VHF
COMMUNICATIONS

13GHz prescaler
Zeljko Bozic, S52ZB

Contents

Good articles again in this issue including an interesting 13GHz prescaler from Zeljko Bozic and a 12mm DX report sent to me by Ted Barnes from New Zealand. I have removed the list of kits and PCBs from the back page because it is getting more and more difficult to supply these items. If you are interested in an item please phone me or send me an email, I will always try my best to find what you want.

73s - Andy

Alexander Meier, DG6RBP

Down Converter for YIG oscillators (10MHz to 2GHz)

Using this downconverter in combination with a YIG tuned oscillator (YTO, 6 - 8GHz) and a 6GHz local oscillator, one can convert the frequency range of the YTO to 10MHz to 2GHz. This is useful when building a sweeper or a signal generator. The output signal is amplitude regulated and harmonics are well suppressed.

1.

Introduction

YIG oscillators are used for Wobblers and signal generators because of their special characteristics, like linear tuning and wide frequency range. Suitable circuits using YIG oscillators have already been published in [1,2].

If you have a YIG oscillator from 6 to 8GHz, it's frequency can be converted to a range of 10MHz to 2GHz with the circuit shown here. For example, with a 2 to 8GHz YIG oscillator, the downconverter and a frequency doubler (like the HMC204 [3]), it is possible to realise a wobbler from 10MHz to 16GHz.

2.

Circuit Description

The circuit diagram of the Down Converters for YIG oscillators is shown in Fig 1. The output signal of the YIG oscillator (6 to 8GHz, approximately 40mW) is mixed with a local oscillator of 6GHz (1mW) e.g. the PLL oscillator [4] and produces an output from 10MHz to 2GHz. Subsequently, a low-pass filter suppresses other mixing products and the oscillator signal. The signal fed to the mixer (U1) is regulated using the attenuator (U6 1-32dB) to reduce unwanted mixing products by using a suitable power level for the mixer. Also it is used for the alc (automatic level control). Two integrated amplifier stages (Gain block) (U2) and (U3) amplify the down conveted signal by approximately 40dB. Part of output is sampled with the diode detector (D1) and used to control the output amplitude (ALC) (U5D) by feeding the attenuator (U6). The trimmer (R17) sets the output power output from approximately 1 to 20mW. The operation amplifier (U5B) keeps the impedance of the attenuator mostly independent of the attenuation level. If the amplitude control loop goes out of it's range of control, the LED (D8) (unleveled) is illuminated.
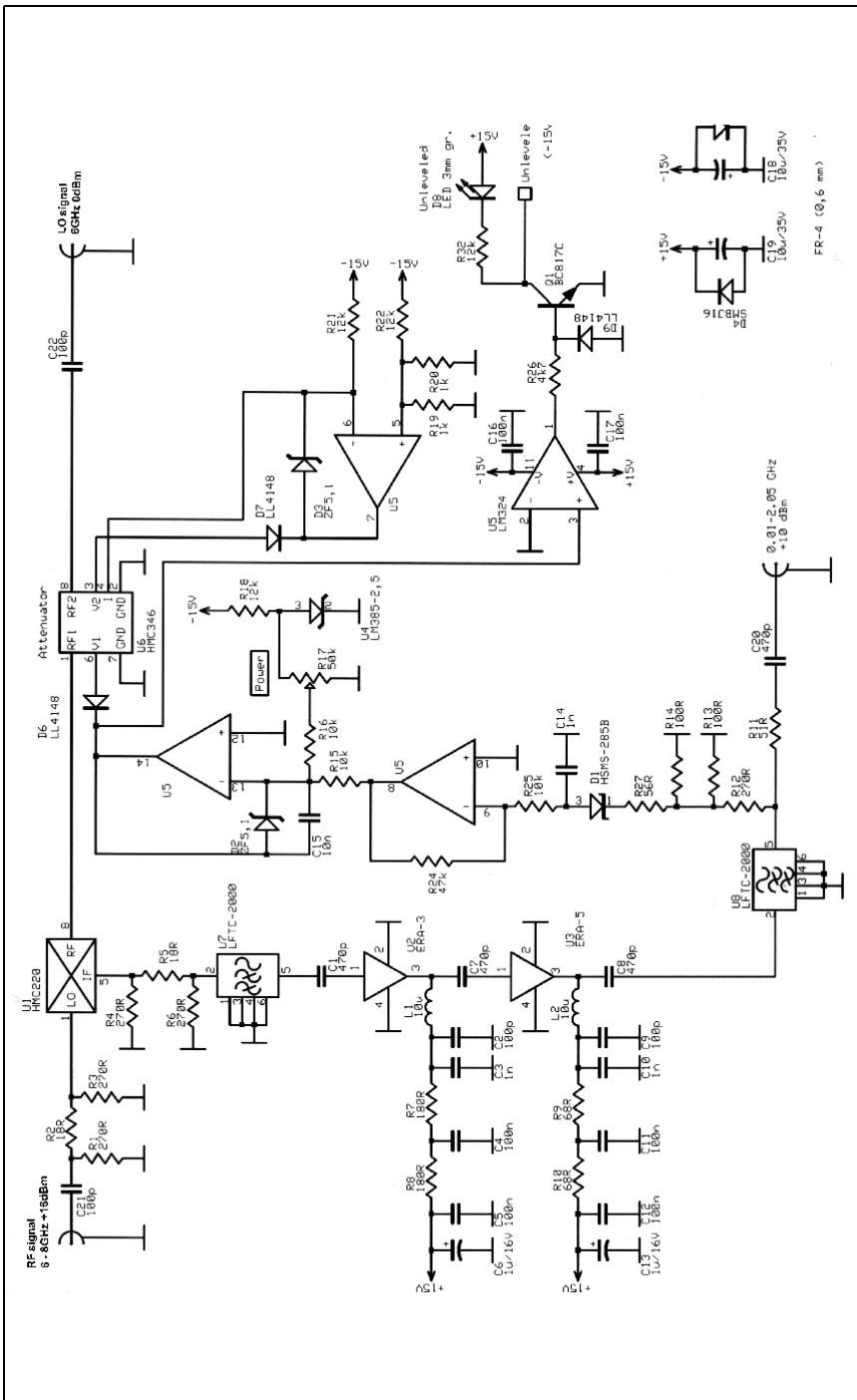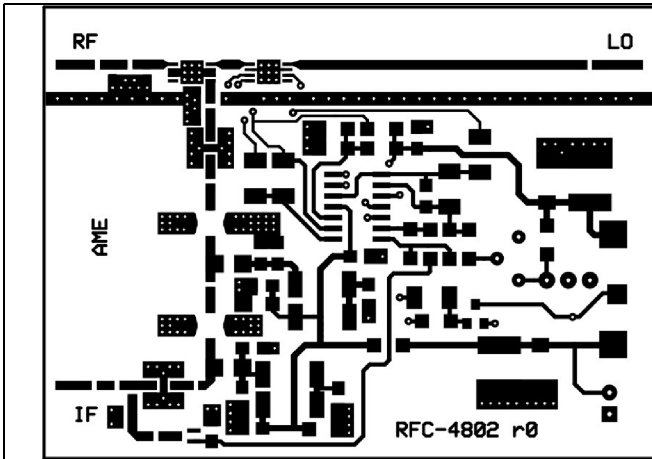
Fig 1: Circuit diagram of the Down Converter for YIG oscillators.

Fig 2: Top side of Printed Circuit Board for the down converter for YIG oscillators.

The technical data of the converter are summarized in Table 1.

3.

Construction

The PCB layout is shown in Figs 2 and 3, and the component layout is shown in Fig 4. The printed circuit board material used is a 0.6mm thick Epoxy material (FR-4).

First put some solder paste onto the component pads of the attenuator (U6) and the filters (U7 and U8). Then place these components and solder them using a hot-air station. Subsequently, all further components are fitted and soldered by hand, according to the component layout shown in Fig 4.

When the printed circuit board is completed and the flux is cleaned off, it should be soldered into the prepared (drilled) tinplate housing. The bottom side of the pcb must be soldered completely on all sides, in particular around the SMA sockets where there should be
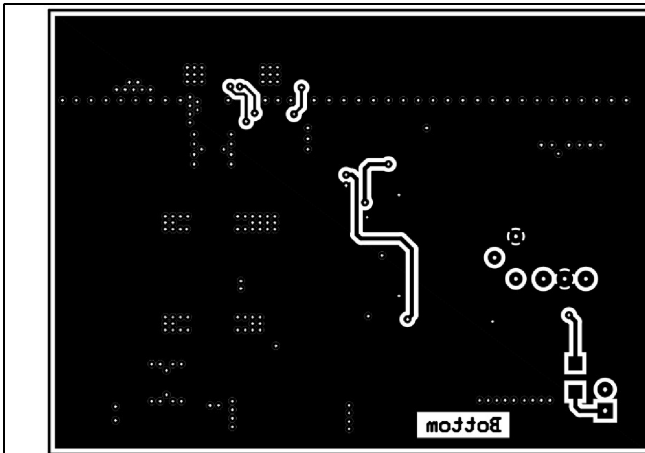
Fig 3: Bottom side of Printed Circuit Board for down conveter for YIG oscillators.
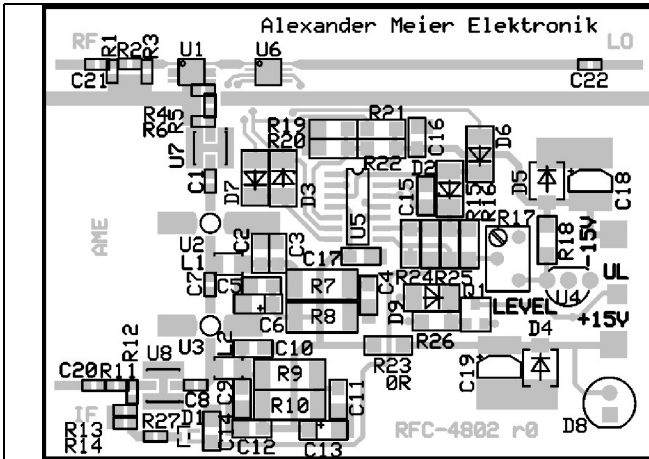
Fig 4: Component layout for the down converter for YIG oscillators.

no gaps. On the top surface the gnd bar is soldered to the housing. The installation of a partition is not necessary.

4.

Parts list

Semiconductors:

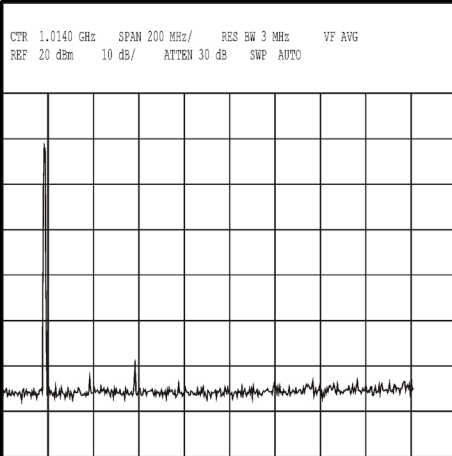|	
Q1	BC817C
U1	HMC 220 MS8 (Mixer,Hittite)
U7,U8	LFTC-2000 (Filter, Mini-Circuits)
U2	ERA-3 (Gain block, Mini-Circuits)
U3	ERA-5 (Gain block, Mini-Circuits)
U4	LM385-2.5
U5	LM324 SMD
U6	HMC 346 MS8G (Attenuator, Hittite)
D1	HSMS-285B
D2,D3	ZF 5.1
D6,D7,D9	LL4148
D4,D5	SMBJ16
D8	LED 3mm green, low current

Other Items:

3x1nF	feedthrough capacitors	
3x SMA Female (microstrip tab)		
1x tinplate housing	55 x 74 x 30mm	
1xPCB DG6RBP	Down Converter	
R1,R3,R4,R6	270R	0603
R2,R5	18R	0603
R11	51R	0603
R27	56R	0603
R9,R10	68R	2512
R7,R8	180R	2512
R12	270R	0603
R13,R14	100R	0603
R19,R20	1K	1206
R26	4.7K	1206
R15,R16,R25	10K	1206
R18,R21,R22,R32	12K	1206
R24	47K	1206
R17	50K	trimmer
C21,C22	100pF	0603
C2,C9	100pF	0805
C20	470pF	0805
C3,C10,C14	1nF	0805
C15	10nF	0805
C4,C5,C11,C12, C16,C17	100nF	0805
C6,C13	1µF/16 V tantalum SMD	
C18,C19	10µF/35 V Elko SMD	
L1,L2	10µH SIMID 1210	

CTR 1.0140 GHz   SPAN 200 MHz/   RES BW 3 MHz   VF AVG
REF 20 dBm   10 dB/   ATTEN 30 dB   SWP AUTO

Fig 5: The spectrum of the output signal at 100MHz (10mW).

CTR 1.0140 GHz   SPAN 200 MHz/   RES BW 3 MHz   VF AVG
REF 20 dBm   10 dB/   ATTEN 30 dB   SWP AUTO
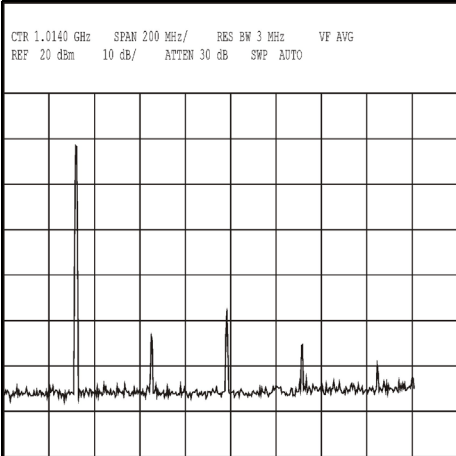
Fig 6: The spectrum of the output signal at 320MHz (10mW).

5.

Results of measurement

The output spectra at 200MHz, 320MHz and 800MHz are shown in Figs 5, 6 and 7. Harmonics and the input signal are suppressed by approximately 50dB, within a small range around approxi-

mately 300MHz the suppression of harmonics is only about 35dB. The spectrum were measured with a power output of 10mW. With lower power outputs the suppression is even better.

Although it would be possible to operate the mixer of the converter with other input frequency ranges (e.g. YIG 4-6GHz and LO 4GHz) this is not recommended. The good suppression of the harmonics is only achieved when the signal of the yig oscillator is mixed with 6GHz.

The finished prototype is shown in fig. 8.

CTR 1.0140 GHz   SPAN 200 MHz/   RES BW 3 MHz   VF AVG
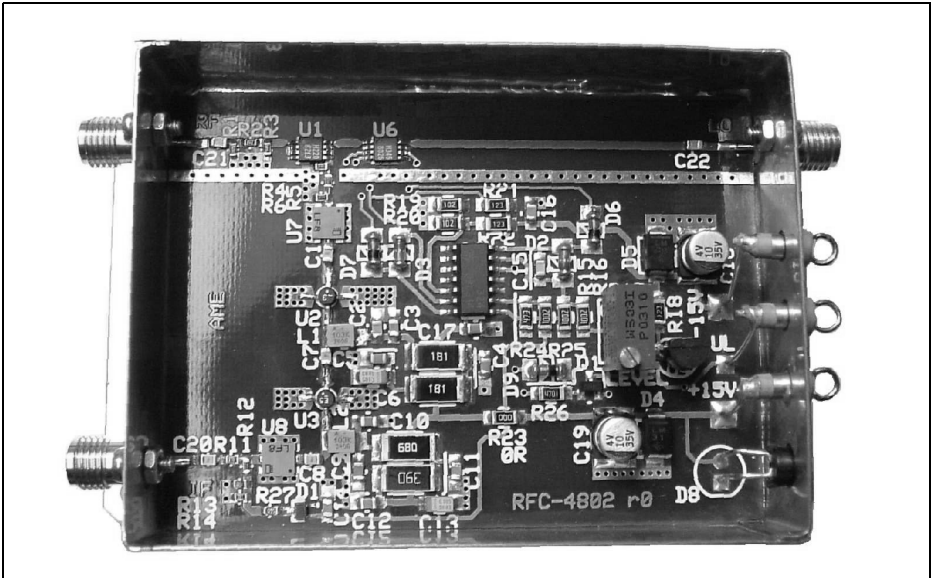REF 20 dBm   10 dB/   ATTEN 30 dB   SWP AUTO
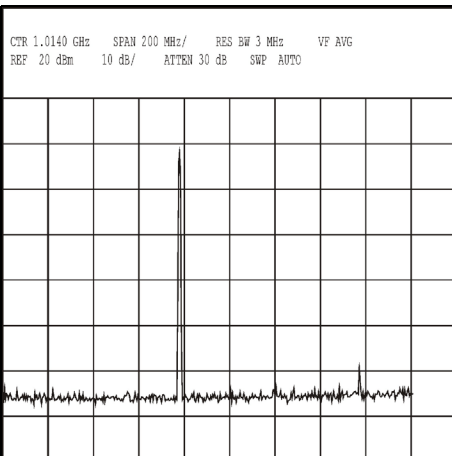
Fig 7: The spectrum of the output signal at 800MHz (10mW).

6.

Literature

[1] Frequency Generator (Wobbler) to 4GHz, Wolfgang Schneider, DJ8ES, VHF Communications Magazine 1/2002 pp. 2 -16

[2] Synthesised signal generator for 10 to 1800MHz, Bernd Kaa, DG4RBF, VHF

Fig 8: Picture of the completed prototype of the down converter for YIG oscillators.

Communications Magazine 2/2004 pp. 66 - 94

[3] Data sheet HMC-204, Hittite Microwave corporation, 20 alpha Road, Chelmsford, MA01824

[4] Universal PLL oscillator module, Alexander Meier, DG6RBP, VHF Communications Magazine 3/2005 pp. 156 - 163

Table 1: Technical data for the down converter for YIG oscillator.

Parameter		Min.	Typ.	Max.
RF input	Frequency range	6GHz		8GHz
	Input power		+16dBm	
LO input	Frequency range		6GHz	
	Input power		0dBm	
RF output	Frequency range	10MHz		2GHz
	Output power *	+5dBm		+13dBm
	ALC		±0.3dB	
	Harmonics	-35dBc	-50dBc	
Impedance			50Ω	
Supply voltage		±13.5V	±15V	±16.5V
Supply current (+15V)			130mA	150mA
Supply current (-15V)			15mA	25mA
Housing			Tinplate	
Dimensions			74 x 55 x 30mm	

* Note, Output power can be set using trimmer R17

Wolfgang Schneider, DJ8ES

Selective wattmeter

To measure the performance of SSB transmitters or amplifier stages the usual method is to use a two tone signal and measure the intermodulation products. To be able measure the individual spectral lines that are close together, requires either a high quality spectrum analyser or a selective measurement receiving device with a suitable level display. The following article describes a narrow band level meter with crystal filter.

1.

Introduction

A selective wattmeter makes it possible to take narrow band measurements of individual sections and/or ranges within a wide frequency spectrum. When it is used with a suitable oscillator, e.g. the DDS oscillator described in [3] the frequency spectrum up to 50MHz can be measured.

2.

Circuit description

The circuit of the selective wattmeter is roughly divided into two parts. One part consists of a wide band ring mixer (Fig 1) and the other part is a crystal filter followed by a logarithmic detector (Fig 2).

The IE500 diode ring mixer (IC1) has good intermodulation behaviour when matched to 50Ω. The RF port has a 10dB attenuator and the LO port has a 3dB attenuator. The IF port feeds a 10.7MHz pass band filter followed by a standardising amplifier for 50Ω using a BFR90a transistor (T1) that gives approximately 23dB gain.

The input signal that has been mixed down to the IF frequency is fed to a narrow band crystal filter. This is a 4 pole filter, this type of filter has already been described in [2].

The logarithmic level sensor uses the familiar AD8307 by Analog Device (IC2). This has a dynamic range of about 90dB. The response is not completely linear with frequency but this is not relevant for this application because only
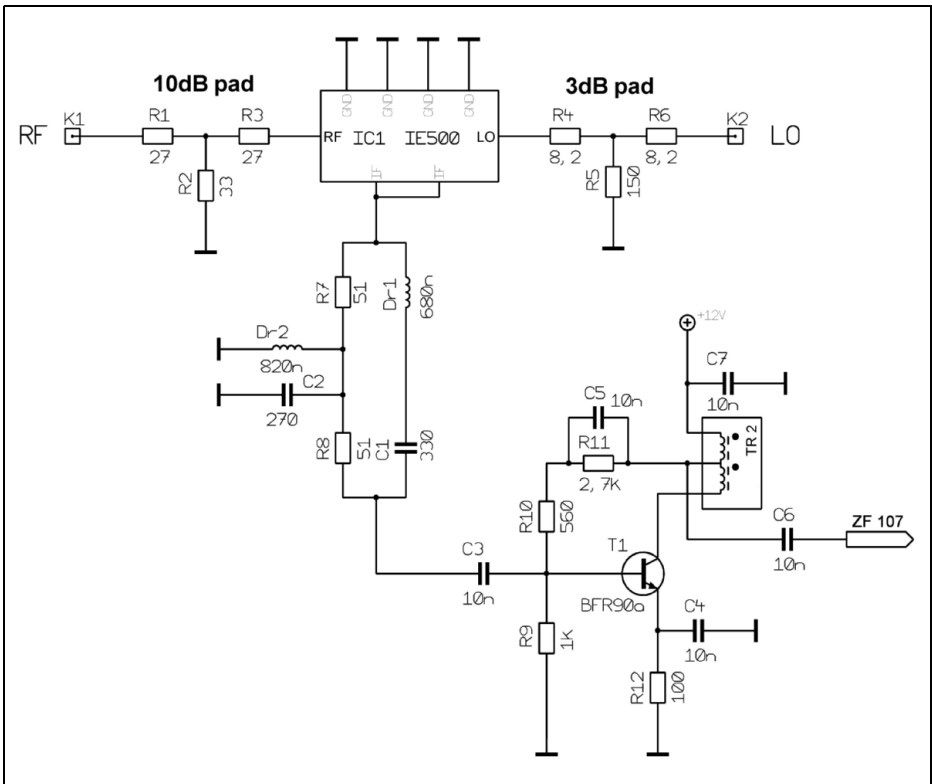
Fig 1: Circuit diagram of the mixer, filter and amplifier.

narrow band measurements are being made at 10.7MHz.

