane, but it's still requires a lot of work if you don't have a programmable calculator or know how to use one. Just enter your data in the spaces provided below and let the web page do the calculation. All input data is converted to inches for calculation purposes, however, for entry purposes, the dimenches can be mixed or matched.



$$L = 0.00508 \cdot b \cdot \left\{ \ln \left[ \frac{2 \cdot b}{w + h} \right] + 0.5 + 0.2235 \cdot \frac{w + h}{b} \right\}$$

Where:

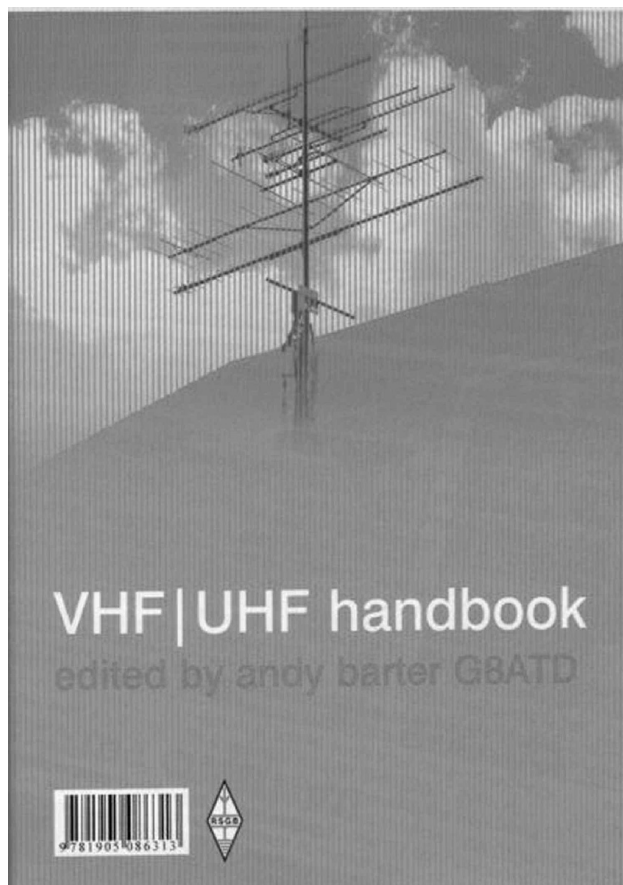
- L = Inductance in uH
- K = 0.00508 (dims in inches)
- 0.0002 (dims in mm)
- b = length (in or mm)
- w = width (in or mm)
- h = thickness (in or mm)
- ln = Natural Logarithm
- 2.303 \* Ln<sub>10</sub>

Strip Width (w) (in, or mm)	<input type="text" value="0.25"/> <input type="button" value="inches"/>
Strip Length (b) (in, or mm)	<input type="text" value="3.000"/> <input type="button" value="inches"/>
Thickness (h) (in, or mm)	<input type="text" value="0.05"/> <input type="button" value="inches"/>
Inductance	<input type="text" value="0.054 uH"/>

**Fig 11: The flat strip inductance calculation page.**



## The new VHF/UHF handbook



Edited by Andy Barter, G8ATD

The VHF/UHF Handbook is the guide to theory and practice in the VHF and UHF bands.

The VHF/UHF Handbook is packed with content from around the world. This book provides the reader with a wide range of interesting operating modes and techniques alongside the VHF/UHF basics. There are many topics covered including getting started, antennas and constructing your own equipment. There is truly something for everyone, from rag

chews on local nets or repeaters to specialist modes such as Amateur Television (ATV), Satellite operation is covered, as is the thrill of DX via one of the rare propagation modes or EME. There is much more in this great book, with its contents offering the best practice in every field.

