VHF
COMMUNICATIONS

76GHz sextupler and amplification
Sigurd Werner, DL9MFV

Contents

There are some very interesting articles in this issue including three from subscribers to VHF Communications Magazine, André Jamet, Primoz Lemut and John Fielding. I am very pleased to receive such articles and in fact I already have some waiting for publication in issue 4/2008 but I always want to receive more articles. If you have a project that you think is worth publishing, please contact me.

I have just finished updating the RSGB/ARRL International Microwave Handbook, it should be published before Christmas, so there is one thing to put on your present list.

The Internet Treasure Trove has some entries sent directly to me, if you spot a useful web site while surfing please send me an email and I can add it to the list.

73s - Andy

André Jamet, F9HX

The harmful effects of local oscillator noise

1

Introduction

The goal of any transmission system is to transfer information from point to point. It is correctly done when the received information is identical to the original one. In practice, the information can be distorted and/or mixed with disturbing signals. Noise of all transmission stages is the main troublemaker. At the limit, they can be so high that the ratio information/noise ratio is too low to get the information available.

2

Local oscillator

A local oscillator, LO, is composed of an oscillator eventually followed by mixers and/or frequency multipliers in order to obtain the desired frequency, because it is not possible or desirable to start with a very high frequency oscillator.

As with all electronic devices, an oscillator makes noise. As long as oscillators were based on transistor and crystal arrangement, their noise was acceptable. The use of lower Q resonators and the modern devices PLL and DDS are quickly showing their weakness where more demanding performances are required.

Noise level is an important limitating parameter for weak signals communications as it decreases the RX sensitivity and TX signal quality.

More, other shortcoming can upset weak signal reception when strong signals are on adjacent channels: it is called reciprocal mixing.

Those defects are restricting for both amateur and professional equipment.

3

Electronic circuits noise

Electronic circuits are composed of active and passive components. It is well known that any conductor generates noise by thermal agitation of electrons, called white noise as it extends from DC up to the highest frequencies.

The power of thermal noise is given by:

$$P_{noise} = k \, T \, B \text{ (watts)}$$

where:

k = Boltzmann's constant ($1.38.10^{-23}$ K/J)

T = absolute temperature (K = 273°C + ambient)

B = bandwidth (hertz)

At 27°C and one hertz of bandwidth, we have:

$$P_{noise} = 1.38 \times 10^{-23} \times 300$$
$$= 4.14 \times 10^{-21} \text{ (watt)}$$
$$= 4.14 \times 10^{-18} \text{ (mW)}$$
$$= 10 \log (4.14 \times 10^{-18})$$
$$= -174 \text{dBm/Hz}$$
(see Appendix)

Oscillators produce other kinds of noise. Most influential are:

- Shot noise occurs when there is a potential barrier, i.e. diodes and transistors junctions

- Flicker noise, called pink noise or 1/f noise as it is decreasing with the frequency up to a cut off frequency and remains constant after that. Metallic resistors produce the lowest and it is more prominent in FET and MOS transistors.

Powers of different noises add.

4

Oscillator noise

An oscillator is composed of an amplifier in conjunction with a high Q resonator. Oscillator behaviour survey comprises:

Whys: various noise voltages and currents are present together with normal DC and HF ones. They are due to various oscillator components, passive and active.

Wherefores: those voltages and currents modulate the normal ones by amplitude and phase modulation.

Mixer and other stages following the oscillator attenuate amplitude-modulated noise. Therefore, amplitude noise is generally treated as immeasurable and ne-

glected. If it were not, any residual AM noise would produce a familiar effect. On the contrary, phase-modulated noise produces less intuitive effects and requires more attention to be well understood. Even after saturated stages phase modulation remains. Moreover, the phase offset is multiplied by any multiplier stage.

Usually, we use the expression: phase noise.

5

Low Noise LO design

To design a low noise oscillator, we need to follow some requirements. It is the aim of this article, to minimise the phase noise, the advice of the KA2WEU/DJ2LR, Pr. Dr Ulrich Rohde check list [5,6] and F5RCT lecture [7] is:

- Maximise the unloaded resonator Q

- Maximise the resonator reactive energy

- Avoid oscillator saturation

- Choose an active device with the lowest possible noise figure at the actual working conditions, especially for flicker noise under 10kHz

- Use low frequency negative feedback to reduce transistor flicker noise

- The energy should be coupled from the resonator rather than another part of the oscillator

- Use high stability and low noise passive components. Be careful with low Q varicap diodes.

More, external reasons can add noise:

- DC supplies from linear and switching regulators
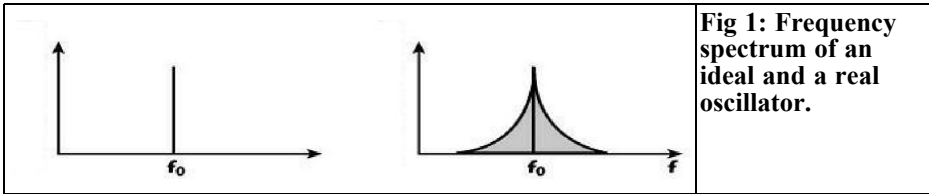
- Vibrations (fan) and mechanical

Fig 1: Frequency spectrum of an ideal and a real oscillator.

shocks (microphonic effect of resonators and some capacitors and coils)

- Digital circuits (computer).

6

Whole oscillator noise

When multipliers stages follow an oscillator the phase noise is increased by the multiplication factor. For an N factor, noise is degraded by 20 log N. In real world conditions, it is a minimum value increased by the multipliers stages own noise. Therefore, a choice should be made between:

- To start from a rather low frequency oscillator that requires a large multiplication

- To start from a high frequency oscillator.

