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TriHelix or Dish for P3E at 2.4GHz

Simulation and Measurements

Paolo Antoniazzi, IW2ACD and Marco Arecco, IK2WAQ

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Nearly all of the articles in this issue are not from the parent magazine UKW Berichte but from new authors who have submitted articles direct to VHF Communications Magazine. Thank you to all of the new authors, I am always pleased to publish articles submitted to me and hope that there will be more to come. If you think you have an article please send it to me, I can always help with formatting, artwork and translation. Why not give it a go?

Franco Rota is working on an exciting article for the next magazine. This will be an easy to build 10Ghz 1W PA. I have seen the first draft of the article and Franco has found a very useful microwave MMIC.

73s - Andy



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Jean Claude Beneche, FI1AIA

Modification of the Philips PRM8060 for Amateur use

1.0

Introduction

The PRM8060 is a VHF transceiver tuned for the 146 - 174MHz frequencies. The front panel contains an LCD display, a potentiometer and four switches as shown in Fig 1. There is no integrated speaker. To use the transceiver on amateur radio frequencies, the internal EPROM must be changed.

2.0

Compatibility

The following EPROMs are available

Version	Firmware	Freq (MHz)
PRM 8060-SA9	prm8060_144.bin	144-146
PRM 8060-SU0	prm8060_430.bin	430-440

Firmware revisions:

- V1.0 First version with 50 channels.
- V2.0 Backup system state when turning off the transceiver. 16 channels added.
- V2.1 Power switching, shift activation, reverse mode and reset feature bugs fixed.
- V3.0 Control from a serial interface. Channel programming by computer. 1750Hz tone command on the front panel.



Fig 1: The front panel of a PRM 8060 FM transceiver.

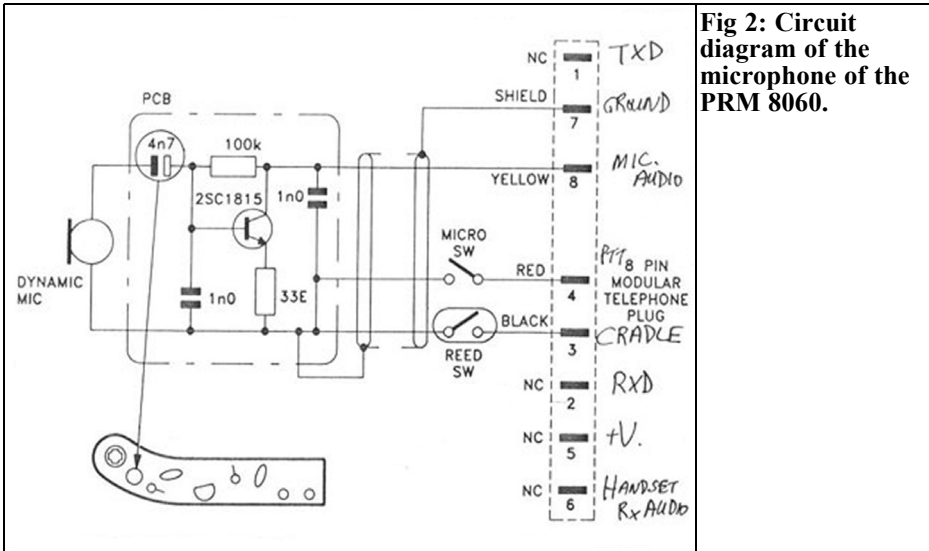


Fig 2: Circuit diagram of the microphone of the PRM 8060.

3.0 Connections

There is a unique RJ45 plug on the front panel for the microphone. This type of plug is used for computer networks.

The original microphone has a transistor

Table 1: Pin connections for the 15 way connector on the rear panel.

Pin	Function
1	Non regulated power 13.8V
2	Regulated power 9V
3	GND
4	Microphone
5	GND
6	Transmit audio in
7	RXD
8	PTT
9	Receive FM out
10	Receive AF out
11	TXD
12	Alarm in (not used)
13	Generation of 1750Hz tone
14	Loudspeaker mute
15	Loudspeaker audio out

preamplifier, a PTT switch and a microphone hang up reed switch. Fig 2 shows the details of the original microphone. The reed switch signal is used for 1750Hz tone burst see modifications.

The back panel has a BNC plug for the antenna. There is Molex connector for the 13.8V power connection and the loudspeaker, see Fig 3. The Molex connector used is the same type of plug used for the power plug of a CD drive or hard drive on a PC. If the original cable is missing it is possible to get one from an old PC power supply.

There is also a DB15 connector for accessories on the back panel. It is used

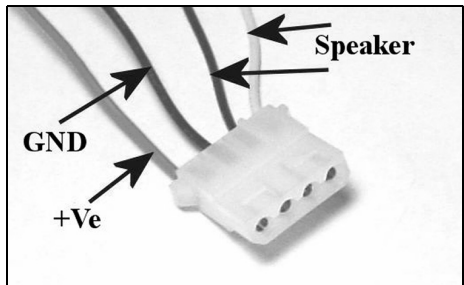


Fig 3: Molex connector for power and loudspeaker.

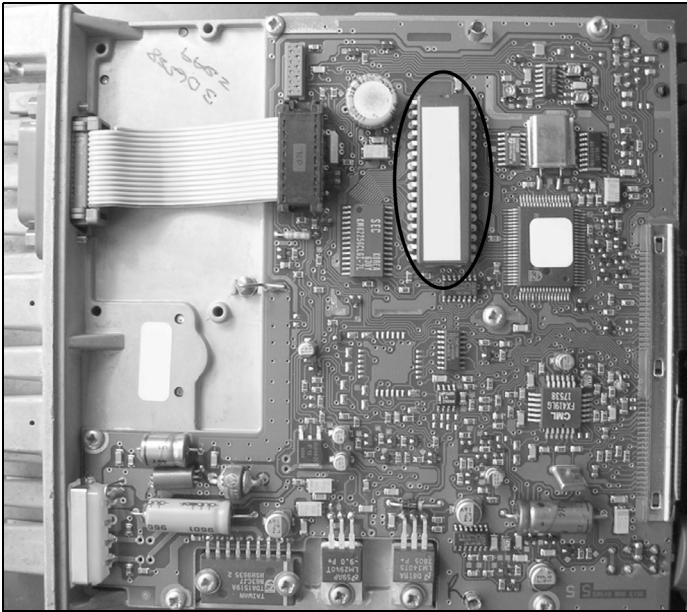


Fig 4: The main board showing the EPROM that must be replaced.

to generate the 1750Hz tone burst and connect to a computer. The pin connections for this plug are shown in Table 1.

4.0

Modifications

4.1 Replacing the EPROM

The first step is to dismantle the front panel. It is fixed by a nut located behind the plastic button of the potentiometer and a screw on top of the microphone plug. The small flat ribbon cable is very fragile so the front should be taken off with extreme caution. A small screwdriver can be used to unlock the plastic tab that holds the flat ribbon cable in place.

The second step is to remove the plastic protection box that is held by screws on the side of the unit. Once this is removed the logic board is visible from above. The board is protected by shielding on the under side. The EPROM to be changed is

shown in Fig 4. It can be extracted with the help of a small screwdriver using a delicate lever action taking care not to bend the pins. Be careful to observe the correct orientation of the EPROM when it is replaced. Now everything must be put back together then the power, loudspeaker and microphone connected.

4.2 Power tuning

The transmitter allows you to choose between two power levels: low or high. In high power mode the PRM8060 can deliver up to 30W. It is safer not to exceed the 25W recommended by the manufacturer.

There are two holes on the side of the radio to adjust the power using a special small isolated screwdriver. Fig 5 shows the position of the two potentiometers:

- R578 : Low power
- R581 : High power

4.3 Setting the microphone gain

If needed it is possible to increase or reduce the microphone gain. This setting

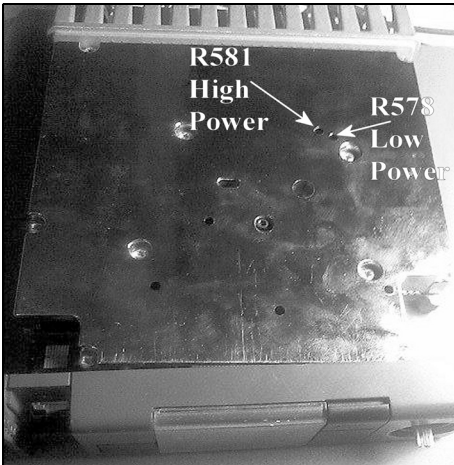


Fig 5: View showing the holes to adjust the RF output power levels.

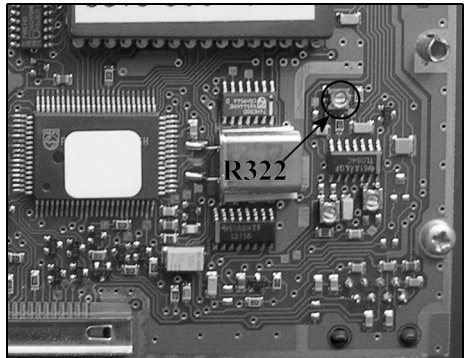


Fig 6: View showing the microphone gain control.

can be set by ear with feedback from someone listening to your transmission. It is better to do it by measuring the deviation with the proper measuring tools. The level is adjusted using R322, see Fig 6.

4.4 Adding the 1750Hz tone burst

To add tone burst it is necessary to connect a small interface on the 15-way male connector on the back of the unit; Fig 7 shows the circuit of the interface.

It is possible to modify the microphone (optional). The original microphone has a reed switch to detect hang up. This contact is removed and replaced with a push button used for the tone burst.

4.5 Serial connection

The link with a computer can be made in two different places. First by the microphone plug on the front panel of the transceiver, Fig 2 shows the connections required. Secondly by the DB15 plug on the back of the unit. In both cases it uses TTL levels (0, 5V). It is necessarily to change this to the RS232 level of the PC (-12V, +12V) to prevent damage to the

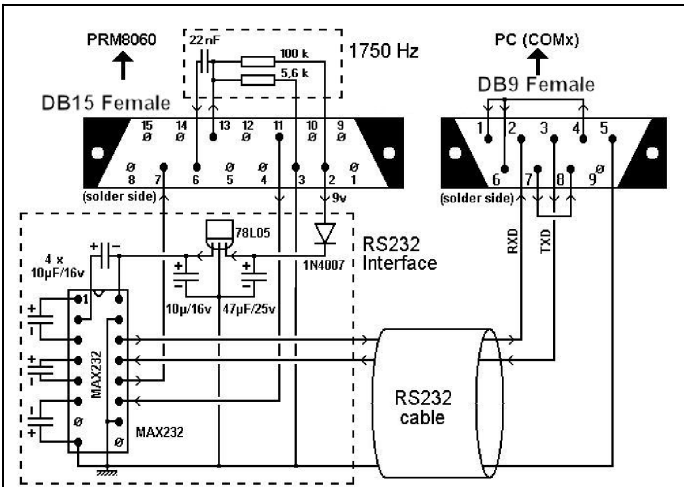


Fig 7: Circuit diagram showing the 1750Hz tone bursts modification and the computer interface.



Fig 8: The display showing adjustment of the squelch level.

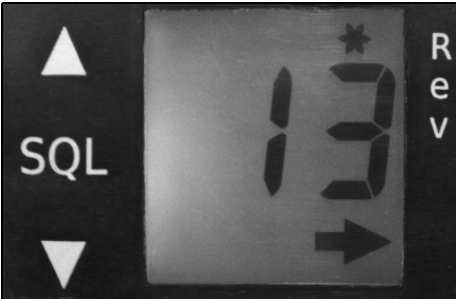


Fig 9: The display showing repeater shift.

radio. A circuit using a MAX232 can be used as shown in Fig 7.

5.0

Use

5.1 Switch on and setting the volume

The transceiver is switched on by pressing the volume potentiometer. The volume level is set using the potentiometer.

5.2 Changing the channel

The keys «+» and «-» (see Fig 1) are used to change the channel. A long press allows a different channel to be selected.

5.3 Adjusting the squelch

The key between «+» and «-» is used to set the squelch level. Pressing this key

makes a cursor appear that indicates the value of the squelch threshold on the screen (see Fig 8). The level can be adjusted using the «+» and «-» keys. Pressing the key again returns to the channels. The level of the squelch can be set between 0 (open squelch) and 15.

5.4 Working with repeaters

When an arrow is shown on the right hand corner of the screen, the selected channel is configured to function on a repeater, see Fig 9. A negative 600kHz shift is used during transmission.

The button on the right of the LCD screen activates the reverse function, it inverts the frequency of transmit and receive. A star is shown when the transceiver is working in reverse mode.

If the modification has been done, a 1750Hz tone can be transmitted by simultaneously pushing the PTT button and the button for 1750Hz on the microphone. It is also possible to activate the tone by pressing the reverse button on the front panel while the radio is transmitting (PTT pushed).

5.5 Changing the power

Continuously pushing the right key (reverse) changes the power output. When the power is on «high», the symbol))) is shown on the screen, see Fig 10.



Fig 10: The display showing adjusting the RF output power level.



5.6 Scanning

Pressing the centre key on the left for two seconds can start channel scanning. When the squelch is opened scanning is halted until the squelch closes. To stop the scanning the scan button should be pushed again. Pressing the PTT also stops channel scanning. Any channel can be locked out of scanning by programming, see computer control below.

5.7 Reset

Resetting the transceiver back to factory parameters can be achieved if necessary. This function is automatically carried out when the new EPROM is fitted. To manually force a reset, simultaneously press the «+» and «-» keys while the transceiver power is connected. There is a signal from the loudspeaker during the reset process.

6.0

Computer control

Thanks to the serial interface it is possible to take control of the radio by a computer. It is also possible to reprogram the channels in the memory. The setting for the serial link is: 4800bps, 7 bits, 1 stop bit, even parity.

A serial terminal (Windows hyper terminal) can be used to control the transceiver. The list of the commands available can be displayed on the screen by pressing «H» and pressing «V» selects the firmware version.

Programming a channel is selected with the «P» option. The channel number between 00 and 99 is requested. The next step asks for the frequency of the PLL, it is a hexadecimal word calculated by the following formula:

$$\frac{Rx_frequency_kHz}{12.5}$$

For example:

- F = 145.000Mhz
- N = 145000 / 12.5 = 11600 decimal = 2D50 hexadecimal.

A scientific calculator can be used to do the conversion to hexadecimal.

The following bits define the state byte for the channel:

Bit n°	Value	Description
0	1	Shift enable
1	2	Reverse mode enable
2	4	Shift +
3	8	Lock out

This byte should be converted to hexadecimal.

7.0

Displaying the receive level

The received signal level can be displayed by pressing the «Squelch» and «reverse» buttons at the same time. The value is in hexadecimal format. To return to normal mode press the two buttons again.

8.0

Information

8.1 Licence

We believe an amateur radio modification should be in the public domain therefore this project is protected by a GPL licence. It should be noted that according to the terms of this licence it is possible to freely broadcast this program, but it is necessary to preserve and broadcast the description of the licence and the author's rights whatever the broadcast



Fig 11: The logo of the third version of the GPL free software licence.

medium. The logo for the third version of the GPL free software licence is shown in Fig 11.

8.2 Authors

This firmware had two main developers:

- Florian Mazen (F4FEZ)
- Pierre Col (F8EGQ)

8.3 Acknowledgments

The following radio amateurs are thanked for their help to realise this project operators:

- Roger Monteix (F4EPQ) who supplied the PRM8060 and different accessories.
- Jean-Claude Beneche (F1AIA) for the information on the settings, the measurements and tests.
- Jean-Michel Manaranche (F5BVJ) for tests and feedback.
- Pierre Mainguet (F8FHC) for the English translation.
- Scott Biggs (KD0FRN) for translation help and proof reading.
- Manuel Jesus (CT1EWT) for the UHF version

8.4 Contact

Via the Website for the project:
<http://prm80.sourceforge.net/>

Appendix

A1 Detailed list of the VHF channels

Channel	RX Freq	Shift
00	R0	145.6000 - 600 kHz
01	R0X	145.6125 - 600 kHz
02	R1	145.6250 - 600 kHz
03	R1X	145.6375 - 600 kHz
04	R2	145.6500 - 600 kHz
05	R2X	145.6625 - 600 kHz
06	R3	145.6750 - 600 kHz
07	R3X	145.6875 - 600 kHz
08	R4	145.7000 - 600 kHz
09	R4X	145.7125 - 600 kHz
10	R5	145.7250 - 600 kHz
11	R5X	145.7375 - 600 kHz
12	R6	145.7500 - 600 kHz
13	R6X	145.7625 - 600 kHz
14	R7	145.7750 - 600 kHz
15	R7X	145.7875 - 600 kHz
16	S8	145.2000
17	S8X	145.2125
18	S9	145.2250
19	S9X	145.2375
20	S10	145.2500
21	S10X	145.2625
22	S11	145.2750
23	S11X	145.2875
24	S12	145.3000
25	S12X	145.3125
26	S13	145.3250
27	S13X	145.3375
28	S14	145.3500
29	S14X	145.3625
30	S15	145.3750
31	S15X	145.3875
32	S16	145.4000
33	S16X	145.4125
34	S17	145.4250
35	S17X	145.4375
36	S18	145.4500
37	S18X	145.4625
38	S19	145.4750
39	S19X	145.4875
40	S20	145.5000
41	S20X	145.5125
42	S21	145.5250
43	S21X	145.5375



44	S22	145.5500	2	430.0750	+ 1.6MHz
45	S22X	145.5625	3	430.1000	+ 1.6MHz
46	S23	145.5750	4	430.1250	+ 1.6MHz
47	S23X	145.5875	5	430.1500	+ 1.6MHz
48	SSTV	144.5000	6	430.1750	+ 1.6MHz
49	APRS	144.8000	7	430.2000	+ 1.6MHz
50	Up R0	145,0000	8	430.2250	+ 1.6MHz
51	Up R0X	145,0125	9	430.2500	+ 1.6MHz
52	Up R1	145,0250	10	430.2750	+ 1.6MHz
53	Up R1X	145,0375	11	430.3000	+ 1.6MHz
54	Up R2	145,0500	12	430.3250	+ 1.6MHz
55	Up R2X	145,0625	13	430.3500	+ 1.6MHz
56	Up R3	145,0750	14	430.3750	+ 1.6MHz
57	Up R3X	145,0875	15	434.6000	- 1.6MHz
58	Up R4	145,1000	16	434.6250	- 1.6MHz
59	Up R4X	145,1125	17	434.6500	- 1.6MHz
60	Up R5	145,1250	18	434.6750	- 1.6MHz
61	Up R5X	145,1375	19	434.7000	- 1.6MHz
62	Up R6	145,1500	20	434.7250	- 1.6MHz
63	Up R6X	145,1625	21	434.7500	- 1.6MHz
64	Up R7	145,1750	22	434.7750	- 1.6MHz
65	Up R7X	145,1875	23	434.8000	- 1.6MHz
			24	434.8250	- 1.6MHz
			25	434.8500	- 1.6MHz
			26	434.8750	- 1.6MHz
			27	434.9000	- 1.6MHz
			28	434.9250	- 1.6MHz-
			29	434.9500	- 1.6MHz
			30	434.9750	- 1.6MHz
			31	433.5000	

A2 Detailed list of the UHF channels

Channel	RX Freq	Shift
0	430.0250	+ 1.6MHz
1	430.0500	+ 1.6MHz

Channel	Repeaters	F(RX)	Channel	F (MHz)	Channel	F (MHz)	Channel	F (MHz)
00	145,6000	R0	16	145,2000	33	145,4125	50	145,0000
01	145,6125	R0x	17	145,2125	34	145,4250	51	145,0125
02	145,6250	R1	18	145,2250	35	145,4375	52	145,0250
03	145,6375	R1x	19	145,2375	36	145,4500	53	145,0375
04	145,6500	R2	20	145,2500	37	145,4625	54	145,0500
05	145,6625	R2x	21	145,2625	38	145,4750	55	145,0625
06	145,6750	R3	22	145,2750	39	145,4875	56	145,0750
07	145,6875	R3x	23	145,2875	40	145,5000	57	145,0875
08	145,7000	R4	24	145,3000	41	145,5125	58	145,1000
09	145,7125	R4x	25	145,3125	42	145,5250	59	145,1125
10	145,7250	R5	26	145,3250	43	145,5375	60	145,1250
11	145,7375	R5x	27	145,3375	44	145,5500	61	145,1375
12	145,7500	R6	28	145,3500	45	145,5625	62	145,1500
13	145,7625	R6x	29	145,3625	46	145,5750	63	145,1625
14	145,7750	R7	30	145,3750	47	145,5875	64	145,1750
15	145,7875	R7x	31	145,3875	48	144,5000	65	145,1875
Shift 1X :- 600 kHz			32	145,4000	49	144,8000	PRM8060-SA9	

Fig 13: Cut this out (or download a copy from the project web site) and stick it on your PRM8060 as a memo for the channel frequencies.