The output of the AD8307 is a linear dB scale (pin 4) with a slope of 20mV/dB and is high impedance. Two buffer amplifiers using the dual OP amp NE5532 (IC3a, IC3b) have a fixed gain set by the feedback resistors. These provide two loadable outputs of 100mV/dB.

The 50KΩ trimmer R19 adjusts the output linearity and the 50KΩ trimmer R20 adjusts the offset of the overall system, details of this are given in the set up section.

3.

Construction

The Printed Circuit Board (DJ8ES 073) is a double sided epoxy board measuring 72mm x 72mm. It fits into a standard 74mm x 74mm x 30mm tinplate housing.
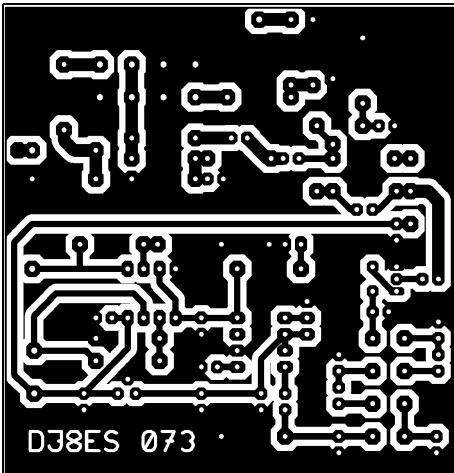
The PCB layout is shown in Figs 3 and 4. The top side (component side) serves as a ground surface and should be soldered, all round, to the housing. The wired components are fitted on the components side according to the layout shown in Fig 5. All ground connections of the wired
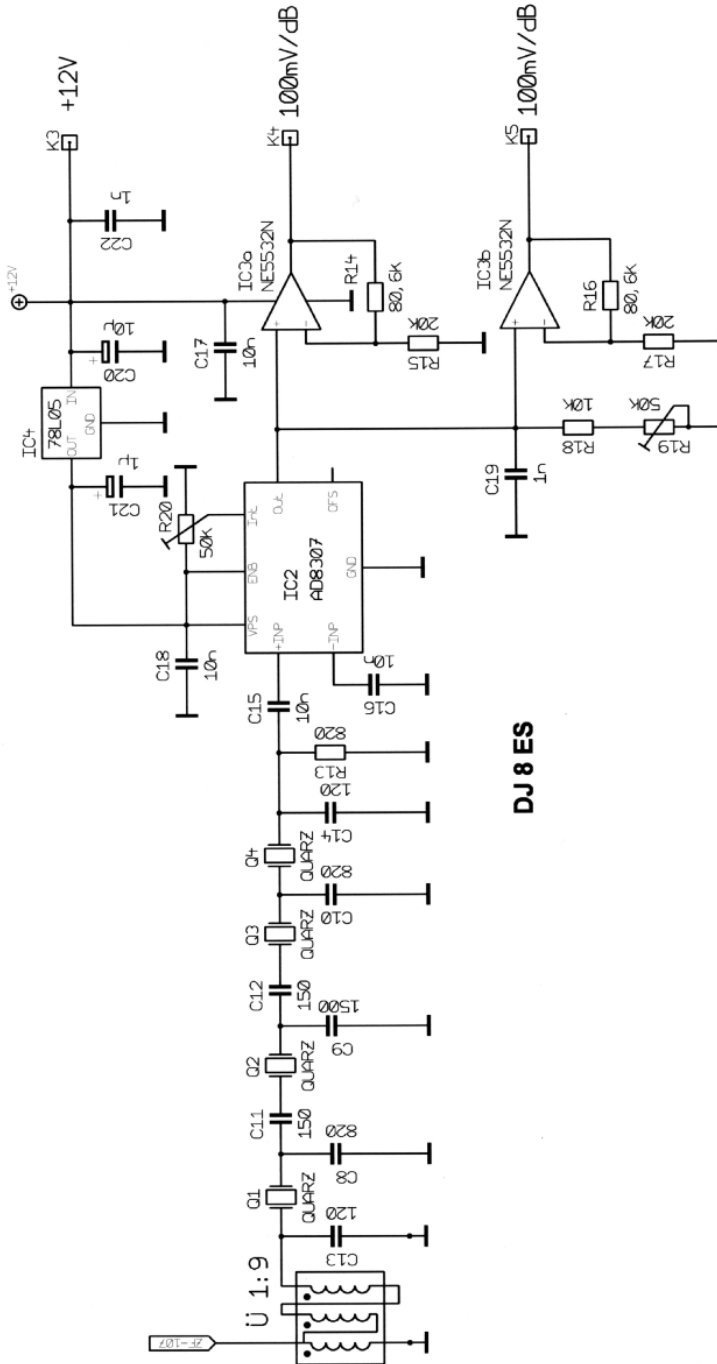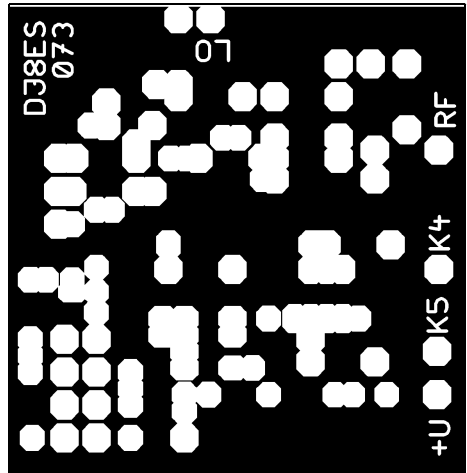
Fig 2: Circuit diagram of crystal filter and logarithmic amplifier.

Fig 3: Bottom side of PCB layout.

Fig 4: Top side of PCB layout.

components are soldered on both sides of the PCB. Also the ring mixer is soldered to the ground surface all around it's housing.

SMA sockets are used for the RF and LO inputs because of their small size but other sockets such as BNC can be used if required. The supply voltage (+12V) is fed in using a solder-in 1nF feedthrough capacitor with the ground connection made where the feedthrough is soldered to the housing. The outputs for the measured voltage use two feedthrough capacitors.

4.

Parts list

IC1	IE500	ring mixer
IC2	AD8307	IC log. Detector
IC3	NE5532N	operation amplifier
IC4	78L05	voltage regulator
T1	BFR90a	transistor
Q1 - Q4		10.7MHz, quartz filter
		See text

TR1	transformer	trifilar
TR2	transformer	bifilar
Dr1	680nH	axial choke, RM 10mm
Dr2	820nH	axial choke, RM 10mm
R19,R20		precision trimmer
		50KΩ
C20	10μF	tantalum
C21	1μF	tantalum
2 x SMA		flange socket
3 x 1nF		solder in feedthrough
1 x solder point 4mm		
1 x PCB		DJ8ES 073

Ceramic capacitors, EGPU, RM 2.5mm:

C1	330pF
C2	270pF
C3 - C7	10nF
C15 - C18	10nF
C8,C10	820pF
C9	1500pF
C11,C12	150pF
C13,C14	120pF
C19	1nF

1/4W resistors RM 10mm:

R1,R3	27Ω
R2	33Ω
R4,R6	8.2Ω
R5	150Ω
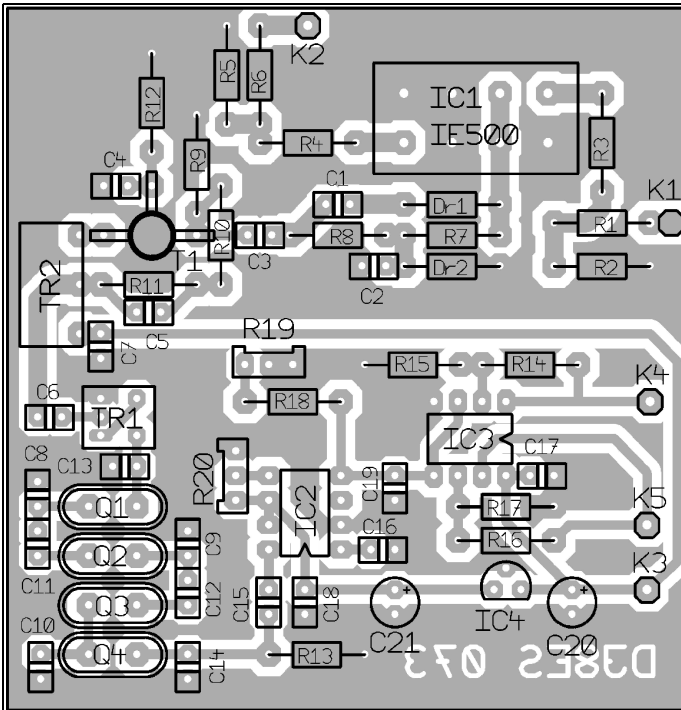
Fig 5: Component layout for the selective wattmeter.

|	
R7,R8	51Ω
R9	1KΩ
R10	560Ω
R11	2.7KΩ
R12	100Ω
R13	820Ω
R14,R16	80.6KΩ
R15,R17	20KΩ
R18	10KΩ

5.

Test and alignment

When the supply voltage of +12V is applied the supply current should be approximately 40mA with no input signal. Use two multi-range digital meters to measure the outputs, these should show a small reading which represents the background noise of the circuit.

Using a suitable variable oscillator, such as the DDS oscillator [3] and a test signal, for example a 0dBm (1mW) 10MHz signal with an external 10dB attenuator. The DDS oscillator is adjusted to 10.697825MHz that is the IF offset, i.e. the centre frequency of the crystal filter, plus the test signal.

With the trimmer R19 (50KΩ at pin 4 of the AD8307) the linearity of the overall system can be adjusted. The voltage difference between the test signal with and without the 10dB attenuator must be accurately set to 1V. This is measured on one of the two 100mV/dB outputs. The easiest method is to use an adjustable attenuator. It is the difference in voltage levels that is being measured, the actual level is not relevant.

The offset of the input signal is adjusted last. With a 0dBm test signal, use the trimmer R20 (50KΩ at pin 5 of the

4 Pole ladder filter (f = 10,697.825Hz)
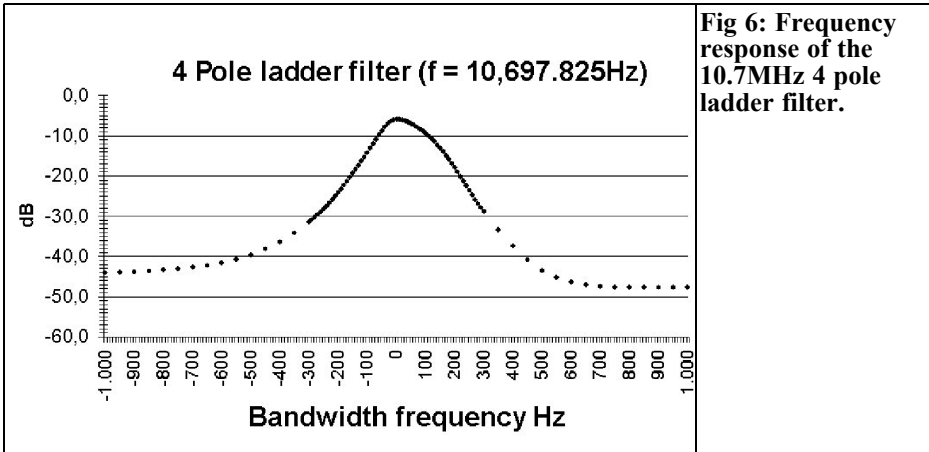
Fig 6: Frequency response of the 10.7MHz 4 pole ladder filter.

AD8307) set the 100mV/dB output to exactly 7.0V.

6.

Operational experiences

The frequency of the suggested DDS oscillator can be adjusted in 1Hz, 10Hz, 100Hz and 1kHz steps. In practice the 100Hz steps worked satisfactorily. The important fact is that the steps must be small compared to the bandwidth of the crystal filter.

For two tone test measurements signals with a separation of 1000Hz can be measured easily. So signals of f1 = 800Hz and f2 = 1800Hz can be used.

With an external attenuator of 10dB on the measuring input, the 1dB compression point is about +9dBm. The dynamic range of the total circuit of the selective wattmeter, including the DDS oscillator, is somewhat more than 45dB.

Despite far better performance of the AD8307 logarithmic amplifier the dynamic range is reduced by:

1 The relatively bad signal of the DDS oscillator:
The alias products produced by the DDS oscillator cannot be completely eliminated by the low pass filter built into the oscillator. The output still has signals with a maximum level of –50dBc. A combination of finely tuned DDS oscillator and a PLL oscillator would be more elegant, however this would increase the complexity of the system.

2 Cross talk within the compact unit:
Using RF separation (e.g. chamber building method) of the individual functional modules such as mixers, crystal filters and detector, the cross modulation could be avoided as far as possible and an improvement around approximately 10dB is realistic.

A 100µA meter can be used to display the output. This is fed via a 100KΩ trimmer used to set the sensitivity. The scale will be linear in dBm. The zero position corresponds to a value of -50dBm and full scale +10dBm.

The background noise of the total circuit is somewhat over -50dBm. It will always be visible even with no input signal.
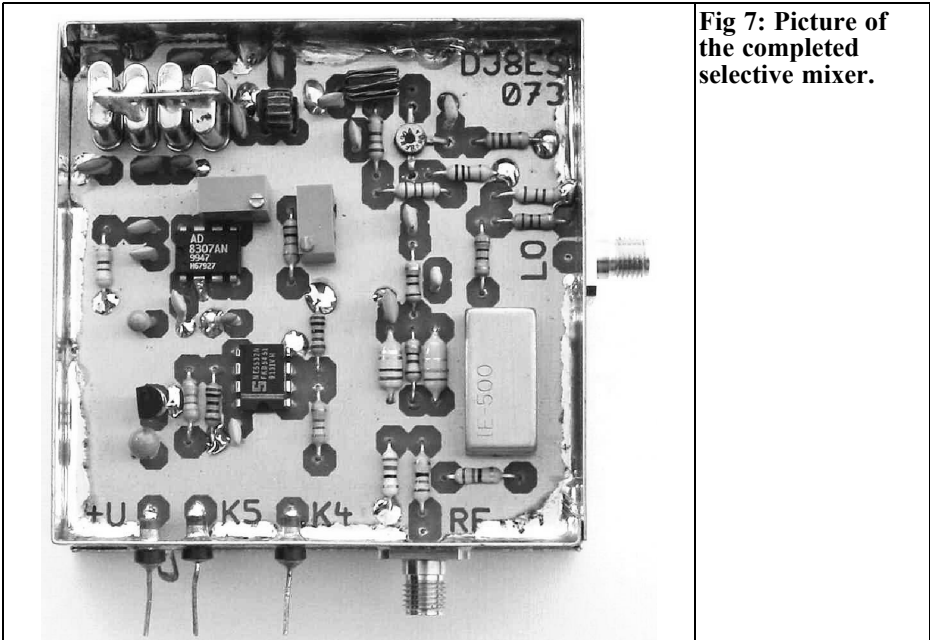
Fig 7: Picture of the completed selective mixer.

7.

References

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

[2] Design and assembly of a simple 4 pole ladder filter, Wolfgang Schneider, DJ8ES, VHF Communications Magazine 1/2006 pp. 2 - 6

[3] DDS oscillator for QRP experiments on short wave and 6m band, Wolfgang Schneider, DJ8ES, VHF Communications Magazine 1/2005 pp. 44 - 52

Professor Dr. Ing. Gerd Janzen, DJ6SJ

Reciprocal effects between antennas and surrounding metal objects, Part 1

Antennas should hang as freely as possible and be located as far away from any disturbing objects as possible. This statement actually applies to any kind of antenna and in particular to amateur radio antennas. However radio amateurs are usually the worst case for complying with these requirements.

In the following article the influence of metal objects on antennas in some simple arrangements is calculated and interpreted with the help of the antenna simulator EZNEC.

1.

Introduction

Every radio amateur knows that an antenna should be mounted as high as possible and in the area far away from disturbing objects. If you examine the environment of a normal amateur radio antenna, then it is quickly apparent that this is very rarely "ideal ". Problems can be: the ground is too close; a roof is too close; a tree is too close; even worse other transmitting or television aerials are too close; a gutter or down pipe is too close; a chimney is too close; the metal frame of a window is too close; the metal

frame of a greenhouse is too close; a lighting conductor is too close, an overhead cable or power line is too close… These are surely not all the possible disturbances that can affect the radiation field of an antenna.

However one should also consider that not all objects have negative affect on antennas. It is well known that soils or flat roofs are to some extent conductive and they can have a very positive affect on the radiation pattern of an antenna providing a welcome lowering of the main beam angle.

In order to get "a feeling" for the affect that metal objects exert on a transmitting
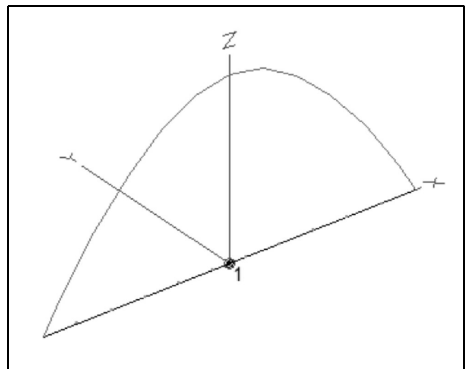
Fig 1: Diagram of the dipole on the x axis showing the current distribution with a maximum on 1A.

antenna and at what distance from the antenna they still have an affect, some simple arrangements from a dipole and/or a directional antenna and a disturbing wire and/or pole were simulated with the help of the familiar antenna simulation program EZNEC.

2.

Basic data for the simulation of a disturbed antenna with EZNEC

The antenna simulator program EZNEC is available as version 3 and, since May 2004, version 4 [1]. Anyone without a good knowledge of this program and its operation is referred to the literature [2, 3] for a compact introduction to EZNEC 3 and/or 4.

The programs EZNEC3 and EZNEC4 allow the use of 500 segments, and since May 2004 the extended plus version EZNEC+4 permits the use of up to 1500 segments. For the simulations shown here 22 segments are sufficient for the directional antennas and 44 segments for the dipole antennas. A good first impression of the program, with meaningful results, can be achieved using the free downloadable version with 20 segment capability [1].

The following basic adjustments should be made in the main menu of EZNEC:

UNITS are adjusted to "wavelengths". This has the advantage that all simulation calculations are to a large extent frequency independent and thus the results are valid for both the 160m Band and the higher frequency bands such as VHF and UHF.

GROUND TYPE is "Free space". The antenna system to be examined hangs in a perfectly free environment. This is necessary in order to decouple the disturbance introduced by metal objects from the conductivity of the soil and from reflections and avoid additional complications.

WIRE LOSS is adjusted too "Zero". The slightly absorbing influence of the wire material will not effect the investigations.

FREQUENCY is entered as 30MHz - however this is more a formality, since "wavelengths" are to be used. It is good to imagine "standardised wavelength data" in metres or millimetres.

The entry of the antenna structure follows. Emitter one defines a simple horizontal hanging wire dipole, which extends from a centre *(*coordinates x/y/z=0/0/0) about a quarter wavelength long in the negative and positive x direction. In the WIRES menu the point of origin is selected as 0.24/0/0, the end point is +0.24/0/0. Because "wavelengths" is selected as the unit, this means that the dipole is 2 x 0.24λ, thus not quite a half wavelength long. The wire size of 0.0005 is entered, which means at 30MHz a wire diameter of d = 0.0005λ = 5mm should be defined. For the 160m Band the wire would be thick according to this definition at 8cm, and for the 2m Band about 1mm.

Next the number of segments is entered; for the dipole wire 11 segments will be used (for reference: 9 or 7 segments could be enough, an odd number of segments should be used because this makes it possible to assign an exact centre).

For the menu option SOURCES we must enter; "Specified Position", enter the feed position; "Amplitude", enter the feed amplitude; "Type", enter the type of feed. This sounds complicated but it is not; use

Table 1: showing the current flow in each of the 11 segments of the dipole.

```
                        EZNEC+ ver. 4.0

   Dipol(0,48 lambda lang)                   13.06.2005    12:49:23

           --------------- CURRENT DATA ---------------

   Frequency = 30 MHz

   Wire No. 1:
   Segment  Conn       Magnitude (A.)  Phase (Deg.)
   1        Open        ,16779          -4,71
   2                    ,44815          -4,04
   3                    ,67808          -3,27
   4                    ,8526           -2,36
   5                    ,96212          -1,19
   6                   1                 0,00
   7                    ,96212          -1,19
   8                    ,8526           -2,36
   9                    ,67808          -3,27
   10                   ,44815          -4,04
   11       Open        ,16779          -4,71
```

50 (50% of wire 1, thus supply in the centre of the dipole wire, into segment 6); 1 (=1A feed amplitude); and I (current supply).

Thus a half wavelength dipole is defined, which is fed at the wire centre with a 1A supply. The feed current could be 7mA or 145A it in no way affects the function of the antenna. 1A has been chosen in order to see later how large the radiation induced current is in the disturbing wire in relation to the excitation current. The dipole antenna is defined along the x axis, it can be seen using the VIEW ANTENNA, Fig 1 shows this.

Now it is time to see the simulation results, to do this select "Src Dat" (SOURCE DATA) in the main menu of EZNEC. A new window appears "SOURCE DATA" with the feed point data of the dipole: apart from different numbers, which reflect the specified input (e.g. the current value 1A), the most important value for us is shown, this is the input impedance of the antenna. It shows that for the half wave dipole in free space the impedance ZA= (71.39 - j2.96)Ω. The real part of this is 71Ω that is the typical input resistance of a dipole [4]. The imaginary part denoted by the letter "j" is small and amounts to approximately -3Ω. This small value indicates that the antenna is nearly resonant, thus the power supplied by the transmitter will radiate well without so-called reactive power on the input causing reflections.