On one hand, high stability, low retrace effect and low noise sources, such as OCXOs, caesium or rubidium generators, usually generate a 10MHz signal. A very large multiplication factor is required for the microwave bands. On the other hand, VHF OCXOs are able to give an acceptable solution if the drift and the retrace effect are cancelled by a permanent

power supply [8].

A 10MHz OCXO can be used as a reference for PLL and DDS devices [9]. A microwave PLL can be based on a DRO.

Those solutions avoid a large multiplication factor, but they are more difficult to achieve.

7

Real oscillator spectrum

Fig 1 shows the frequency spectrum of an ideal and real oscillators. The spectrum spread is due to the phase noise. This is common for analogue signals, like those from an LO. Fig 2 shows the time domain. In this case, the word jitter is used in digital applications.

From the Fig 1, we can see two sidebands around the central frequency called the carrier. Fig 3 shows the complete spectrum (DSB) and the half-sideband spectrum (SSB). As the phase noise spectrum is symmetric around the carrier, it is commonly characterised by the SSB spectrum. The symbol is $L_{\varphi}(f_m)$. It represents the ratio of the SSB noise power in a one hertz bandwidth centred at f_m hertz away from the carrier, to the carrier power. It is called SSB spectral density;
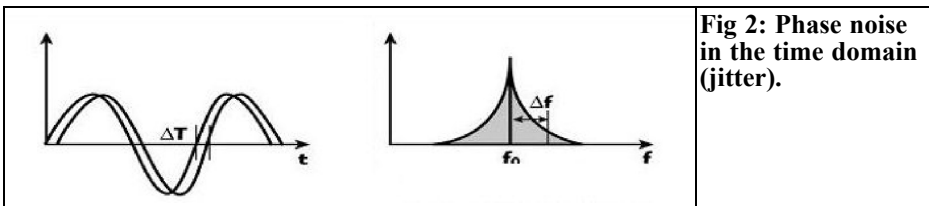

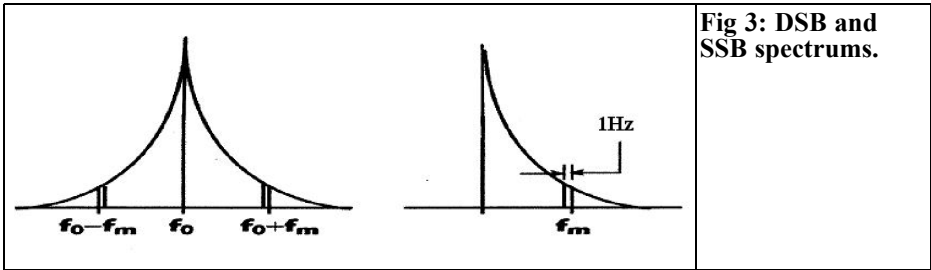**Fig 2: Phase noise in the time domain (jitter).**

Fig 3: DSB and SSB spectrums.

the units are dBc/Hz.

If we assume that the phase noise is approximately constant over the bandwidth of interest, at a distance f_m from the carrier and for a bandwidth B, noise power is given by:

$$P_{\text{in B hertz}} = B\, L_\varphi(f_m)\, P_c \quad \text{(watt)}$$

where P_c is the carrier power.

It is generally more convenient to work in dBs. Then:

$$P_{\text{in B hertz}} = 10 \log(B) + L_\varphi(f_m) + P_c \quad \text{(dBm)}$$

That offset (f_m) can be 10Hz, 100Hz, 1kHz, 10kHz or 100kHz.

Spectral density can be calculated from Leeson [1] who established a formula using the oscillator's parameters. A number of surveys [2,3,4] added corrections and complements.

Fig 4 shows the general look of a calculated curve. We can see several regions with different slopes. Near the carrier, up to a few tens of hertz, the slope is following an f^3 law, so a -30 dB/decade slope. Within the kilohertz region, the law is f^2 equals to -20 dB/decade. Then, follows the thermal noise region with a constant value.

A more detailed analysis shows other regions in f^4 and f^1. After reading of several surveys, it is questionable, so the above is mentioned.

8

LO noise measurement

To measure an LO noise spectral density, we need equipment able to handle megahertz and even gigahertz signals. In general, its own noise near the carrier would be at the same level or even more than we want to measure.
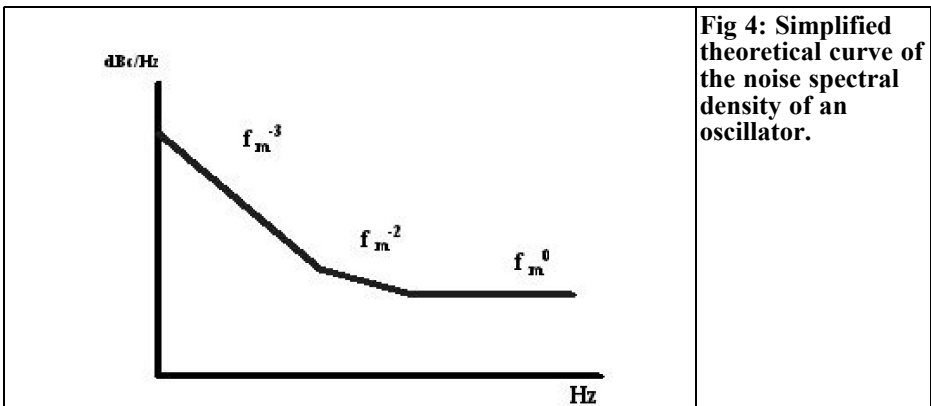