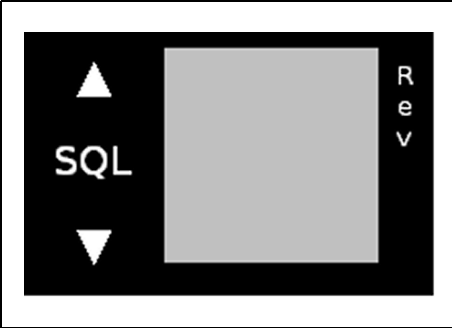


Fig 12: Cut this out (or download a copy from the project web site) and use it a sticker for the front panel of your PRM8060.

A3 Sticker for the front panel

Fig 12 shows a sticker for the front panel to replace the original PRM8060 legend. A small flat screwdriver can be used to dismantle the transparent plastic cover.

A4 Table of the channels to be glued

Fig 13 shows a summary of the channels and frequencies to be cut and glued into the transceiver box

1 nHenry → 1 Henry

USB inductance meter

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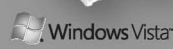
Each device is individually calibrated, and thanks to high quality construction and innovative software it delivers a measurement precision in the order of 0.1%.

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- HENRYTEST inductance meter
- USB 2.0 cable
- CD-ROM with software and drivers

€ 49,90





Gunthard Kraus, DG8GB

An Interesting Program: Simulation of RF circuits with LTspice IV. Part 1

1.0

Introduction

The title “An Interesting Program” is true for the SPICE Simulation software LTspice IV because it has been mentioned and used in many articles published in VHF Communications magazine. Therefore it is time to present its potential in more detail, in particularly for RF simulation. It already has a large worldwide following and is free of charge.

1.1 Characteristics

It is a free SPICE simulator using the complete SPICE Kernel without any restrictions (genuine full version). Originally developed and made available by Linear Technologies as a tool for development engineers of switch mode power supplies that use components from the companies product portfolio. Therefore the program was also known as “LTSwitcherCAD” for a long time.

It is now offered free for all interested circuit developers and SPICE users and is maintained accordingly. It has found a large circle of enthusiasts who have formed an LTspice Internet Community.

1.2 Facilities

In principle nothing is missing: from

analogue simulation in the time domain, to frequency response measurement using AC sweeping, up to noise and S parameter simulation as well as Fourier transformation, everything is available at no cost. It is very stable but of course there are some problems or questions but appropriate assistance can be found from the worldwide Internet forum. The integration of new component models from the Internet is completely problem-free and the “symbol editor” is easy to use to produce new components for the library.

1.3 Peculiarities

It is strange: because the software is free of charge in this case any “knick-knacks” are missing or any incentive to attract customers. In plain language it has a very simple Windows interface but “quirky” operation in the form of a mixture from mouse clicks, key commands and command line input. It is an acquired taste to use but once the tricks have been learned it is quick and easy to use. It is helpful if you have some experience of SPICE simulation.

1.4 Desirable improvement

Big transitions are not allowed so the simulation of digital circuits is difficult: there are only basic modules (called “primitives”) available and in addition a special LT-version that uses the defaults of the SPICE Kernel. It is difficult to



examine components from the TTL or CMOS series because the models from other component providers usually do not work. Again the Internet is useful because individual LTspice lovers have taken the trouble to close this gap and provided suitable models.

1.5 Download, file size and support

The download from the Linear Technology web site [1] is approximately 8Mb. Download with or without registration can be selected. Registration offers the opportunity to become part of the user base and membership of the LTspice forum.

The possibility of the Internet contact with other users has already been mentioned and should be always used. There is a very active LTspice Yahoo Group that is free to join after completion of a questionnaire; it gives fast competent answers to questions. It is absolutely recommended and almost inevitable! In addition the author currently has a 140-slide LTspice tutorial in German or English on his web site [2]. This can be downloaded and is continuously updated. This tutorial should be read before starting your own work. It uses many practical examples to introduce the operation and additionally represents a reference book for many questions for RF and system engineering simulation that emerge during the development of your own ideas.

2.0

Why use SPICE in the RF range?

This question seems justified, if one thinks of the good and economical linear S parameter simulation programs like PUFF or Ansoft Designer SV but the answer is actually very simple:

If a SPICE simulation of an RF circuit

e.g. an LNA, is made and gives the same results for the same operating point as a linear simulation with the S parameter files made available by the manufacturer, then unexpected new possibilities open:

The model very accurately represents the actual component (and its measured S parameters). Thus the operating point can be played with without dependance on a few S parameter files available from the manufacturer that are dependant on a certain operating point. Additionally one can really trust the simulation results in the time domain e.g. when fed with signal forms other than a sine wave or with overloading.

Changes can be made after consideration to see the effects immediately; these will always supply the correct results using the SPICE model to adjust the operational conditions. The large signal behaviour and all overloading effects can be analysed and displayed in the time domain. The associated spectra can be displayed by means of the FFT, this becomes interesting for output stages.

The frequency responses of amplification, the noise figure and the S parameters are available within the circuit thus the effect of each change of circuit is quickly determined to reflect all DC voltages and direct current changes. The optimum can be looked for with a "Parameter Sweep".

S parameter simulation is not redundant by any means because it is always based on semiconductor components with accurately measure characteristics (a crucial touchstone for reference). SPICE works differently with a model of the component that approximates to the total characteristics; there can sometimes be some "bumps". There are no models for certain components e.g. microstrip lines. These require some improvisation to laboriously adapt the general SPICE model for the lossy Tline (lossy transmission line).

A block diagram for an RF system will be used to show the exact procedure and

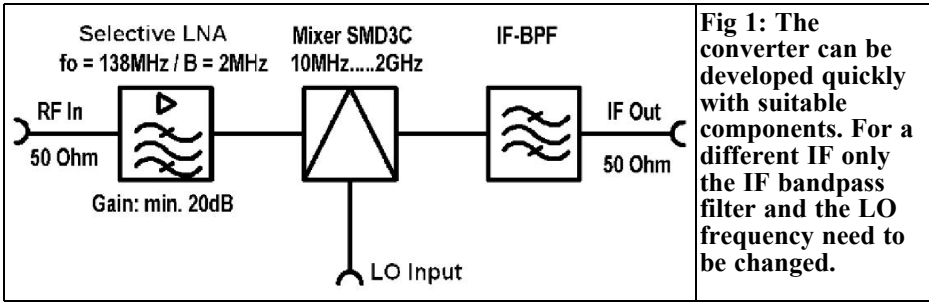


Fig 1: The converter can be developed quickly with suitable components. For a different IF only the IF bandpass filter and the LO frequency need to be changed.

the validity of the above statements. In addition some steps and procedures are necessary. The readers must install LTspice on their PC.

3.0

The project: 137MHz converter

The project selected is a 137MHz converter for NOAA meteorological satellite reception. It was chosen because an article about a low-noise preamplifier using a BF998 dual gate MOSFET for the frequency range from 136 to 138MHz, written by the author, was published in UKW Berichte [3] in 1998

and 1999. This has been used again and again by the author.

The article showed the results of PUFF simulations and measurement on several differently manufactured prototypes. Now the LTspice simulation of the amplifiers will be compared to see if it can keep up with the other simulations.

The remainder of the converter comprised of a SMD3C commercial double balanced mixer on its own PCB in a milled aluminium box with an IF band pass filter on the output. The local oscillator used was an old professional signal generator set to 37MHz. The 99 to 101MHz IF signal was fed to a suitable FM receiver (it was a high-quality car radio) Fig 1 shows the block diagram of this arrangement.

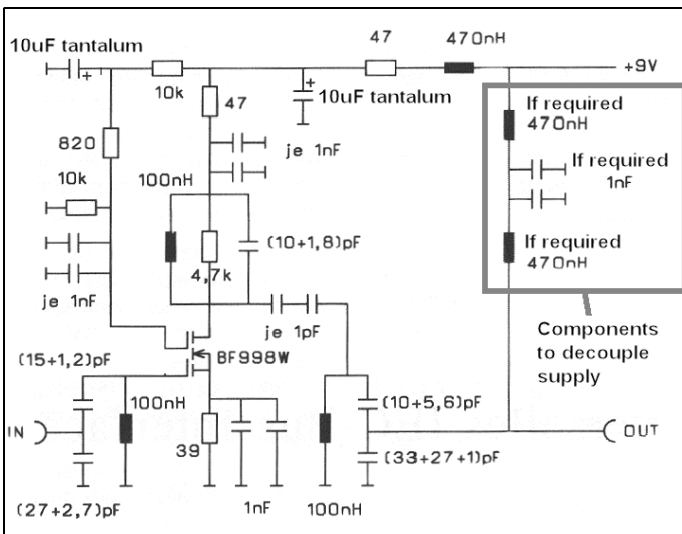


Fig 2: The circuit of the LNA. The remote supply was not included. But anyway: a dual gate MOSFET always needs a high supply voltage.



```

* BF998 SPICE MODEL OCTOBER 1993 PHILIPS SEMICONDUCTORS
* ENVELOPE SOT143 (R)
* 1.: SOURCE; 2.: DRAIN; 3.: GATE 2; 4.: GATE 1;
.SUBCKT BF998 1 2 3 4
L10      1 10      L=0.12N
L20      2 20      L=0.12N
L30      3 30      L=0.12N
L40      4 40      L=0.12N
L11      10 11     L=1.20N
L21      20 21     L=1.20N
L31      30 31     L=1.20N
L41      40 41     L=1.20N
C13      10 30     C=0.085P
C14      10 40     C=0.085P
C23      20 30     C=0.085P
C24      20 40     C=0.005P
D11      42 11     ZENER
D12      42 41     ZENER
D21      32 11     ZENER
D22      32 31     ZENER
RS       10 12     R=1.00
MOS1     61 41 11 12 GATE1 L=1.1E-6 W=1150E-6
MOS2     21 31 61 12 GATE2 L=2.0E-6 W=1150E-6

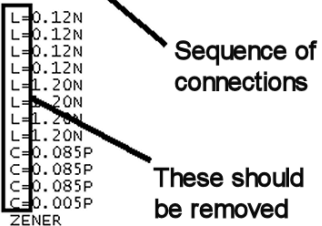
.MODEL ZENER D BV=10 CJO=1.2E-12 RS=10

.MODEL GATE1
+ NMOS LEVEL=3 UO=600 VTO=-0.250 NFS=300E9 TOX=42E-9
+ NSUB=3E15 VMAX=140E3 RS=2.0 RD=2.0 XJ=200E-9 THETA=0.11
+ ETA=0.06 KAPPA=2 LD=0.1E-6
+ CGSO=0.3E-9 CGDO=0.3E-9 CBD=0.5E-12 CBS=0.5E-12

.MODEL GATE2
+ NMOS LEVEL=3 UO=600 VTO=-0.250 NFS=300E9 TOX=42E-9
+ NSUB=3E15 VMAX=100E3 RS=2.0 RD=2.0 XJ=200E-9 THETA=0.11
+ ETA=0.06 KAPPA=2 LD=0.1E-6
+ CGSO=0.3E-9 CGDO=0.3E-9 CBD=0.5E-12 CBS=0.5E-12

.ENDS BF998
    
```

Fig 3: The SPICE model of the BF998 was easy to find on the Internet, but it has a problem.



4.0

Low-noise preamplifier with a BF998 dual gate MOSFET

Fig 2 shows the complete circuit diagram of this preamplifier from the original article [3]. A capacitive divider performs input matching and the output uses a capacitively coupled double tuned band pass filter. The coupling capacitor in the output filter uses a 0.5pF interdigital capacitor. The output is matched to 50Ω with a capacitive divider. The gain of the stage is approximately 22dB with a noise figure somewhat under 1dB.

4.1. The BF998 SPICE model

This model can be found on the Internet, enter “bf998 spice model” into a search engine and the result will be as shown in Fig 3. It looks good and very well documented but there is a problem:

- The model author probably meant to

show the inductances and capacitances by inserting the legends “L=” and “C=”. However SPICE rejects each simulation abruptly. Therefore: These legends should be deleted from the file and the result saved as “BF998.lib” in the LTspice library “lib/sub”.

Now the symbol for a dual gate MOSFET must be set up. Go to the menu and click on “new symbol”. The tools necessary for drawing e.g. line, circle, rectangle etc. are under “Draw” with a grid on the screen to help placement. Before starting to draw, load the symbol for a MOSFET from the library, the path is “LTspiceIV\lib\sym” and the symbol is “nmos4.asy” as shown in Fig 4. This shows how it is constructed and the size so that the new symbol can be made the same size so that it fits into the family of symbols, not too large or too small. Now there is a choice: either delete this symbol from the screen (Press F5 then click the left mouse button on the symbol; back with “Escape”) and draw the dual

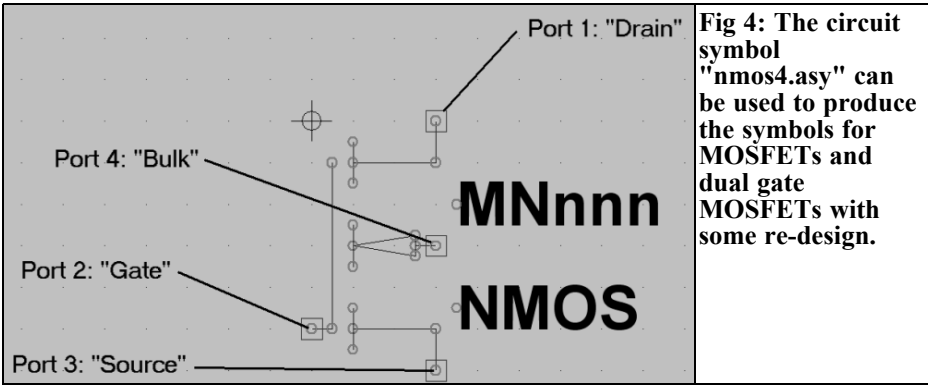


Fig 4: The circuit symbol "nmos4.asy" can be used to produce the symbols for MOSFETs and dual gate MOSFETs with some re-design.

gate MOSFET from scratch, or you modify the symbol "nmos4.asy" as shown in Fig 5 and save it as "BF998.asy". In both cases three references are necessary:

- The small red circles (they will be red on your screen but not in the black and white Fig 5) mark the ends of lines and are for indication and points to connect into the component.
- The actual component connections (ports) are small rectangles with a red circle on the inside. These cannot be drawn they have their own function under "Edit" called "Add Pin/Port". Clicking this displays the "Property menu" as shown in Fig 5. The data to

be entered is the port identification, the pin number for the SPICE model and the position of the port identification beside the symbol (Left/offset = 5). For the BF998: Source = port 1, Drain = port 2, Gate2 = port 3, Gate 1 = port 4. After setting up each port it is connected to the symbol with a simple line.

- To shift something on the screen, press F7. Click on the appropriate part, it hangs on the cursor and can be moved to its new position. A further mouse click fixes it and ESC exits the function.

Now the finished symbol is stored in the

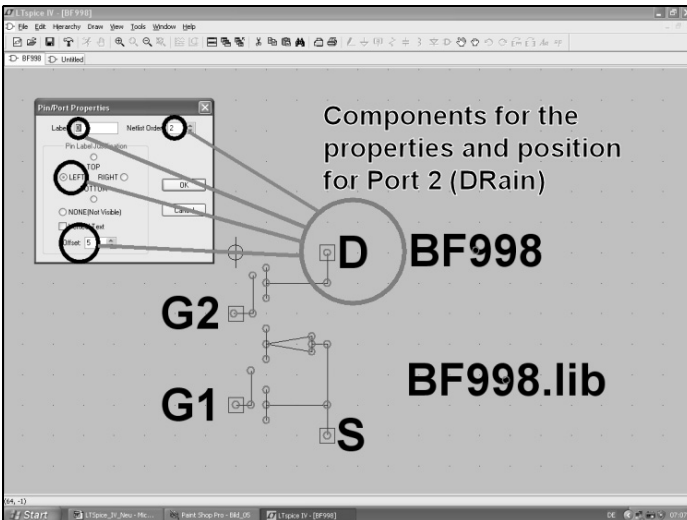


Fig 5: A MOSFET can be simply modified to make a dual gate MOSFET. See the text to position the new port.

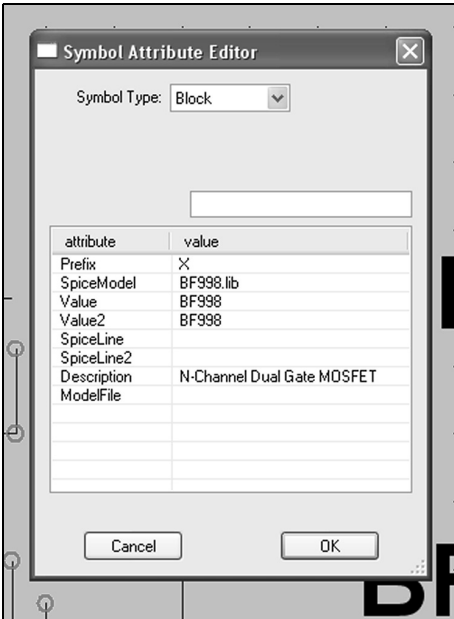


Fig 6: Pay attention to the text when filling out of this Property list for the BF998 symbol.

library path “SpiceIV/lib/sym” as “dgate_MOSFET.asy” for later use. Then make a genuine BF998 from it by opening the menu under “Edit” and “Attribute/Edit of attributes” and change

the attributes as shown in Fig 6. This is sophisticated; you must click each entry and make the input in the window above the complete list. But the input is only accepted when the next line is clicked. To complete the process, open the “Edit” menu again and select “Attribute/Attributes Window” (this shows additional information on the screen). From the list that emerges, click on “Value” and confirm with OK. Suddenly the designation “BF998” appears beside the symbol. Repeat the procedure and ensures that the SPICE model designation “BF998.lib” is visible.

A new file in “lib/sub” named “Dual gate MOSFETs” can be created to hold the “BF998.asy” – now you are finished.

4.2. SPICE simulation of the S parameters with ID=10mA

4.2.1. Determining the operating point

On the Internet only S parameters for drain currents of 5mA and 10mA can be found. So a SPICE simulation with ID = 10mA will be examined to see if the results are the same. The first circuit shown in Fig 7 ensures that the operating point adjusts itself to the same values as the S parameter file (ID = 10 mA, VDS = 5V, VG2S = 3.5V).

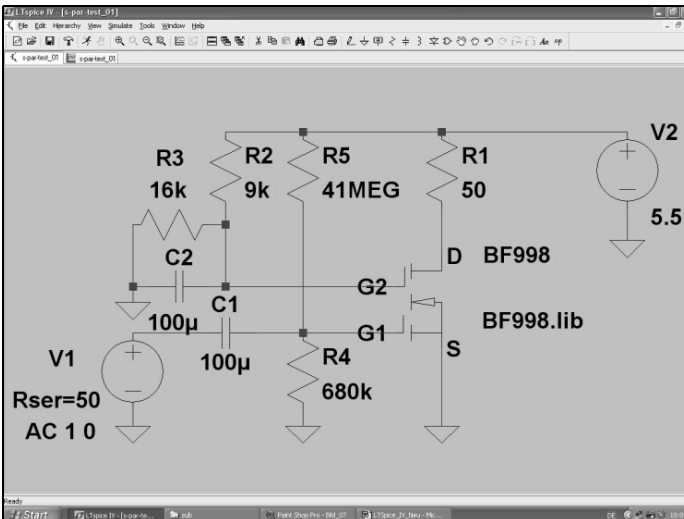


Fig 7: The first step for the SPICE simulation is to draw the circuit to control the operating point. The S parameters for the simulation of the MOSFET are for an operating point with ID = 10mA.



```

--- Operating Point ---

V(n001):      5.5      voltage
V(n003):      5.00077  voltage
V(n002):      3.52     voltage
V(n005):      0.0897312 voltage
V(n004):      4.48656e-016 voltage
I(C2):        3.52e-016 device_current
I(C1):        8.97312e-018 device_current
I(R5):        1.31958e-007 device_current
I(R4):        1.31958e-007 device_current
I(R3):        0.00022   device_current
I(R2):        0.00022   device_current
I(R1):        0.00998461 device_current
I(V2):        -0.0102047 device_current
I(V1):        8.97312e-018 device_current
Ix (u1:1):    -0.00998461 subckt_current
Ix (u1:2):    0.00998461 subckt_current
Ix (u1:3):    3.99684e-012 subckt_current
Ix (u1:4):    6.93895e-014 subckt_current
    
```

Fig 8: This table contains all the DC voltages and direct currents in the circuit. For those who would like to have more details, close the table and measure the required point on the circuit.