The minus sign on the imaginary part means that the antenna feed impedance is "capacitive", thus a small coil is needed with an inductive reactance of +j3Ω, in order to give the ideal value j0 and real resonance. Simply increasing the length of the antenna wire a few fractions of a wavelength would reach "natural" resonance. Since that is not so important for use here, the dipole length 2 x 0.24λ will be used.

The source data is shown for a 50Ω system with s = SWR = 1.433 but it can be read off for a 75Ω system s_{75} = 1.066 using the main menu option "alto SWR Z0" (alternative SWR for different reference resistance). In Fig 1 the current distribution is shown over the dipole wire with a maximum of 1A.

Under option "CURRENTS" a table is available which shows the current flow in
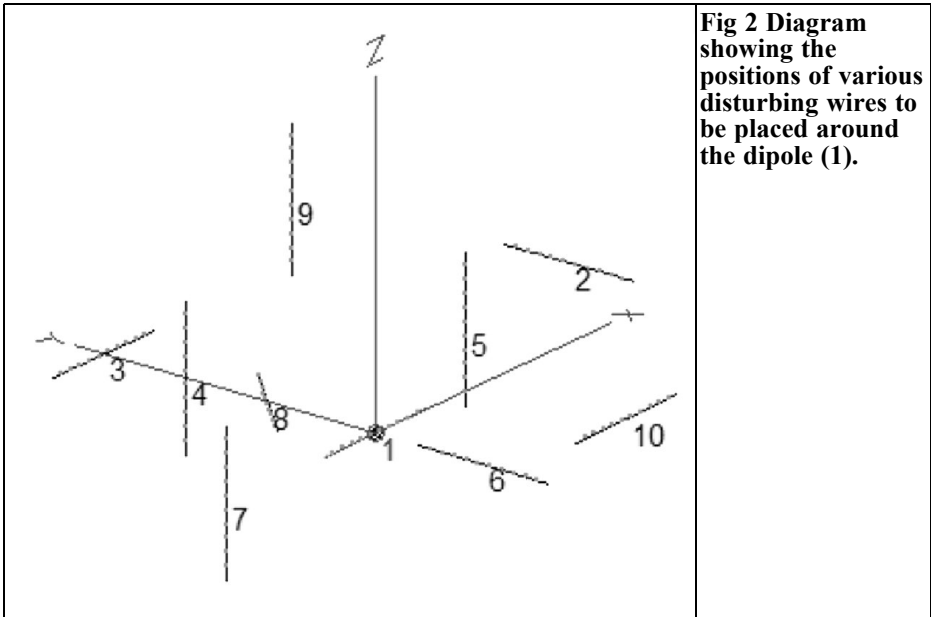
Fig 2 Diagram showing the positions of various disturbing wires to be placed around the dipole (1).

each of the selected 11 segments of the dipole, see Table 1. As expected the flow into segment No. 6, which is in the centre of the dipole wire, is exactly 1A, because it is the selected feed current. Towards the ends of the antenna the flow drops. Theoretically the current value would have to be zero at the wire ends. Here the final segment 1 and 11 show 0.168A because the current flow is always computed for the segment centre and not for the segment end.

3.

Dipole with disturbing wires in different arrangements

Now we will investigate the effects of fixed length wires 2 x 0.24λ = 0.48λ long placed at various places around the dipole in free space. A diagram of how these disturbing wires will be oriented is

shown in Fig 2. The centre fed dipole is "1" in the diagram with the y axis in the centre of the coordinate system, the x axis to the right and the z axis upward. The first disturbing wires examined will be arranged parallel to the dipole; wire 3 directly in propagation direction of the antenna; wire 10 parallel, but shifted in x-direction.

The next wires to be examined will be at right angles to the dipole, wires 2 and 6. Wire 8 is rotated 45° to the y axis as an example of this type of disturbance. All of the disturbing wires so far are in the x-y plane, at the end of this series of investigations vertically arranged wires are examined, wire 4 (directly in the propagation direction of the dipole), wire 7 (in the direction of the wire dipole) as well as wires 5 and 9.

The details of these wires is shown in Table 2 which is produced using the EZNEC menu option OUTPUTS - show Description. It is to be stressed that these 9 disturbing wires are not simulated all at once in the following description because

Table 2: The x/y/z co-ordinates of the disturbing wires used. The wires are all 0.48λ long.

```
                    EZNEC+ ver. 4.0

                               12.06.2005    11:07:34

             -------------- ANTENNA DESCRIPTION --------------

Frequency = 30 MHz
Wire Loss: Zero

                 -------------- WIRES --------------

No.       End 1   Coord. (wl)           End 2   Coord. (wl)        Dia (wl)  Segs  Insulation
     Conn.      X        Y       Z   Conn.      X        Y       Z                       Diel C  Thk(wl)
1         -0,24,     0,      0            0,24,     0,      0       0,0005   11      1        0
2          0,4,   -0,16,    0,5           0,4,   -0,64,   0,5      0,0005   11      1        0
3         -0,24,     1,      0            0,24,     1,      0       0,0005   11      1        0
4            0,    0,7,   -0,24             0,    0,7,    0,24      0,0005   11      1        0
5          0,8,    0,3,   -0,24           0,8,    0,3,    0,24      0,0005   11      1        0
6            0,  -0,16,      0             0,   -0,64,      0       0,0005   11      1        0
7         -0,7,     0,   -0,24          -0,7,     0,     0,24      0,0005   11      1        0
8         -0,17,   0,23,     0           0,17,   0,57,     0       0,0005   11      1        0
9          0,5,    0,7,    0,16           0,5,    0,7,    0,64      0,0005   11      1        0
10         0,3,   -0,5,      0            0,78,   -0,5,      0       0,0005   11      1        0

Total Segments: 110

                 -------------- SOURCES --------------

No.       Specified Pos.    Actual Pos.       Amplitude    Phase      Type
     Wire #  % From E1   % From E1  Seg        (V/A)      (deg.)
1         1      50,00       50,00     6         1           0          I

No loads specified

No transmission lines specified

Ground type is Free Space
```

it would not be possible to see the effect of a single wire.

Fig 3: Diagram of the current distribution in the dipole with a parallel disturbing wire 2 x 0.24λ long. The disturbing wire is shown at a distance of 0.3λ along the y axis. The maximum current in the dipole is 1A which induces a maximum of 0.628A in the disturbing wire.

3.1. Dipole with a parallel disturbing wire

First the centrally fed antenna wire is to be supplemented by a parallel disturbing wire with a length of 2 x 0.24λ = 0.48λ. Thus the disturbing wire is a resonant length, and its effect on the dipole will be a maximum, which will be shown in the following investigation. The coordinates of this second wire read:

End1: -0.24/y/0; End2: +0.24/y/0, where the value y ranges from very large (e.g. y = 100λ to small values (e.g. y = 0.005λ). The value y = 100λ means a distance of 100 wavelengths, in the 10m Band this gives a distance of the parallel wire from the dipole of 1km. The value y = 0.005λ represents the distance 0.005 x 10m = 5cm in the 10m Band. Wire sizes and number of segments are selected as with the dipole.

Fig 3 shows the disturbing wire, which is

Table 3: Impedance and SWR for the dipole with varying distance (y) from a parallel disturbing wire plus the maximum current in the disturbing wire with a constant 1A feed current to the dipole.

y distance in λ	Impedance Z_A in Ω	SWR for 50Ω	Max current in A
100	71.39 - j2.96	1.43	0
10	71.44 - j2.97	1.43	0.026
5	71.58 - j2.99	1.44	0.052
4	71.68 - j3.02	1.44	0.065
3	71.91 - j3.08	1.44	0.086
2	72.51 - j3.31	1.47	0.128
1	75.21 - j5.13	1.52	0.247
0.8	69.23 + j3.24	1.39	0.303
0.7	63.17 - j2.62	1.27	0.339
0.6	66.9 - j12.5	1.44	0.383
0.5	80.45 - j13.59	1.68	0.441
0.4	90.08 + j1.85	1.80	0.520
0.35	87.47 + j13.55	1.81	0.570
0.3	77.86 + j24.03	1.79	0.628
0.25	61.81 + j29.31	1.75	0.691
0.2	42.13 + j25.76	1.78	0.763
0.15	23.55 + j11.61	2.27	0.840
0.1	11.2 + j11.9	4.73	0.921
0.05	8.33 - j39.84	9.88	1.0
0.02	10 - j53.1	10.74	1.046
0.01	9 - j51.26	11.49	1.048
0.005	6.56 - j43.77	13.53	1.035

not fed, shifted in y the direction. This wire is sometimes called "passive", although its role will be not at all passive in the overall system.

In Fig 3, for a distance y = 0.3λ, a clear current distribution on the disturbing wire is shown, although this is not fed and is unconnected.

Note: EZNEC represents the current curves independently of the selected design size in a full-scale picture, so that relative current differences can be read off immediately.

The results for different y distances are summarised in Table 3. It shows; Impedance Z_A of the dipole, its standing wave ratio "s" in a 50Ω system and, most important, the current flow in the centre segment No. 6 of the disturbing wire. There is an effect on the dipole data when the disturbing wire distance is less than 2 wavelengths (y = 2λ), and current flows in the disturbing wire even at distances of 10 wavelengths. An appreciable current of 0.026A, 2.6% of the feed current, flows in the disturbing wire even with y = 10λ (in the 10m Band that is 100m distance, in the 160m Band it would be 1.6km!). Approximations under 1 wavelength (y < 1λ) the current flow rises to very high values, that can exceed even the dipole feeding current of 1A. It should be pointed out that the default of a constant 1A feeding current into the dipole would not be achieved in practice because of the varying impedance Z_A. Since in this investigation current flows are compared with current flows, this is unimportant.
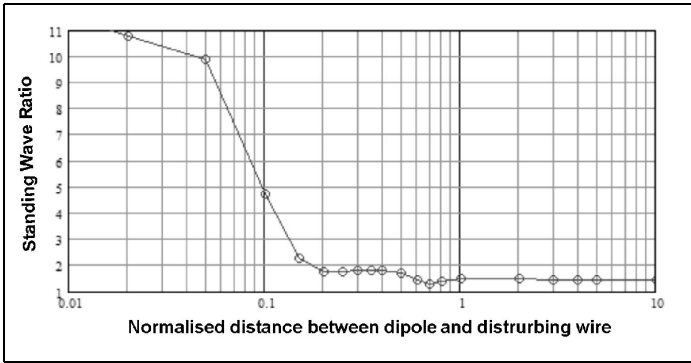
Fig 4: Graph of SWR for the dipole against distance from parallel disturbing wire.

A diagram shows the effects of the resonant disturbing wire better than a table. The standing wave ratio "s" over the distance y standardised by wavelength is shown in Fig 4. To make it easier to read off the effects a larger distances the y axis is logarithmic. The influence on standing wave ratio is not very large for distances from 10λ to 1λ, it increases from 1λ to 0.2λ and shows a clear influence less than 0.2λ.

The maximum current in the disturbing wire for different distance is shown on a similar logarithmic scale in Fig 5. The induced current is small with a large distance from the dipole and continuous increases as the distance reduces rising to over 1A. The energising current in the dipole is a constant 1A.

The influence of the disturbing wire on the radiation pattern is even more interesting than it's effect on the impedance and standing wave ratio of the dipole. With no parallel disturbing wire the familiar radiation pattern with a gain G = 2.15dBi = 0dBd results transverse to the dipole. dBi is the gain relative to an isotropic radiator, dBd is related to the gain of the simple dipole. Here the more universal measure dBi is used.

If the parallel wire is brought nearer to the dipole, then the symmetrical radiation diagram pattern deforms in the forward and reverse directions, when $y = 0.25\lambda$ a 2 element beam with a gain G = 6.39dBi in forward direction results, with a gain in the reverse direction Gr = 1.28dBi. From this we get the relationship V/R = 5.11dB, the parallel wire works as reflector. Fig 6 shows the radiation patterns of the simple dipole (symmetrical curve) and the beam curve when $y = 0.25\lambda$. The pure dipole radiation pattern is symmetrical with a gain G = 2.12dBi and has the same radiation forwards and backwards
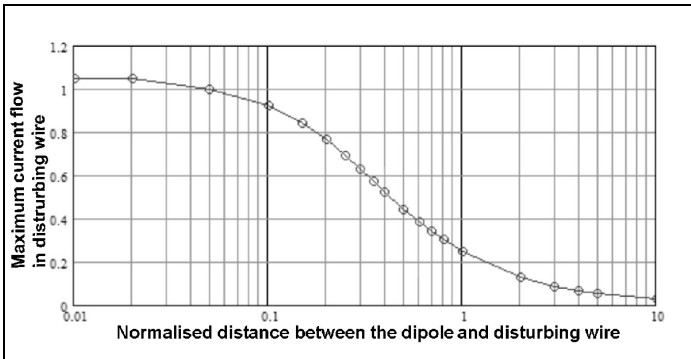
Fig 5: Graph of induced current in a parallel disturbing wire against the distance to the driving dipole which is driven with a constant 1A.
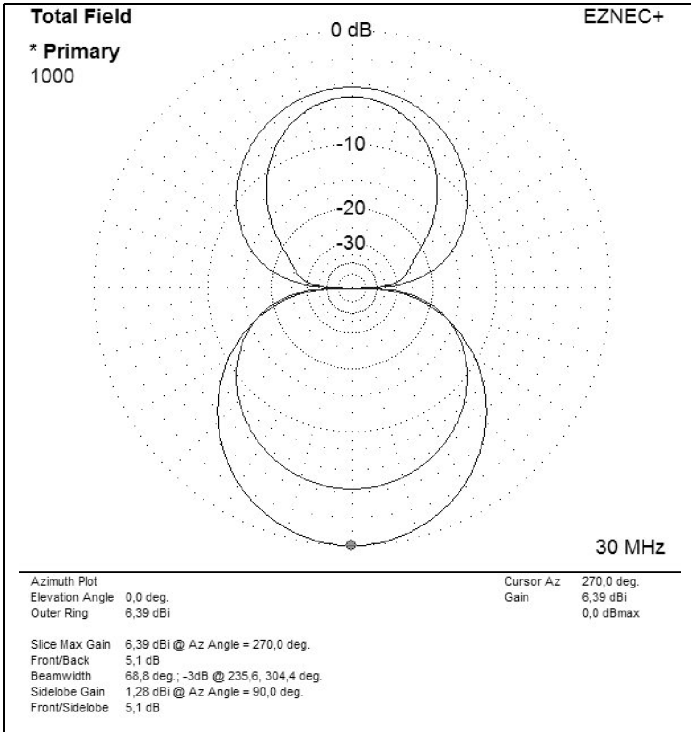
Fig 6: Radiation pattern of a standard dipole and of the dipole with a parallel disturbing wire at a distance of 0.25λ.

(V/R = 0dB). The system with two wires (Dipole + reflector) represents a simple 2 element beam, which is not optimised for V/R ratio but that is not the task of this investigation.

During further approximation of the disturbing wire on the dipole, e.g. with y = 0.1λ, the radiation direction moves toward the other side, the disturbing wire now acts as a director. The data for this is G=6.81dBi and V/R=2.59. With even more reduced distance with y = 0.05λ coincidentally gives a quite good 2 element directional antenna with G = 6.75dBi and V/R=24.45dB (!), Fig 7.

3.2. Dipole with perpendicular disturbing wire

What happens, if the passive disturbing wire is not parallel to the dipole but at right angles to the dipole wire (similarly to example wire 6 in Fig 2)?

Enter a wire into to the WIRES menu, it is to be quite close to the dipole. The data is: End1: 0/0.16/0; End2: 0/0.64/0 again wire 0.48λ long, which is a distance of 0.16λ from the dipole feed point at it's closest end. The antenna feed impedance is ZA= (71.39 - j2.96)Ω which is unchanged despite the proximity of the resonant wire.

A current flow in the disturbing wire would indicate a reaction to the dipole field, but the current flow is zero and the radiation pattern is uninfluenced. That means that a wire symmetric to the axis of the dipole in the direction of propagation is completely decoupled from the dipole. This fact is well known and is used for the boom of an antenna.

However what happens if the wire on y axis is off centre, not in the direction of propagation (similar to example wire 2 in Fig 2)? With the wire moved 0.1λ along
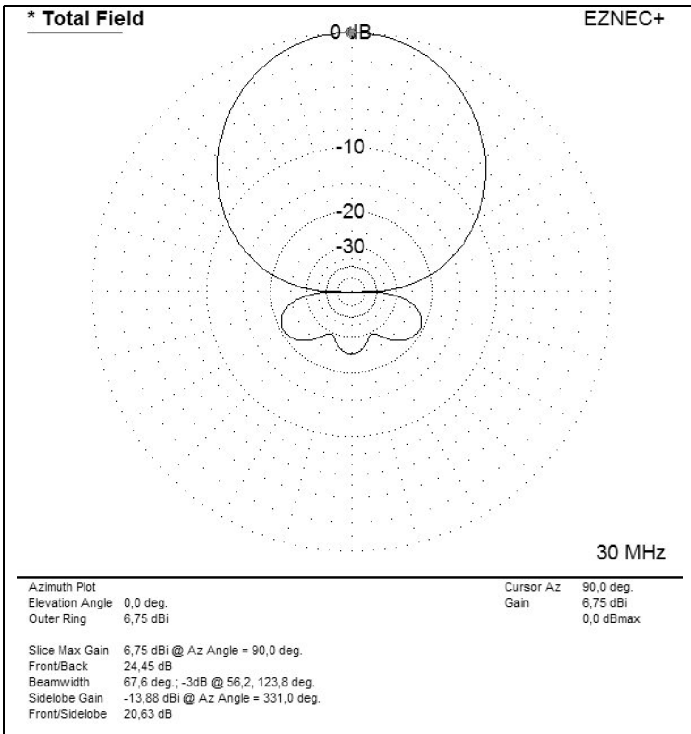
Fig 7: Radiation pattern of the dipole with a parallel disturbing wire at a distance of 0.05λ.

the x axis gives: End1: 0.1/0.16/0; End2: 0.1/0.64/0, Fig 8.

Fig 8: Diagram of the current distribution in the dipole with a perpendicular disturbing wire 2 x 0.24λ long. The disturbing wire is shown at a distance of 0.1λ along the x axis.

The antenna feed impedance is now ZA= (72.18 - j4.65)Ω and is only changed slightly. The current flow in the disturbing wire is no longer zero, but has a maximum value of 0.164A. The radiation pattern shown in Fig 9 is clearly deformed towards the wire ends.

If the disturbing wire is moved still further e.g. x = 0.24λ (End1: 0.24/0.16/0; Ende2: 0.24/0.64/0), then the antenna impedance is ZA= (71.64 - j7.82)Ω and the current flow in the disturbing wire rises to a maximum of 0.265A. which is not negligible.

To be continued ….
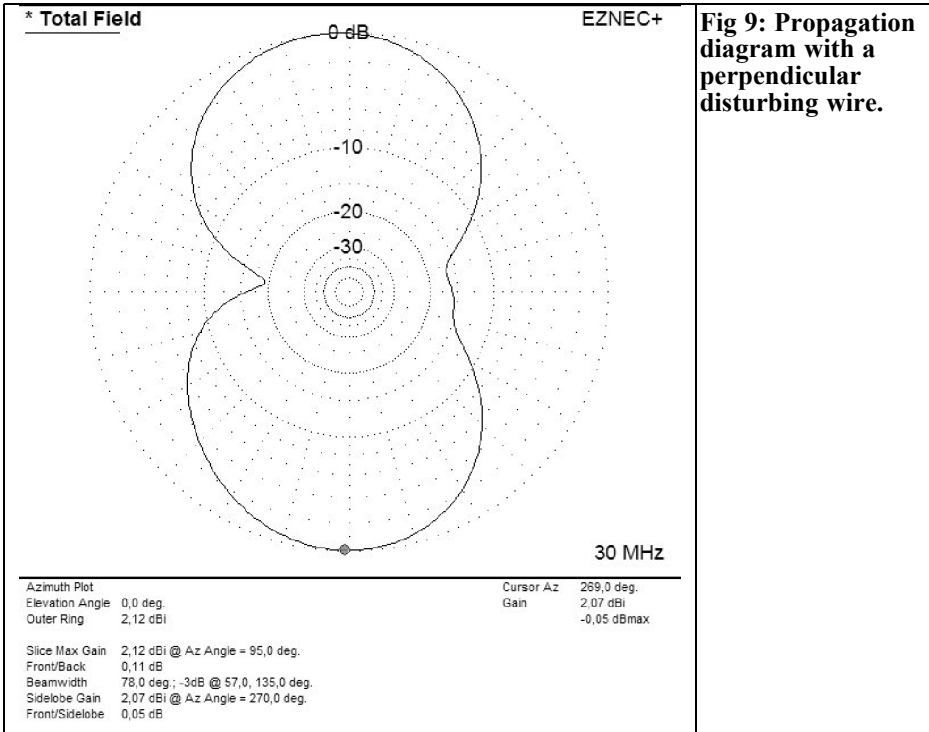
Fig 9: Propagation diagram with a perpendicular disturbing wire.

X. Literature

[1] R Lewallen, W7EL, P.O.Box 6658, Beaverton, OR, 97007, the USA; http://eznec.com

[2] EZNEC 3.0 - antenna simulation under Windows, Gerd Janzen, DF6SJ, RADIO AMATEUR 49 (2000) booklet 10:1110 - 1112, booklet 11:1238 - 1241, booklet 12:1362 - 1365.

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[3] Gerd Janzen, DF6SJ, EZNEC 4 - antenna simulation, RADIO AMATEUR 49 (2000) booklet 7:696 - 699.

[4] Short Antennas, G. Janzen, DF6SJ, Franckh publishing house, Stuttgart, 1986,

Zeljko Bozic, S52ZB

13GHz prescaler

1.

Introduction

A simple and inexpensive option for expanding the range of a frequency counter to the GHz range is to add an external prescaler, which divides by 1000. So, with a 10MHz counter with 1Hz resolution we can measure frequencies up to 10GHz with a resolution of 1kHz. Because the resolution is reduced by a factor of 1000 it is recommended that the counter should have at least 1Hz resolution or even better 0.1Hz.