Fig 4: Simplified theoretical curve of the noise spectral density of an oscillator.
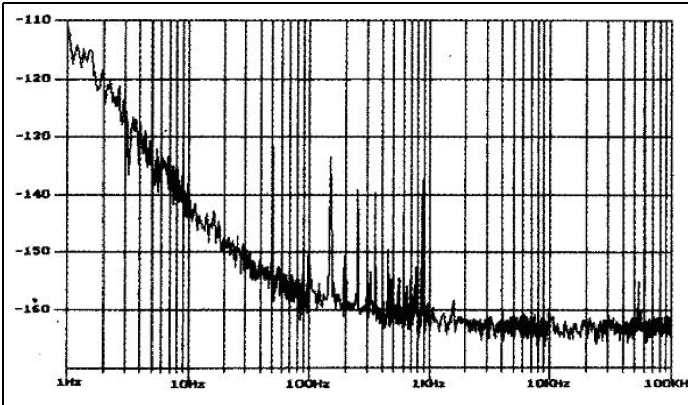
Fig 5: Spectral density of a very low noise OCXO spoiled by a poor power supply.

Specialised equipment is not common for a radio amateur. For example, the E 5052 from Agilent cost is $85,000! The Aeroflex PN 9000 is able to do the test at a quite high cost.

An affordable method is to use a very clean auxiliary signal, a detector or a phase discriminator to get a signal acceptable for use with standard equipment

Fig 5 shows the spectral density of a very low noise OCXO. Nevertheless, we can see large undesirable spurious signals between 100 and 1000Hz. They come from the power supply (main frequency harmonics and variations).

However, we can show a poor quality oscillators spectrum with a common spectrum analyser. That will be detailed later with added noise oscillators. Moreover, we can listen an oscillator signal with a receiver to "hear" its noise but not

Fig 6: Narrow frequency modulation device.

be able to quantify the noise. That is very useful for UHF and microwaves.

9

Test of deliberately added noise oscillators

It is useful to know the effect of a noise polluting a sine wave signal.

To test amplitude modulation effect, a 1000Hz signal from an audio generator is mixed using a dual gate FET (BF 988) with noise. The noise is produced by a device comprising of a 78L09 regulator followed by two stages of amplification [11]. On an oscilloscope, we can distinctly see the noise above 1% modulation. With headphones, we can hear much lower levels.

To test phase modulation, an 8038 deliv-
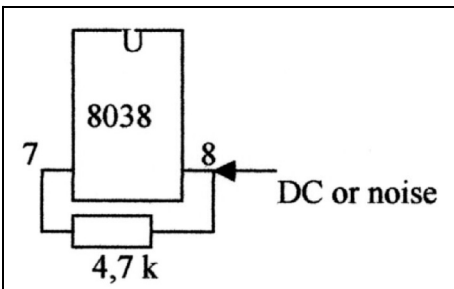
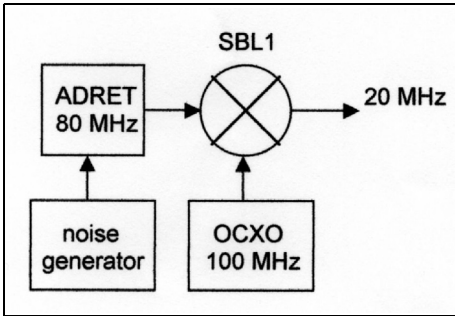| Table 1: Fig 6 calibration.			
F	U_8	Δf (Hz)	U_{dc} (mV)
1039	9,600	-23	50
1058	9,560	-4	10
1062	9,550	0	0
1080	9,510	18	-40
1103	9,460	41	-90

Fig 7: HF device to produce a noisy HF signal.

ers a 100Hz signal with a narrow frequency modulation (NBFM produces the same effect as phase modulation). Fig 6 shows the block diagram. First, the device is calibrated with a DC signal (see Table 1). We have:

$$\Delta_f / \Delta_{Udc} \text{ (average)} = 1Hz / 2.2 \text{ mV}_{dc}$$

NB: there is no amplitude modulation, as the amplitude remains constant.

The noise generator already described is used with this modulation. The threshold is 40mVpp that is to say about 18Hz or a 1.8% frequency variation.

NB : the audio effect is not the same when the signal is amplitude or phase modulated by the noise. It is difficult to explain this! It seems deeper and more disturbing even when the bandwidths are

Fig 8: Original 20MHz spectrum.

the same.

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TX signal purity versus LO noise

LO noise spoils TX signals even when there is no mixer, only frequency multipliers. The received signal spreads, as a large part of the power is lost in undesirable sidebands. The noise appears around the received carrier. It is really a jammer!

F1GHB and F5EFD "listened" to two LOs at 10GHz. One is an OCXO followed by multipliers. Its measured noise is about -90dBc/Hz. The other is a synthesiser based on a LMX 2326 having noise of about -63dBc/Hz. With the first one, nothing was wrong but audible noise was heard on the second.

Radio amateurs working the microwave bands above 76GHz know how difficult it is to obtain a pure note. SSB and CW can be upset by the phase noise. WA1ZMS has made a presentation at MUD 2004 [10]; we can "hear" noise of several oscillators at 241GHz.

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Noisy HF TX

To known how noise upsets a HF TX, the device in Fig 7 is able to produce a noisy 20MHz signal.