Operating point
 UDS
 UG2S
 Drain current
 Source current

The resistors, capacitors and the ground character can be found in the toolbar. For the voltage sources press the component button at the right hand side of the diode symbol button and search for "voltage" in the list. Two sources are required in the circuit connected to the ground. The BF998 can now be found in the library. Everything is connected together with the "Wire" symbol to produce the final circuit.

Programming the two voltage supplies is quite simple: right mouse click on the circuit symbol and then on "Advanced" to open the "Property menu". For the source V1 (input signal) set an amplitude of "1" and a phase of "0" as well as an internal resistance with 50Ω necessary for the S parameter simulation under "Small signal AC analysis". The source V2 is the DC supply voltage; therefore it is only necessary to enter 5.5V under "DC Value". The values of the resistors and capacitors are just as easy, right mouse click on their symbol to enter the value.

Looking at the circuit, the drain resistor R1 gives the correct output impedance of 50Ω for the S parameter definition while the remaining resistors are voltage dividers to adjust the operating point. Capacitor

C1 isolates Gate 1 from the 50Ω internal resistance of the source V1 and capacitor C2 grounds Gate 2 over the frequency range.

Only a simulation command for the DC analysis is missing. Under "Simulate" there is an option "Edit simulation cmd" that gives a selection menu, select "DC op. Pnt.". Pressing OK makes the ".op" command hangs on the cursor so it can be placed on the circuit – and you are finished.

Now click on the running man next to the hammer on the toolbar and a table like Fig 8 is displayed with all the voltages and currents listed. It is too laborious to decipher this so close it and move the cursor onto an interesting point in the circuit: the appropriate voltage and current are displayed at the bottom left hand corner of the screen. Because all values are in the green range and agree with the values in the S parameter file it is OK to start the SPICE S Parameter simulation.

4.2.2. Determination the S parameters at the selected operating point

Under "Simulate" use the option "Edit simulation cmd" select "AC Analysis". Select a decade Sweep starting at a

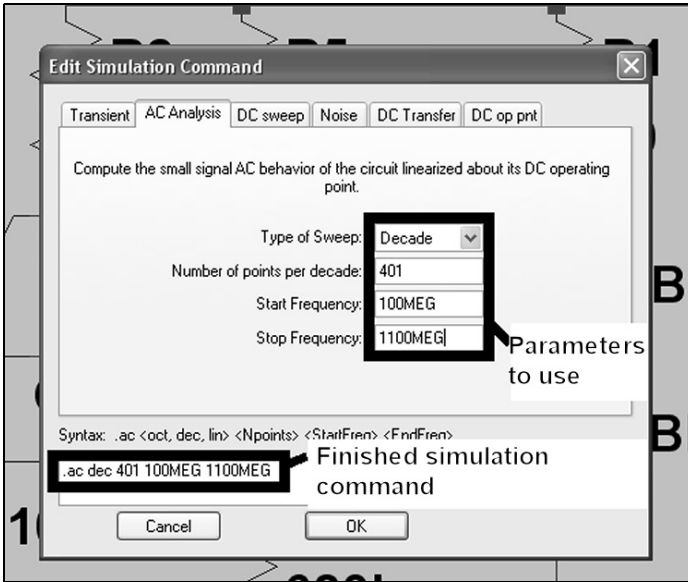


Fig 9: To get more information use the simulation command for an AC sweep from 100 to 1100MHz.

frequency of 100MHz (the S parameter file starts at 50MHz) and stopping at a frequency of 1100MHz with 401 points per decade, see Fig 9.

Now only the "SPICE Directive" for determining the S parameters is missing. Under "Edit" and "SPICE Directive" use this option to register in the free field:

.net I(R1) V1

That is easy to understand:

- ".net" calls the network simulation
- "I(R1)" gives in the voltage across the output resistor R1

- "V1" means that the voltage source produces the incident wave

The previous directive ".op" is now automatically switched off by replacing the point in the directive (".op") by a semicolon (".op;"). An AC Sweep can now be selected, starting from 100MHz up to 1100MHz with a tick of 100MHz and the simulation started.

If the result diagram is empty after the simulation, right click on the diagram and select "Add Trace".

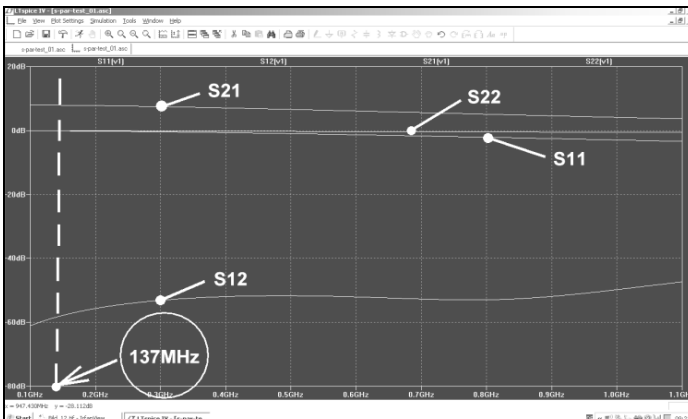


Fig 10: This is interesting for the circuit developer: the simulated S parameters for a drain current of 10mA.



```

! SIEMENS Small signal Semiconductors
BF998
! S1 Dual Gate MOSFET Tetrapole in SOT143
VDS = 5 V      VGS = 3.5 V      ID = 10 mA
! Common Source S-Parameters
February 1991
# GHZ S MA R 50
! F
! S11      S21      S12      S22
! GHZ      MAG      ANG      MAG      ANG      MAG      ANG      MAG      ANG
0.050      0.9992      -4.0      2.230      174.5      0.0009      89.0      0.9923      -2.1
0.060      0.9991      -4.8      2.225      173.2      0.0011      86.1      0.9915      -2.5
0.070      0.9988      -5.6      2.218      172.0      0.0013      85.9      0.9917      -3.0
0.080      0.9974      -6.4      2.221      170.8      0.0015      84.8      0.9917      -3.4
0.090      0.9969      -7.2      2.228      169.6      0.0016      83.6      0.9914      -3.7
0.100      0.9968      -8.0      2.232      168.4      0.0018      83.1      0.9908      -4.1
0.150      0.9918      -12.0      2.212      162.4      0.0026      78.6      0.9895      -6.2
0.200      0.9849      -15.9      2.211      156.5      0.0035      74.5      0.9865      -8.4
0.250      0.9769      -19.8      2.190      150.5      0.0043      71.2      0.9848      -10.4
0.300      0.9680      -23.6      2.159      144.6      0.0050      67.6      0.9806      -12.4
0.400      0.9451      -31.2      2.095      133.1      0.0062      60.8      0.9749      -16.4
0.500      0.9199      -38.4      2.021      122.1      0.0072      54.4      0.9681      -20.3
0.600      0.8945      -45.4      1.945      111.6      0.0077      48.8      0.9608      -24.0
0.700      0.8686      -52.2      1.861      101.1      0.0079      45.5      0.9527      -27.7
0.800      0.8382      -59.0      1.769      91.2      0.0088      39.9      0.9403      -31.4
0.900      0.8085      -65.2      1.694      81.7      0.0086      30.6      0.9315      -35.2
1.000      0.7824      -71.5      1.622      72.3      0.0079      26.6      0.9230      -38.9
1.200      0.7256      -84.0      1.467      53.0      0.0069      21.1      0.9071      -46.9
1.400      0.6697      -95.2      1.309      35.4      0.0061      4.9      0.8853      -54.1
1.500      0.6464      -100.2      1.243      26.9      0.0043      -5.0      0.8781      -57.8
1.600      0.6265      -105.1      1.187      18.5      0.0034      -19.2      0.8724      -61.6
1.800      0.5887      -115.0      1.086      1.2      0.0029      -78.7      0.8643      -69.0
2.000      0.5586      -124.4      1.001      -16.9      0.0064      -122.3      0.8494      -77.0
!
! f      Fmin      Gammaopt      rn/50
! GHZ      dB      MAG      ANG      -
0.010      3.00      0.80      1      3.40
0.050      1.30      0.80      3      2.00
0.100      0.80      0.80      6      1.40
0.200      0.70      0.80      12      1.10
0.450      1.00      0.78      27      1.00
0.800      1.75      0.73      48      0.96
0.900      2.00      0.71      53      0.94
1.100      2.35      0.63      65      0.84
! SIEMENS AG Semiconductor Group, Munich
    
```

Operating point

Fig 11: The genuine S parameter from the manufacturer as found on the Internet.

In the list click on the four S parameters; S11, S21, S12, S22 and press OK. Move the mouse on the number division of the horizontal frequency axis until a ruler emerges and left click. Now a linear sweep from 100 to 1100MHz with a tick of 100MHz can be selected. In the same way the amplitude range can be changed to the range from +20dB to -80dB with a tick of 10dB.

The program plots the phase curve as well as the amplitude curve. If you are not interested in the phase curve, use the mouse on the right axis to locate the option "Don't plot Phase".

Success can be seen in Fig 10 but it is most important to compare it with a genuine S parameter simulation. I have one from the Internet on my own computer stored as S2P-Datei for ID = 10mA, it is shown in Fig 11. The simulation shown in Fig 12 can be recognised

immediately as a PUFF simulation. Comparing Fig 10 (LTspice) and Fig 12 (PUFF) your heart beats quickly but the curves agree to a great extent up to the non-critical deviation with S12. Therefore the creator of the BF998 SPICE Model did a good job.

Table 1 shows the exact simulation results obtained for a frequency of 137MHz. These are read off by moving the cursor to the correct points.

There may be a reason for a furrowed brow because of the differences: the SPICE model came from Philips, the S parameter file came from Siemens. There are quite large differences between MOSFETs so all may not be spoken.

Now the complete LNA circuit shown in Fig 2 can be simulated in exactly the same way.

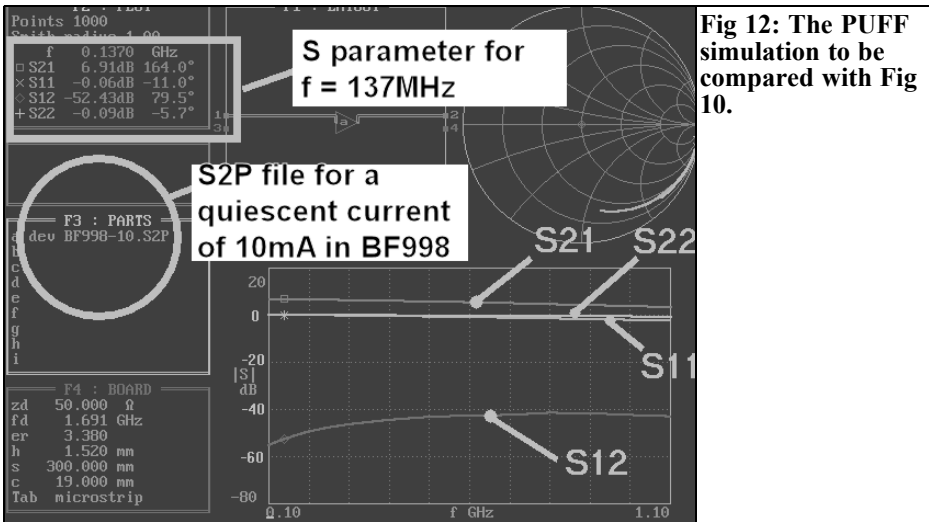


Fig 12: The PUFF simulation to be compared with Fig 10.

4.3. LTspice simulation of the 137MHz LNA

4.3.1. Determining the operating point

Some work is required to convert Fig 2 into a suitable diagram for the SPICE simulation shown in Fig 13. The resonant circuit coil with has a value of $L = 100\text{nH}$ set in the “Property menu” and a series resistance of 1Ω for a quality of $Q = 60$. The simulation command “.op” is set as described in 4.2.1, the result table gives the operating point:

VDS = 8.1V
VG2S = 4V
IS = ID = 5mA

Those are exactly the values shown in the S parameter file for a drain current of 5mA.

Table 1: S Parameters determined by LTSpice and PUFF.

S Parameter	LTSpice	PUFF
S11	-0.085dB	-0.06dB
S21	+7.95dB	+6.91dB
S12	-58dB	-52dB
S22	-0.05dB	-0.09dB

4.3.2. Determining the S parameters for the LNA

Following the exercise in the preceding chapter this is now easy using the circuit from Fig 13:

Use the “Edit simulation command” to set a decade AC Sweep from 120 to 160MHz with 4001 points per decade. Or use the command:

```
.ac dec 4001 120MEG 160MEG
```

Use the SPICE directive for S parameter simulation:

```
.net I(R1) V2
```

R1 represents the output resistance of the stage at 50Ω and V2 is the signal source. Make sure that the attributes are set “AC amplitude AC = 1” and “AC phase = 0” as well as the internal resistance of 50Ω

After the simulation display all S parameters with the phase not displayed, a linear frequency axis from 120MHz to 160MHz with a check mark every 5MHz and the amplitude range of +30dB to -80 dB with a tick of 10dB.

The result should be as shown in Fig 14 and the direct comparison from the previous publication [4] is shown in Fig 15.

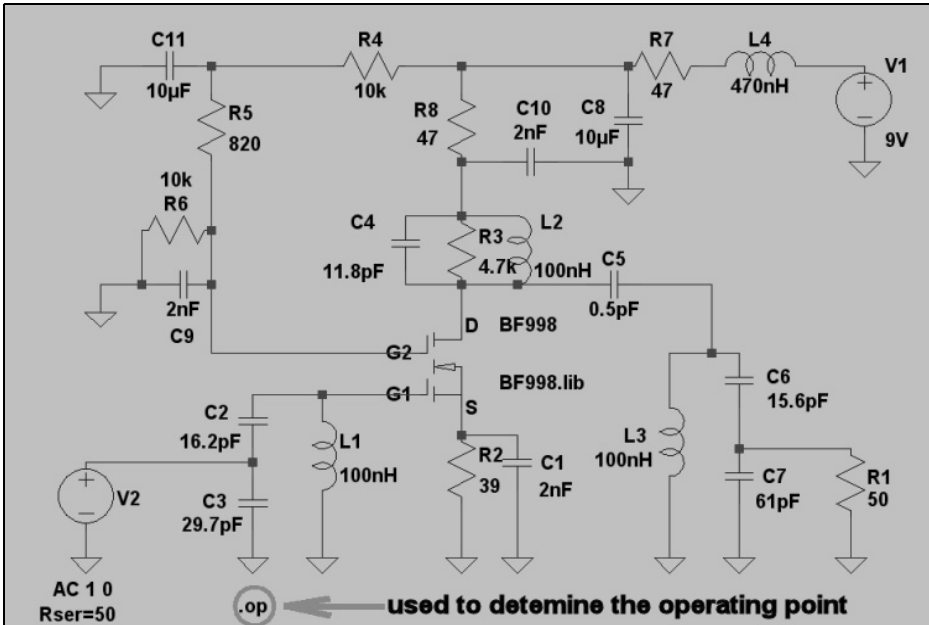


Fig 13: The SPICE diagram prepared from Fig 2.

The simulation at that time used the ARRL radio designer showing the S21 Curve. This showed a gain of approximately 22.5dB at the centre frequency (this was also measured). That is about 2.5dB less than the SPICE simulation shows. In addition the measured centre frequency of the transmission curve is 138MHz while SPICE shows 139MHz.

Using the circuit shown in Fig 16 with Ansoft designer SV to simulate the circuit with the S parameter file for ID = 5mA gives the results shown in Fig 17. The comparison of the SPICE and Ansoft results is really informative:

Ansoft gives less gain of approximately 23dB at 138MHz - thus 1MHz less than SPICE. Otherwise both simulations are nearly identical for all S parameters.

The difference of centre frequency is harmless:

- Related to 137MHz the deviation with the Ansoft simulation is less than 0.7%. With SPICE there is then

approximately a 1.3% error. But the coils have alignment cores to cater for this.

Now we can lean back really calmed down.

4.3.3. Simulation of noise figure (NF in dB) using LTspice

A study of different data sheets for the BF998 that have an indication of the typical noise figure and S parameter files for ID = 10mA (looking for the noise parameters) gives a feeling for the order of magnitude that can be expected. Depending upon the manufacturer, values of 0.6dB (Siemens/Infineon), 0.7dB (Philips) and up to 1dB (Vishay) can be found for the range between 100 and 200MHz. It is interesting to see what SPICE will predict.

The circuit shown in Fig 13 must be changed to provide suitable commands. Fig 18 shows the necessary changes:

- The internal resistance of 50Ω in the

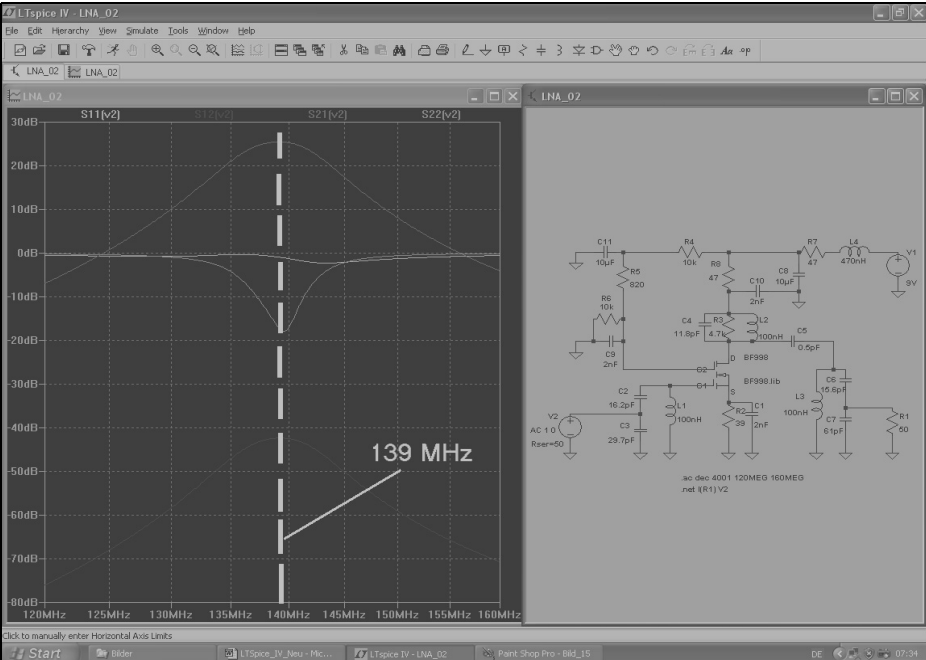


Fig 14: Here the goal of the work: the S Parameter simulation of the complete circuit.

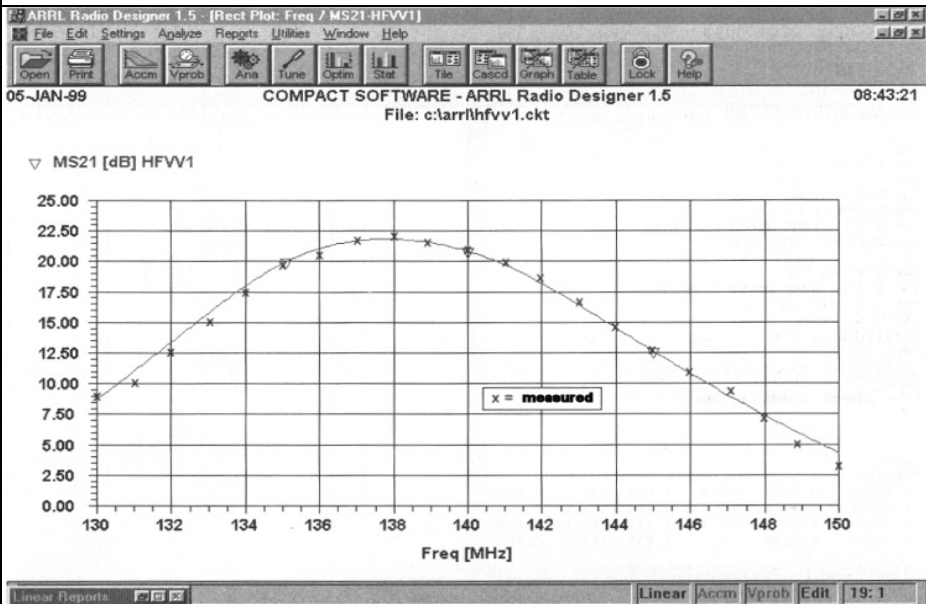


Fig 15: This was from a while ago, the simulation at that time included measurements of the prototype for the comparison with Fig 14.