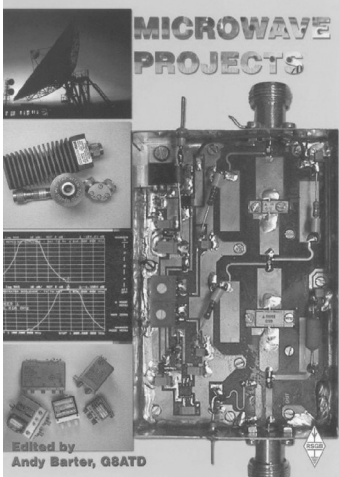
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## Microwave Projects Books edited by Andy Barter, G8ATD



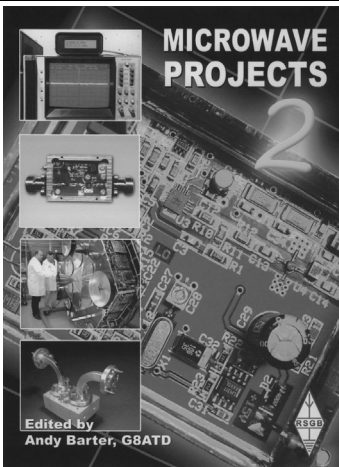
Microwave Projects is aimed at those who are interested in building equipment for the amateur radio microwave bands.

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

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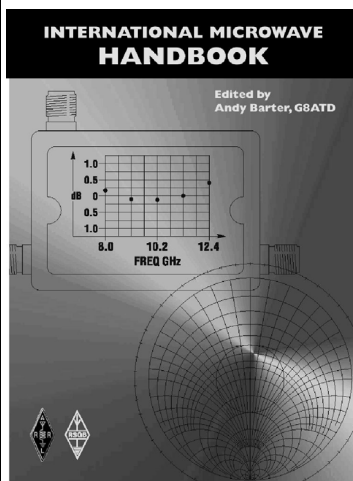
- Transverters and transmitters
- Receive amplifiers
- Filters
- Miscellaneous, including rare articles on using YIG oscillators and microwave absorbers.
- Power amplifiers

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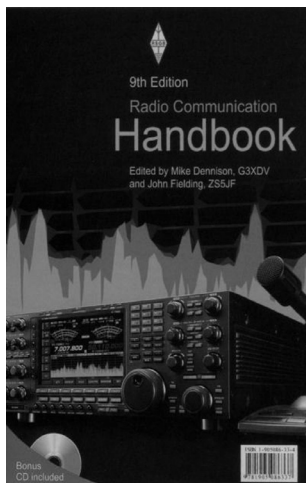


The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area, along with excellent designs from around the world 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.

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*Gunthard Kraus, DG8GB*

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## Microwaves & RF

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This is an RF magazine with interesting articles and the occasional review. The “White Paper” and other documents are worth registration.

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

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## Design Tools

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The Linear Technology web site is always worth a look . Under the Design Support section there are Application Notes and support. There is also a section for Design Simulation that contains a range of software.

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

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## Circuit simulator “Qucs”

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This is an open source circuit simulator with the nice name “Qucs”; it stands for Quite Universal Circuit Simulator. The simulation tool is still being developed but should be no competition for PUFF

but is an alternative with som potential.

Address:

<http://sourceforge.net/projects/qucs>

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## The four metres web site

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This site has all you want to know about the four metre band. This is becoming more available as the band II TV stations are switched off in favour of the digital channels.

Address:

[http://www.70mhz.org/history\\_g5um.htm](http://www.70mhz.org/history_g5um.htm)

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## Echolink

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This site has a introduction to Echolink the new digital mode for radio amateurs. There are some rigs being produced now with Echolink facilities so it is as well to know what it is all about.

Address:

<http://www.g3vfp.org/interlink.html>



## ADT-200

The ADT-200 is an all-digital transceiver designed using the principles of a software defined radio (SDR). The only exceptions to digital components are the preselector and the transmitter output stage that uses high voltage MOSFETs. It was developed by the Swiss amateur Hans Zahnd, HB9CBU and is probably the world wide first all digital amateur transceiver to work independently (no PC!).

The ADT-200 has four independent receivers and eight internal VFOs for many split frequency options. Only three rotary controls and 18 keys switches are used to operate the transceiver. All relevant information appears on two high visibility displays. There are options for 2m or 70cm modules.