Thanks to the fast development of the satellite communications and point-to-point/multi-point radios, the offer of prescaler ICs is continually growing and they can be obtained for increasingly higher frequencies. Two new 13GHz prescalers from NEC: the uPB1512TU-A and uPB1513TU-A are both housed in a miniature 8 pin lead-less minimold package only 2.2 x 2.0 x 0.5mm in size. They are designed to serve as low cost alternative to GaAs prescalers. They are available at a favourable price under $4 and good performance makes them interesting for radio amateurs and professionals.

2.

Block diagram

The block diagrams are shown in Figs 1 and 2. The difference between first and second version are in U1, which divides by 8 or 4, and in the division factor of U2 that is 64 or 128.

3.

Circuit diagram

The circuit diagram of the 13GHz prescaler is shown in Fig 3. The input signal passes through the coupling capacitor C1 to pin 2 (IN) of U1. The unused input (IN-) is connected to ground through capacitor C2 and 51 ohms resistor R1. Free running oscillation of the first divider stage is typical for this kind of divider. After switch on the prescaler oscillates at about 10GHz.

The signal is divided by 8 in the uPB1512TU-A or by 4 in the uPB1513TU-A. From pin 7 (OUT) this is connected through capacitor C4 to the pin 1 (IN) of the uPB1507GV. The unused output is connected to ground through capacitor C3 and 51 ohms resistor R2. Also the unused input pin 8 (IN-)
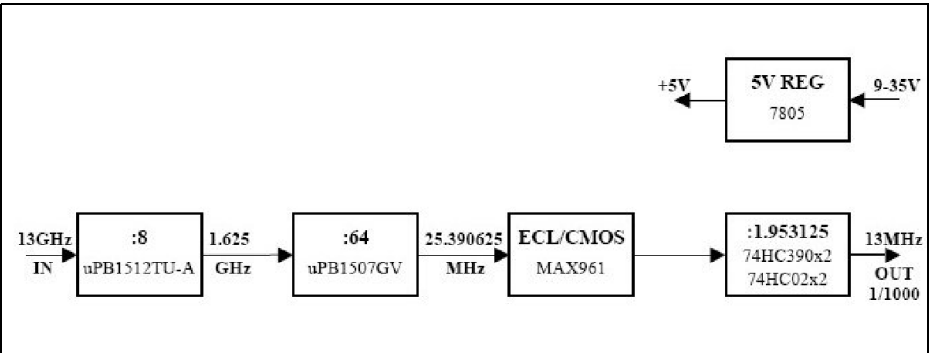
Fig 1: Block diagram of the 13GHz prescaler with uPB1512TU-A.

of the uPB1507GV is connected to ground through capacitor C9.

Depending on the type of IC used for U1, either an uPB1512TU-A or uPB1513TU-A, we set the division factor of U2 (uPB1507GV) to 64 or 128. The PCB is designed for a division factor of 64; for a division factor of 128 cut the connection between +5V and pin 6 (SW2) before fitting the uPB1507GV.

The output signal from U2 is ECL, 1.6Vpp with a DC offset at about +3V. This is amplified to CMOS levels by the ECL/CMOS converter MAX961. The MAX961 is a high-speed comparator from Maxim. It compares the output signal from U2 with the threshold voltage from the resistive divider R3/R4 and generates a clean square wave signal at

the output that is then divided by 1.953125. This divider is constructed from U4, U5, U6 and U7 (74HC390 x 2, 74HC02 x 2).

Division by non-whole numbers such as 1.953125 is realised using three 1.25 dividers (1.25 x 1.25 x 1.25 = 1.953125), dividing by 1.25 is realised by removing every fifth pulse. With this division factor the overall division factor is 1000 (8 x 64 x 1.953125 = 1000 or 4 x 128 x 1.953125 = 1000). The operating voltage of the prescaler module (approximately 9 to 35V) is fed through the feedthrough capacitor C20 and the reverse voltage protection diode D1 to the voltage regulator U8. This generates a regulated voltage of 5V for the prescaler module.
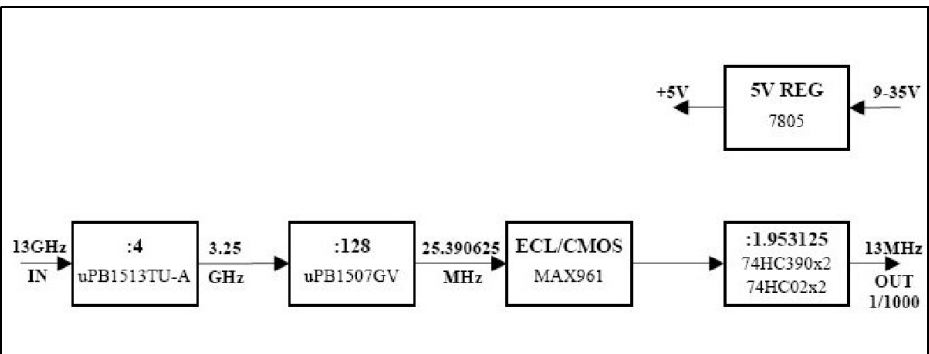
Fig 2: Block diagram of the 13GHz prescaler with uPB1513TU-A.
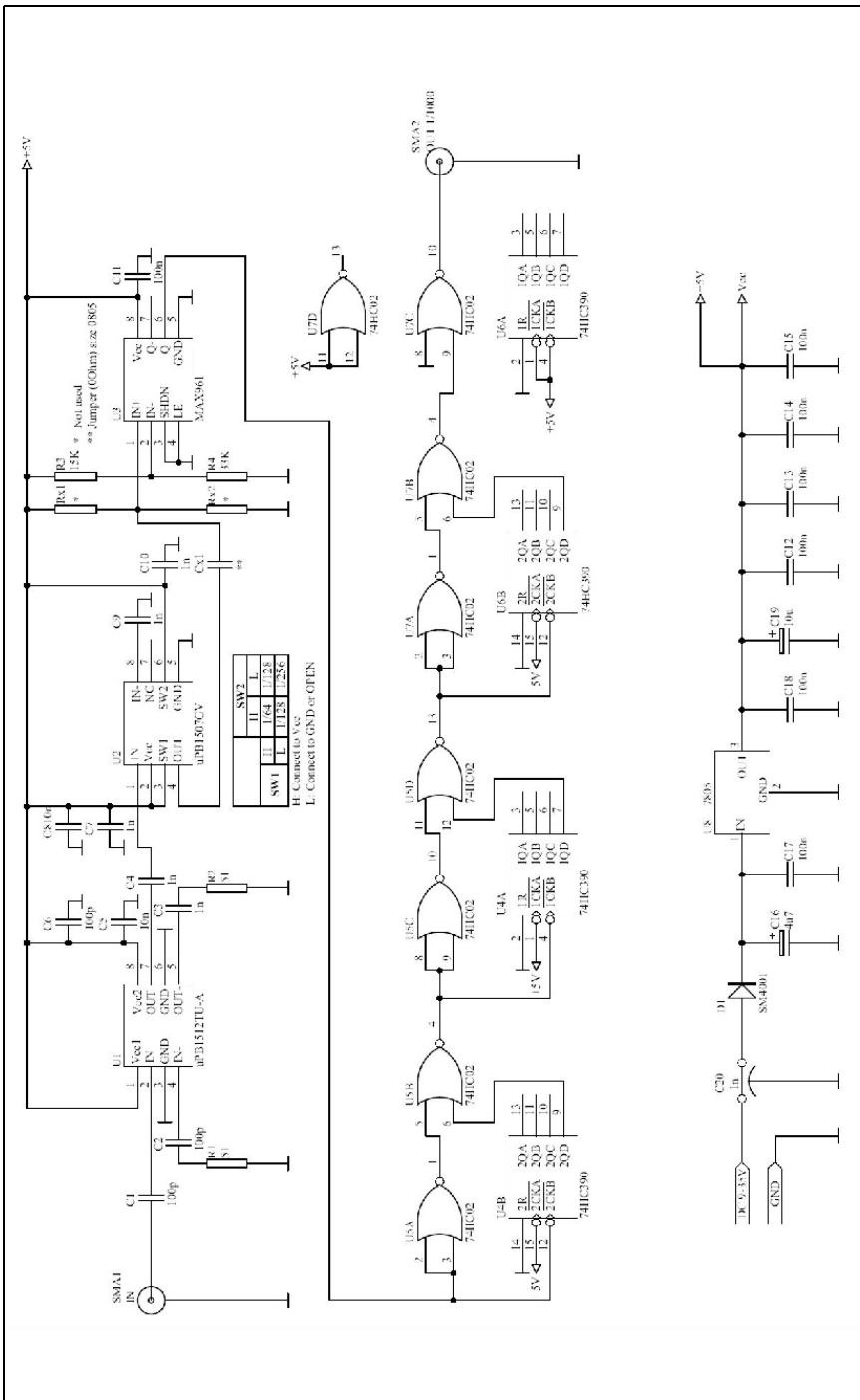
Fig 3: Circuit diagram of the 13GHz prescaler.
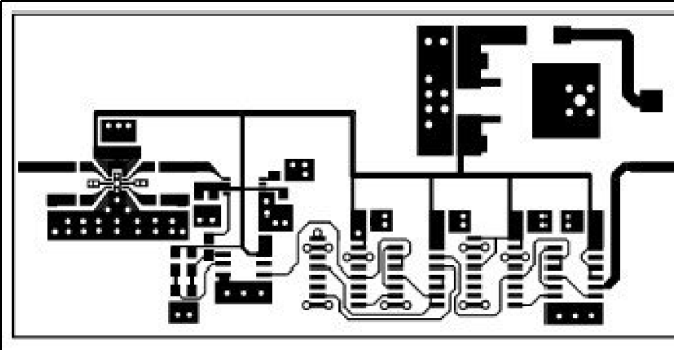
Fig 4: Top side PCB layout for the 13GHz prescaler.

4.

Construction

The prescaler is constructed on a double-sided PCB made from a 90mm x 44mm piece of 0.80mm thick FR4 laminate. The PCB layouts are shown in Figs 4 and 5, the component layout is shown in Fig 6. Special attention is required to assemble U1 (uPB1512TU-A or uPB1513TU-A).

Because of it's extremely small size (2.2 x 2.0mm) the distance between the centres of the pins is only 0.5mm; the pins are 0.2mm wide so the distance between them is only 0.3mm! A still bigger problem is the pad on the bottom side of the package, which must be soldered to ground. Because of this, construction is very difficult and it is only recommended for very experienced SMD constructors.

We had the help of magnifying glasses, a

soldering iron with a fine tip and SMD solder paste to assemble U1. It would have been even better if we had the use of a reflow soldering unit or hot air station.

All other components are normal SMD size. The PCB has positions for Rx1 and Rx2 but they are not used. Instead of Cx1, fit a 0805 size jumper (0Ohm). When all components are assembled, fit the prescaler into the housing made from 0.5mm thick brass plate.

I tested some samples of the uPB1512TU-A and uPB1513TU-A and every one showed the same result for input sensitivity and minimum and maximum frequency. Although U1 is specified as a 13GHz prescaler, they operate from 0.6 to 17.2GHz. The input sensitivity curve versus frequency is shown in Fig 7. The prescaler draws about 102mA. Depending on the interested in this project, PCBs, Kits or ready made and
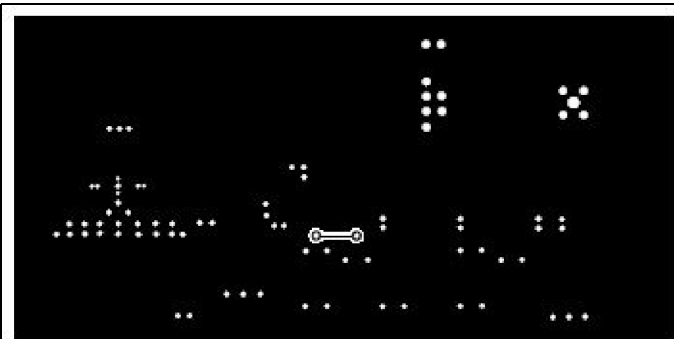
Fig 5: Bottom side PCB layout for the 13GHz prescaler.
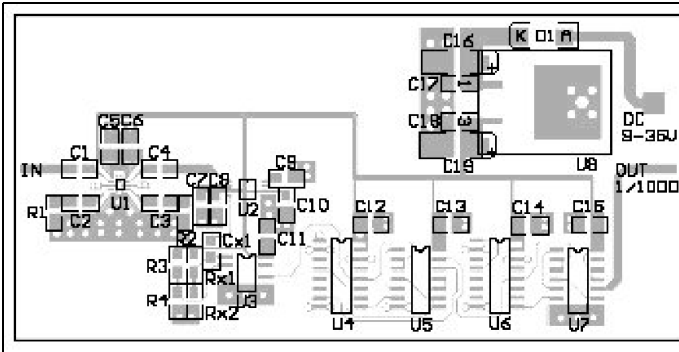
Fig 6: Component layout of the 13GHz prescaler.

tested prescaler modules will be available from the author. Figs 8, 9, and 10 show the prototype of the prescaler module.

Finally I wish to thank Mr Thomas Sporkmann of NEC Electronics (Europe) GmbH for samples and support. Also I wish to thank Matjaz Vidmar (S53MV) and Stojan Kuret (S51WI) for their help.

5.

Literature

[1] Zeljko Bozic S52ZB: 6 GHz preskaler, CQ ZRS 2/1998

[2] Alexander Meier DG6RBP: 12GHz divide by 1000 prescaler, VHF Communications 4/2003

[3] NEC data sheet: uPB1512TU-A, uPB1513TU-A, uPB1507GV

[4] MAXIM data sheet: MAX961

Table 1: Parts List.

R1,R2	51Ohm	0805
R3	15k	0805
R4	33k	0805
C1,C2,C6	100p	0805
C3,C4,C7,C9, C10	1n	0805
C5,C8	10n	0805
C11,C12,C13 C14,C15,C17,C18	100n	0805
C16	4u7	SMD tantalum 35V
C19	10u	SMD tantalum 35V
C20	1n	Feedthrough
D1	SM4001	MELF (DO-213AB)
U1	uPB1512TU-A (uPB1513TU-A)	MINIMOLD-8
U2	uPB1507GV	SSOP-8
U3	MAX961	SO-8
U4,U6	74HC390	SO-16
U5,U7	74HC02	SO-14
U8	7805	D2PAK
Cx1	0Ohm	0805 jumper
Rx1	-	0805 not used
Rx2	-	0805 not used
SMA1, SMA2	-	SMA panel socket with 4-hole flange
PCB	90 x 44mm/FR4 0.80mm	
Housing	-	Brass plate 0.5mm

Fig 7: Input sensitivity vs. frequency.
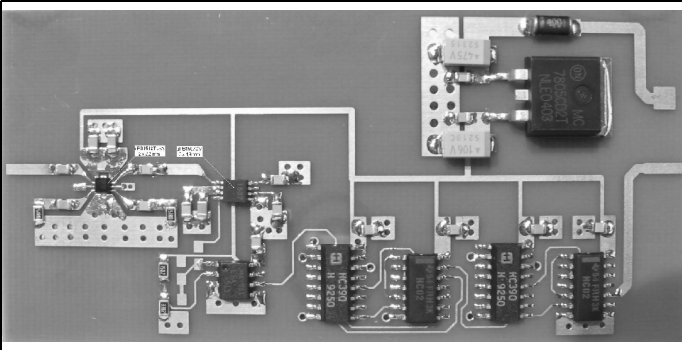
Fig 8: Prototype of the 13GHz prescaler.
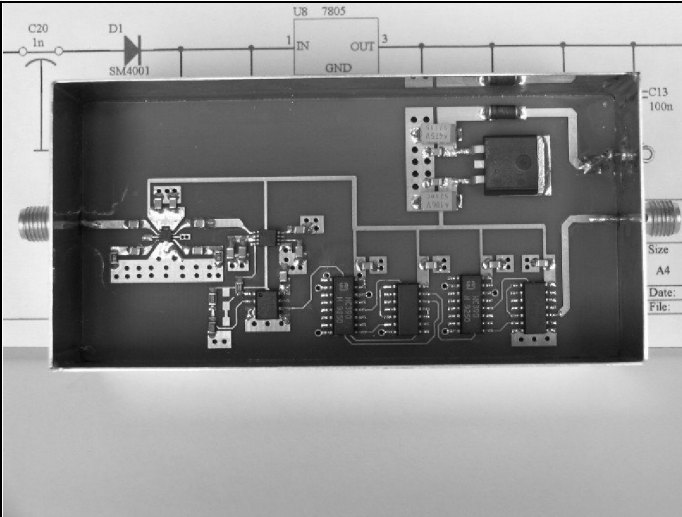
Fig 9: Completed prototype of the 13GHz prescaler.
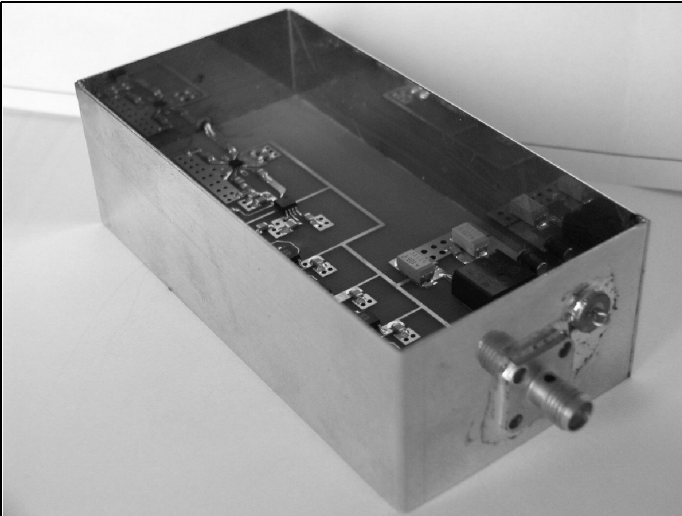
Fig 10: Rear view of the completed prototype of the 13GHz prescaler.

Gunthard Kraus, DG8GB

Determining S-parameters with PSPICE

An introduction to PSPICE was presented in VHF Communication Magazine [1], this gave the arguments for and against simulation in the time domain. Despite the fascinating possibilities of this simulation method, a true RF and microwave enthusiast will long, again and again, for the familiar world of S-parameters. That is what following article will discuss.

1.

Introduction

A simulation with PSPICE shows the voltages and current flows in a circuit that can be loosely described as oscillograms. There is another option offered by this program; if it is accepted that the behaviour of the circuit at the selected operating point is linear, it examines the frequency response for the signals of interest for this special case. This mode of operation is called "AC-Sweep" and it is the key to the determination of the S-parameters by showing the voltage ratios and frequency response. Using a trick with a Twoport, all four S-parameters (S11/S21/S12/S22) can be determined easily in dB over the frequency range shown in graphical form i.e. "Rectangu-

lar Plot" (also see [2] and [3]).

2.

Determination of S11

Fig 1 shows the basic circuit to be considered. A pulse generator with an internal resistance of 50Ω feeds a load Z1 over a long coaxial cable, which has a characteristic impedance Z = 50Ω. For the pulse this cable initially represents a real resistance of 50Ω therefore there is a simple voltage divider of two 50Ω resistors at the cable entrance. Therefore the instantaneous amplitude measured is:

$$V_{in} = \frac{V_o}{2}$$

The power is:

$$P_{inc} = \frac{\left(\dfrac{V_o}{2}\right)^2}{50\Omega} = \frac{V_o^{\,2}}{4 \cdot 50\Omega}$$

An Incident Wave travels down the cable, at the defined speed in the cable, towards the load Z1. This is the maximum amount of power that can be delivered to the load by the generator (power
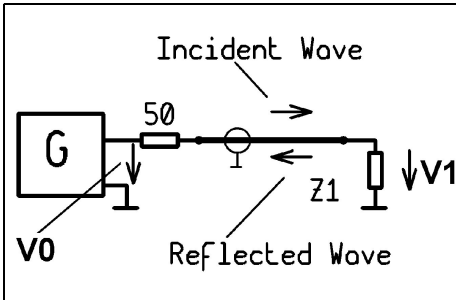
Fig 1: The basic circuit, a generator feeding a load via a coaxial cable.

matched). When this wave reaches the load resistance Z1, it can only be completely absorbed if a power match exists, thus the value of Z1 should be:

If that is not the case, the "surplus" makes it's way back to the generator in

$$Z_1 = R_i = Z = 50\Omega$$

the form of an echo (Reflected Wave). This is expressed as the "reflection factor", with correct adjustment it is zero (nothing is reflected) and for a complete mismatch (no-load operation or short-circuit) nothing is absorbed and everything must go back to the generator. In this case the reflection factor rises to 100%.

For practical purposes this is simplified by just considering the voltages of these power waves. The formula for the power is shown above and is related to the square of the voltage. Taking the square root of the power gives:

Incident wave:

$$a_1 = \sqrt{P_{incident}} = \sqrt{\frac{(V_o)^2}{4 \cdot Z}} = \left(\frac{V_o}{2}\right) \cdot \frac{1}{\sqrt{Z}} = V_{incident} \cdot \left(\frac{1}{\sqrt{Z}}\right)$$

This formula shows again: the incident wave can simply be described as half the no-load generator voltage.

Reflected wave:

$$b_1 = \sqrt{P_{reflected}} = \sqrt{\frac{(V_{reflected})^2}{Z}} = (V_{reflected}) \frac{1}{\sqrt{Z}}$$

The reflection factor "r" is simply the relationship between the reflected wave and the incident wave:

$$r = \sqrt{\frac{P_{reflected}}{P_{incident}}} = \frac{b_1}{a_1} = \frac{V_{reflected}}{V_{incident}}$$

and is the same as the S-parameter S11, if the input of a Twoport is used instead of the load R1 connected to the cable output. (But the output of this Twoport itself must be accurately terminated with the system resistance to avoid any echos inside this Twoport!). The logical sum of the voltages appears across R1:

$$V_{R1} = V_{incident} + V_{reflected}$$

When R1 = Z we measure half $V_{incident}$ because no reflection occurs and $V_{reflected}$ is therefore to zero.