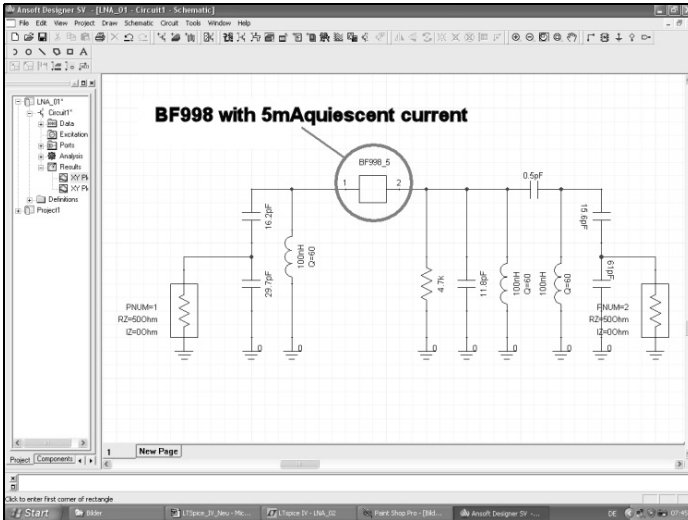


Fig 16: With modern simulation programs it is still better. This is the simulation circuit for the Ansoft Designer SV.

voltage supply V2 must be removed and replaced by an external resistor R9.

- A label “out” must be attached to the output resistor R1. It is beside the ground character on the toolbar.

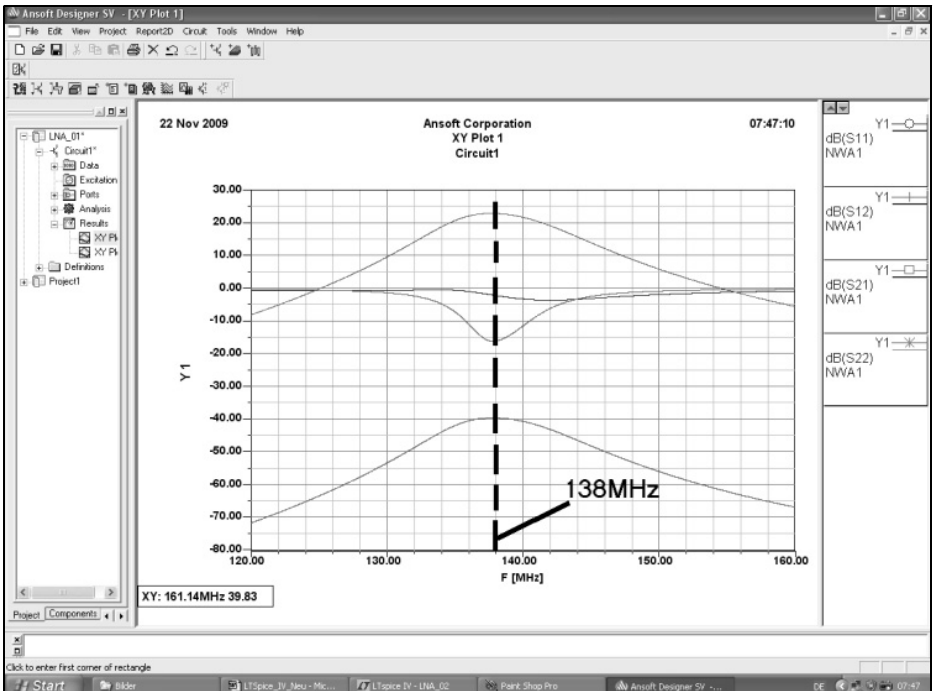


Fig 17: The new simulation for comparison with Fig 14. Using Ansoft Designer SV.

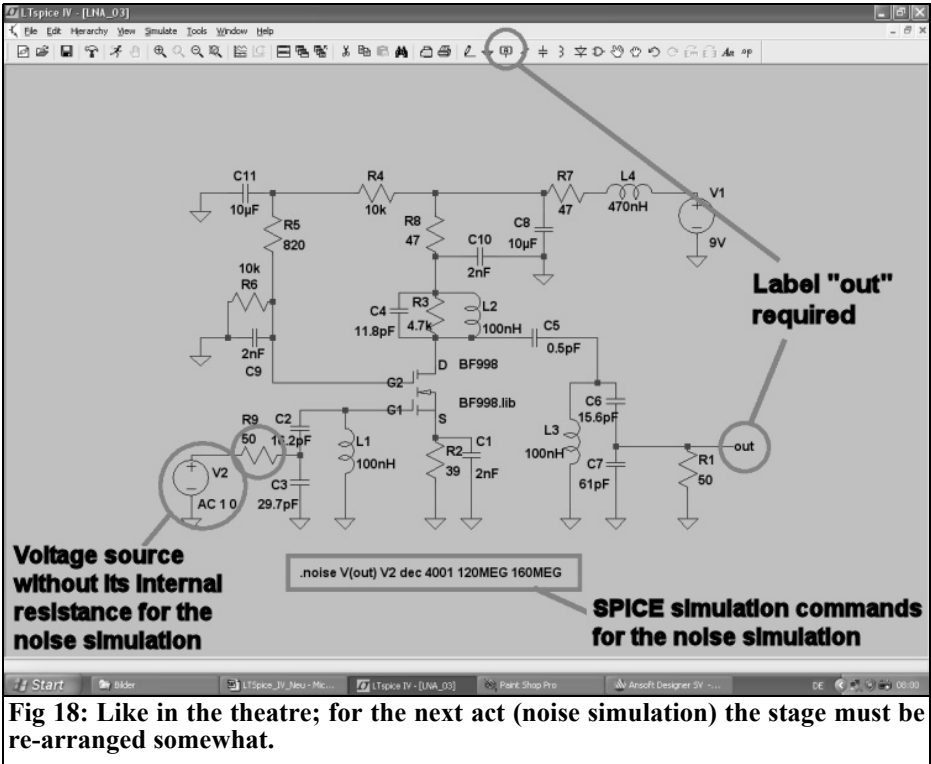


Fig 18: Like in the theatre; for the next act (noise simulation) the stage must be re-arranged somewhat.

- Under “Simulate” and “Edit simulation Command” enter the command necessary for the noise simulation. It

is for a decade Sweep with 4001 points from 120 to 160MHz:

.noise V(out) V2 dec 4001 120MEG 160MEG

Pointing to the input resistor R1 shows the spectral noise performance density for the output. Simply clicking on the wire between input voltage source V2 and the resistor R9 shows the spectral noise performance density referred back to the input.

To simulate the noise figure “NF in dB” some additional work is necessary. Open an empty sheet in a text editor (e.g. Notepad) and type the following line:

.func NF(50) 10*log10(V(inoise)* V(inoise)/(4*k*300*50))

Store this file as “plot.defs” in the LTspice start-up directory as restart LTspice to let the change become effective. This technique gives a new function

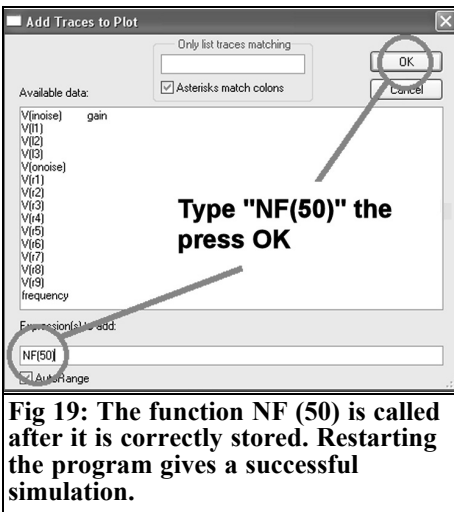


Fig 19: The function NF (50) is called after it is correctly stored. Restarting the program gives a successful simulation.

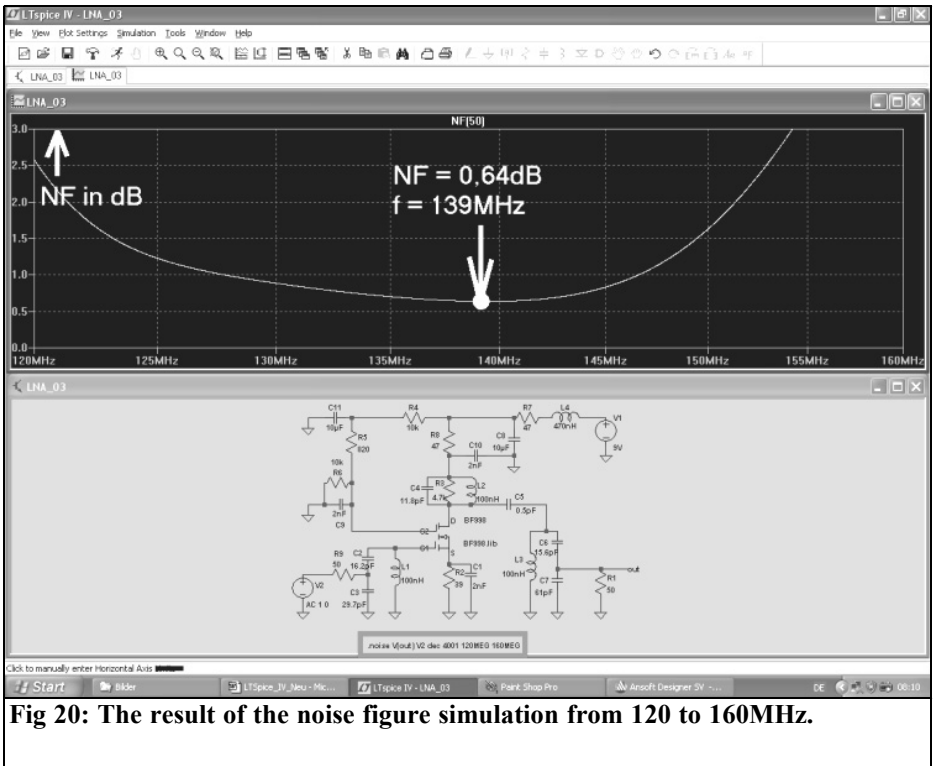


Fig 20: The result of the noise figure simulation from 120 to 160MHz.

(“func”) with the name “NF(50)” that can be used to compute the noise figure, NF, for a 50Ω system with a temperature of 300° Kelvin and plotted in the units of dB.

So to continue:

Simulate the circuit. When the simulation is complete the circuit is displayed with an empty result diagram on the screen. Right click and open the menu “Add trace”. In the empty lower input field type the function call “NF(50)” as shown in Fig 19, after pressing OK the simulated noise figure NF in dB is displayed. This is shown with reasonable axes in Fig 20, these compare favourable with the values from the datasheets mentioned.

X. Literature to part of 1

- [1] Internet: <http://www.linear.com>
- [2] The author’s home page: www.elektronikschule.de/~krausg
- [3]: Puff project: Low-noise preamplifier for 137MHz or 145MHz, Gunthard Kraus, DG8GB and Andreas Zimmermann, DG3SAZ, UKW Berichte 4/1998 pp 233 - 245, 1/1999 pp35 – 52 and 3/1999 pp 167 - 182. Unfortunately this article was never published in VHF Communications Magazine.
- [4] UKW Berichte 1/1999, page 43



André Jamet F9HX

Harmonic or overtone ?

1.0

Introduction

A crystal oscillator can work in fundamental mode or overtone mode. It is very important to know the behaviour in both modes to order a crystal and avoid disappointments.

2.0

Crystal resonator

A crystal made of quartz (Fig 1) shows a piezoelectric effect.

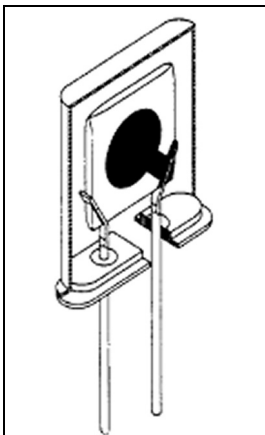


Fig 1: Internal view of a crystal.

Mechanical stresses applied on it produce electric charges. Conversely, the crystal is deformed by the application of an electric charge. That is used to make resonators [1,2,3,4,5].

Fig 2 shows the equivalent electrical circuit of a crystal resonator. C_0 represents its parasitic capacity including the case.

Fig 3 shows the reactance variation versus frequency. We can see very high variations including nulls at marked frequencies f , $\approx 3f$ and $\approx 5f$. They corresponds to the Fig 2 tuned circuits. They are called overtone. Higher order overtones can be generated 7 or 9. However,

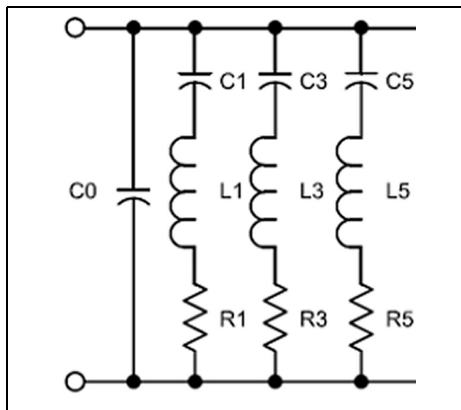


Fig 2: Equivalent circuit of a crystal.

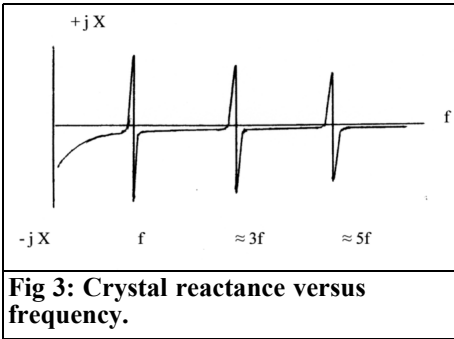


Fig 3: Crystal reactance versus frequency.

Q is reduced and the oscillator design is more difficult.

AT-cut crystals are thickness shear vibrators. Fig 4 shows in dotted lines the nodal lines. In both cases, the edge effects acts twice whatever the node number in the crystal thickness. The exact frequency of an overtone is a little bit higher than the integer multiple of the fundamental because there are several mechanical $\lambda/4$ in series and only the outer are detuned by metallisation and contact with the air [3].

Only odd-numbered overtones can be generated. With an even-numbered overtone, the electrodes should show identical polarity and no field will be developed.

There are some spurious frequencies above each overtone.

3.0

Oscillators

Owing to the crystal resonators high Q, it is well advised to use it as a resonator for a high stability oscillator. Fundamental mode and overtone are commonly employed in well-known devices like Colpitts, Pierce, Hartley, Vackar, Butler, Driscoll and so on.

To obtain a defined frequency overtone oscillation, it requires forcing to prevent a fundamental or other order overtone to take place. An auxiliary circuit tuned at

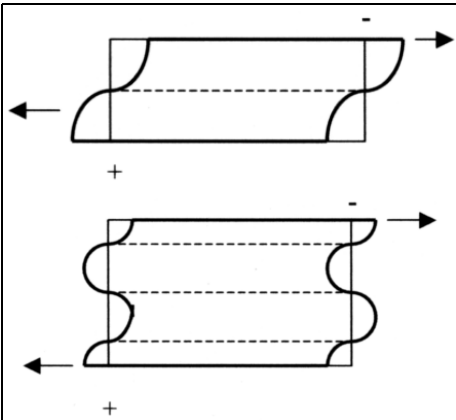


Fig 4: Top: Fundamental frequency, thickness proportional to $\lambda/2$.

Bottom: Third overtone, thickness proportional to $\approx 3\lambda/2$.

the overtone frequency plays that role.

It is very important to notice that overtone frequencies are not exact integers of the fundamental frequency.

4.0

Experimentation

To confirm the above theory, two checks have been done with the same crystal, first in fundamental mode, second in overtone. The crystal used is designed for 106.5MHz in fifth overtone. It is intended for a 10.224MHz LO of a 10GHz/144MHz transverter.

The first circuit (Fig 5) is a Pierce followed by a multiplier stage to get the fifth harmonic. The measured frequency is 21.274MHz for the oscillator and we verify: $21.274 \times 5 = 106.370\text{MHz}$ at the multiplier output. A series capacitor modifies the crystal frequency by approximately 10^{-5} .

The second circuit (Fig 6) is a Butler giving a VHF frequency of 106.49995 to 106.50287MHz by changing the tuning

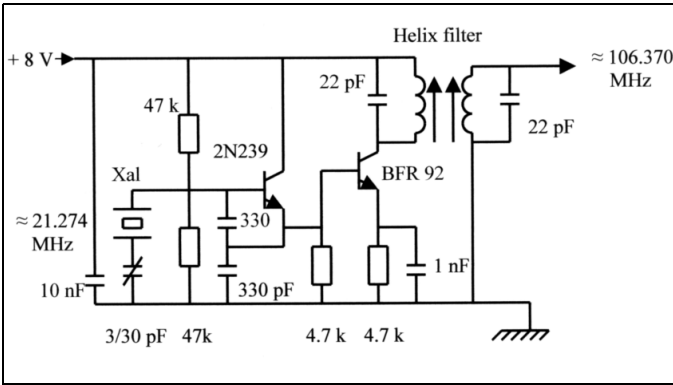


Fig 5: Fundamental oscillator plus a multiplier.

before the oscillation stops. The overtone frequency is higher by about 1.2×10^{-3} than the multiplied frequency.

It was not possible to make the two circuits closer in frequency.

behaviour of an oscillator well to order the right crystal and to avoid problems. The manufacturer will be able to cut and trim the blank to get the right frequency and a high Q.

5.0

Conclusion

The theory is verified: an overtone frequency derived from a crystal oscillator is higher than the fundamental frequency produced by the same crystal.

It is very important to know the actual

6.0

References

- [1] Le résonateur à quartz en HF et VHF, F9HX, Radio-REF 2/2003 (numerous references about crystals oscillators)
- [2] Design of Crystal Oscillator Circuits, Neubig, B, DK1AG, VHF Communications 3,4/1979

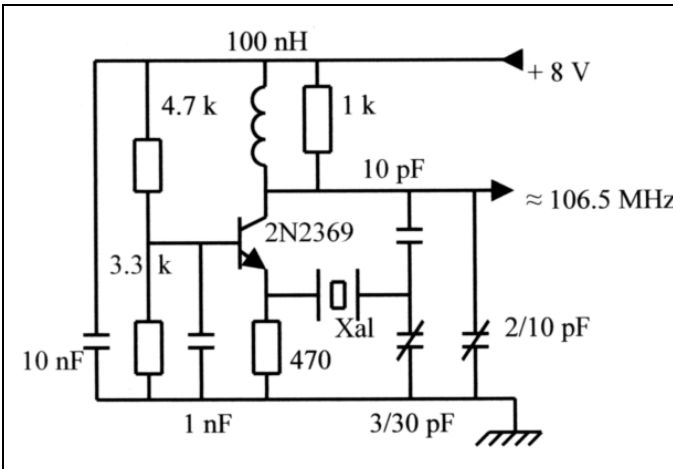


Fig 6: Overtone oscillator.



Table 1: Extract from the International Electromechanical Vocabulary for piezoelectric devices.

German	English	French
02 Grundfrequenz	fundamental frequency	fréquence fondamentale
03 Frequenz der Oberschwingung	overtone frequency	fréquence partielle Obertonfrequenz
04 Obertonordnung	overtone order	ordre d'un partiel
08 Resonanzfrequenz	resonance frequency	fréquence de résonance
09 Serienresonanzfrequenz	series resonance frequency	fréquence de résonance série
11 Parallelresonanzfrequenz	parallel resonance frequency	fréquence de résonance parallèle
13 Nennfrequenz	nominal frequency	fréquence nominale
14 Arbeitsfrequenz	working frequency	fréquence de fonctionnement

[3] TELE QUARZ Group, Catalogue 1999

[4] Crystal Testing, Detlef Burchard, VHF Communications 3/93

[5] Low Noise VHF-Crystal Oscillators, DK4XP, DUBUS TECHNIK VI

7.0

Appendix

The International Electrotechnical Vocabulary gives the official terms to use for piezoelectric devices in Section 561. The 561-02 chapter gives the Characteristic of Piezoelectric Resonators. Table 1 is an extract.



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Paolo Antoniazzi, IW2ACD, and Marco Arecco, IK2WAQ

TriHelix or Dish for P3E at 2.4 GHz?

Simulation and Measurements

At 13cm, satellite activity is confined to the "bad zone" created by 802.11x WiFi, Bluetooth, cordless phones, microwave ovens, etc. The experience with the AO-40 and AO-51 S-band downlinks has shown that the QRM is certainly a problem with the 13cm activity that so far has been receive-only. In the near future we will see 13cm used for uplinks, and we can hope that this will be a true solution. This has been noted particularly by Tom Clark (K3IO, ex W3IWI) and Bob McGwier (N4HY) in an interesting paper in 2006 ARRL Proceedings [1]. At the moment, and particularly with the coming P3E satellite, however we must survive with a 13cm downlink and a good helix or dish antenna is needed. The ranges of frequency involved with P3E are: 2400.600 - 2401.000MHz (multimode) and 2400.275 ÷ 2400.425MHz (Linear pass-band).

In the past we have simulated, realised and tested with very good results many helix antennas with 5 and 16.7 turns [2, 3] and also a small 1.7 turn version [4] used as a feeder for a 24 inch dish. The new antenna that we are discussing, named TriHelix, is an array of three helices with the layout of an equilateral triangle with rounded angles. A 3-port coupler model SC-2403 from HyperLink with N connectors is

used (see Fig 3). A comparison between Helices and Dish Antennas at 2.4 GHz is reported in this article.