The ADT-200 is going to be available

from the end of the 3rd quarter of 2007. Further information is available from

Address: [www.adat.ch](http://www.adat.ch)

## Notification

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

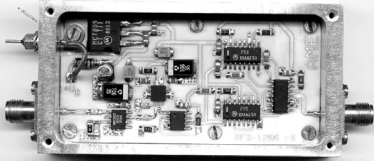
We wish to point out that neither the compiler nor the publisher has any liability for the correctness of any details listed or for the contents of the sites referred to!



**The ADT-200 all-digital amateur radio transceiver.**



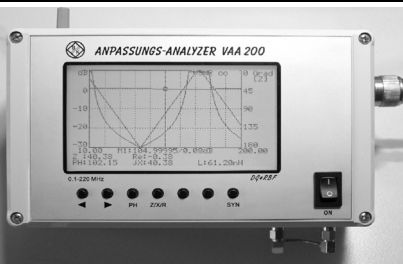
## Ready built units, Kits and PCBs available for recent projects



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

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

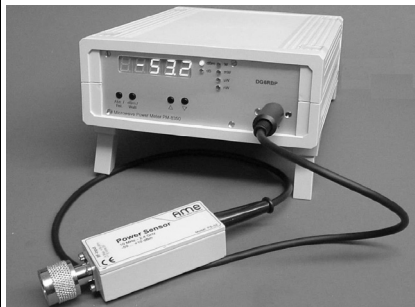
Also available an 18GHz **£230.00**  
version of the divide by 1000 prescaler



**Vector analyser VAA 200 for the 0.1 to 220MHz range with graphical display by Bern Kaa, DG4RBF from issue 3/2007**

PCBs **£31.50**

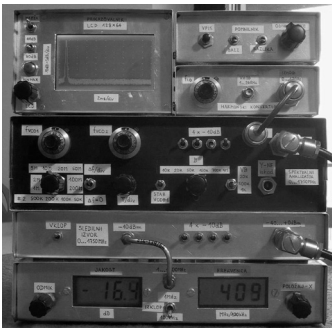
Micro-controller with software **£95.00**



**Display unit for the power detector by DJ8ES, Alexander Meier by DG6RBP from issue 1/2007**

PCB **£25.00**

Micro-controller with software **£65.00**



**Spectrum Analyser by Matjaz Vidmar, S53MV**

This project was first published in issue 4/1998 to 4/1999 (now included on CD1, see back cover) The LCD oscilloscope is in this issue.

Complete set of PCBs for all modules including the LCD oscilloscope **£70.00**

To order use either the web site [www.vhfcomm.co.uk](http://www.vhfcomm.co.uk) or contact  
K M Publications: Tel +44 (0)1582 581051, email [andy@vhfcomm.co.uk](mailto:andy@vhfcomm.co.uk)



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*Especially Covering VHF, UHF and Microwaves*

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## Blue Binders

These binders hold 12 issues (3 years) and keep your library of VHF Communications neat and tidy. You will be able to find the issue that you want easily. Binders are £6.50 each + postage. (UK £1.40, Surface mail £1.60, Air mail to Europe £2.00, Air mail outside Europe £3.60)

## PUFF Version 2.1 Microwave CAD Software

This software is used by many authors of articles in VHF Communications. It is supplied on 3.5 inch floppy disc or CD with a full English handbook. PUFF is £20.00 + postage. (UK £0.60, Surface mail £1.30, Air mail to Europe £1.50, Air mail outside Europe £2.50)

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Each CD is £10.00 which includes 2<sup>nd</sup> class postage in The UK and surface mail overseas. Air mail postage is £0.60 for Europe and £1.00 outside Europe.

## VHF Communications Web Site [www.vhfcomm.co.uk](http://www.vhfcomm.co.uk)

Visit the web site for more information on previous articles. There is a full index from 1969 to the present issue, it can be searched on line or downloaded to your own PC to search at your leisure. If you want to purchase back issues, kits or PUFF there is a secure

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