If R1 is larger than Z, the total voltage is larger than $V_{incident}$ and the echo is in phase with the incident wave.

If R1 is smaller than Z the voltage is smaller than the incident wave and the echo is in anti phase to the incident wave.

Now back to PSPICE. The determination of the S-parameter S11 now becomes child's play. Using a small additional circuit and interpreting the formulae above correctly we can choose between the two cases:

- Using the "AC-Sweep" S11 in dBs as a function of the frequency can be determined, in the same way that it can be done using a microwave CAD program or measured with a network analyser.
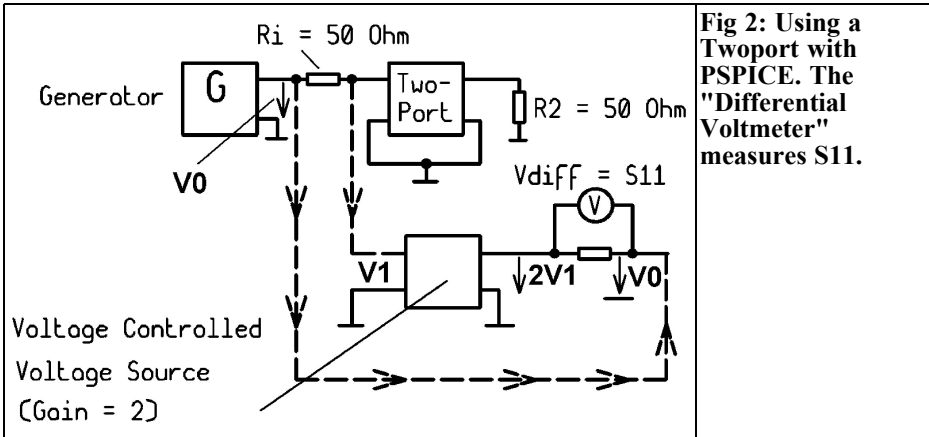
Fig 2: Using a Twoport with PSPICE. The "Differential Voltmeter" measures S11.

- Working in the time domain, the value of the reflected signal at the input can be measured at certain frequencies and varying operating points. So the reflections, and thus S11 can be determined.

Changing the formula:

$$V_{reflected} = V_1 - V_{incident} = V_1 - \frac{V_o}{2} = \frac{1}{2} \cdot (2 \cdot V_1 - V_o)$$

Using this in the relationship for S11:

$$S11 = \frac{b_1}{a_1} = \frac{V_{reflected}}{V_{incident}} = \frac{\left[\frac{1}{2} \cdot (2 \cdot V_1 - V_2) \right]}{\left(\frac{V_o}{2} \right)} = \frac{2 \cdot V_1 - V_o}{V_o}$$

Using PSPICE this means:

- The voltage V1 at the input of the Twoport must be doubled.

- Then the incident voltage V0 is subtracted from it (with a "differential marker")

- The result, in dBs, of this operation is the desired S-parameter (because PSPICE automatically makes the calculation of the relationship between V1 and V0 to obtain dBs).

The practical modifications to the PSPICE simulation are shown in Fig 2.

The input voltage of the Twoport (V1) is sampled, by a "voltage-controlled voltage supply" with a gain of 2 and fed to the left hand end of a resistor $Z = 50\Omega$. The other end of this resistor is connected to the incident voltage V0.

Measuring the voltage across the resistor with a "differential marker" with the result in dBs gives S11. An example is given later.

3.

Determination of S21

This is trivial to measure in dBs using PSPICE, if V2 is used in the following formula:

$$\frac{V_2}{V_0} = V_u$$

Then take the logarithm of the result.

Measuring the S-parameter S21 with the "half incident voltage" as an input signal is incorrect by a factor of 2 (6 dBs). A voltage-controlled voltage supply with a gain of 2 is used to make the measure-
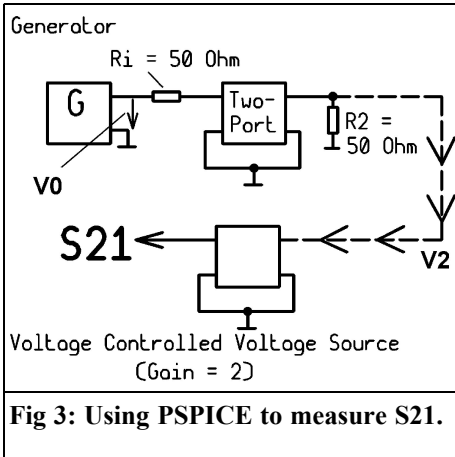
Fig 3: Using PSPICE to measure S21.

ment of S21 correct, it only needs V2 to be increased by a factor of 2. The practical modifications to the PSPICE simulation are shows Fig 3.

By the way:

The determination of S21 and S11 can be achieved with one simulation using just two additional circuits.

4.

Determination of S12 and S22

Things become much simpler, because as with a genuine network analyser only input and output of our Twoport must be interchanged: S22 can now be measured with the circuit of Fig 2, connected to the output (Port 2) of the Twoport.

The last S-parameter, S12, can be measured using a voltage controlled source, as used in Fig 3, by treating port 1 as the output.

The following sections show practical examples

5.

First practical example: 110MHz low pass filter

A familiar circuit will be used to confirm the measurements discussed above. The Chebeyschev low pass filter used in earlier articles will be used. It exhibits the following data:

- Cut off frequency f_g = 110MHz

- Series inductor circuit

- Degree of filtration n = 5

- Characteristic impedance Z = 50Ω

- Ripple in pass band = 0.1dB

- Ripple at maximum value of S11 = -16.4dB

The trusted DOS program "fds.zip" is used to calculate the component values:

- C1 = C3 = 33.2 pF

- C2 = 57.2pF

- L1 = L2 = 99.2nH

The complete test circuit as it is entered using the program "5Spice" is shown in Fig 4. This should be fairly easy, it was described in detail in [1]. The position of the voltage controlled sources required is shown as well as the place where they can be found in the 5Spice menu.

The results of a simulation are shown in Fig 5. This was obtained by pressing the F8 key and selecting; a linear AC-Sweep from 1KHz to 250MHz using 1000 measuring points; output as a Graph/Table in dBs.

To see the Chebeyschev ripple, the normal range of zero to –0.5dB was stretched as shown in Fig 6 which compares favourably with other S-parameter
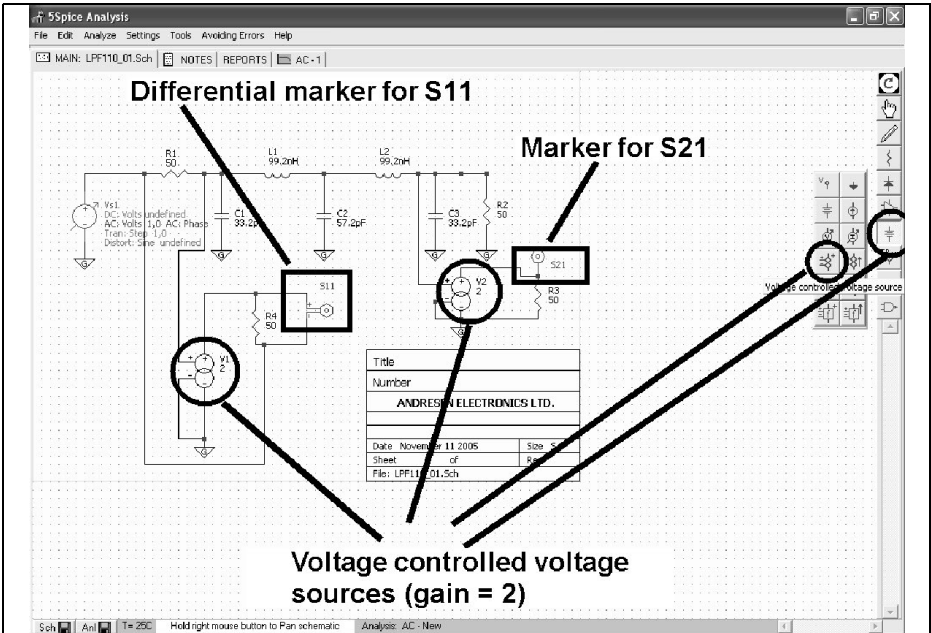
Fig 4: PSPICE simulation circuit for a 110MHz low pass filter with circuits for measuring S11 and S21.
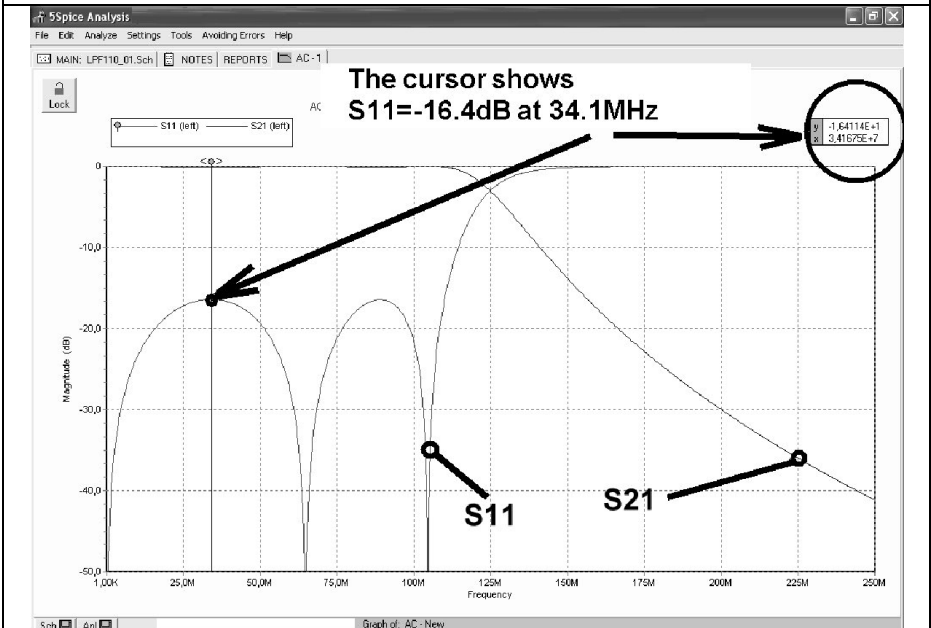
Fig 5: Simulation output from PSPICE, similar to that obtained from PUFF or APLAC.
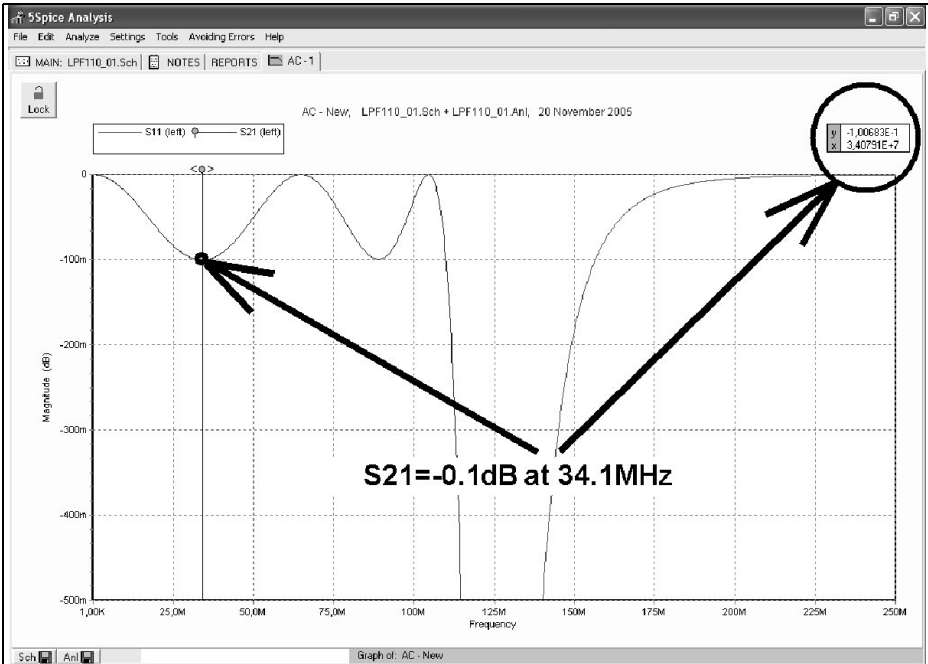
Fig 6: Enlarged view of the simulation output of the 110MHz filter showing the Chebeyshev ripple.

programs (like e.g. PUFF).

Since the circuit is symmetrical, S22 = S11 and S21 = S12 there is no need to repeat the analysis for the output.

Fig 7: Circuit diagram of the gain block with input and output Z=50Ω.

6.

Second practical example: Gain block using two transistors

This example shows a discrete component DC coupled amplifier using an NPN and a PNP transistor (Fig 7). Negative feedback via a 470Ω resistor gives input and output resistances of 50Ω.

The simulation modifications required for "5Spice" to measure S-parameters are as follows:

- A voltage source including external 50Ω resistor at the input

- A 50Ω resistor on the output

- A voltage controlled voltage source at the input to directly measure S11.
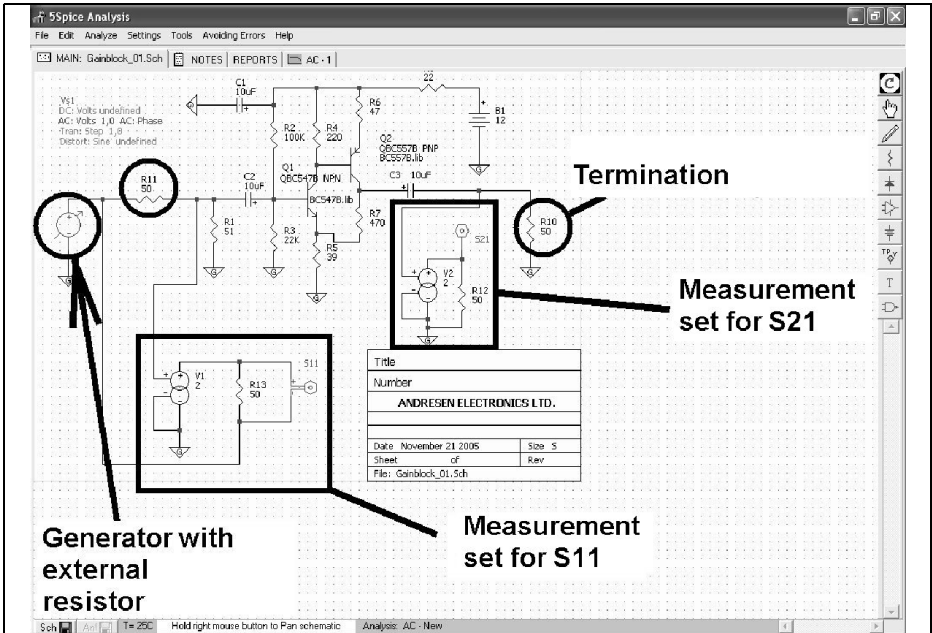
Fig 8: Simulation diagram of the gain block set up using the "5Spice" editor.
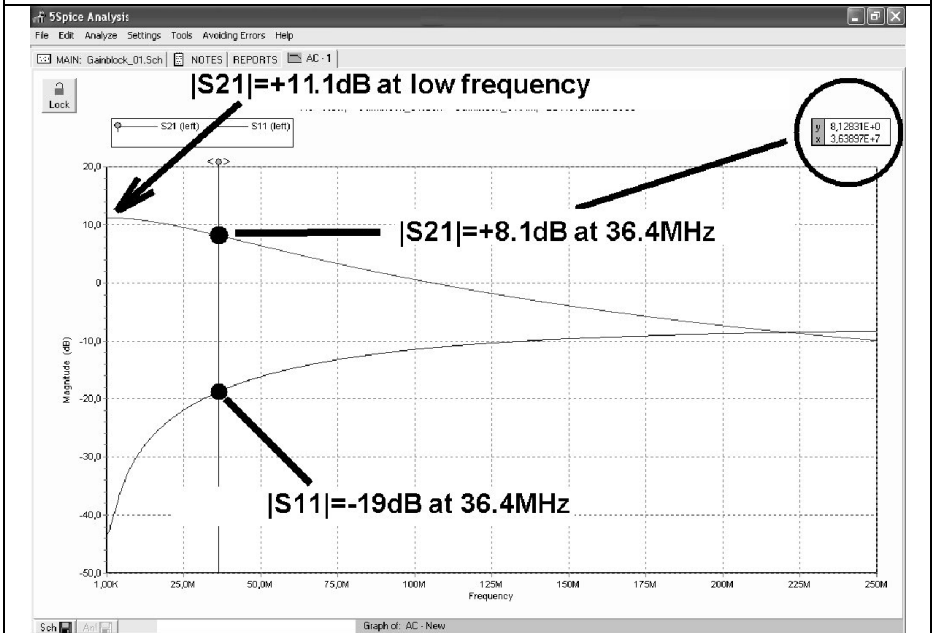
Fig 9: Simulation output showing S11 and S12 up to 250MHz.

Fig 10: Modifications to the simulation cicuit to measure S12 and S22.
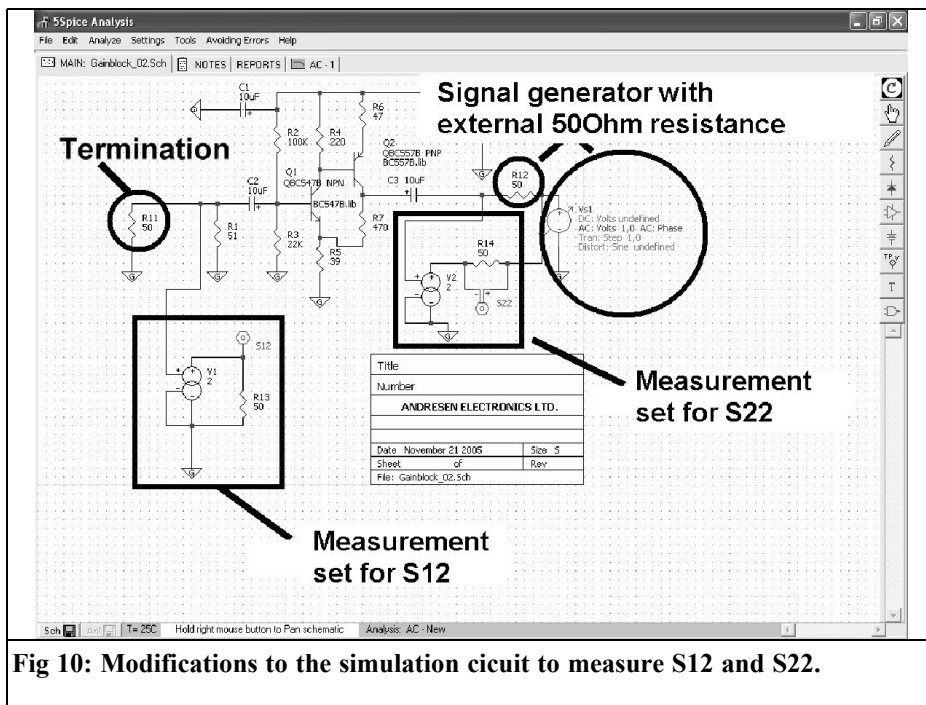
- S21 is measured at the output.

Fig 8 shows the simulation circuit where all of the components can be recognised. The SPICE models were found on the Internet in the download section of the "Philips SPICE Library". (The author can supply a CD with more than 200Mb of personally collected SPICE and S-Parameter libraries). These must be merged with the already existing files by clicking "Tools/Rebuild SPICE Library" and the correct file assigned to each transistor (after one click on the circuit symbol). The connection point and direction of the voltage controlled voltage sources are shown. These are found under the Sources button, after the placing them do not forget to set the gain to 2 (double click on the circuit symbol).

Now things get serious. Pressing the key F8 calls up the analysis set up menu. Simply use the attributes used for the low pass filter simulation above (linear AC-

Sweep from 1kHz to 250MHz with 1000 measuring points). The result output should be set as "graph/Table" with "Autoscale" selected and "Magnitude in dBs"; then assign the two markers S11 and S21 to the left hand axis of the diagram. The result of the simulation is shown in Fig 9, the cursor is positioned at the cut off frequency, the -3dB point on the S21 curve at 36.4MHz. S21 decreases with frequency, to find out where the -11dB point lies (for example), move the cursor with the mouse until the value appears in the window.

With that simulation complete save the project under a new name. Now the input voltage with it's external resistance is moved to the output together with the circuit for measuring reflection. The voltage controlled voltage source should be named "S22". The measuring circuit at the input should be terminated with 50Ω and the marker named S12. The changes are shown in Fig 10.
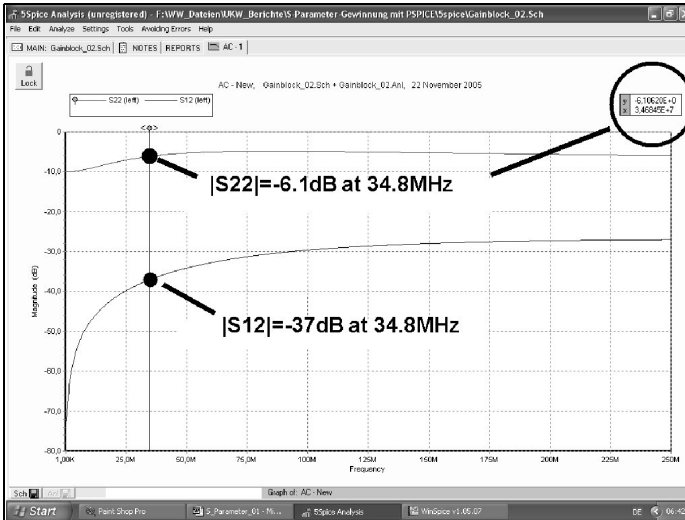
Fig 11: Simulation output showing S12 and S22.