This device comprises a HF generator (ADRET 6315) to deliver an 80MHz signal. The noise generator can modulate that generator. A mixer mixes the output signal and a 100MHz produced by an OCXO. A spectrum analyser shows the output at 20MHz. Scan is one kHz/division, bandwidth 100Hz and scale 10dB/division.
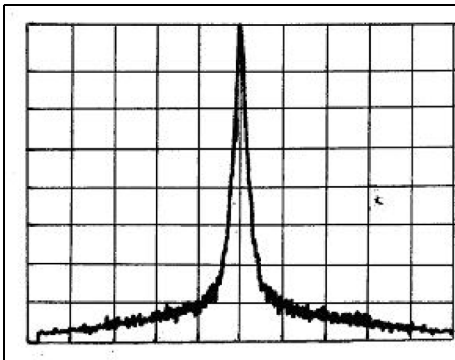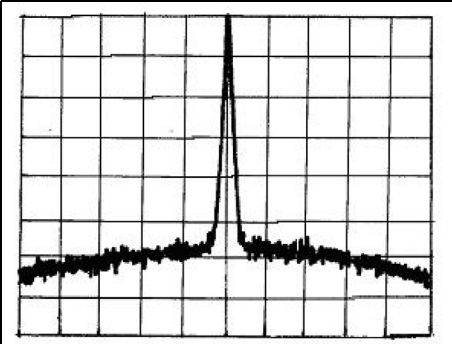
Fig 9: 20MHz spectrum with added noise.

Figs 8 and 9 show the 20MHz spectrums before and after noise is added. The floor noise, noise without signal, increases by more than 15dB. With a RX in AM mode, we can hear the noise on each side of the tuning frequency, as we hear a NBFM signal. In NBFM mode, noise is at the centre frequency. In SSB mode, we can find the noise on one side. For those three cases, listening seems not to be affected by this noise level.

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Deliberately noisy microwave signal

Fig 10 shows the equipment. It starts with a XO at 108MHz, phase modulated by a varicap. The noise generator is as above. Several multipliers stages follow to produce a 10,368MHz signal. Because of the large multiplication factor and the heavy saturation, the output signal is easily phase modulated without noticeable AM.

With a 10GHz RX, we can hear the sound effect. In SSB mode without added noise, tuning gives the usual tone variation, low to high frequency. With an increase added noise the tone become "rougher and rougher"

My "poor mans lab" does not allow me to quantify precisely that effect. Nevertheless, my old 141T spectrum analyser with an 18GHz plug-in is able to show me the spectrum variations during the test. The floor noise of that old spectrum analyser is -110dBm for a 100Hz bandwidth in the 8.23 to 14.35GHz range. Figs 11 and 12 clearly show the noise effect. We can see that the reports by F1GHB and F5EFD are roughly confirmed. A 20dB of noise increase changes a signal from correct to unacceptable.

By interpolation, we have approximately: Fig 11

- -50dBc at 1kHz offset; then:

- -50 -log 100 (B in Hz) = -70dBc/Hz

Fig 12

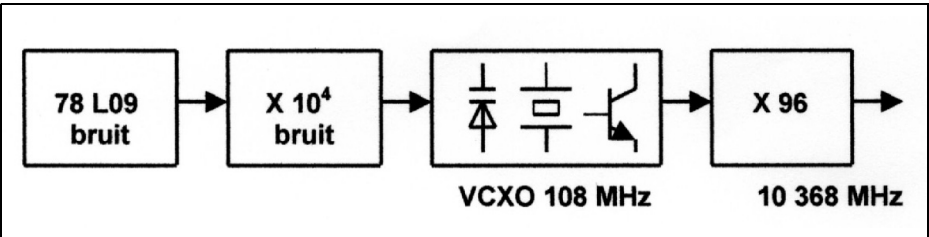- -30dBc at 1kHz offset; then:

- -30 -log 100 = -50dBc/Hz

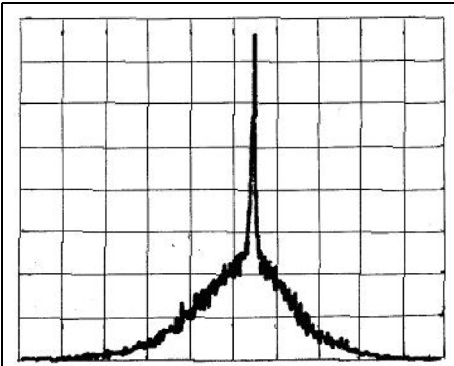**Fig 10: Equipment to produce a noisy microwave signal.**

Fig 11: Correct 10,368MHz spectrum, BW=100Hz, scan=5kHz/division, scale=10dB/division.
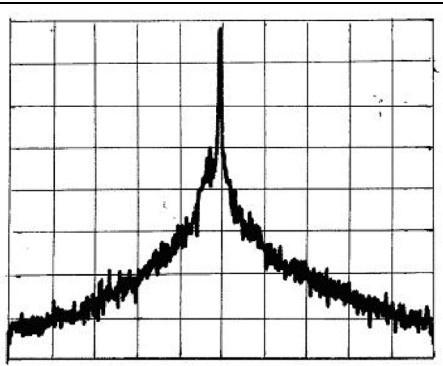
Fig 12: Noisy 10.368MHz spectrum, BW=100Hz, sacn-5kHz/division scale=10dB/division.

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Noise LO effect on a RX sensitivity

The 144MHz transceiver [F9HX 12] is based on an IQ mixer and a zero intermediate frequency. A 24MHz VFO with a frequency variation by a varicap is followed by multipliers stages to get an LO output at 144MHz. Figs 13 and 14 show the 144MHz spectrum before and after deliberately added noise.