1.0

Introduction

As known, it is not very difficult to design and make a helix antenna for a specific frequency and gain. The difficulties are in making precise measurements on the real antenna. The SWR measurements are not difficult if a High Directivity Directional Coupler such as the Narda model 6093 with more than 40dB Directivity from 1.7 to 4.2GHz is used. Luckily, also the measurement of radiation diagrams to obtain the directivity data is a more conventional task with circular polarisation (both helices or a parabola with helix feeder). In effect, the big errors related to direct-reflected wave encountered in Yagi tests [5] are not a problem with circular polarisation because the sense of the reflected wave is reversed. If the receiving antenna is of opposite sense (or the signal is received via a reflective path) a signal loss of 20dB or more results from the cross polarisation. It is important to remember that a plane electromagnetic wave is

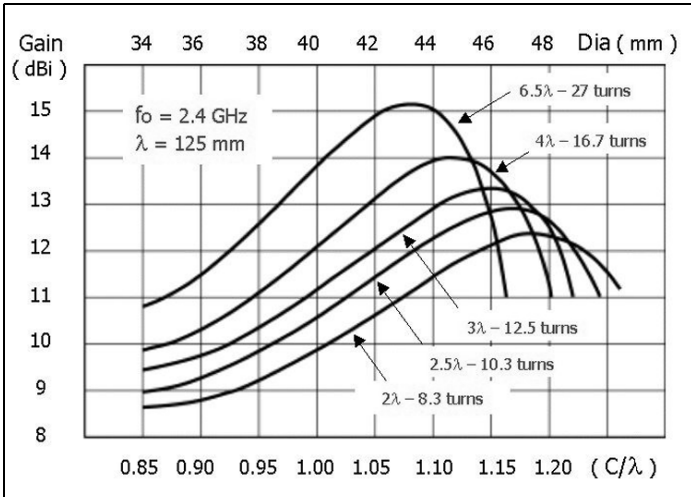


Fig 1: Simulation work by D. T. Emerson on the Helix.

linearly polarised when the electric field lies wholly in one plane containing the direction of propagation. The polarisation is vertical when that plane is perpendicular, horizontal when it is parallel to the earth's surface. A plane electromagnetic wave is circularly polarised when the extremity of the electric vector describes a circle in a plane perpendicular to the direction of propagation, making one complete revolution in one period of the wave. The polarisation sense used by the Institute of Electrical and Electronic Engineers (IEEE) is a de facto standard: When viewing the antenna from the feed-point end, a clockwise rotation results in right-hand circular polarisation (RHCP), and a counter clockwise rotation results in left-hand circular polarisation (LHCP). When two stations use circularly polar-

ised antennas over a non-reflective path, both must use antennas with the same polarisation sense. The sense of the helix must also be reversed when feeding a parabolic dish reflector.

2.0

Trihelix simulation with NEC-Win-Pro

Using the powerful NEC-Win-Pro software [7] as a simulation tool and starting from the important results obtained by the simulation work of D. T. Emerson (Fig 1 and [8]) we analysed two different

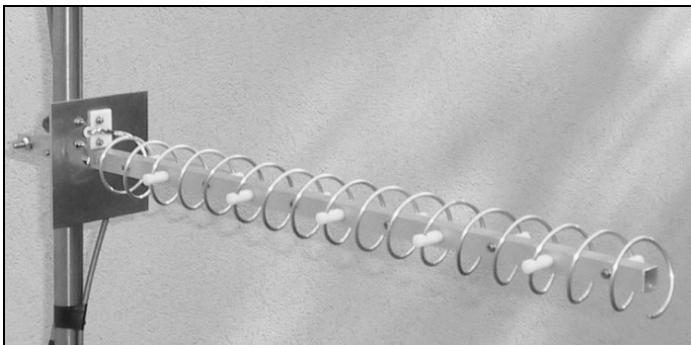


Fig 2: 16.7 turn helix antenna.



Fig 3: A 3 x 16.7 turn TriHelix antenna.

antennas. One is a simple unit of five turns (gain = 12dBi) and the other a 16.7 turn with 14.7dBi gain (Fig 2). The more complex antenna named TriHelix, a 19dB array of three 16.7 turn helix is discussed in this article (Fig 3)

It is easy to simulate the TriHelix by means of Nec-Win-Pro using the “GH” (Generate Helix) instruction that needs only eight parameters to define it independently by the dimensions and its complexity:

- Number of segments (10 segment per turn x number of turns). The number of segments chosen is a good trade-off among simulation software rules and the need to describe with enough accuracy the circular shape of the antenna.
- Spacing between turns (0.24λ or 30mm @ 2.4GHz)
- Total length of the helix (500mm for 16.7 turns helix)

- Helix radii in two perpendicular directions, normally to the helix symmetry axis, both at the beginning and the end of the winding (this feature allow us to also simulate elliptic and conical helices. This last characteristic has been very useful to simulate the helical low profile conical feeder of the dish described later)

The square ground plane (λ side or 125 x 125mm with a grid pitch of 0.05λ @ 2.4GHz), near the feed point, has been performed using the “GM” instruction that, by means of only few parameter, gives the facility to translate or rotate a structure with respect to the coordinate system or generate new structures translated or rotated from the original one. The great flexibility of this instruction has been confirmed during the optimisation of the distance among the three helices in order to have the maximum power gain at working frequency of 2.4GHz: 2.08λ or 260mm

To minimise the current in the ground

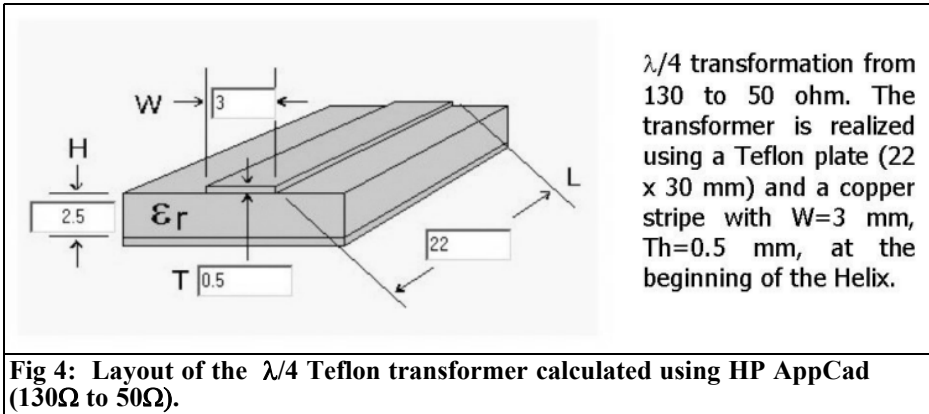


Fig 4: Layout of the $\lambda/4$ Teflon transformer calculated using HP AppCad (130 Ω to 50 Ω).

plane, a 6mm long stub has been added (suggested) between the helix and the grid plate.

The simulation of the TriHelix array needs about three times the number of segments required by the single helix (3024):

- Helix = 16.7 turns x 10 segments per turn = 167 segments
- Stub = 1 segment
- Ground plane = $2n(n+1) = 840$ segments ($n = 20$ the number of parts that the plate is divided by the wires that compose it).

We also tried to simulate the TriHelix using a perfect ground instead of a ground plane but differences detected in power gain, radiation angle and feed impedance are not significant. The only big difference is the loss of the radiation pattern at the back of the antenna when using the perfect ground. The simulated radiation pattern of TriHelix antenna is shown in the Fig 6.

Because of the layout of the three helices at the vertex of an equilateral triangle (260mm side), the radiation pattern is not perfectly symmetric as for the single helix antenna.

The simulated value for the input impedance (referred to 130 Ω) of the 16.7 turn helix is shown in Fig 8. The matching

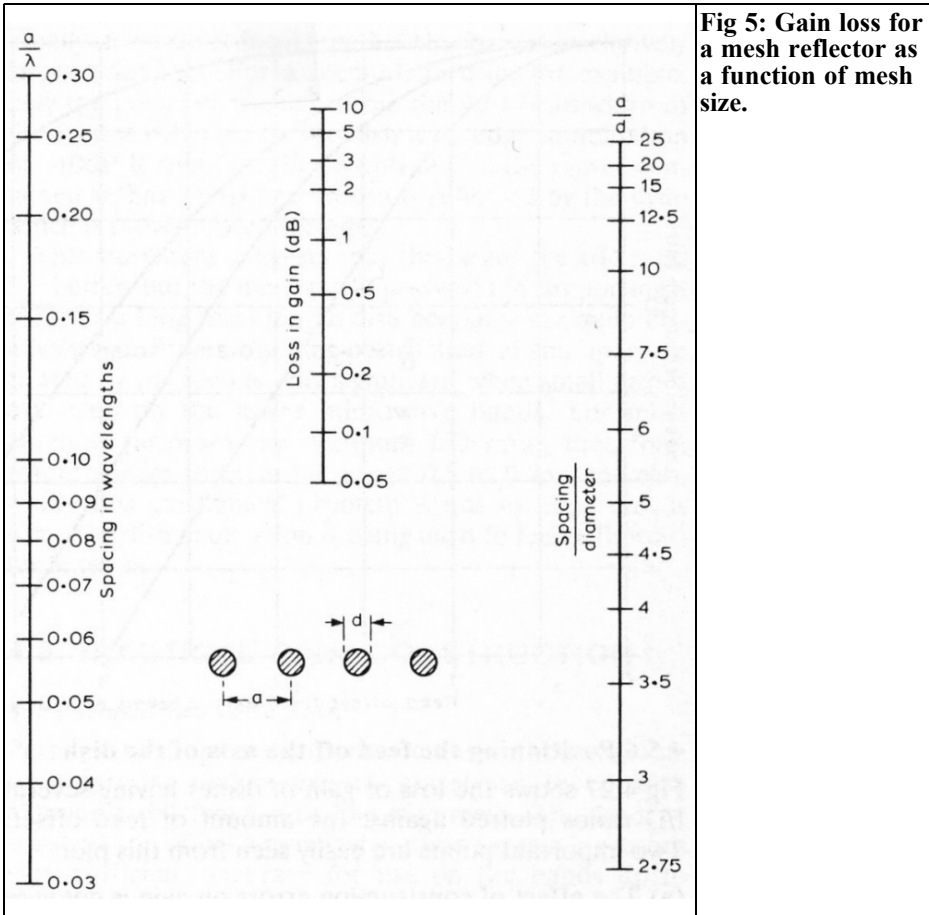
between the 130 Ω nominal input impedance of the single helix and 50 Ω is obtained using a $\lambda/4$ transformer (Teflon support with $h = 2.5$ mm and line width $W = 3$ mm, $Z = 81\Omega$). The transformer layout is shown in Fig 4 (calculated using a good HP tool, AppCad [9]). See also the photograph in Fig 12 where the Teflon impedance transformer is shown at the beginning of the smaller helix. With the multiple-helix arrays the mutual impedance of adjacent helices has to be considered, but when separated by a wavelength or more, as is typical in helix arrays, the mutual impedance is only a few percent or less of the helix self impedance (about 130 Ω).

In designing the feed corrections for a helix array, the effect of the mutual impedance can often be neglected [10] without significant consequences.

3.0

Dish simulation using NEC-Win-Pro software

A dish is a parabola: a parabolic line rotated around the axis that passes through both the focus and the centre of the curve. The choice of a parabola with a central focus instead of one with offset



has been used because it is easier to find the exact centre of the feed point without problems.

The physical dimensions of the reflector we considered are the diameter of the dish $D = 600\text{mm}$ and its depth at the centre $c = 94\text{mm}$. Using these two dimensions we can calculate the focal length "f" and the "f/D" ratio that is an easy way to describe our parabola, and the power gain "G" in dBi (referred to the isotropic gain):

$$f = D^2/(16c) = 239.4\text{mm} \text{ and } f/D \approx 0.4$$

We can also calculate the f/D ratio if we know the aperture angle (2Ψ) of the

parabola seeing the dish from its focus:

$$f/D = \sin\Psi/4 (1 - \cos\Psi)$$

$$G = 10\log_{10} [(\pi^2 D^2 \eta) / \lambda^2]$$

where λ is the wavelength in millimetres and $\eta =$ aerial efficiency (< 1)

This last equation comes from a re-definition of the capture area of an antenna. It is a simple geometric computation considering a dish antenna pointed at an isotropic transmitter at some distance away. The assumed isotropic antenna is a point source that radiates equally in every direction. The efficiency is the ratio between the power received, at the electrical connection to the feed,

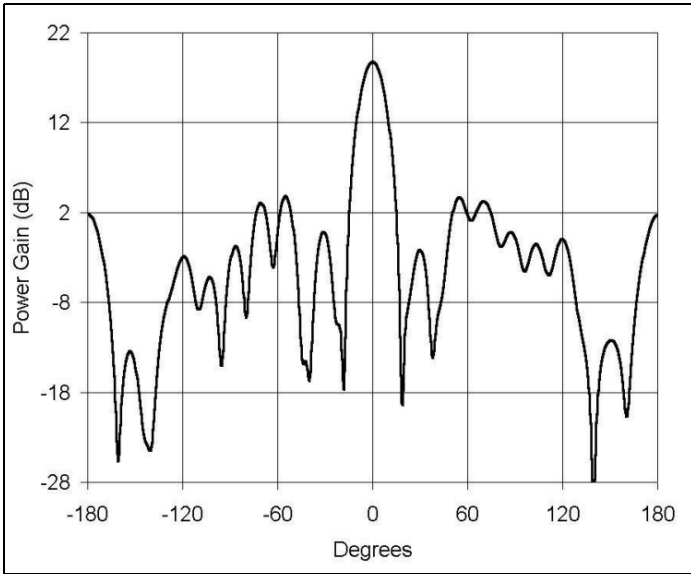


Fig 6: Radiation diagram of the TriHelix Antenna with 19dBi Gain (Simulation).

and the power that really arrived. By hypothesis, considering the non-realistic case of a perfect lossless antenna ($\eta = 1$), the power gain can grow until $G_{max} = 23.6\text{dBi}$. A more practical value of the gain should be $G = 20.6\text{dBi}$ assuming that its efficiency drops to 50% ($\eta = 0.5$). This point will be analysed better, after

the evaluation of the simulation results of the complete system. As already described in the introduction, to avoid the error problems, related to the model included in NEC-Win Synth [11] software, we decided to use Autocad to prepare the file for the simulation program. The equation, that has been used,

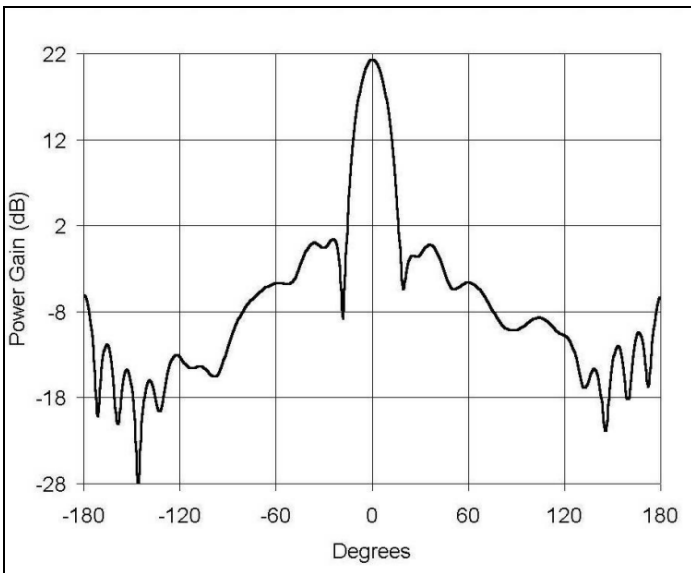


Fig 7: Radiation Diagram of the 21dBi Dish (Simulation).

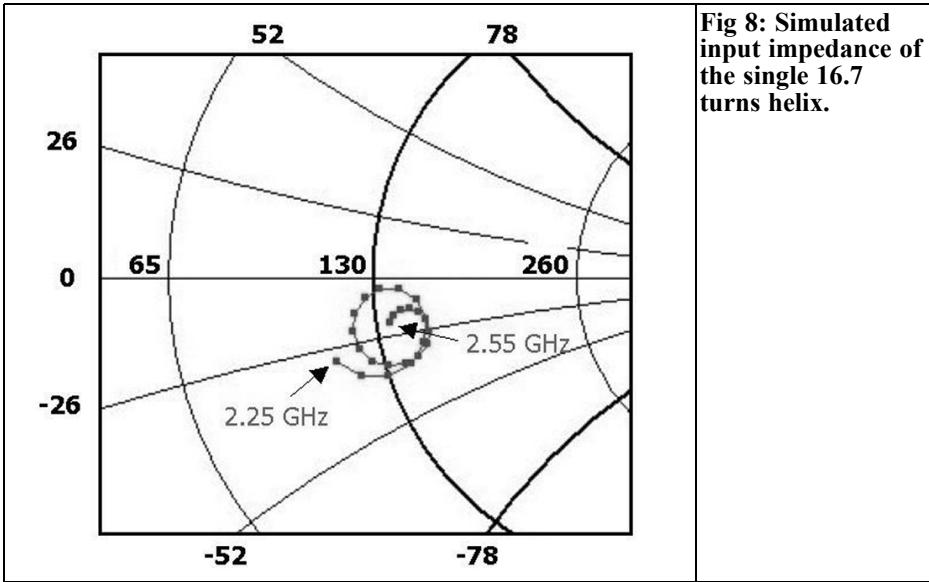


Fig 8: Simulated input impedance of the single 16.7 turns helix.

was the classic one of the parabola that is easy to find in every manual of analytic geometry:

$$y = (x^2 + z^2)/4f$$

where: x, y and z are the three coordinates in the space of the reflector expressed in millimetres.

At this point we use the NEC-Win Synth software with a double purpose: to translate the DXF file to an NEC file and to remove the possible segmentation errors performed using Autocad. It is mandatory to perform this kind of translation because the NEC is the only file format usable by NEC-Win-Pro. Unfortunately this software has not been updated and so

it is possible to perform the translation of the DXF file only using versions of Autocad older than release 9. To have a good trade-off between the simulation accuracy and the computation time (the number of segments of the complete system is 2102) it has been decided to use a mesh reflector pitch $a = 20\text{mm}$ combined together with a wire diameter $d = 6\text{mm}$. This matching, with the aid of the Fig 5 allow us to calculate the gain loss due to the use of a non-solid reflector. The values to be entered in the graph are “ $a/\lambda = 0.16$ ” and “ $a/d = 3.33$ ” giving an equivalent gain loss of 0.05dB (out of the scale).

Table 1: Simulated and measured Helix and Dish Gain.

Type of Antenna	Simulation				Measurements	
	Power gain (dBi)	Radiation Angle -3dB	Front to back Ratio (dB)	Lobes (dB)	Radiation Angle -3dB	Gain Vs. Single Helix (dB)
Dish	21.6	16.0	27.8	24	14	+7.0
TriHelix	19.0	15.0	17.0	15	16	+3.5
Single Helix	14.5	26.0	18.5	20	30	0.0



Fig 9: The complete 60cm dish antenna with the feeder and the support arm.