The simulation results are shown in Fig 11, these show the same information that you would obtain from a program such as PUFF. However we must not forget that these apply for "small signal values". The assumption is made that the parameters are linear at the operating point, even the genuine S-parameter CAD programs make this assumption.

With different controlling circuits we can take a quick look at the practical data. First save the current project with another name.

7.

Simulation in the time domain

The simulation circuit (see Fig 12) must now be modified:

- The measuring circuit for S21 at the output is removed and a simple marker "Vout" for the voltage across the 50Ω load is added.

- The input voltage must be a sine wave voltage with a peak value of 0.2V at f = 10MHz.

- The S11 measuring circuit is not touched.

- Press the F8 key and change the simulation defaults to; "Transient New /T = 0… 10μs/time step = 0.005μs".

- The results are produced by showing S11 and Vout using the option "Autoscale" in "graph/Autoscale" mode.

The simulation results are shown in Fig 13. The shape of the output signal is very well simulated and can be checked for distortions and/or over driving. If no distortions can be seen, the linearity of the circuit is sufficient and the simulation results for S-parameter, both for the "AC-Sweep" and "frequency range", are the same for time step at a single frequency! In addition the display range is sufficient to be able to see the "reflected wave" at the input and make a better investigation (e.g. wave shape/phase position/amplitude).

It is interesting with larger input or even over driving, because the "linear simula-
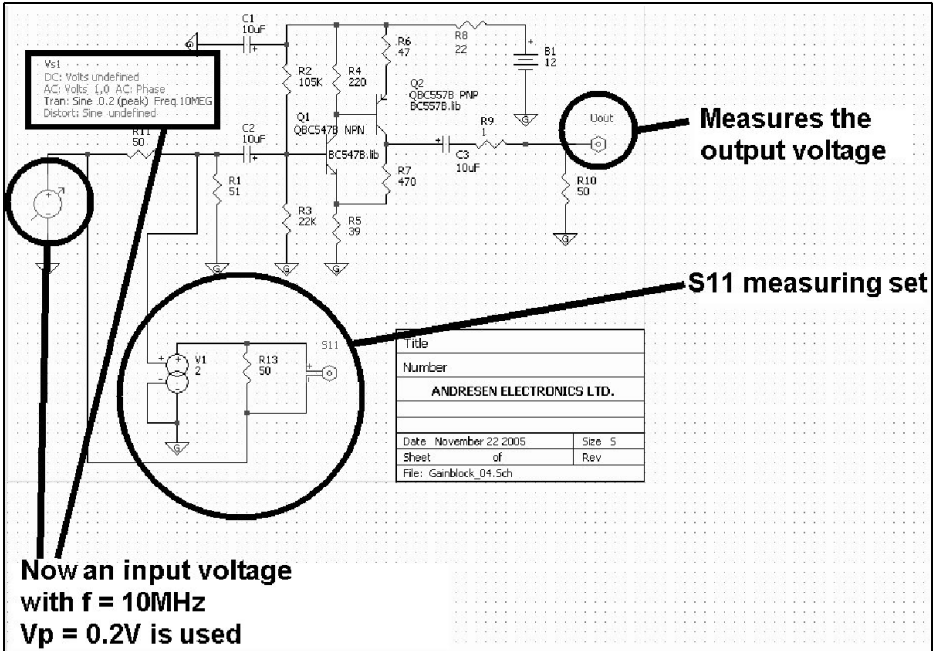
Fig 12: Simulation circuit for a large signal test in the time domain with an input voltage of Vp = 0.2V / f = 10MHz. The S11 measuring set is used as before but the S21 measuring set is replaced by a simple voltage marker.

tion in the frequency range" from the preceding section does not apply. Fig 14 shows the result for an input voltage with the peak value of 0.5V. It is interesting to see that distortion begins for the positive half wave of the output voltage, while the
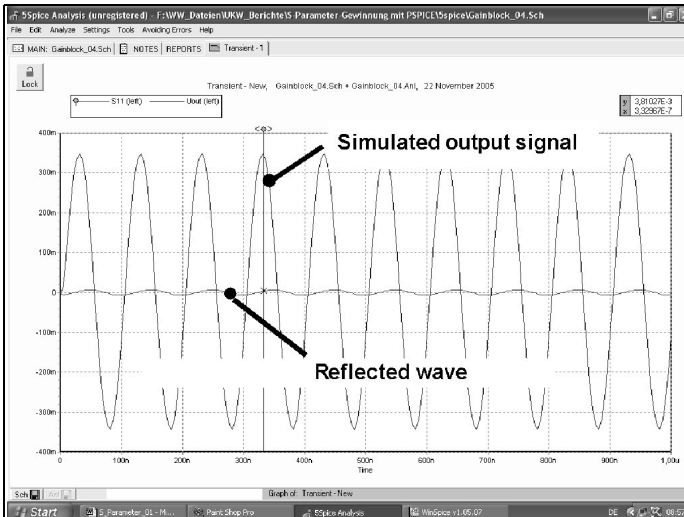
Fig 13: Simulation output with 0.2V input showing that the output is linear at 10MHz.
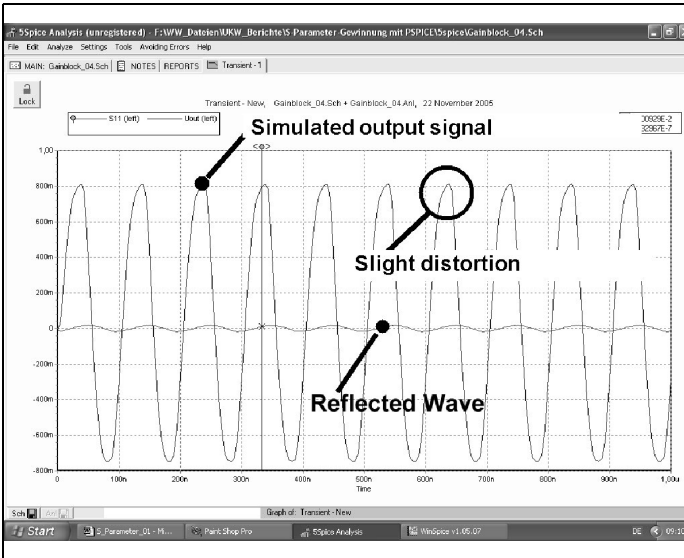
Fig 14: Simulation output with 0.5V input showing that the output is becoming non linear at 10MHz.

reflection at the input still looks straight. Finally Fig 15 shows the results with an input of 1V peak, the output is badly distorted and the reflected wave is also distorted. In such a case we cannot rely on the linearity of the small signal parameters.

8.

Conclusions

It is possible to determine all S-parameters of an N port using the "AC-Sweep" option of PSPICE and some simple techniques but extra effort by the user.
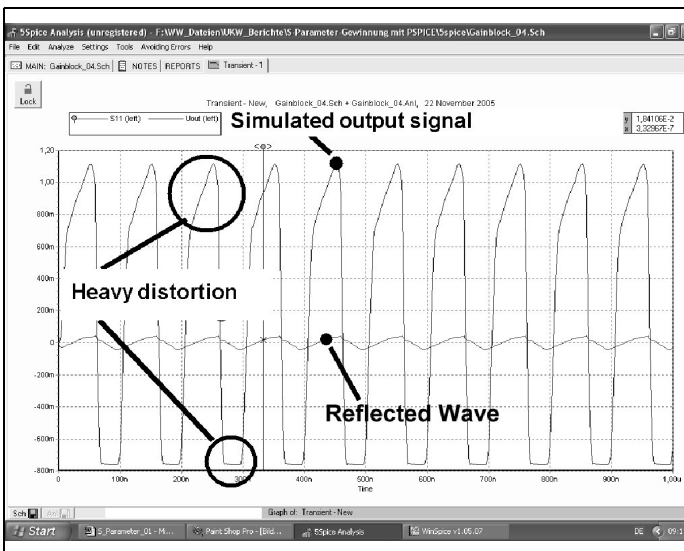
Fig 15: Simulation output with 1.0V input showing that the output is very distorted which means that small signal parameter do not apply.

Certain programs (e.g. electronics Workbench) offer similar options. Finally the theory to determine S-parameters with PSPICE is not too complicated but the programming of results to display in a Smith Chart is more complex.

The PSPICE simulation in the time domain is helpful for developers to see the performance at the extremes, since it concerns the determination of distortions and harmonics and saturation effects. Thus you get the benefit of the latest developments plus the practical behaviour.

It is possible to determine S-parameters even for the exotic non linear cases (see again Fig 15). The work required is considerable because it requires the use of the Fourier transformation to develop the spectra of the signals involved for the incident and reflected wave portions, to find out their values at the desired fundamental frequency and then define the S-

parameters. This is repeated for different input voltages then you know exactly what "the thing" is. However this is beyond the scope of "5Spice" so a more capable program would be required.

9.

Literature

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

[2] Microsim-Application-Note: Obtain S-Parameters from Probe

[3] Microsim-Application-Note: Create S-Parameter subcircuits for Microwave and RF Applications

Johannes Schad, DG6NDS

Design of a Quad Yagi: Part 1

If the advantages of a Yagi array are combined with those of a Quad antenna it is still an optimised antenna, for short, a Quad Yagi, becomes a Quagi. Design, simulation and construction of Quad Yagis is described in the following article.

1.

Quagi antennas (Quad Yagis)

The Yagi array is probably the best known antenna. These antennas can achieve a very high gain, depending upon size (length). Only parabolic antennas or large antenna arrays can achieve a few more dBs gain. A further advantage of Yagis antennas is that they have good mechanical and electrical characteristics and above 1.5GHz they are so small that they become inconspicuous.

My special interest is high gain Yagis for 2.4GHz that are suitable for WLAN application. Naturally I had the question: how well do the antennas described in numerous publications perform in practice and how good are their electrical characteristics.

There are many Yagis for 2.4GHz/WLAN use advertised on the Internet, mostly on EBAY with some from commercial suppliers. There are often serious errors in the descriptions of the antennas, probably due to lack of expertise. Often the gain of 2.4GHz WLAN Yagis is given in dBm, the antenna radiation pattern is missing or 45cm long antennas are described with a gain of 18dBi.

First attempts with simple long Yagis were not very successfully. Thus the idea was formed to develop a 2.4GHz antenna, a cross between a Quad antenna and a long Yagi. It will be seen later that this antenna has substantial advantages over a "normal" Yagi array.

An Internet search for a manufacturer of Quagi antennas was unsuccessfully. Radio amateurs rarely use this type of antenna but from my conclusions later, this is an injustice.

The publications by Wayne Overbeck [3] and [4] plus the examples in Rothammels Antennenbuch (Red Sheep antenna book) [2] were used as the basis of development.
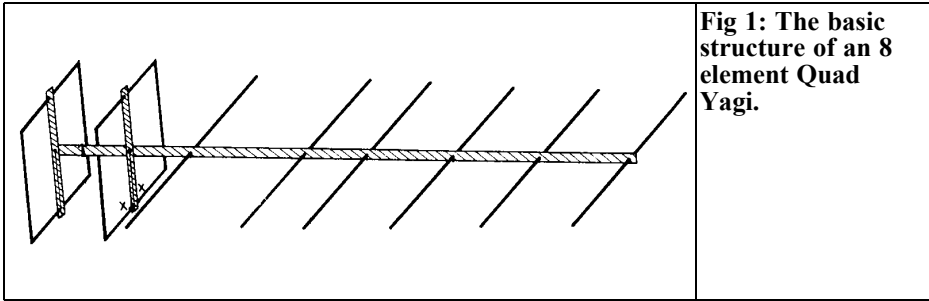
Fig 1: The basic structure of an 8 element Quad Yagi.

2.

Basics

As the name implies a Quagi antenna consists partly of a Cubical Quad antenna for the central exciter and partly a long Yagi. Wayne Overbeck, N6NB, described this form of antenna for the first time in 1972. The Cubical Quad was developed the first time 1942 in Quito (Ecuador) by Clarence C. Moorlands, W9LZX and patented in 1951. The long Yagi is a development of the famous Yagi-Uda antennas, invented in 1926 by two professors Hidetsugu Yagi and Shintaro Uda (both from Japan). An 8 element Quagi is shown in Fig 1.

The Quagi is like most Yagi arrays, a multi-element antenna with a fed element and radiation-coupled parasitic elements. This means that the antenna consists of several single antenna elements with only one element fed, the electromagnetic wave is propagated in a preferred direction (directional antenna). This is confirmed in [1, pp53]. Since Quagi and Yagi arrays have very similar electrically characteristics the characteristics of an optimised Yagi can be used as the basis for development.

Optimised Yagis have different lengths and distances of the parasitic elements (directors). Compared to non-optimised Yagis they have a higher gain as well as better radiation pattern and side lobe

suppression. [2, pp352.]

2.1. Basics of the Yagi structure

2.1.1. Mutual coupling of the parasitic elements

It has already been mentioned that the individual elements are not all fed but are energised by mutual coupling. The individual elements of the antenna are fed by the electrical field, propagated as a wave, from the driven element that is connected to the transmitter. Since the parasitic elements are not attached to a load, they can radiate the energy absorbed from the wave field without loss into the ether.

In order to achieve a directive effect; the parasitic elements must be the correct length and spacing so that the field radiated in the preferred direction can reinforce the existing field. In the opposite direction the fields cancel and thus reduce the effect. [1, pp63.] shows the complete mathematical derivation and proof of this behaviour.

The following statement is explained by the information shown in Fig 2. The amplitude of the current flow in the parasitic element drops rapidly with the distance *a* from the fed element. However the phase shift of the current flow increases. In order to fulfil a defined phase shift between the current flow in both elements e.g. to achieve a directional effect, the element distance *a*
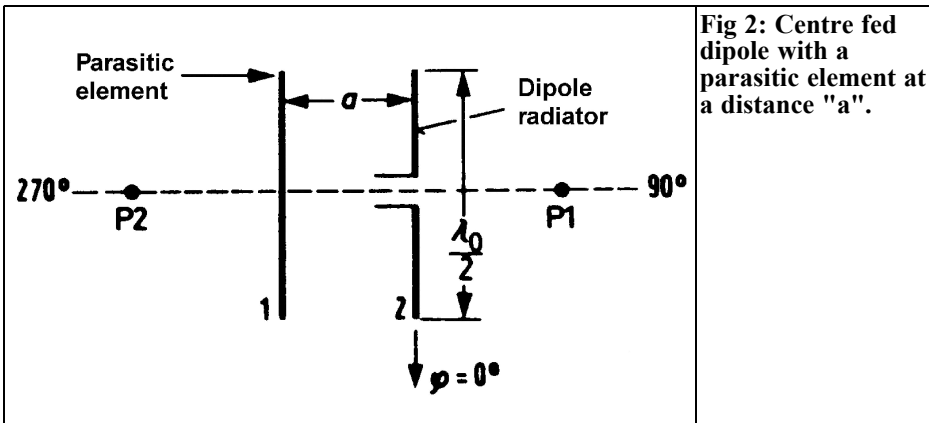
Fig 2: Centre fed dipole with a parasitic element at a distance "a".

should be between $\lambda_0/4$ and $\lambda_0/2$. It is also possible to achieve a defined phase shift by changing the length of the parasitic elements when the distance to the dipole antenna is smaller. Extending the reflector or shortening the director compared to the fed element can increase the directivity and gain (principle of the super gain antenna).

Thus there are many variations with just two elements for a directional antenna. Naturally the complexity increases with an increasing number of elements.

In addition the antenna feed impedance, S, is greatly affected by number an position of the parasitic elements, P.

Different combinations of half wave dipoles feeds and $\lambda_0/2$ parasitic elements are shown in Figs 3a – 3c.

The antenna feed impedance of the half wave dipole is somewhat lowered in Fig 3a. In Fig 3b the impedance increases and in Fig 3c it corresponds to the input impedance of a half wave dipole in the free space. [1, pp65]

Additionally the gain of these arrangements is different. In general it can be said that with an increasing number of directors the gain of an antenna increases and the 3db radiation angle decreases.
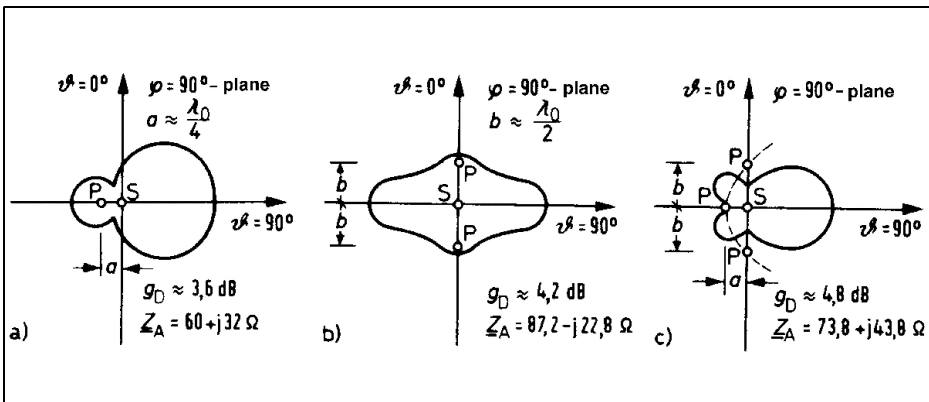
Fig 3: The effects of different parasitic element on the input impedance: a) impedance reduced, b) impedance increased, c) normal situation.
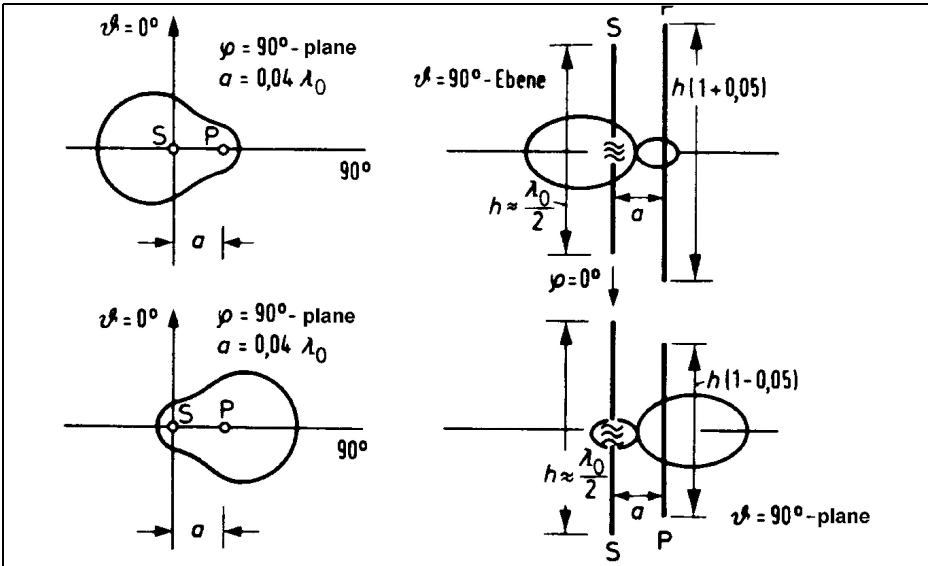
Fig 4: The effects of changing the length of the parasitic element.

2.1.2. Behaviour of the radiation pattern, gain and impedance as a function of the parasitic element

It was suggested above that changing the length of the parasitic elements changes other parameters. This is most noticeable in the radiation pattern and gain of the antenna. The conclusions from [1, pp66] are shown in Fig 4:

If a parasitic, P, element about 5% longer is placed behind the feed dipole, S, then the result is the radiation pattern shown in Fig 4a. This element is called a reflector and is not connected to the radiator. This effect is caused by the fact that parasitic elements that are longer than the fed element cause an inductive resistance according to the following formula taken from [1, pp65]:

$$Z = -jZ \cot \frac{\pi h^*}{\lambda_0}$$

$$h^* = h(1 + 0.05)$$

Elements that are shorter by x% are known as directors and introduce a capacitive resistance. They obey the above formula with $h^* = h (1-x)$. The radiation pattern is shown in Fig 4b.

The gain of these antenna combinations is still dependant on the distance between the radiator and the parasitic element. The effect of the relative distance a/λ_0 on the gain for the dipole and reflector, a, and the dipole and director, b, is shown in Fig 5.

It can be seen that a theoretically higher gain can be achieved with a dipole and director rather than a dipole and reflector. Since the antennas can be used for different purposes, one must weigh up whether a high gain is best. The distances for maximum gain are different for each type. The gain of a Quagi antenna as a function of the number of directors is considered later.

The relationship of an antenna system as a function of the length of the parasitic
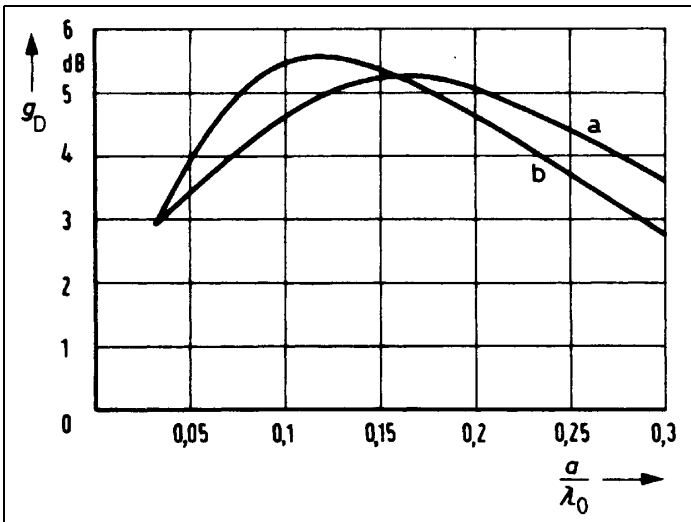
Fig 5: Graph showing the effect of changing the distance of the parasitic element on gain.

a) Dipole with reflector

b) Dipole with director.

elements is shown in Fig 6. A field strength relationship of 3 (9 times current flow) corresponds to about 9dBs from the reflector arrangement.