Those spectrums are displayed with a 100Hz bandwidth, a 1kHz/division scan and 10dB/division. We obtain approximately:

Before added noise:

- -80dBc at 1kHz offset; then:

- -80 - log 100 = -100dBc/Hz

After added noise:

- -40dBc/Hz at 1kHz offset; then:

- -40 - log 100 = -60dBc/Hz

In the last case, in SSB mode, signals are heavily spoiled. Weak signals are unintelligible.
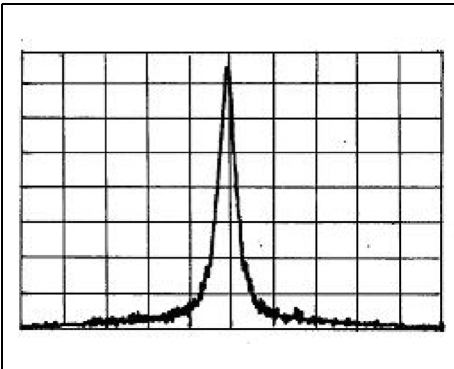
Fig 13: Correct 144.300MHz spectrum.

Fig 14: Noisy 144.300MHz spectrum.

Fig 15: Reciprocal mixing.

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Reciprocal mixing

That is the second LO noise effect. It limits the low signals reception in the presence of loud adjacent signals (Fig 15). When the LO noise mixes with an interfering signal, a signal is produced in the IF bandwidth B. It can perturb or even damps it.
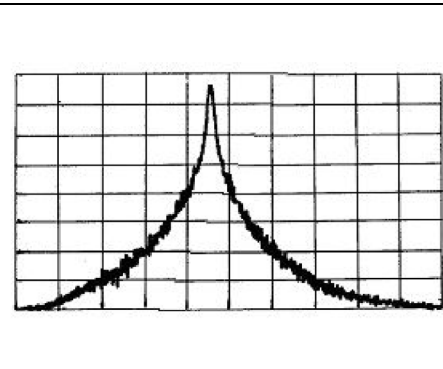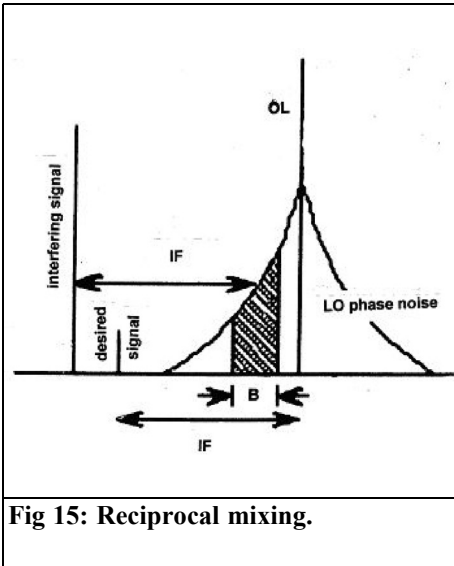
This effect is very awkward on the HF and VHF bands during the contests. A loud station can be near a weak one. The RX seems to go "deaf".

From [13 and others], we can calculate the reciprocal mixing effect. For example, a 2m band RX with a 2700Hz bandwidth 9MHz IF and a -90dBc/Hz noise LO. It is tuned for a S3 = -129dBm signal. An S9+20dB = -73dBm interfering station is 10kHz away.

Approximate calculation gives:

- LO noise power in 270Hz: - 90 + 10 log 2700 = -90 + 34 = -56dBm

- Mixer output from LO noise + interfering signal = -73 - 56 = -129dBm.

We notice that the noise level produced by the interfering signal equals the desired signal.

A test verifies that calculation. Two RF generators feed simultaneously a RX while the LO is added noise or not. Without, a weak signal is audible; with, it is almost lost in noise.

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Conclusion

As we can see, LO noise is an important parameter for communication equipments. In the HF and VHF bands, it can upset weak signals reception by loud adjacent signals; that is currently the case during contests.

For very low signals communications, it is a limiting factor. Microwaves are very demanding owing to the very large multiplication factor in LO.

For any more information contact: F9HX agit@wanadoo.fr

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Appendix

Thermal noise comprises two orthogonal components not correlated: amplitude and phase. Each is half of the total. Consequently, from [18]:

$$P_{phase\ noise} = P_{amplitude\ noise} = -174 - 3 = -177dBm/Hz$$

Nevertheless, -174dBm/Hz remains the currently used value.

17

References

It is easy to find tens of references on the Internet, professional and amateur topics. Here is an extract of those I read and assimilated.

[1] A Simple Model of Feedback Oscillator Noise Spectrum, D.B. Leeson, Proceedings of the IEEE, February 1966

[2] Oscillator Phase Noise: Theory and Prediction, K.V. Puglia, Microwave Journal, 9/2007

[3] All About Noise in Oscillators, Ulrich L. Rohde, KA2WU/DJ2LR/HB9AWE, QEX 12/1993, 1/1994, 2/1994

[4] Oscillators: A New Look at an Old Model, Stan Alechno, Microwave & RF, 12/2001, 1/2003

[5] Comprendre les performances des récepteurs radio, F4BUC, Radio-REF juin 2003

[6] Phase Noise :Theory versus Practicality, John Esterline, Microwave & RF, April 2008

[7] Le bruit de phase, F5RCT, REF 67, 3/2007

[8] Difficultés de pilotage par OCXO en SHF, F9HX, Radio-REF 12/2004

[9] Un pilote VHF stabilisé par PLL pour votre transverter SHF, F9HX, Radio-REF 11/2003

[10] Millimetre-Wave LO References & Phase Considerations, WA1ZMS, Microwave Update 2004