4.0

Low profile helical dish feeder simulation

Considering our previous experience with the helix design and manufacture we decided to use a low profile conical feeder to supply the dish, placed at the parabola focus, having a radiation angle given by the following:

$$\theta = 2 \arctang [D/2(f - c)] \approx 128^\circ$$

Almost all feeders will provide less energy at the edge than in the centre of the dish (see W1GHZ, [12]). The difference, in power, at the edge of the dish is defined as the "edge taper". Changing the shape of the feeder radiation pattern, we can change the edge taper illuminating the dish. Different edge tapers will produce different amounts of illumination and spill over losses: a small edge taper

results in a larger spill over loss, while a large edge taper reduces the spill over loss at the expense of an increase of the illumination loss. A good tradeoff in order to have the maximum efficiency of the antenna, is obtained with an edge or illumination taper of about 10dB. Looking closely to the parabolic surface, we find that the focus distance ($f = 239.4\text{mm}$) at the centre is smaller than the distance "d" between the focus and the dish edge:

$$d = \text{sqr}t [(D/2)^2 + (f - c)^2] = 333\text{mm}$$

so the wave that follows the edge has been submitted to an additional " ΔA " attenuation:

$$\Delta A = 20\log_{10}(d/f) = 2.9\text{dB}$$

This equation comes from the free space attenuation formula where the variables frequency and gain are constant (also defined as inverse square law) and so do not appear in the formula. Considering the additional attenuation, just analysed,

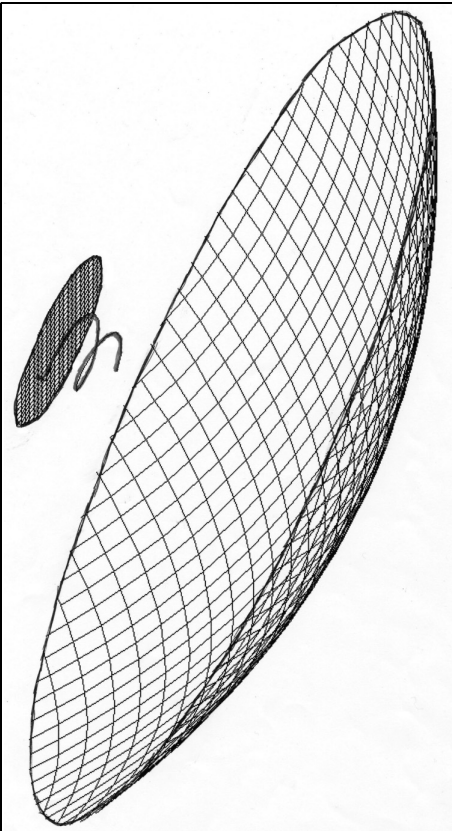


Fig 10: Mechanical design (DXF) of the complete dish system.

the optimum illumination taper becomes: 7.1dB (10dB - 2.9dB). The feeder can be any microwave device able to concentrate the beam, on the reflector, within the already calculated radiation angle

"θ": dipole, horn, waveguide, helix and so on. We decided to use the helix considering our previous experience [2, 3 and 4] in simulating and manufacturing this kind of antennas. The behaviour of the current versus length of a typical helix shows three different regions:

- Near the feed point where the current decay is exponential
- Near the open end with visible standing wave
- Between the two helix ends where there is a relatively uniform current an small SWR (transmission line)

There are two ways to obtain a good circular polarisation helix:

- Tapering the helical turns near the open end, to reduce the reflected current from the arm end
- Using only the first helical turns where the decaying current travels from the feed point to the first minimum point (Fig 11).

Starting from these considerations our final Low-Profile Helix uses a pitch $a = 0.16\lambda$ (20mm @ 2.4GHz) and is both conically wound with a cone 62/41mm diameters and very short (only 1.7 turns - see photograph in Fig 12). Let me summarise the mechanical dimensions of the feeder chosen to supply our 60cm dish:

- Number of turns: 1.7 (10 segments per turn)

Table 2: SWR measurements using the Narda High Directivity Directional Coupler.

Type	Return Loss (dB)	SWR	Note
Anne-50 by Minicircuits	30	1.06	Microwave SMA termination 50Ω + N/SMA adapter
HP model 909A	33	1.04	50Ω N termination
11.7 turn helix	20	1.22	$\lambda/4$ adapter = 23mm
TriHelix + 3 port	23	1.15	N.3 x 30cm cables

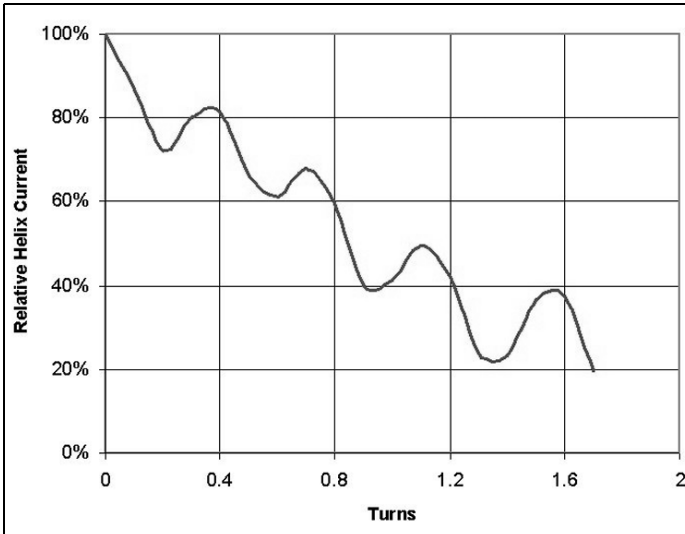


Fig 11: Current simulation in the dish feeder.

- Pitch between contiguous turns: 20mm
- Starting diameter: 62mm
- Ending diameter; 41mm
- Helix wire diameter: 3mm
- Stub: 6mm, 2mm wire diameter, only for simulation purposes to contain the current in the screen
- Screen diameter 125mm, wire diameter 1mm

The simulation of the feeder has been carried out using 674 segments: 18 for the helix and remaining 654 to define the screen with a 6.25 segmentation pitch. The simulation [4] of the feeder, using NEC-Win-Pro software, gave the following results at the frequency of 2.4GHz:

- Power gain: $G = 9.6\text{dB}$
- Radiation angle: $59^\circ @ -3\text{dB}$, $91^\circ @ -7.1\text{dB}$, $107^\circ @ -10\text{dB}$, $128^\circ @ -14\text{dB}$, $146^\circ @ -17\text{dB}$
- Front to back ratio: 16.6dB

As can be seen from the simulation results our 21dB gain dish has a high

illumination loss, but this is an advantage from the F/B, Secondary Lobes and Spill over point of view.

5.0

Dish antenna complete simulation

During the simulation phase we made some changes in the original file to optimise the performance of our antenna.

The first one involved is the diameter of the feed reflector that had been reduced from 125 to 100mm in order to reduce its masking effect on the beam reflected by the dish surface. The advantage of this operation was insignificant both for Power Gain and for the Front to Back Ratio so we decided to come back to the original screen: 125mm diameter. The second action involved the distance between the feeder ground plane and the parabola surface from 220 to 290mm. The best distance, from the Power Gain point of view that is the most important parameter for the amateur, was 270mm Theo-

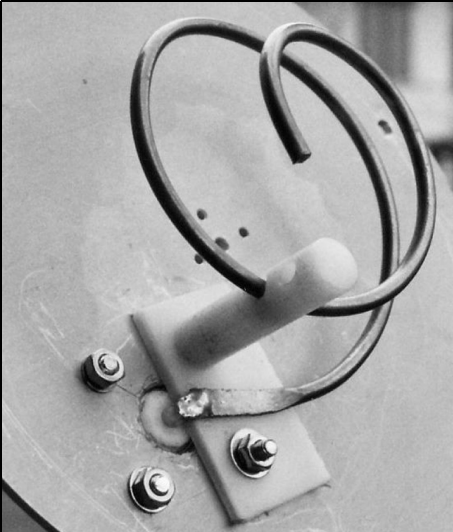


Fig 12: The helix feeder (the $\lambda/4$ transformer is realised using a Teflon plate, 22 x 30mm, and a copper strip with $W = 3\text{mm}$ and thickness 0.5mm at the start of the helix).

retically, the dish receiving antenna works as follows: the electromagnetic waves arriving on a parallel path from a distant source is reflected by the parabola surface to a common point: the focus. A transmitting antenna reverses the path: the radio wave generated by a point source placed at the focus is reflected into a beam of rays parallel to the axis of the parabola. The differences between the theory and practice are related to the following reasons:

- The focus is only a geometric point and the feeder is bigger than that point
- Defining the distance of the feeder is simpler to detect practically: it is not so easy to find the electrical centre of a helix having a truncated cone shape. Considering the simulation results and the calculated focal length it can be put about at a percentage of the total feeder length (34mm).

After the implementation of the optimisation described the results of the simulation are summarised at a frequency of 2.4GHz:

- Power Gain: 21.6dBi
- Beamwidth: $16^\circ @ -3\text{dB}$
- Front to back ratio: 27.8dB

Fig 7 shows the power gain graph.

Now we are able to calculate the antenna efficiency " η " that is the ratio between the gain, obtained from the simulation, and the maximum possible gain G_{MAX} , already computed using the equation of aerial capture area:

$$\eta = (10^{G/10})/[(\pi D/\lambda)^2] \approx 60\%$$

The efficiency loss of our complete system (-2.2dB) can be explained by the dish illumination. We consider the uniform illumination of the dish and so the increased attenuation caused by the different path that the electromagnetic wave covers at the periphery of the reflector with respect to the centre. The spill over and side-lobe losses can be neglected.

6.0

SWR and impedance measurements

One of the critical points in low SWR measurements is the extremely high quality required of the cables and adapters [6]. The time and money spent on high-quality cables can be wasted if there are large impedance mismatches within the connectors, at the connector-cable interface and with the adapters. David Slack of Times Microwave Systems writes: "... a microwave cable assembly is not just a wire. It is a passive, TEM mode, microwave component and an integral part of a system" Assuming a high quality cable is used; the predominant contributor to the SWR of a cable assembly (on a 10 -

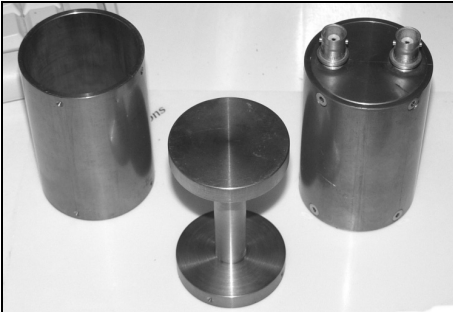


Fig 13: Brass $\lambda/2$ cavity used to filter out possible interfering signals.

50cm short assembly) is the connector. Improperly compensated geometry changes in the connector interface will exhibit very poor SWR characteristics. In previous eras, this design was considered a "black art," and trial and error was a key component of high performance design. Today, computer simulation of discontinuities in connectors is an art, and the practical results are visible when the SWR performance of a very good cable assembly (N male connectors) as that of the Times Microwave Systems is shown [6]. Only N and SMA connectors are used in our tests.

7.0

Antenna power gain measurements

To reduce the measurement errors, the distance between transmitting and receiving antennas must be considered and the signal level measured with a filtered RF voltmeter having a 30 - 40dB dynamic range. Also, the wave reaching the receiving antenna should be as planar as possible. The first condition can be easily established starting with the received power and calculating the attenuation experienced by the wave in the open space:

$$A = 32.4 + 20\log_{10}(f) + 20\log_{10}(d) - G_t - G_r$$

Here, "A" is the attenuation in decibels, "f" is the frequency (megahertz), "d" is the distance (km),

"G_t" is the gain of transmitting antenna (dBi) and "G_r" is the gain of receiving antenna (dBi) obtained by simulation. There is also a simple, easy to remember method of calculating the free-space attenuation by considering the distance between the two antennas in terms of wavelengths. When $d = \lambda A$ is always 22dB between isotropic antennas.

This equates to 12.5cm at 2400MHz. The attenuation increases by 6dB for each doubling of the path distance. This means that the free space attenuation is 22dB at 0.125m, 28dB at 0.25m, 34dB at 0.5m, etc. To make the wave reaching the receiving antenna as planar as possible, the capture area of the receiving antenna in square metres and the maximum acceptable phase error are needed:

$$A_c = G_r \lambda^2 / 4\pi$$

This expression is valid for an antenna with no thermal losses and was certainly useful for our experiments. Assuming that the capture area is circular, the minimum distance in metres between the antennas will be:

$$d > nG_r \lambda / \pi^2$$

For a maximum phase difference of 22.5°, which is usually enough, $n = 2$.

If a phase error of only 5° is required, $n = 9$. In the case where one dimension prevails over the others, the maximum length, instead of the capture diameter, is used. In this case, the minimum distance in metres becomes, [13]:

$$d > nL^2 / \lambda$$

where L is the maximum length in meters (50cm for the TriHelix).

The site we selected (a 200m² garden) is particularly useful for our measurements. For the tests we used $d = 4m = 32\lambda$ at 2400MHz.

The directivity of an antenna is based



entirely on the shape of radiated power pattern and does not consider the minor lobes. In this parameter, the antenna efficiency is not involved. The gain of a transmitting antenna is defined as a ratio between a maximum radiation intensity of the antenna and a maximum radiation intensity of a reference antenna with the same power input in the gain the antenna efficiency is involved. The helix antenna power gain (in decibels) can be computed using the following:

$$G = 10 \log_{10} \frac{(360^2 \eta) / (\pi \theta^2)}{(41253 \eta) / \theta^2} \approx 10 \log_{10}$$

where: θ is the main-lobe radiation angle at half power in degrees, η is the efficiency (<1).

The measured radiation lobes (-3dB) at 2400MHz agree within about ± 1.5 degrees with the calculated values on both the helices.

8.0

Measuring the dish

The complete system is shown in Fig 9. In our experience it is not very difficult to design and make helices for different working frequencies and gains. The true difficulties, for the serious experimenter, are in the precise measurements of the antennas. When starting with measurements it is very important to define the polarisation sense of the Helices used in the test setup. The IEEE definition, that has become the standard, is that in viewing the antenna from the feed-point end, a clockwise rotation results in right-hand circular polarisation (RHC), and a counter clockwise rotation results in left-hand circular polarisation (LHC). This is important, because when two stations use helical antennas over a non-reflective path, both must use antennas with the same polarisation sense. If antennas of opposite sense are used, a signal loss of at least 20 - 30dB results from the cross

polarisation alone. Remember that the complete dish produces circular polarisation with a sense opposite to that of the feed helix (because of the reflection of the wave from the parabolic surface).

For almost all the tests we used a generator from 2.2 to 2.7GHz using a simple VCO module followed by a 20dBm amplifier. Using an HP 435B Power Meter as a detector, we have good sensitivity and the system is perfectly stable for all the tests. To limit the measurement errors, the distance between transmitting and receiving antennas has to be considered. To determine this distance, you need to be able to measure the signal level easily with a filtered RF voltmeter having a 20 - 30dB dynamic range. Also, the wave reaching the receiving antenna should be as planar as possible. The filtering action for FM TV and other interfering signals is achieved with a simple $\lambda/2$ cavity (Fig 13). Some attention will be reserved for the omnipresent WiFi signals.

9.0

Conclusions

Helical antennas are relatively easy to build, but they have less performance in terms of side lobes and front to back ratio. This behaviour is very important when receiving signals from satellites avoiding noise and interference caused by various types of RF sourced in 2.4 - 2.5GHz band, as described previously. The 600mm dish is simple to build and clearly better than an array of helices.

Adding a shoulder cylinder at the dish periphery and absorbing material to reduce the edge effects can improve the dish antenna performance, but this is not so easy to apply at 2.4GHz.



10.0

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Andrea Daretti, IZ2OUK, andar3@gmail.com

Correlation function: What is that? And what is it for?

When I was a student I was supposed to learn and understand the correlation function and all of the associated mathematics. Instead I didn't understand it at all. To pass the test I did the same old "tricky tricks", as all students in the solar system still use: learning it all by heart and crossed fingers.

Fortunately (very fortunately) I just skimmed the test, I passed it and promised myself to forget it forever. But life is strange sometimes; I began working for HP and became a sales engineer for medical applications. One day my boss called me and said: "Our factory of South Queensferry (Scotland) designed a new system, excellent to make the signal correlation function of signals. This year you should sell one of this wonders (some several thousand dollars each)". "Oh no" I said "I hate correlation. Give me some instrument easy to sell and that make me rich!". The answer was the typical sales manager answer: "Don't bother me: go and sell and don't come back without an order!". I can't report what said, but then in deep depression, I took the training manual and I started reading.

1.0

Miracle!

I was deeply stricken by a vision! I perfectly understood it all. Maybe it was the right timing. Even more, I not only understood it all, but also started to love it. Together with a sales engineer in Rome, where I was based at the time, we became the Italian "Champions" of this product. That year we sold three of them; one to University of Medicine Department of Human Physiology for a very interesting neuroscience application. A second was sold to Fiat for vibration analysis. The discovery of Pulsars could have been much simplified with such a system.

Now I'll try to strike you too, if you have the curiosity to learn. I don't intend to give you a brilliant lesson of mathematics. Just skimming, to prevent the real experts from criticising this little article. Instead, what I do intend to do is to give you just the "feeling", so that if I'll be able to triggered your interest, you can run to professionally written books and feed your curiosity.

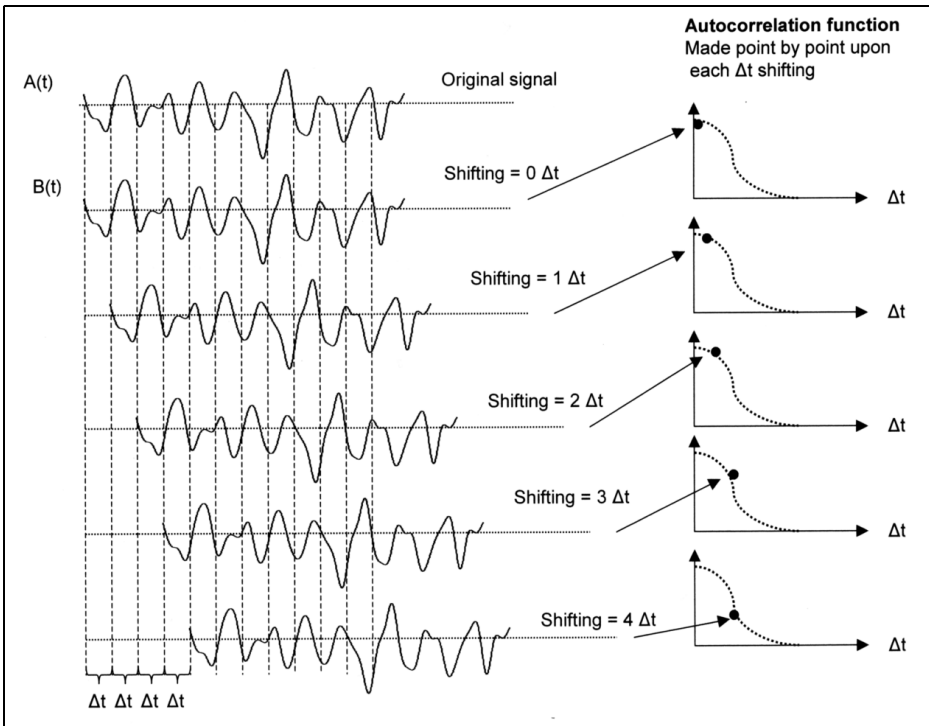


Fig 1: The process of plotting an autocorrelation graph point by point.

2.0

And now, “Back to basics”

Question: What’s this strange and intimidating thing called “Correlation functions, Autocorrelation and Crosscorrelation”?

It is a method to evaluate similarities among signals (or set of data).

Let’s see how it works and maybe there are some chances I’ll transfer a bit of my enthusiasm to you. Let’s take a “whatever” signal (more specifically: the time footprint of a signal): like the noise output of an amplifier. At the same time (and this is important) take a second signal, that can even be the same (this is called Autocorrelation) or a different one

(this is called Crosscorrelation).

To start, just imagine that the second is the same signal of the same length. Take the first copy; slice it into many small slices and measure each slice amplitude; exactly the process of “sampling” that any A/D converter does. You will come out with a series of numbers taken after slicing and measuring (sampling) all the “slices” of your signal. Then do the same for the second signal to be correlated. In total we will get two series of data (amplitude of “slices”) of the same length. Multiply each corresponding pair of samples, and sum together all the multiplied results. Then divide this final result by the number of samples, so to normalise. This will generate one single number, maybe big, and this single number will be the first point of a new graph: the Autocorrelation plot.



3.0

Here comes the magic.

Take the same second signal: delay it by a little time (Δt) and do the same process (slice, measure, multiply each pair of data, sum all of them and normalise). It will still be a big number but not as big as before, because the shifting has reduced the coincidence between the two signals. Still it will be a good value. This value will be the second point into the Autocorrelation plot. Then do this same process once more shifting the second signal by one more Δt . Again you will have a good value, not as big as the other two, but still a value. This is the third point of the Autocorrelation plot. Repeat this process again and again along all the signal string, delaying the second signal each time by one Δt time interval. And point-by-point you will draw the complete Correlation graph plot, Autocorrelation in this case. See Fig.1

This is the process of the plotting of Autocorrelation graph point by point. You will understand that if the two signals have nothing in common (casual), like noise is, pretty soon the value of the graph will go close to zero. But imagine that hidden in the noise there is a periodic signal.

Surprise, surprise!

As soon as the shifting (delaying) of the second signal is equal to the hidden signal period there will be something more in common, and an evident value, will pop up because the multiplication of common value will give a good value. That's it! The correlation process allowed you to spot periodic hidden signal. And it will also provide the same results for a negative value of Δt as well: that's why you will see the sign " \pm "

Remember: it will tell you about periodicity, but not necessarily about the signal's shape. As mentioned, all the num-

bers creating the correlation graph will be "normalised"; i.e. divided by the total number of slices (samples): so to get an average value

For those needing mathematical symbolism, we started with a big sum of numbers $x(t)$ and $x(t\pm\Delta t)$ (same signal shifted by Δt) multiplied and normalised

$$(Auto)Correlation(\Delta t) = \frac{1}{T} \sum_0^T x(t) \cdot x(t \pm \Delta t)$$

T is the total number of samples

The correlation plot will have the calculated value on vertical axes, and Δt , the time shift (and not time!) on the horizontal axis. See Fig 1. If we now make Δt very, very, very small we will move into an integral function. Don't get scared: the famous (or infamous) symbol \int comes from middle ages and stands simply for "Sum" and in fact it looks like an S (so I was told...). And the formula will become:

$$(Auto)Correlation(\tau) = \left(\frac{1}{T}\right) \int_0^T x(t) \cdot x(t \pm \tau) dt$$

T is extended to the full length of the signal trace and Δt becomes τ . Don't worry: that's all mathematics you will see!