As described in 2.1.1, the input impedance of the fed element changes as a function of the number, arrangement and the distances of the parasitic elements. If a parasitic element is made closer to the emitter, then the input impedance is reduced. The behaviour of the input

impedance as a function of the element distance, related to the wavelength (a/λ_0) as well as the kind of the parasitic element is shown in Fig 7.

The impedance, R_A, is mainly dependent on the distance of the appropriate element. It is possible to have input impedances as low as 20Ω. Loss free matching networks or different radiation elements are used to transform this to the normal feeder impedance of 50Ω. If the input
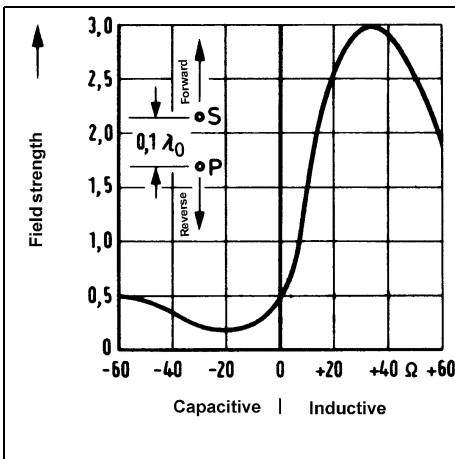
Fig 6: The effect of changing the length of the parasitic element.
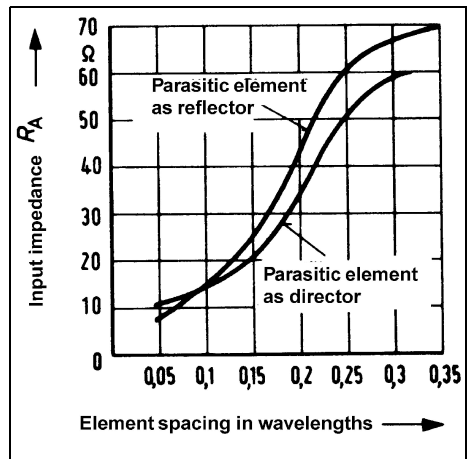
Fig 7: The effect on input impedance of changing element spacing.
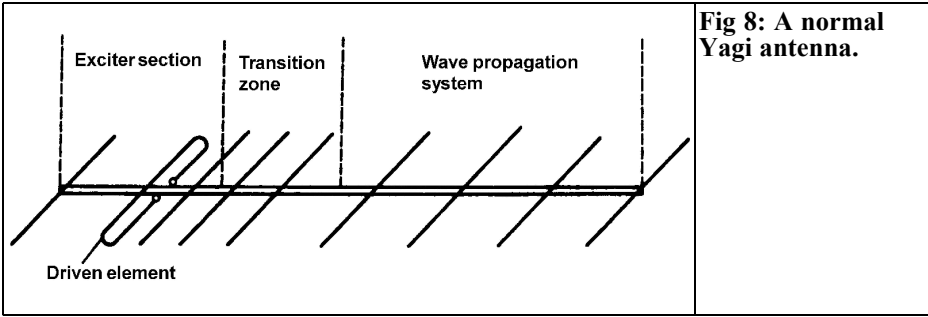
Fig 8: A normal Yagi antenna.

impedance of the antenna becomes too small either a folded dipole (240Ω in free space) or a Quad element (120Ω in free space) can be used.

Apart from the familiar variables, the element thickness must be considered. The bandwidth increases with the thickness of the elements because the Ohmic loss is reduced. The shortening factor *V*, as a function of wavelength to element thickness, must be considered. An appropriate diagram is shown in the appendix.

2.1.3. Exciter section of a Quagi

The preceding explanations concerning the wave behaviour and the measured variables referred to a Yagi like the one shown in Fig 8.

As already described, the electrical behaviour of a Quagi is almost identical to that of a Yagi and therefore it has been assumed that their performance is the same.

However the exciter section must be modified as shown in Fig 9. Instead of the folded dipole aerial and a reflector, two Quad elements (Cubical Quad) are used. Thereby substantial improvements can be obtained.

2.2. Quad loop and Cubical Quad

2.2.1. Quad loop

Beside the normal dipole antennas there are also loop antennas, Fig 10 shows different kinds of loops. The advantage of these is that they electrically continuous making them immune to static disturbances. This also makes them substantially less sensitive to environmental and polarisation issues. There are also many more design variation possible than for a simple dipole.

Square and rectangular full wave loops will be used for the following designs. A detailed explanation of the different kinds of loop is included in [2, pp318].

The Quad loop (square loop) can be
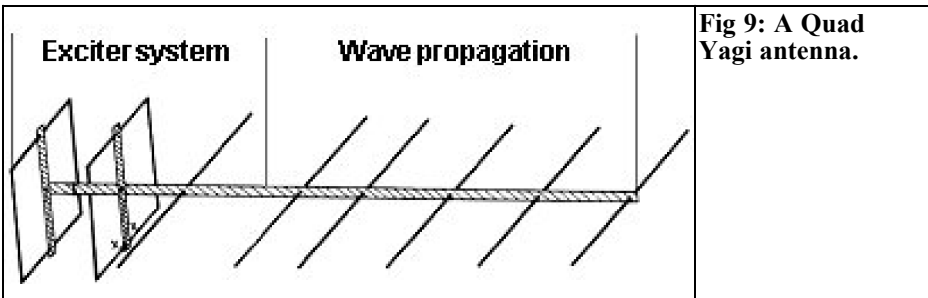
Fig 9: A Quad Yagi antenna.
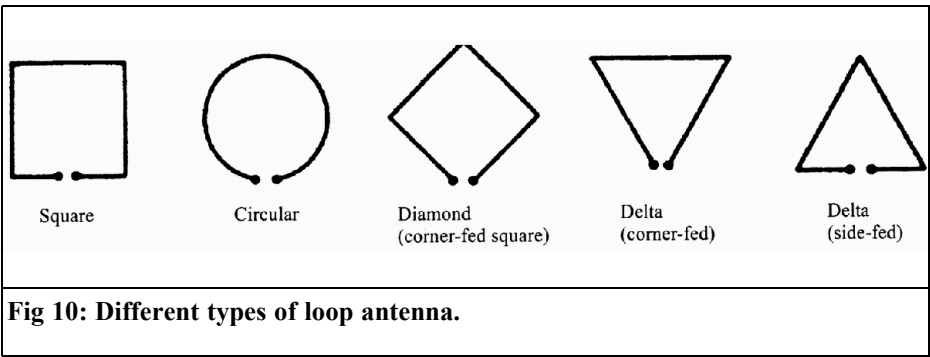
Fig 10: Different types of loop antenna.

considered like a folded dipole used on a Yagi. If a folded dipole aerial (240Ω input impedance) is opened out on the long side, then it forms a square loop. The current distribution in the loop does not change from the folded dipole aerial, as Fig 11 shows.

The square full-wave loop $l=\lambda$ is theoretical in resonance, where l is the actual length of the square plus or minus a factor U. In practice it was found that loop antennas do not require a shortening factor, but rather need an extension factor. This effect cannot be attributed to the capacitive effects that arise with dipole antennas because loops have a closed form. Bending the feeder wires causes

this extension effect. For Quad loops a shortening factor of approximately 1.02 should be used. The total can be determined using the general calculation formula:

U/m = 304 • MHz/   [2, pp323]

Fig 14 shows how the radiation resistance varies as a function of the separation factor. Since the square loop has a fixed separation factor of 0.25λ, the radiation resistance is approximately 115 to 120Ω (in free space). Later we will see that this radiation resistance proves as extremely favourable for the overall construction.

The gain of such a square loop is approximately 1dB higher than a half wave dipole. This can be explained by considering the Quad loop as two half wave
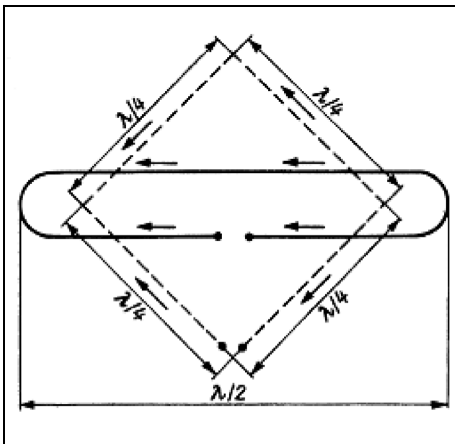
Fig 11: Comparison of a folded dipole and a Quad loop element.
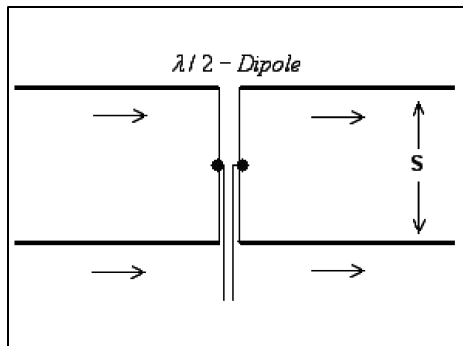
Fig 12: A Quad loop viewed as a pair of half wave dipoles.
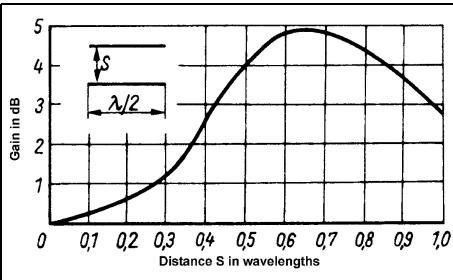
Fig 13: Relationship of dipole spacing on gain.
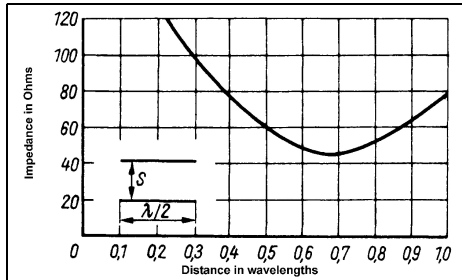
Fig 14: Relationship of dipole spacing on input impedance.

dipoles stacked at a distance of 0.25λ as shown in Fig 12.

Fig. 13 shows the gain of an array of parallel dipoles as a function of the stacking distance, S, the gain is relative to a half wave dipole. With 0.25λ spacing the gain shown on graph is 1dB relative to a half wave dipole. The gain can also be determined by the effective area of the antenna. The literature [2, pp319] describes that the effective area of a Quad loop antenna is $A_w=0.164λ^2$. From [5, pp17] gives the familiar formula for the gain over an isotropic emitter:

$$G_{global} = A_W \cdot \frac{4\pi}{\lambda^2} = 0.164\lambda^2 \cdot \frac{4\pi}{\lambda^2} = 2.06$$

$$G / dBi = 10 \cdot \log G = 10 \cdot \log 2.06 = 3.12$$

The maximum possible gain is achieved a stacking distance of 0.65λ (rectangular or oblong loop). However this gain optimisation causes a reduced radiation angle and disturbing side lobes.

2.2.2. Polarization of a Quad loop

The polarization of square loop antennas depends exclusively on the feed point. If this is on the horizontal side, as shown in Fig 15a, it gives horizontal polarization. This can be seen from the current flow arrows, the sides A and C are fed in phase and the sides B and D are fed in anti-phase. Clearly the polarization is horizontal. When the feed is on the vertical sides this gives vertical polarization.

If a square loop stands on it's points and is fed there, then it is called a pointed
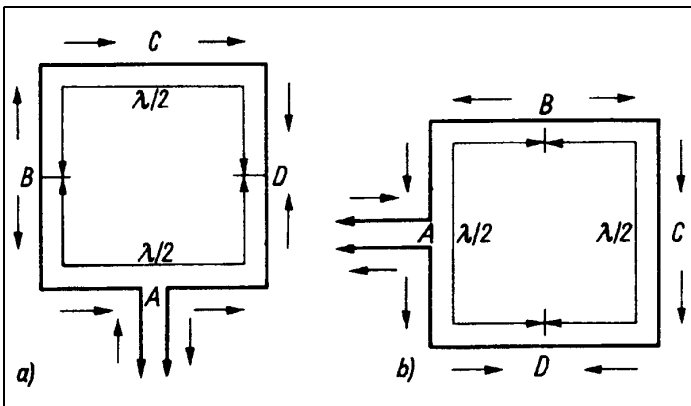
Fig 15: Polarisation of a Quad loop with different feed point.

a) Horizontal polarisation.

b) vertical polarisation.
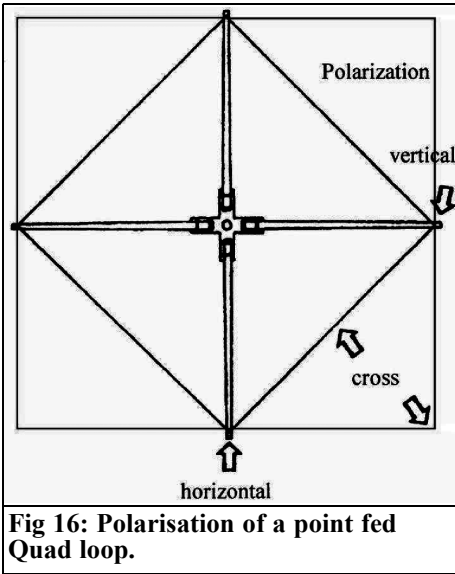
Fig 16: Polarisation of a point fed Quad loop.
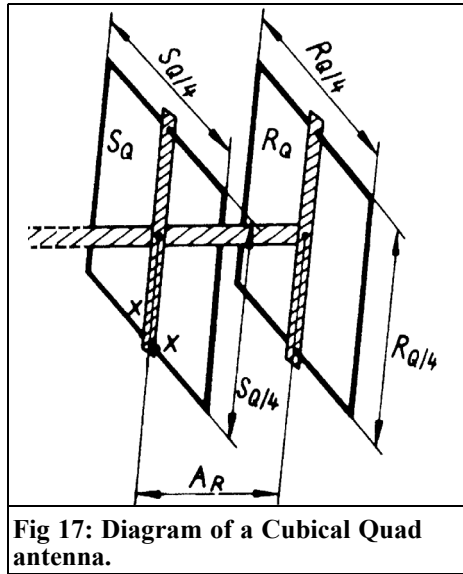
Fig 17: Diagram of a Cubical Quad antenna.

quad (Fig 16). At first sight it is not clear what polarization will be generated. The behaviour is well-know, it will be cross polarization when fed between the usual feed points for horizontal and vertical polarization. The characteristics of a single loop antenna are used to develop the antennas in the next chapter.

2.2.3. Cubical Quad

The Cubical Quad is an arrangement of two Quad elements with one fed as a parasitic element as described above for the Yagi antenna. The word "cubical" describes the appearance of this arrangement (Fig 17). Radio amateur like to use this antenna because of it's advantages.

The values for maximum gain and forward to back relationship already described are still valid. Also the antenna can be optimised depending the intended use.

The general dimensions for a cubical quad antenna are: a fed 1λ Quad loop element with an additional Quad element spaced at a distance of approximately 0.1 to 0.2λ from the fed element. If the

parasitic element is somewhat larger than the fed loop, then it works as reflector (inductive phase shift).

As already described, a simple Quad element has a shortening factor of 1.02 in relation to the resonance wavelength. For a Cubical Quad the shortening factor for the driven element should be from 1.01 to 1.015 and from 1.05 to 1.07 for the parasitic element (reflector). [2, pp388f]

The formulae taken from [2, pp388] are:

Overall size of the reflector element:

$$U_R / m = 316.3 \cdot MHz / f$$

Overall size of the fed element:

$$U_S / m = 303.3 \cdot MHz / f$$

Distance between both elements:

$$A_R / m = 44.4 \cdot MHz / f$$

The maximum gain of this arrangement is when $A_R = 0.125λ$ and is approximately 7.3dBi. The impedance varies

between 60 and 120Ω and depends on A_R (0.1 to 0.2λ).

To be continued ……..

X.

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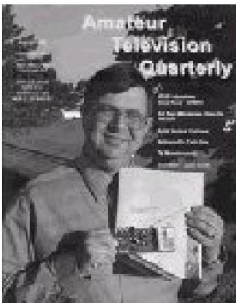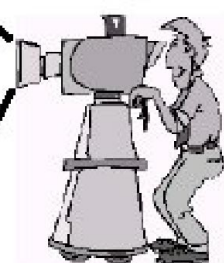
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[3] The Quagi Antenna Turns30, Wayne Overbeck, N6NB

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Carl G. Lodström SM6MOM & KQ6AX

The Noble Art of Piping DC to the LNA

1.

Background

The nice presentation of the potent MGA 62563 Low Noise Amplifier (LNA) by Franco, I2FHW, in the 3/2005 issue of VHF Communications Magazine became the inspiration for this article. That amplifier must be the ideal LNA for "In The Antenna" applications. I have not tried it yet, but a sample shipment of five is on its way! (Fantastic! I called the local HP for samples last Friday. They were going to be sent to me in California from Germany, and here they are on a Wednesday afternoon! Good job!) Now, where did I put the microscope?

The MGA-62563 is almost microscopic! Fig 1 shows it compared to a 1206 (a common size in normal electronics) surface mount capacitor. Yes, I know there are even smaller components that look like not very coarsely ground black pepper!

Like most of you have surely done, I too have put pre-amps in the antenna for the old TV, and even to the ham radio. Once I even designed a bi-directional amplifier for 2m. A low noise FET one-way and a 20W PA module the other. It has automatic T/R switching. There is an article about it in VHF Communications Maga-

zine [1].

While I was waiting for the MGA 62563 I tried to kick it around in Eagleware, GENESYS, to see if something could be done to improve match, gain or NF. I must say that, amazingly, all parameters were almost as good as can be! Only S_{11} could be improved a little, a less wideband application, with a very small inductance in series with the input. But it is good enough as it is anyway. So this is truly the "50Ω in - 50Ω out" amplifier! Fig 2 shows the MGA-62563 built up as a General Purpose Amplifier with a 5V regulator and BNC connectors.

The past decade, or few, I have usually employed the pretty good MAR-6 from Mini Circuits [2]. It has a reasonable match to 50Ω and very good match, in
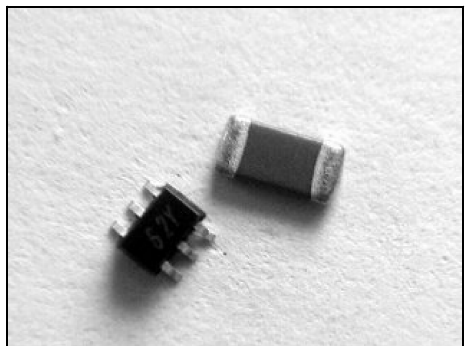
Fig 1: The MGA-62563 compared to a standard 1206 SMD capacitor.
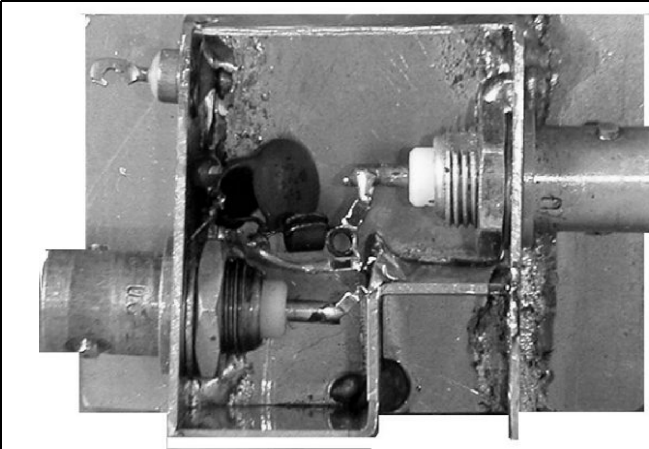
Fig 2: Picture of an MGA-63563 as an LNA.

and out, to 50Ω and 75Ω as well, it needs only 1mA, it is low cost and easy to employ. 20dB gain and 3dB NF are not too bad either. It has almost become the Standard Gain Element in my toolbox. But the MGA 62563 is a far better LNA than the MAR-6. The most attractive way to supply DC power to an antenna LNA is to feed it via the coax. We will look at how to do this for a MAR-type amplifier and for the MGA type, with its extra connection to V+, as well. This practice has a few traps that I will touch on later.

2.

Considerations

In a city, and sometimes in rural environments, there are often some very strong signals floating around. TV and FM transmitters, pagers… and if we use an LNA with a wide band antenna the LNA may start to cause intermodulation.

Adding an LNA that makes the signals another 20dB stronger can create real problems. So, before adding an LNA, make sure that you look at what you are getting from your antenna. If the antenna is tuned, like a single band Yagi, you may not have many strong out of band signals. If it is just a vertical stick you may have all kinds of strong signals.

If this describes your situation you will have to fix it before putting an LNA in line. Possible actions are:

- Using a tuned antenna for the frequency of interest
- Making the antenna directional using a rotator.
- Adding resonant traps in the feeder.

All amplifiers (not to mention receiver inputs!) have a limited ability to deal with many strong signals at once. The MAR-6 has 1dB output compression at +2dBm, also indicating an output level where the signals will begin to mess up each other. With about a 20dB gain, this is −18dBm on the input. This level can occasionally be reached with an antenna on the roof in an urban setting. Intermixing of the signals, strong and weak, will lead to the distortion of them all.

The 1dB compression point for MGA-62563 is 17.8dBm. As this amplifier also has about 20dB gain, it corresponds to an input level of -2dBm. You will have to be close to a quite strong transmitter to get this! This amplifier ought to do fine in most city environments. Already the MAR-6 does fine in many cases. It is limited to about 2GHz, whereas the MGA goes to 3GHz! The better NF (0.8dB) of MGA compared to MAR-6

(NF ~ 3dB) can make quite a difference though.

The MGA amplifier can handle about 16dB more signal than the MAR before the output starts to compress or acting as a mixer for all input signals. This ability is expressed as the IM3 point, referred to the output. For the MAR-6 it is 14.5dBm and 32dBm for the MGA-62563. Both have ~ 20dB gain so the numbers can be compared.