[11] Le bruit des régulateurs de tensions continues, F9HX, Radio-REF 3/2008

[12] 2 m Direct Conversion Transceiver, F9HX,VHF-Communications 1/2003

[13] Low-Noise VHF-Oscillator with Diode Tuning, DJ7VY, VHF Communications 2/81

[14] Theory of Intermodulation and Reciprocal Mixing: Practice, Definitions and Measurements in Devices and Systems, Ulrich L. Rohde KA2WEU/DJ2LR/HB9AWE, Part 1, QEX 11-12/2002

[15] Local Oscillator Phase Noise and its Effect on Receiver Performance, C. John Grebenkemper, Watkins-Johnson Company, Tech-notes 11/12/81

[16] Noise Reference, Applied Radio Labs, http://www.radiolab.com.au

[17] The Receiver Noise Equation: A Method for System Level Design on a RF Receiver, Hyung JounYoo & Ji-Hoon Kim, Microwave Journal, 8/2002

[18] Additive (Residual) Phase Noise Measurements of Amplifiers, Frequency Dividers and Frequency Multipliers, Jason Breitbarth & Joe Koebel, Microwave Journal, June 2008

Gunthard Kraus, DG8GB

Practical project: Noise factor measurements with older spectrum analysers, part 2

Continued from 1/2008

4.2. The preamplifier

4.2.1. Starting the development

MMICs with at least 20dB gain and an upper cut-off frequency of at least 2.5GHz are used in the amplifier stages. This means that different interesting ranges can be covered like "government bands", "Meteosat" or the 13cm amateur band but not the lower amateur radio frequencies. The amplifier must be absolutely stable and oscillation free, which can surely be achieved using simulations.

Three such building blocks can be connected in series in a long enclosure to realise the minimum gain of 60dB required. Extensive precautions must be employed to suppress the inevitable self-oscillation.

Finally the Agilent ABA-52563 broadband amplifier was chosen, the data sheet begins:

Agilent ABA-52563: 3.5GHz Broadband silicone RFIC Amplifier

Features:

- Operating frequency: DC ~ 3.5GHz

- 21.5dB gain

- VSWR < 2.0 throughout operating frequency

- 9.8dBm outputs P1dB

- 3.3dB noise figure

- Unconditionally stable

- Single 5V supply (ID = 35mA)

4.2.2. Development of the single stage amplifier prototype

The S parameters for the ABA-52563 are required, but unfortunately they are only printed as table in the data sheet. Thus they must be copied from the table, deleting unnecessary data, into the required S2P form as "ABA52563. s2p " in the project directory, this is shown in Table 1.

The result of the first PUFF simulation, only using the S parameters above, is shown in Fig 3 (it is clearer and faster with PUFF rather than with the Ansoft designer student version). The result looks good in the intended frequency range from 500 to 2500MHz; S21 is approximately constant at 21db, S11 and S22 at -15dB.

Fig 4 shows the practical circuit diagram on the left. It has extensive decoupling of the supply to prevent noise, particularly low frequencies on the supply line (e.g. self-noise of the power supply) from reaching the RF output via the 33nH choke. The right hand side of Fig 4 shows the simulation diagram that the CAD programs (PUFF or Ansoft designer student version) are content.

Table 1: The S parameters for the MMIC ABA 52563 from Agilent.

! ABA-52563 S PARAMETERS

ghz s ma r 50

0.05	0.01	146.6	12.10	-2.6	0.03	0.3	0.15	-2.4
0.10	0.01	134.0	12.11	-4.8	0.03	-0.3	0.15	-5.1
0.20	0.01	-40.6	12.16	-9.6	0.03	0.1	0.15	-9.6
0.30	0.01	-53.2	12.19	-14.5	0.03	1.2	0.15	-13.0
0.40	0.02	-56.7	12.19	-19.5	0.03	2.4	0.14	-15.7
0.50	0.03	-141.5	12.26	-24.8	0.03	1.0	0.15	-15.7
0.60	0.03	-128.1	12.24	-29.8	0.03	3.1	0.15	-17.6
0.70	0.04	-127.5	12.21	-34.9	0.03	4.3	0.15	-20.3
0.80	0.04	-126.7	12.18	-39.8	0.03	6.1	0.15	-22.5
0.90	0.05	-123.9	12.16	-44.7	0.03	7.4	0.15	-24.2
1.00	0.05	-125.0	12.13	-49.7	0.03	11.7	0.15	-26.4
1.20	0.05	-123.4	12.10	-59.6	0.03	10.8	0.15	-29.4
1.40	0.06	-127.4	12.05	-69.4	0.03	12.4	0.15	-32.4
1.60	0.06	-133.8	12.04	-79.6	0.03	13.0	0.15	-35.3
1.80	0.06	-136.7	12.00	-89.8	0.04	14.7	0.15	-37.8
2.00	0.07	-142.5	11.94	-100.4	0.03	14.3	0.15	-38.3
2.20	0.07	-143.9	11.87	-111.2	0.04	16.7	0.15	-37.8
2.40	0.08	-146.1	11.75	-121.9	0.04	16.2	0.15	-37.3
2.60	0.09	-148.4	11.56	-133.2	0.04	17.3	0.14	-36.9
2.80	0.09	-149.5	11.33	-144.5	0.04	15.6	0.14 -	36.4
3.00	0.10	-152.7	10.95	-156.1	0.04	15.8	0.13	-35.9
3.20	0.10	-158.7	10.51	-167.5	0.04	15.6	0.13	-35.4
3.40	0.11	-163.2	9.97	-178.7	0.04	15.5	0.13	-34.9
3.50	0.11	-167.6	9.67	175.9	0.05	16.0	0.13	-34.6
4.00	0.12	165.9	8.25	150.6	0.05	12.0	0.13	-33.4
4.50	0.16	138.3	6.98	126.3	0.05	12.7	0.14	-37.1
5.00	0.19	122.8	5.71	105.0	0.06	9.5	0.12	-48.4
5.50	0.25	112.3	4.85	86.7	0.07	6.0	0.12	-63.0
6.00	0.30	99.3	4.14	70.4	0.07	1.0	0.11	-83.5
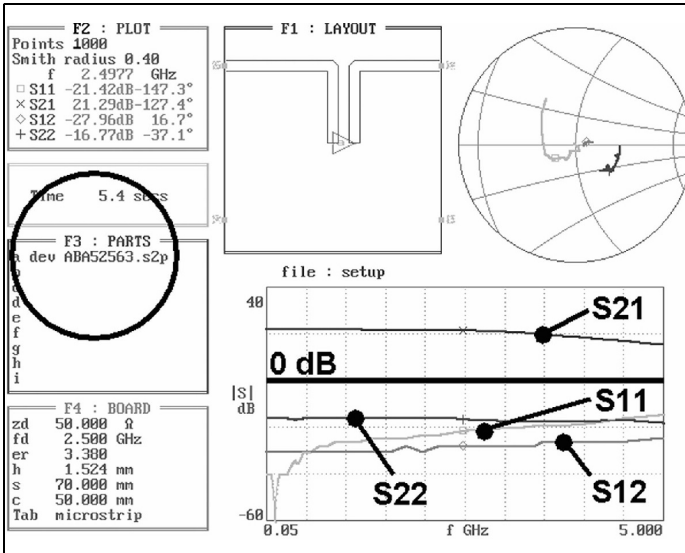
Fig 3: All S-parameters of the ABA52563 showing their dependence on the frequency.