4.0

Good story teller?

And now, let's see if I have been a good story teller (I hate the term "teacher"). We have just seen that a random (white) noise (or casual signal) has no repetitive components in it. For this reason the Autocorrelation will have an initial peak quickly zeroing and stay close to zero. Now: let's think of a sine signal: what will its Autocorrelation graph be?

Let's see it step by step. As long as shifting will be 0 the first point in the

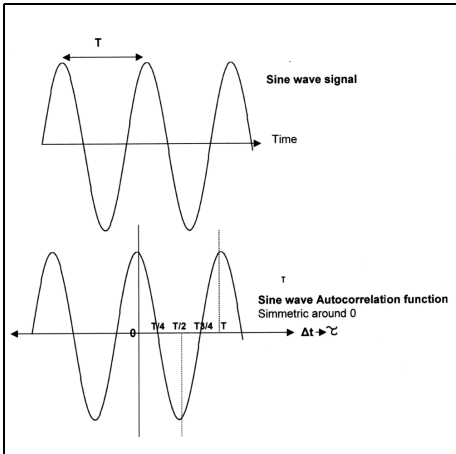


Fig 2: Autocorrelation of a sine signal.

graph will have quite a good value, because we multiply it by itself. As soon as we start delaying the second copy of the signal, the value will be decreasing. Then after a delay of $\frac{1}{4}$ of the sinusoid period, the correlation becomes 0 and then becomes negative. Then up again up until it draws a sinusoid of the same

period up until it is a cosine, no matter what the initial phase of the sine. Note that autocorrelation is symmetric around for both shifting: positive (delaying) or negative (anticipate). See Fig 2. Note the symmetry around 0 as all autocorrelation graphs

More: what's the autocorrelation of a square wave (symmetric amplitude around 0)? Again: $\tau = 0$ at it is a high value, but as soon as we start shifting the second signal copy, there is a linear reduction in value as τ increases, up until the increasing shift will make the two waveforms 180° out of phase. Then it starts to rise, still linearly, up until the two are back in phase. In a nutshell, the autocorrelation of a square wave is a triangular wave of the same period. See Fig 3

Now comes a juicy thing. What will the autocorrelation of a pseudorandom binary sequence (pbs) be? Pbs is a random digital sequence repeating itself after a while (repetition period). Again: the correlation, beside the peak for quickly goes linearly to 0, it will be zero for as long as

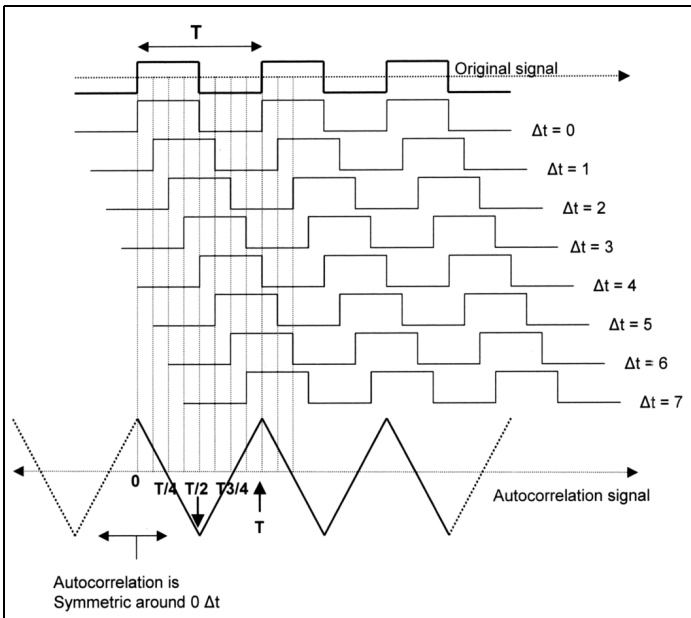


Fig 3: Autocorrelation of a square wave.



is the period of the repetitive digital sequence, and as the example of the autocorrelation for square wave indicated, it will show triangular peaks of the same repetition period.

Don't believe? ...Think harder!

Then, when you have discovered the repetition rate of a signal hidden in a noise even a hundred times bigger, you can make a triggered average and recover it (easy to say, not so easy to do, but the tool is here). With all of this I hope I've given you the feeling that:

“Correlation is a method to check the similarity of two signals that can be the same (autocorrelation) or different (crosscorrelation)”.

5.0

What are the real application of this method?

- Imagine you are an engineer prototyping a car: the inside seems to be noisy, and you suspect that the noise mostly comes from the water pump. So you place a transducer on the pump and a second one on the car inside. Then you do a crosscorrelation among the two and if you are right...
- Imagine you are Jocelyn Bell (discoverer of the Pulsar stars). Instead of analysing some kilometres of strip chart recordings to find minuscule pulses hidden in noise you could have done it in a snapshot.... if this system existed six years before. In fact today correlation in general is very much used in astrophysics and radio astronomy.
- Imagine you are neurosurgeon and want to know if a certain nervous path of your patient is damaged....

- Imagineshould I continue?

- We have seen this process applied to signals in terms of “signal amplitude”, but of course it can be applied to other parameters, like phase, frequency etc.

Are you still alive? I congratulate you!

6.0

Is it any good for Radio Hams?

Unfortunately not much today.

As I said this algorithm doesn't provide information on the “shape” of the signal, but only on some hidden components. The features emerging from this technology are of particular interest to Radio astronomy, first bringing in a huge increase in antennas angular resolution, so to make them (from radio astronomy point of view) as good as an optical telescope, or, in the near future, even better.

7.0

Interferometric radio telescopes.

An interferometer is a radio receiver (often wide band) with two (at least) antennas with identical RF and IF placed far away from each other (Fig 4). The real signal will reach the antennas with a travel distance different by L . This means a phase difference: if in phase the two signals will be added, if out of phase will be subtracted. Noise from the two preamps, being random, will be reduced. The received signal, once added and detected, will have a theoretical shape as shown in Fig 5.

The peaks will come when the signals

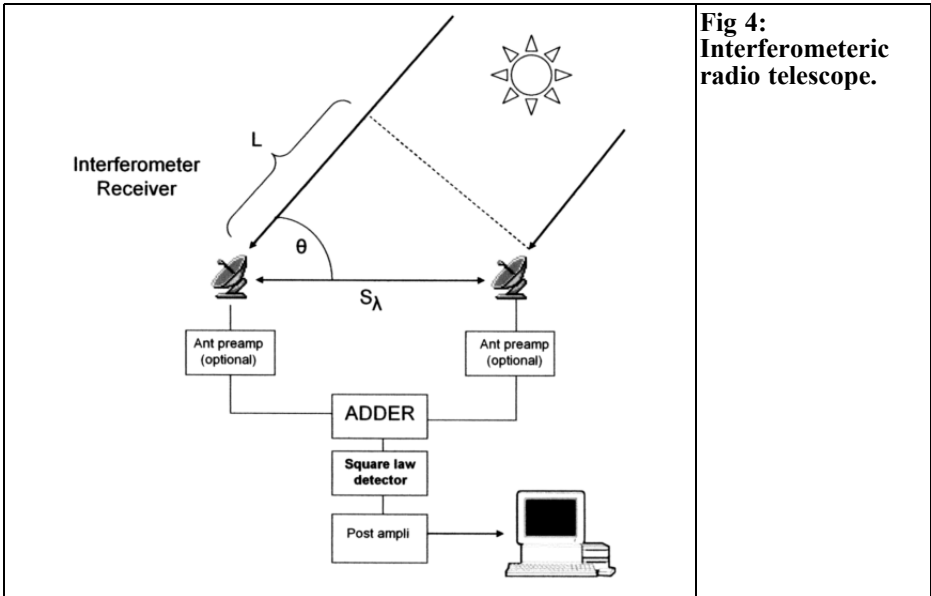


Fig 4:
Interferometric radio telescope.

arrive in phase at the two antennas and then added: the nulls will come when the two signals are out of phase by 180 deg. The peak width represents the angular resolution of our interferometer. It is increased greatly compared to the lobe of each single antenna. In fact if the two antennas are far enough, the peak width (Beam Width between First Null =

BWFN), i.e. the angular resolution, becomes:

$$BWFN = \frac{1}{S_\lambda} (rad) = \frac{53.7}{S_\lambda} (deg)$$

Where S_λ = Antenna distance in wavelengths

This can be a huge increase if distance is big enough! This is the reason why antennas are placed as far apart as possible, even across continents. The problem then is the perfect synchronisation of recording the incoming signal, but this is another story. It is with this technique that today we can admire the beautiful radio sky maps. An added benefit is that by receiving signals from two antennas, the sensitivity is doubled (theoretically!).

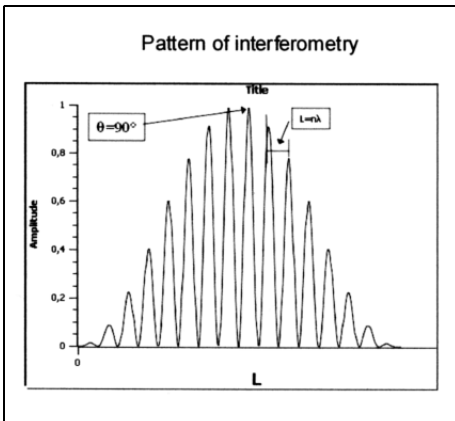


Fig 5: Peaks come every time $L = n$. Peaks are as wide as per increased resolution.

8.0

Correlation radio telescopes.

If we multiply each other the two antenna signals, instead of simply adding, we

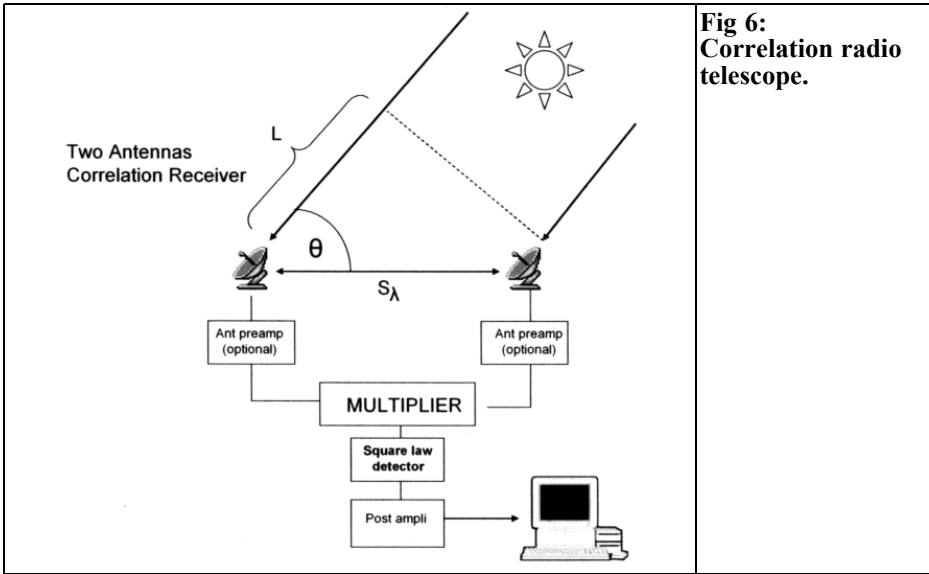


Fig 6:
Correlation radio telescope.

perform exactly an operation of correlation. The time shifting of the two signals, as required by correlation operation, is performed by the earth's rotation for fixed radio telescope (like Arecibo) or by scanning for steerable ones. The product of two random signals like noise (external noise, preamplifiers noise and gain insta-

bility) gives a zero DC component, while sensitivity to correlated signals is positive and is $\sqrt{2}$ times the sensitivity of the same receiver operating in standard (total power) mode. This is the pay-off for such an increase of complexity.

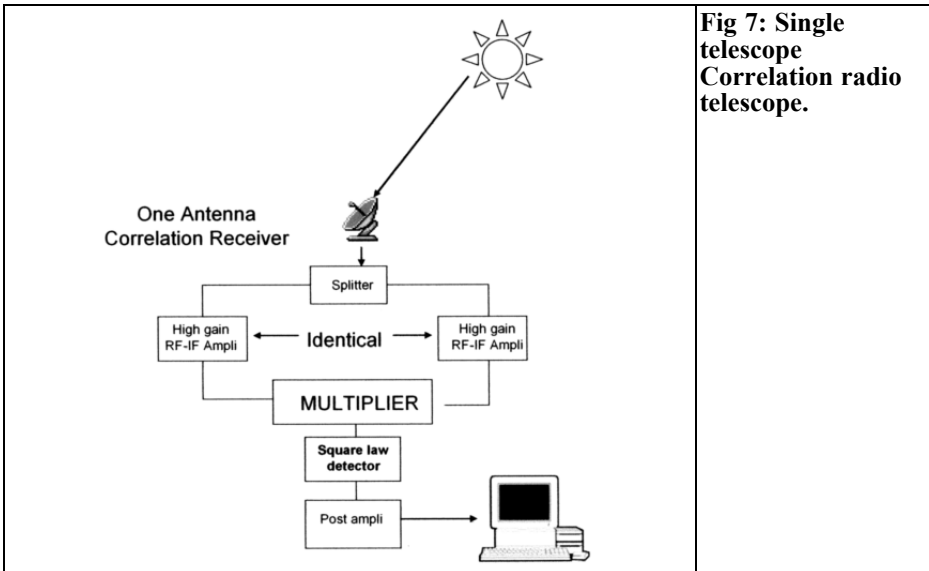


Fig 7:
Single telescope Correlation radio telescope.



9.0

Further evolution.

As a simplification, a correlation receiver can also be realised with one single antenna, loosing of course the advantage of increased angular resolution (Fig 6). This new configuration is indicated to reduce the noise of RF-IF amplifiers, and particularly amplifier gain instability, which, in radio astronomy, is a very important problem. It can't help to reduce the sky or man made external noise reaching the antenna. That can be cured with additional algorithms after detection (e.g. averaging).

Looking ahead.

Today we are more and more speaking of satellites, the Moon, EME and even signals from Mars. All of this implies

increase of sensitivity and huge increase in quality. Digital data encoding can be one of the ways, becoming more powerful and less expensive every day, so that I wouldn't be surprise if in the future we will see some of these technologies downloaded into our radio receivers. And if you have been able to follow me up to the end, I should really congratulate you tough guys!

10.0

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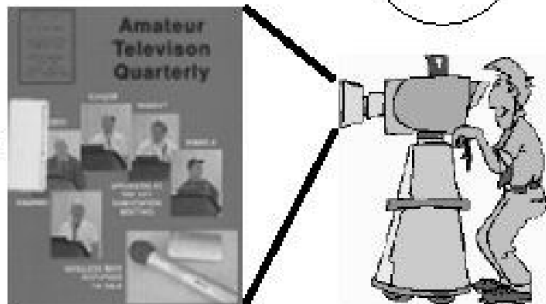
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Antonio Iuliano, IZ1ANS

CTCSS Oscillator

1.0

Introduction

CTCSS is an acronym for “Continuous Tone Coded Selective Squelch”. It is based on the recognition of a sub tone present in the modulation of a transmitted signal by the squelch circuit of a repeater or another station thus permitting access only to the stations that have the desired tone. In the transmitter there is an oscillator at the specified audio sub tone frequency that is selected by a switch to low level modulate the carrier during the whole transmission. The low frequency and the low modulation index allow the squelch to decode the tone without it being audible. The sub tone signal must be injected into the transmitter modulator after the input filters and pre-emphasis circuits.

Some time ago I made a homebrew CTCSS oscillator for an old KDK FM 2033 144Mhz transceiver. The transceiver had an internal socket with all the necessary connections and an 8V DC supply only present during transmit and when selected by a switch on the front panel. There was also a trimmer to adjust the modulation level.

2.0

The circuit

The circuit (Fig 1) is built around a CD40103B 8bit programmable divider; this provides division by a number between 2 and 256. The divider needs an oscillator to give the input frequency. I decided to use a 32.768khz clock crystal, because there were no spare logic elements available I used a transistor for the oscillator.

I found a BC584B in my junk box but any low frequency high beta transistor is useful. It may be useful to verify the output signal and vary the collector resistor to obtain a rail-to-rail swing. The goal was to obtain a good sine wave output by filtering; this requires a constant input duty cycle. This is obtained by dividing the output of the 40103B by two using half of a CD4013 dual Flip Flop.

If N is the dividing modulus entered into the 40103B it divides by N+1. Because the output is divided by two the formula is:

$$N = (16384 / F_{out}) - 1$$

16384/F_{out} is rounded to give an integer number. The output frequency is not exact but presents an acceptable small error of around 0.3%. Table 1 shows the

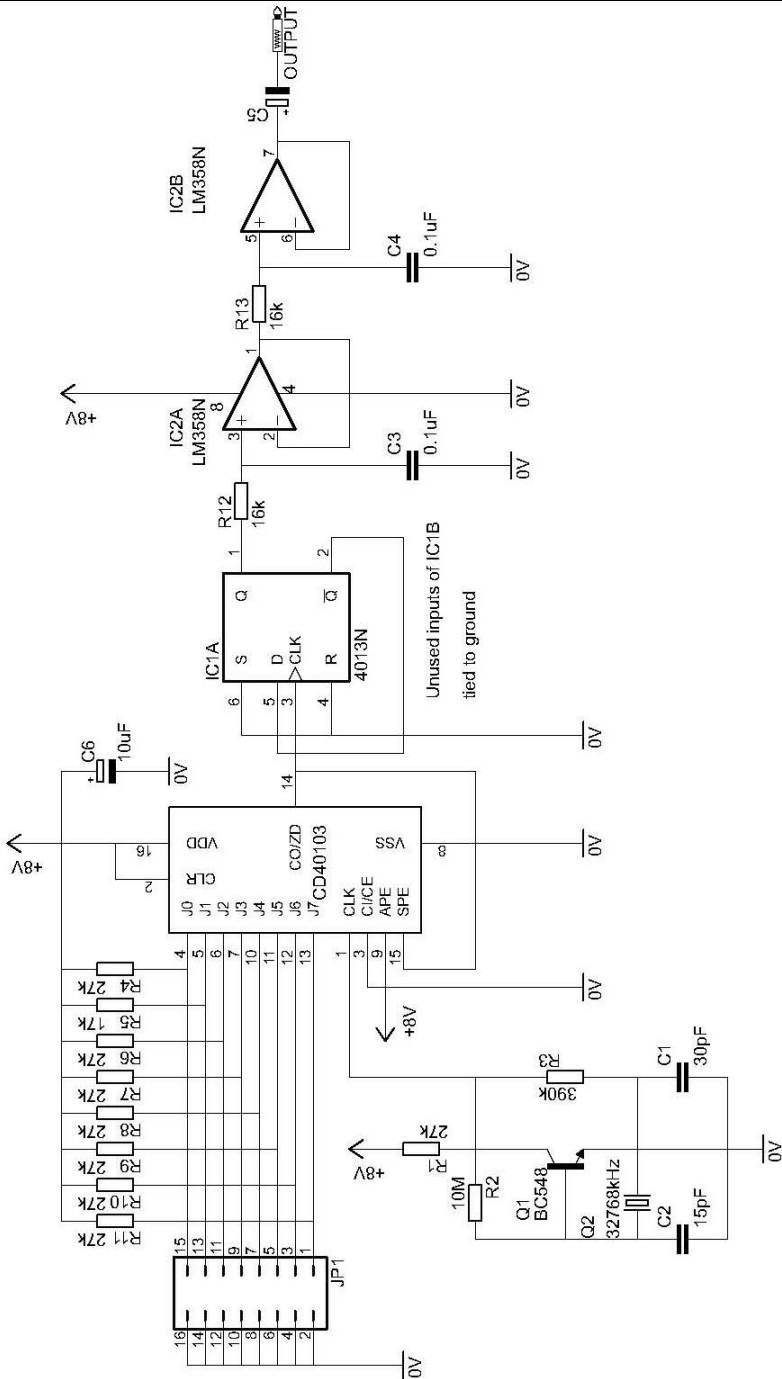


Fig 1: Circuit diagram of the CTCSS oscillator.