With all this nice gain and low NF we could even have a tunable band pass filter, a "Pre-selector" on the receiver input, to help it cope, and be ahead of the game, even if the filter has a terrible insertion loss, like 10dB. A more realistic value, for a reasonable band pass filter, may be 2-3dB.

As I received the LNA samples, I have built up one. It has made me realise that just putting one on a receiver antenna input is not all good. With a simple 40cm long indoor whip antenna the VHF signals got stronger with the amplifier, previously noisy signals got clear, as expected, but the noise floor came up to S5 (on a IC-R7000 set to narrowband FM reception)! Putting an amplifier like this on the receiver input makes a lot of "no-signal" stuff audible!

An attenuator, of some 10-12dB, on the RX input, would not ruin the system NF by much, but it would get the noise floor down. Not a solution really. The best solution is probably a band pass filter. This can be done, but is outside the scope of this article. As I began to look into it I have some ideas that may be interesting, so I think I may write a separate note about what I find.

At VHF and UHF the feeder to the antenna is usually a coaxial cable. They may present a considerable loss if they are long and the frequency high. On reception the loss is added directly to the NF of the receiver. So if a particular cable has a 5dB loss and the receiver has a 3dB NF (not stellar but OK) the whole setup has a NF of 8dB! Not stellar, not even good, it is pretty bad! A LNA on the shack-end of the cable, by the receiver input, would not help at all. If we used the MGA-62563 (NF= 0.8dB!) the 5dB cable loss would add to the 0.8dB NF in the amplifier => 5.8dB NF. The NF property of the amplifier would be wasted.

With the LNA in the antenna feed point we get a tremendous advantage from the apparent "multiplication of antennae" because of the 20dB gain. One antenna, and an LNA with 20dB gain, is basically equivalent to a 100 antennae stack of the same antennae! Even after a 5dB cable loss it is like a stack of 32 antennae! The price to pay is the NF of the amplifier. In the MAR-6 it is 3dB, meaning that the noise power (of the thermal background noise) is doubled. So we are still ahead, compared to the one antenna we started out with. With a 5dB loss in the cable, even a perfect receiver would get a NF of 5dB. With MAR-6 in the antenna, cable loss would be more than made up for by the gain. Even a few dB of NF in the receiver (5 or so) would be cancelled. Although the MAR-6 adds 3dB to the input (on the antenna level) noise floor, thus still limiting the weakest detectable signal, the better antenna location (on a roof) outweighs this.

The excellent 0.8dB NF of the MGA-62563 amp is almost un-touched, if placed at the antenna end of the cable. Even with 25m of RG-58, at 450MHz, the total NF on the receiver input is still only 1dB, with an associated gain of 14dB! With RG-8 the cable length can be doubled, to 52m, for the same data.

3.

A very useful relay

We have determined that the advantages of placing an LNA at the very feed point
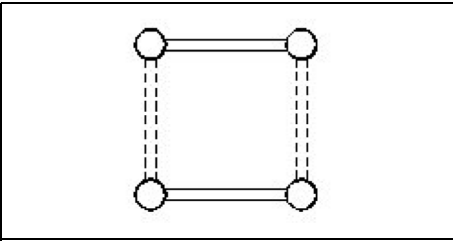
Fig 3: The layout of pins on a transfer relay.
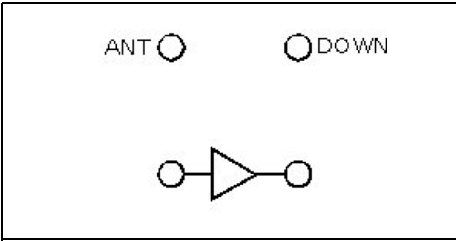
Fig 4: Pass/bypass connections for the transfer relay.

of the antenna are too great to be disregarded! If the antenna is also used for transmission the LNA has to be bypassed. For VHF – UHF the best solution is a coaxial relay of the Transfer type. It can either bypass or engage the LNA. The Transfer relay is probably familiar to the readers of VHF Communications Magazine, but just to make sure: It has four connectors, often arranged as a square (Fig 3). In one state the solid lines connect the connectors. In the other state they are connected by the dotted lines. Some are "latching", meaning that they stay in the position they last switched to. Else they are called "fail safe".

Fig 4 shows the pass/bypass connection. While bypassed one can transmit via the upper path and the isolation in this type of a transfer switch is usually very good, so 100W will probably not damage the LNA. There is the possibility that the shorted LNA (when bypassed) oscillates. Its output is connected to its input. Check this before putting up the relay and LNA! All I can think of for to cure this is to either change the cable lengths to the LNA a little or use the same voltage for

the LNA as for to activate the relay. Disabling the LNA when not switched in. Connect a diode across the coil to absorb the inductive kick!

With the potent output capacity of 18dBm (for 1dB compression) the MGA-62563 could even be of interest as a PA for QRP! A Transfer relay can also reverse the direction (Fig 5) of the amplifier! Placed in the antenna, a much lower input at the shack-end of the coax will be satisfactory. With a cable loss of for example 5dB, a +4dBm input to the cable will drive the amplifier fully, sufficient for 18dBm out. Without the amp on top you would have to feed +23dBm at the shack end of the same cable for the same antenna output.

At 1.3GHz the gain is some 17dB and at 2.4GHz it is ~12dB. This may still save one or two gain stages in the PA.

4.

Connecting the DC
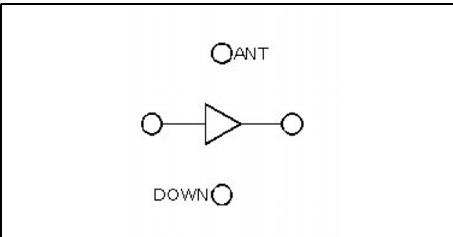
Fig 5: Transfer relay used in reverse direction.

First, let us see how the "MAR-amplifiers" are supposed to be powered! Most amplifiers of this kind like the ERA, MAV. RAM and many more families, are to be fed a *current*, not a voltage! I have seen "designs" in which this was misunderstood! Look at the data for MAR-6! They are listed as 16mA and 3.5V. This does not mean that you shall put 3.5V on the amplifier! It means that
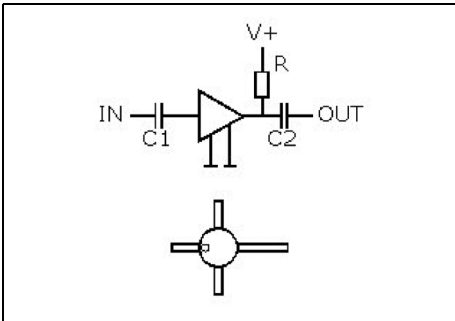
Fig 6: The basic circuit for MAR-6 amplifiers.
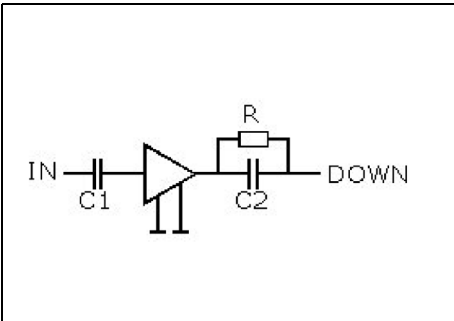
Fig 7: Simplified circuit when feeding via coax cable.

you feed it from a higher voltage *via a resistor*. 16mA will pass and 3.5V will be dropped over the amplifier, so if you feed it from a 12V source: $12 - 3.5 = 8.5$V to be dropped in the resistor. 8.5V / 16mA = 531Ω = R. In reality: maybe 510Ω or 560Ω. As V+ is de-coupled for RF, place R as close as possible to the short connection between the amplifier output and the coax upper end. The R will then act almost like an RF choke and radiate nothing (Fig 6).

If V+ were to be 5V => $R=94\Omega$. Its load of the output would be a concern, so now we add a RF choke between R bottom end and the output line.

A DC voltage will develop on the input from the internal biasing of the amplifier. A blocking capacitor is needed on both input and output, C1 and C2 in the schematic diagram above. One subtle point here concerns a DC open filter or antenna. It would be in series with C1. Imagine that both the antenna and the C1 has each $100M\Omega$, they will divide the ~ 1.5V bias voltage between them. As the antenna is out in the open and C1 hopefully is not, galvanic corrosion will develop on the antenna until the resistance drops and C1 carries all the voltage. This may be of no importance, but a 1M resistor to ground, outside C1, will prevent any DC to develop over the antenna and the corrosion from ever getting started.

Fig 6 shows a picture of what this amplifier can look like, a little windmill. The output can have a longer pin or the input can be marked with a dot. The two pins up and down are to be grounded. It is important to use as short part of the leads as possible! It may be good to let the ground plane come in as a ' V ' from either side, all the way up to the body. Use pieces of copper tape or solder wick braid if you have to extend an existing ground plane.

Looking at Fig 7; the DC for the LNA is coming through the coax, we can simplify our schematic considerably: DC is coming up the cable from "DOWN", drops in R, feeds the LNA, while freshly amplified RF can pass C2 from the amplifier output to the coax input! Now we will never have to use any RFC in series with the R! With regular 1206 surface mount components, R and C2 can be stacked on top of each other, or side-by-side, and the whole project becomes
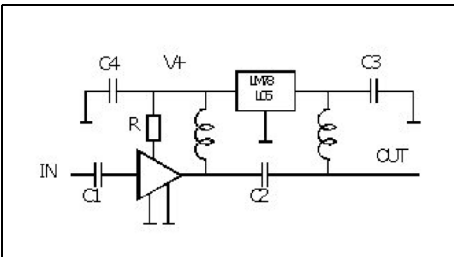
Fig 8: Circuit for powering the MGA-62563.

121
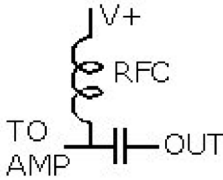
Fig 9: Bias Tee for feeding the coax cable at the shack end.

the size of a bean!

The other kind of LNA

Well, there may be many other kinds, but here we are interested in the MGA-62563, a performer that is not likely to be surpassed for some time to come.

Even here the power can come up via the coax (Fig 8). It is probably easier than to feed a separate wire for +5V. Up in the antenna is now a fixed 5V regulator. I happen to like the LM78L05 as it survives 35V in, is small and has excellent regulation, but there are others. A little $0.33 \sim 1\mu F/35V$ tantalum (C3) on its input and a $0.1\mu F$ disc cap (shortest possible leads!) on the output (C4) completes the de-coupling on the 5V side, here called V+. The inductors are RF chokes, and as the likely frequency is >100MHz, they can be small. A few turns through a ferrite bead plus a few turns in air may well make a fine wide band choke. R is here for the biasing of the LNA, se the data sheet [4]. See also the Application Note [5] for more about how to use of this amplifier! We are now operating with a truly constant voltage (5V) regardless of input voltages in the $6.5 < Vin < 35$ range. This version, with a regulator, can of course be used for the MAR-type of amplifiers as well, and the regulated voltage does not have to be 5V either.

The other end.

So, the LNA in the antenna is ready and installed. Since any reasonable supply

voltage can be accommodated in the MAR-type, by a good choice of R, almost any wall plug transformer with a DC output can be used. 9 – 12 – 15 or even 24V does not matter much. Select your R accordingly! How are we to apply the power to the cable? Connect the amplifier via a Bias Tee (at the shack end of the coax) like the one in Fig 9! For a few hundred MHz C can be a few hundred pF.

The forest trap!

You get it all together and after some listening, you find that the amplifier must be dead! How could this happen, you were so careful in calculating the R! Misread the colour code? No!

We have at least two factors to consider:

- The wall plug transformer may read "12v/300mA" but at the small load of 16mA (in the MAR-6 case) the voltage may be closer to 15 - 18V.
- Before use: put a resistor load on the supply and measure what voltage you will really get! Thereafter: calculate the bias R for the MAR-6.

Much more malicious is a pulse travelling up the cable! It was generated when you connected the cable to the Bias Tee with the power supply already turned on! The sharp 18V ramp travels up the cable. As it reaches C2 it goes right through to the LNA output, which immediately burns up!

The excellent LNA transistor BFP420, that only can handle a V_{CC} of 5 or 6V, as a LNA in the antenna, died every time I forgot this! It was a lot of work to take down the antenna and re-build the LNA each time!

The way to connect it all is to plug in the wall transformer last! The voltage rises then in a slow ramp that will not go through C2 (which is less than 1nF). Also beware of that a length of coax, with open ends, can acquire and store static electricity! Keep some kind of load (a termination or an attenuator for example)

or short on the other end of the cable while installing to the LNA. Leave it on until you are ready to connect the transformer. Grab each end (cable and transformer) with your hands before mating them! Let the cable grounds touch first! Cables with F-connectors, used for RG-59 cable, can be tricky. When mating they can make contact on the centre conductor before they connect the shields.

The concern for killing the MGA amplifier with intermittent connection of the power supply is still real! A sharp step function, of 18V for example, would still go right through C2. Even though the regulator survives, the amplifier may very well not!

So watch out for intermittent connection! It is the same as disconnecting and reconnecting the cable with DC power on! If you have a coax switch in the shack, see to that the Bias Tee and the DC feed are on the LNA-side of this switch.

5.

Keeping dry

It is difficult problem to keep something dry up on the mast. Closing it in seems not to work; it will only ensure that the water that got in anyway will not get out!

Vaseline or silicone grease, applied generously direct on the circuitry, helps. Beware that if the amplifier have operating voltages applied, and if humidity enters, galvanic corrosion will be fast and deadly! Greases may prevent this, or at least delay its onset. It is good to enclose it well, but to leave some openings in the bottom! Make them small (it limits the size of the spiders who can get in!) or cover them with a fine metal mesh. If the compartment is well sealed you may consider a plug of sintered bronze spheres, as used for mufflers on pneu-

matic tools. They will readily let air and humidity through but just about nothing else. Place them in the floor of the box!

Another trick is to let power dissipate in, or on, the box. The box can be really small, so already a few hundred mW will warm it up enough for to keep it nice and dry. If a MAR-6 is driven from an idling 24V wall plug there may well be 30V. => 424mW in the resistor and 56mW in the amplifier! The same for the MGA amplifier, it can run on as much as 80mA at 5V and develop some heat in itself. If a LM78L05 is used it may also have to be heat sunk; it can then dissipate up to 750mW. Use the enclosure for a heat sink! Another source of heat is the relay. It can be left energised over night, dissipating a few Watt!

6.

References

[1] A Bi-Directional Amplifier for 2-m, Carl G. Lodström, SM6MOM & KQ6AX, VHF Communications Magazine 1/95 pp 18 –21.

[2] http://www.minicircuits.com/amplifie.shtml

[3] See what transfer relays there are to find under:

http://search.ebay.com/search/search.dll?cgiurl=http%3A%2F%2Fcgi.ebay.com%2Fws%2Ffkr=1&from=R8&satitle=transfer+relay&category0=

[4] MGA-62563 data sheet http://cp.literature.agilent.com/litweb/pdf/5989-1812EN.pdf

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Ted Barnes, ZL2IP

12mm DX in New Zealand, March 2006

One of Steve, ZL1TPH's missions in life is knocking off as many microwave distance records as possible. Because of the very few people operational on 24GHz in ZL he built a pair of real neat transverters, as he does, and sent one to me, knowing I had a big hill in the back yard that should allow a good long distance to the horizon, and the possibility of beating the existing 126km mark.

The transverters are assembled with modules and components from a variety of sources, including the Wellington VHF Group trading table. The local oscillator and mixer modules are from Kuhne electronic, the PA and LNA are from mm-wave and millitech. Power output for one unit was 1.5W, the other 0.5W. The noise figures were 2db and 5db. The antennas are 330mm diameter Cassegrain feed units attached to the front of the transverter chassis. See photo. At 24 Ghz the 3db beamwidth for these antennas is 2.5deg.

The IF radios are 2 meter IC202's, modified to give +DC volts on the IF coax on transmit to work the T/R sequencer in the transverter. Normal operation is on 24048.100MHz USB. The choice of .1 rather than .2 is because the IC202 is a VXO rig, and .2 comes on the boundary of two tuning ranges. The exact frequency of the transverter LO frequency may be 10kHz out due to the inherent accuracy of the crystal set up, and also will vary a few kHz over a period of minutes due to the crystal drifting. These crystals are temperature stabilized, but even so multiplying a 5th overtone crystal at about 100MHz by some thing like 240 you get to use the RIT a lot!

With a beam width of a degree or two, and a frequency that can drift outside the 3kHz pass band of the IF radio, to have a successful DX contact a bit of prior planning is helpful. The first thing you need to know is where north is. Accurately. I planned to operate from North Mt Egmont, so I went there on a fine day with a telescope mounted on a tripod with azimuth calibrations and a list of features I might see in the distance along with their azimuth angles. From these measurements I

reckoned I found north to +/- 0.5 deg. I then used a long aluminium bar to transfer this direction to the tar seal of the roadway. A couple of washers were hammered onto the road surface with 100mm nails so if I came back and was in the cloud I had North nailed as you might say. As a cross check the GPS was used as a compass by walking about 20m to and from the tripod, but this is not really accurate enough for this application but a good independent reality check. A proper GPS compass would be cool, and eliminate all the above fluffing about, but I don't have one of those.

The path to be attempted for the record attack was North Mt Egmont to a site near Raglan, a distance of 169km. This looked like a true line of sight path, with the Egmont end being at 962m and the Raglan end at 226m. I estimated the path loss + 32 db of water vapour absorption would give 30 db S/N margin.

In the event, on the day, it was too easy! I heard Steve's carrier straight off, and we exchanged 5,5 signal reports. He then discovered that was on a side lobe, so when lined up signals were 5,9++, so we rattled on to each other for half an hour or so. Ray ZL2TAL was with me, and Kevin ZL1UJG was with Steve, so everybody had a good chat. A CW contact was made to chalk up that mode. Brian ZL1AVZ came up on 24GHz but we were unable to work him. At 270km you need some help with the propagation or some very high hills!

All enthused, the next weekend Steve went to Klondyke, and I went back to North Mt Egmont. 221km and a bit problematic path wise and weather wise. About 20km at sea level in the middle and humidity 70%. But he picked up my carrier OK and we had a contact 5,5. There was slow QSB with a period of a minute or so, which gave us an estimated 5db margin on peaks and not quite readable in between. Harry ZL1BK was with him so exchanged numbers with him as well, then the rain came in at their end and the signal faded off into the mush. No CW contact was attempted.

One day we will strike an opening, then who knows how far the signal will go on 24048MHz!

Gunthard Kraus, DG8GB

Internet Treasure Trove

AVAGO

This is a new name to learn and get accustomed to looking at their home page from time to time. Following the change from Hewlett Packard to Agilent there have been more changes making a completely separate company for all semiconductor activities. It will take time to get accustomed to the new homepage until you find familiar things (Application Notes, data sheets etc.).

Address: http://www.avagotech.com/

OH2AUE

Right at the top of Finland there are some people who are unbelievably active on the microwave bands and are not frighten of anything (EME attempts soon). Looking around the homepage of Michael Fletcher you have to ask yourself, when did he learn all this and when does he build it all. Respect!

Address:
http://www.kolumbus.fi/michael.fletcher/index.html

Tonne of software

The personal site of James A. Tonne. It offers some interesting filter and antenna programs, these are partly for downloading and partly interactive versions.

Address: http://tonnesoftware.com/

ON-Semiconductors

This is a joy for anyone who works with PSPICE and will probably be repeatedly visited. It offers thousands of SPICE models for all components and not just "standard Spice" versions. Also modern things are to be found such as "ILIB, IsSPICE, OrCAD template "and other exotic things. These can be found very simply using the desired keyword in the search machine.

Address:
http://www.onsemi.com/PowerSolutions/supportDoc.do? type=models&category=503

MMT

MMT stands for Millimetre Wave Technology Group. A surprising insight into where modern technology is going. There are; ideas, still crazier production technology for unbelievable things. Simply interesting to spend some time browsing

Address:
http://www.mmt.rl.ac.uk/projects/mrffe/mrffe.html

Matjaz Vidmar

For readers of VHF Communications Magazine, Matjaz Vidmar, S53MV, is well known. Many of his interesting articles have been published over the last few years. He is a professor in Slovenia and has an enthusiasm for technology. His projects are very interesting and worth looking at again.

Address:
http://www.ljudmila.org/hamradio/notune.html

TINA-TI

The free simulation software TINA-TI from Texas Instruments: TINA is an easy-to-use, but powerful, circuit simulation program based on a SPICE engine. TINA-TI™ is a fully functional version of TINA, loaded with a library of TI macromodels plus passive and active models. TINA-TI is limited to circuits with two ICs and up to 20 additional nodes.

Address: www.ti.com/tina-ti

Filter pro

The program "filter pro" is an active filter design program from Texas Instruments and is described by them:

FilterPro™ Multiple Feedback (MFB) and Sallen-Key Design Program is a Windows application. This application designs MFB and Sallen-Key low-pass and high-pass filters using Voltage Feedback Op Amps, resistors and capacitors. It also supports a fully-differential version of the MFB circuit. This program includes Bessell, Butterworth, Chebychev, and linear phase filter types and can be used to design filters from 1 to 10 poles. The capacitor values in each stage can be either selected by the computer or entered by the designer. An "always on" prompt window provides context-sensitive help information to the user. The response of the filter is displayed on a graph, showing gain, phase and group delay over frequency.

Address:
http://focus.ti.com/docs/toolsw/folders/print/filterpro.html

TPS40K

The program "TPS40K - designers software" is power supply design program from Texas Instruments and is described by them:

The TPS40K Designer software provides a simple design tool for both novice and experienced power supply designers

using the Texas Instruments TPS40xxx family of controllers. The software includes advanced analysis tools for the experienced designer to do worst-case loop response analysis and component characterisation. A customisable component database is included. It allows the user to add or remove parts to any of the TI Designer series of power supply software (including SWIFT Designer, V3.3 or Higher).

Address:
http://focus.ti.com/docs/toolsw/folders/print/tps40k-sw.html

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.

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VHF Communications Web Site www.vhfcomm.co.uk

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