Fig 4:The circuit diagram and simulation diagram for a single stage amplifier.
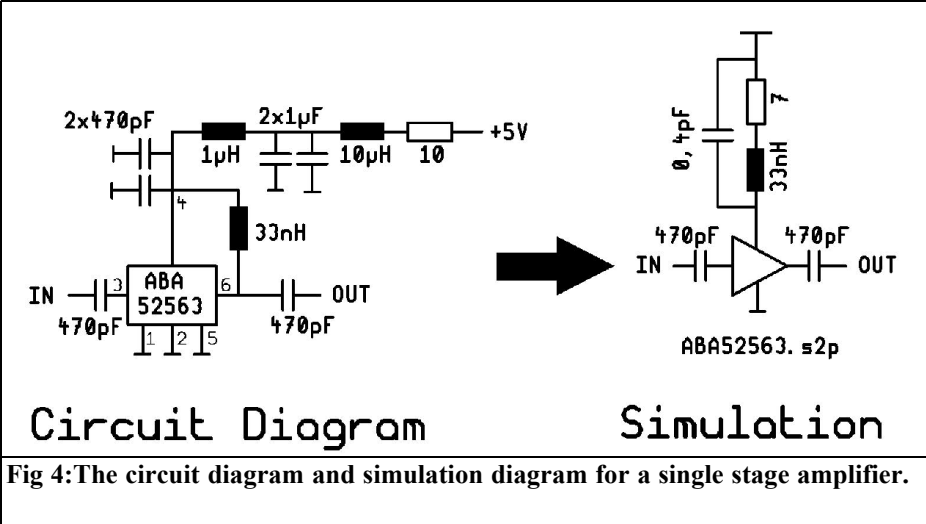
The choke was purchased as a finished 0805 SMD component and it should be represented correctly in the simulation with its self-resonant frequency of 1.5GHz and Q of 30 at 1GHz. PUFF requires some preliminary work to generate a replacement equivalent circuit (series resistance = 7Ω, parallel capacitance = 0.4pF). Ansoft designer student version is much easier to use: there is the construction unit of "chip Inductor" with adjustment possibilities for the individual data and even three different types of representation to select. This is a further
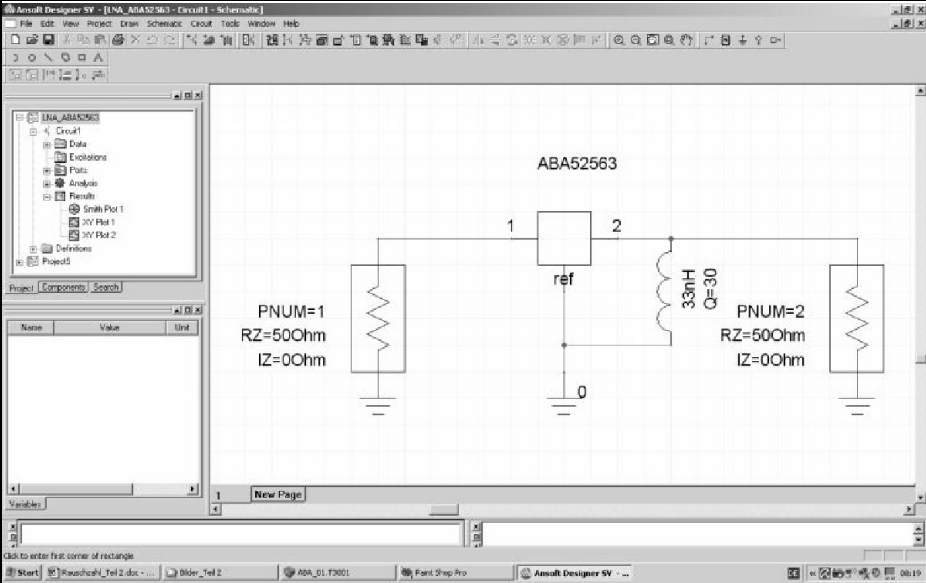