Fout	N1	N+1	N	J7	J6	J5	J4	J3	J2	J1	J0
67	244.5373	245	244	1	1	1	1	0	1	0	0
71.9	227.872	228	227	1	1	1	0	0	0	1	1
74.4	220.2151	220	219	1	1	0	1	1	0	1	1
77	212.7792	213	212	1	1	0	1	0	1	0	0
79.7	205.5709	206	205	1	1	0	0	1	1	0	1
82.5	198.5939	199	198	1	1	0	0	0	1	1	0
85.4	191.8501	192	191	1	0	1	1	1	1	1	1
88.5	185.1299	185	184	1	0	1	1	1	0	0	0
91.5	179.0601	179	178	1	0	1	1	0	0	1	0
94.8	172.827	173	172	1	0	1	0	1	1	0	0
97.4	168.2136	168	167	1	0	1	0	0	1	1	1
100	163.84	164	163	1	0	1	0	0	0	1	1
103.5	158.2995	158	157	1	0	0	1	1	1	0	1
107.2	152.8358	153	152	1	0	0	1	1	0	0	0
110.9	147.7367	148	147	1	0	0	1	0	0	1	1
114.8	142.7178	143	142	1	0	0	0	1	1	1	0
123	133.2033	133	132	1	0	0	0	0	1	0	0
127.3	128.7038	129	128	1	0	0	0	0	0	0	0
131.8	124.3096	124	123	0	1	1	1	1	0	1	1
136.5	120.0293	120	119	0	1	1	1	0	1	1	1
141.3	115.9519	116	115	0	1	1	1	0	0	1	1
146.2	112.0657	112	111	0	1	1	0	1	1	1	1
151.4	108.2166	108	107	0	1	1	0	1	0	1	1
156.7	104.5565	105	104	0	1	1	0	1	0	0	0
162.2	101.0111	101	100	0	1	1	0	0	1	0	0
167.9	97.58189	98	97	0	1	1	0	0	0	0	1
173.8	94.26928	94	93	0	1	0	1	1	1	0	1
179.9	91.07282	91	90	0	1	0	1	1	0	1	0
186.2	87.99141	88	87	0	1	0	1	0	1	1	1
192.8	84.97925	85	84	0	1	0	1	0	1	0	0
203.5	80.51106	81	80	0	1	0	1	0	0	0	0

Table 1: Table showing jumper settings for the common CTCSS frequencies.

value of N for the common CTCSS tones.

At this point we have a square wave at the desired frequency but we need a filter. After some simulation I selected a simple circuit based on two simple RC filters using two sections of an LM358 as buffers. The output signal is practically sinusoidal.

The frequency programming inputs of the 40103B are tied to Vcc by pull up resistors. A “1” is programmed by leaving the jumper open; a “0” is pro-

grammed by shorting it to ground by inserting the jumper.

3.0 Prototype

In the prototype I used SMD resistor and capacitors where possible (Fig 2).

For selecting the frequency I used 8 jumpers to save space but a dip switch

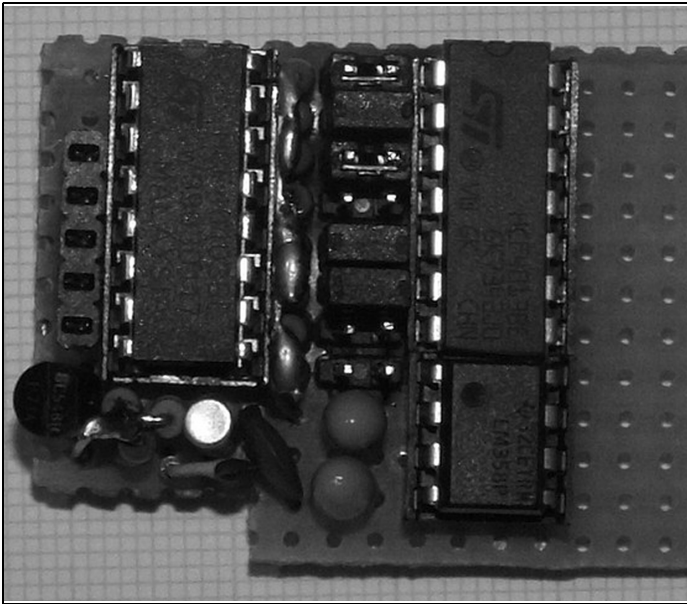


Fig 2: Picture of the prototype CTCSS oscillator.

may be more practical.

It is possible to insert a flat cable connector in place of the jumper and place a dip switch at the other end of the flat cable or switch between some desired frequencies

with a rotary switch and a diode matrix.

Please feel you free to contact me directly by email (antonio.iuliano2@tin.it) or via the editor.

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John Fielding, ZS5JF

John's Mechanical Gem No. 10

Attaching hubs to shafts

Many amateurs construct their own rotator or elevation systems using various items that are often better than commercial rotators or elevators. Usually the commercial items are too expensive or not suitable for large antenna arrays.

the shaft D shaped. For high torque loads the screw point is in shear and if the torque generated is high enough the material can be severely stressed and the screw end can be broken off. Shock loads can be many times the static loads and if the shear stress is high enough it can leave the two parts disconnected. This is an unsafe situation.

1.0

Simple method

In a rotator or elevator system the need often arises to attach a sprocket or gear to a shaft with positive locking. For light loads a method is to use a screw tapped into the collar or bush on the sprocket or gear that bears on the shaft to prevent movement. Often a dimple hole is drilled a little in the shaft for the screw to engage with or a flat machined making

2.0

Using a key

For high torque interfaces the best choice is a form of keyway. This can also be used as a "weak-link" to prevent damage. Common types of keys are feather keys (Fig 1), gib-head keys (Fig 2) and Woodruff keys (Fig 3). For the feather and gib-head key a slot is machined in the



Fig 1: Feather key.

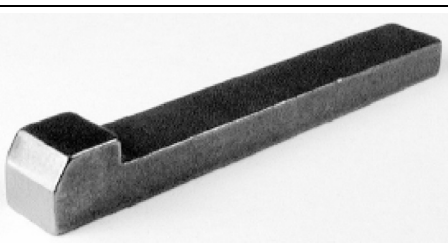


Fig 2: Gib-head key.



Fig 3: Woodruffe key.

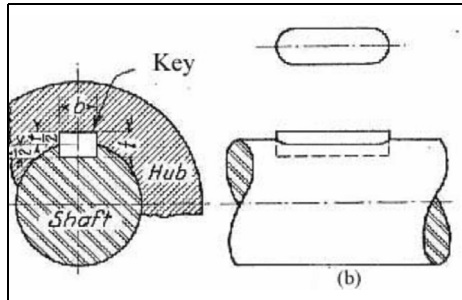


Fig 4: Feather key application.

shaft and a corresponding slot in the sprocket or bush. (For the feather key a slot drill makes the rounded ends so the key seats correctly). The square space is filled with a piece of key material that is a close fit between the two. The feather key normally has a screw in the collar bearing on it to give extra security. The gib-head key has a slight taper and is driven into the square formed between the shaft and collar. For Woodruffe keys the shaft has a semicircular slot machined into it and a Woodruffe key inserted to fill the space fitting into the square slot machined in the sprocket or bush (Fig 4).

Note: Half the key height is in the shaft and slightly less than half is in the collar allowing a small clearance for the top of the key. For a 8mm wide key the shaft depth is 4mm and the hub keyway is typically 3.3mm deep, so a 8mm x 7mm key is used, giving 0.3mm clearance. These dimensions are standardised in either an ANSI standard for imperial keys or in a DIN standard for metric keys.

Cutting the keyway in a shaft is relatively easy using a milling machine and a slot drill or other types of cutters, however cutting the corresponding keyway in the hub or sprocket requires sophisticated slotting machinery or a broach made to the exact size, which is an expensive item.

3.0

Feather key application. Formula to calculate safe torque load

Where a key is used to interface between a shaft and collar the following formula can be used to calculate the approximate safe torque load.

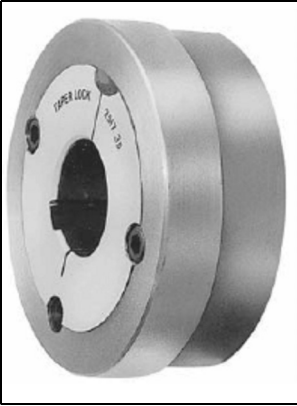
$$\text{Torque (Nm)} = 0.065 \times \text{Key Length} \times \text{Key Width} \times \text{Shaft Diameter} \quad (\text{dimensions in mm})$$

For an 8mm square key 25mm long in a shaft of 30mm diameter the safe torque load is 390 Nm (39.75 Kgf-m = 287.4 ft-lb).

4.0

Using a Taper-Lock

An alternative method is to utilise "Taper-Lock" hubs. Several manufacturers make them, (typically Fenner) and these can be purchased as "bolt-on" hubs or "weld-on" hubs for sprockets etc. They have the internal keyway already machined to close tolerances as well as the split taper collet that is drawn up tight by grub screws (Fig 5). They are available in imperial and metric versions to suit common shaft diameters. The



**Fig 5:
Taper-Lock-
Weld-On
hub.**

combination of the key and the taper collet, which grips the shaft tightly, is a simple method of providing an adequate high torque interface. (To release the taper collet the two grub screws are removed and placed in the two unused holes where they serve to jack the collet out of the hub). The advantage of the taper lock hub is that the key does not have to transmit all the torque, the collet grips the shaft tightly and the key only comes into play once the torque is sufficient to cause the hub to slip on the shaft. Hence, a smaller key than normal can often be used.

An example of the use of taper-lock hubs is shown in Fig 6 showing the elevation mechanism for the proposed new Durban

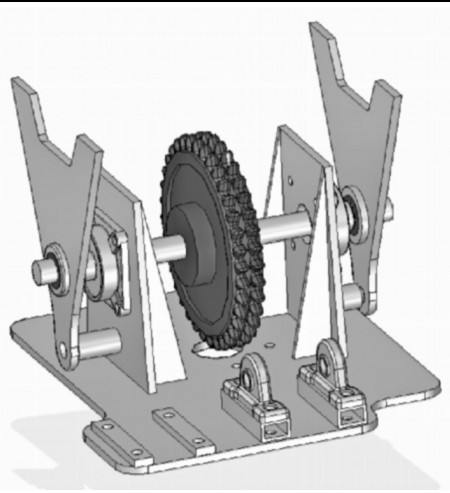
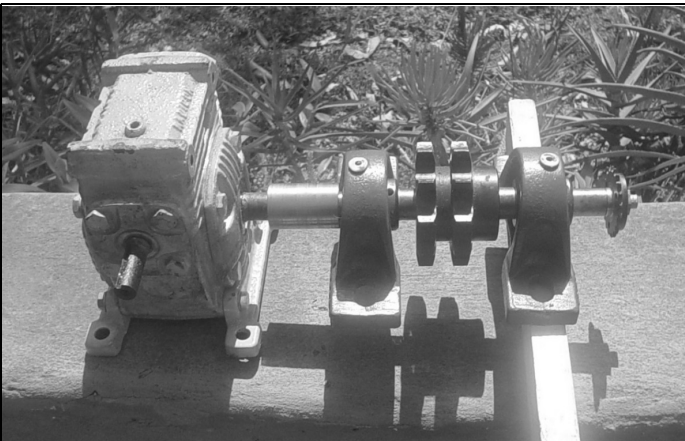


Fig 6: 3D CAD drawing of the elevation assembly.

University of Technology radio telescope known as Indlebe-2. This is a 7.5m diameter parabolic antenna weighing 150kg. The mass of the antenna needs to be lifted through 90°. The elevation mechanism uses a horizontal cross shaft supported in bearings and two wishbones that attach to the rear of the dish. The cross shaft is rotated by a step-down chain drive driven by a worm gearbox. The taper-lock weld-on hubs are fitted to the wishbones and the chain sprocket with keys to prevent slippage.



**Fig 7: Indlebe-2
elevation axis
secondary shaft.**

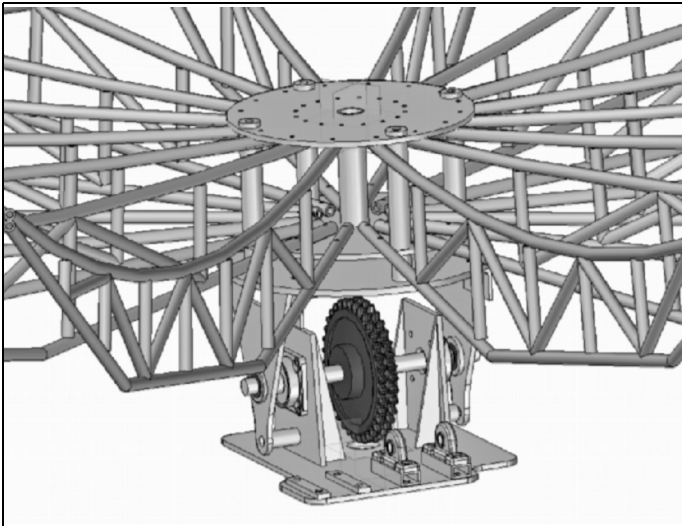


Fig 8: 3D CAD drawing of the elevation assembly.

Fig 6 shows a 3D CAD drawing of the elevation assembly. (The secondary shaft and sprocket are not shown). The bottoms of the wishbones have holes lining up with hollow sleeves welded to the bearing support brackets to accept the

stowing pins. The worm gearbox is mounted adjacent to the secondary shaft bearings. The secondary shaft and gearbox are mounted at the rear of the elevation assembly. The duplex sprocket is 20mm pitch and 240mm in diameter.

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Electrical Specifications at T_{AMB} = 25°C

MODEL NO.	FREQ. (MHz)	GAIN (dB)				MAXIMUM POWER (dBm)	DYNAMIC RANGE			VSWR (:1) Typ.		ACTIVE DIRECTIVITY (dB) @ Pin V+ Isolation-Gain	DC VOLTAGE @ Pin V+ (V)	DC OPERATING CURRENT @ Pin V+ (mA)	
		Flatness					Output (1 dB Comp.)	NF (dB)	IP3 (dBm)	In	Out			Typ.	Max.
	f _c - f _h	Typ.	Min.	Typ.	Max.	Typ.	Typ.	Max	Typ.			Typ.		Typ.	Max.
ZX60-1215LN +	800-1000	16.5	14.0	±0.7	±1.3	12.5	0.4	0.7	26.0	1.65	1.40	13	12	42	50
	1000-1400	14.5	11.0	±1.2	±2.5	12.5	0.4	0.7	27.5	1.70	1.40	13	12	42	50

Fig 9: Data sheet from Mini Circuits for the LNA used on the Indlebe radio telescope.

**Typical Performance Data & Curves at 25°C ZX60-1215LN+**

FREQUENCY (MHz)	GAIN (dB)	DIRECTIVITY (dB)	VSWR IN (:1)	VSWR OUT (:1)	POWER OUT @1dB COMPRESSION (dBm)	IP3 (dBm)	NF (dB)
800	17.25	13.60	1.63	1.76	12.00	24.51	0.42
850	16.90	13.47	1.61	1.62	12.16	25.00	0.41
900	16.59	13.34	1.59	1.49	12.33	25.49	0.41
920	16.46	13.28	1.59	1.44	12.40	25.68	0.41
940	16.30	13.35	1.59	1.40	12.46	25.86	0.41
960	16.15	13.30	1.60	1.36	12.53	26.05	0.40
980	16.02	13.25	1.60	1.32	12.59	26.23	0.40
990	15.94	13.23	1.61	1.30	12.62	26.32	0.40
1000	15.90	13.30	1.61	1.29	12.64	26.41	0.40
1020	15.73	13.28	1.62	1.26	12.70	26.59	0.40
1050	15.53	13.34	1.63	1.23	12.77	26.83	0.40
1080	15.32	13.39	1.65	1.21	12.85	27.13	0.40
1100	15.15	13.39	1.67	1.21	12.89	27.32	0.40
1120	15.00	13.40	1.68	1.22	12.93	27.50	0.40
1150	14.79	13.43	1.69	1.25	12.96	27.73	0.40
1180	14.60	13.44	1.72	1.29	13.01	28.05	0.41
1200	14.45	13.51	1.74	1.31	13.03	28.23	0.41
1250	14.15	13.65	1.77	1.40	13.06	28.65	0.42
1300	13.73	13.78	1.81	1.51	13.05	29.11	0.43
1400	13.01	14.07	1.87	1.74	12.93	29.91	0.45

Fig 10: Typical performance data for the LNA used on the Indlebe radio telescope.

Fig 7 shows a picture of the Indlebe-2 elevation axis secondary shaft, worm gearbox and final drive sprocket. The small sprocket on the end of the shaft drives the feedback pot via a chain.

Fig 8 shows a 3D CAD drawing of the elevation assembly and hub of the parabola with the ribs attached. The wish-bones attach to the bottom plate via angle iron welded to them and bolts onto the circular plate. Four 16mm diameter studs pass through the vertical tubes in the central hub to bolt the hub to the elevation circular plate.

telescope uses these and the noise performance is lower than it was possible to achieve with discrete devices. The LNA shown is very compact and has a noise figure of 0.45dB. Figs 9 and 10 show the data sheet from Mini Circuits (www.minicircuits.com)

5.0

LNA Used for Indlebe radio telescope

Recently (2009) a new range of packaged low noise amplifiers has been introduced by Mini Circuits that offer good performance at an affordable price. The new receiver front end of the Durban Indlebe



Gunthard Kraus, DG8GB

Internet Treasure Trove

Ken's web site

Ken is an accomplished professor from Alamba (a great acknowledgement from Gunthard who is a accomplished professor from Tettngang). This homepage has a wide variety of subjects it is well worth a read

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

A very high dynamic range LNA for 144MHz

Something for constructors and developers of low-noise preamplifiers. A very well written, detailed and informative magazine article in the pdf format; worth a close look.

Address:

<http://www.radiowrench.com/sonic/dubusvhfpreamp.pdf>

Falstad

It is difficult to imagine how the author has gathered so many articles for this web site. There is a surprise when open-

ing this site: innumerable Java animations (applets) on the most diverse topics from electronics and physics and more to be examined and tried out e.g. the antenna simulation with facilities for varying the parameters. In addition the link list is first class.

Address: <http://www.falstad.com/mathphysics.html>

Teslamania

From time to time you may need some entertainment (not only by the cinema) or have a view of another world. Therefore you should take a look at this site.

It is about Tesla currents and makes everything look very wild with metre electric arks hissing and sizzling in the air. It is not really relevant to electronic engineers but it is quite interesting.

Address: <http://capturedlightning.com/>

Spread Spectrum Scene / S Parameter tutorial

This is a favourite site to be visited from time to time. It is regularly extended and reorganised. The author found the fol-



lowing special article while searching for some S Parameter files for a new development. The tutorials are very useful.

Address: <http://sss-mag.com/spara.html>

Antenna Basics

Again and again it is interesting to read are such introductions, particularly if they are easy to read and understand. In addition this is a visible pdf file - an interesting read.

Address:

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Small Spacecraft Antenna Selection Tutorial

Particularly interesting for lovers of horns; not those used for music make, but those that can radiate and receive microwave signals. This is a good overview with an accurate technical background. This is a fascinating subject.

Address: <http://www.antdevco.com/ADC-0807201305%20-%20R0%20Small%20Spacecraft%20Antenna%20Selection%20Tutorial.pdf>

How radio Telescopes work

This title does is not completely fair because the contents clearly go beyond that. There is extra information under "Learn more" that is really interesting.

Address: <http://www.nrao.edu/index.php/learn/radioastronomy/radiotelescopes>

Scanned old books

Many interesting old books have been scanned; among other things the books of Sturley (about receiver engineering) are recommended for the fans of valve technology. According to the site there are about 1043Mb available.

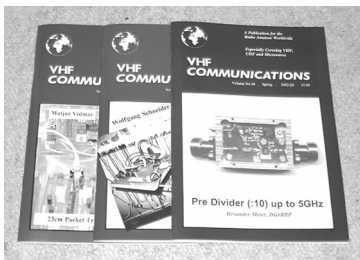
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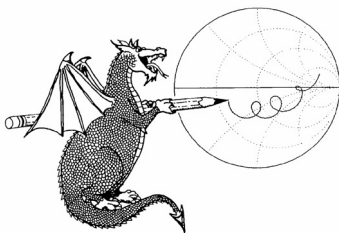


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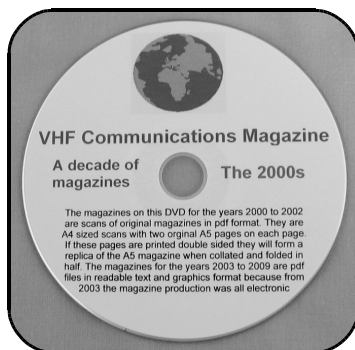


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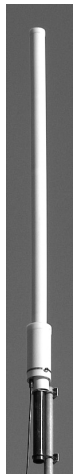
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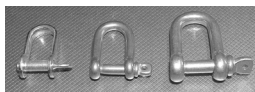
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ASPIS 6 — ASPIS 7
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COBRA 30
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