Fig 5: Ansoft designer simulation, the 33nH choke uses the Ansoft model.
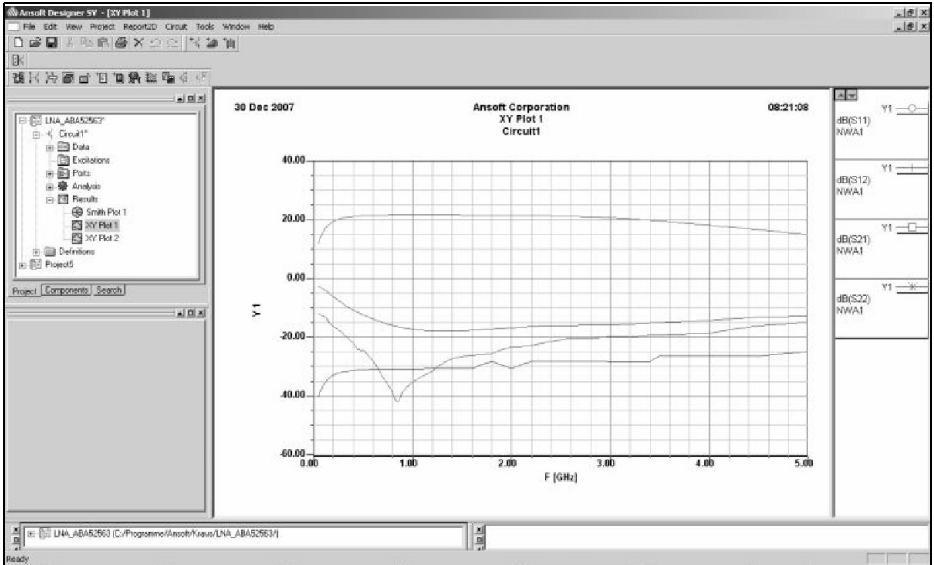
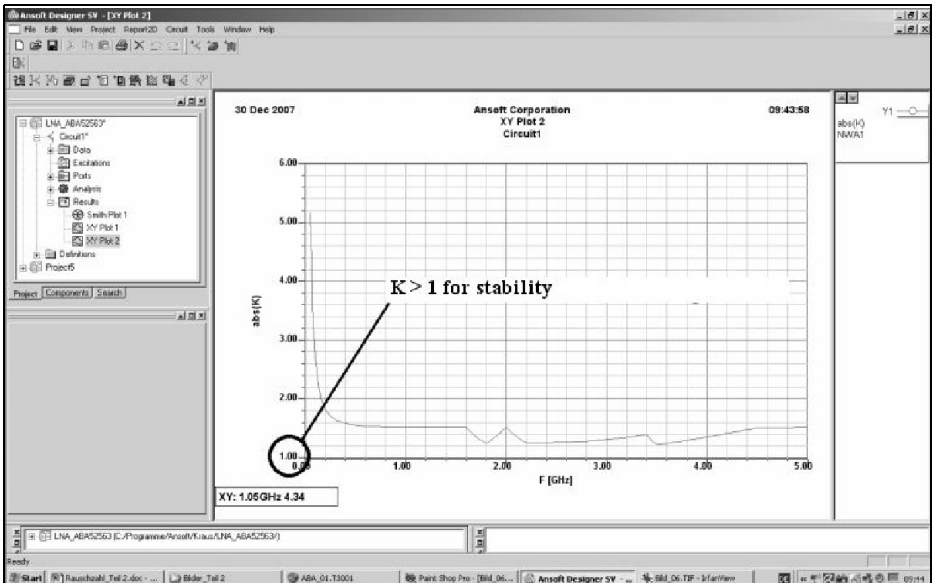**Fig 6: The S parameter simulation of the circuit with a realistic model of the choke shows no surprises.**

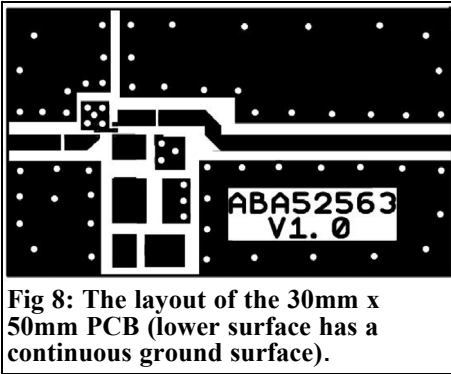**Fig 7: There are no problems with stability!**

Fig 8: The layout of the 30mm x 50mm PCB (lower surface has a continuous ground surface).

reason to only use PUFF as an "RF-pocket calculator" for fast derivation of fundamental solutions and to make the detailed analysis or finished development of the circuit (up to the gain simulation and stability control) with the Ansoft designer. Therefore only Ansoft designer was used for further development. Fig 5 shows the editor screen for the simulation of the circuit from Fig 4.

The simulation of the S parameters for the intended frequency range from 500MHz to 2500MHz is shown in Fig 6. As expected there are no bad surprises, the fall off in gain below 200MHz is inevitable because of the decreasing impedance of the choke used.

Following a successful check on the stability of the amplifier (Fig 7) the first attempt to design the PCB was carried out using double-sided 32MIL (0.82mm) thick Rogers RO4003 material. The layout for a 30mm x 50mm PCB is shown in Fig 8. Who is surprised at the apparently unnecessary length of the PCB in relation to the small space required of the circuit? This is one of the precautionary measures to prevent self-oscillation when three amplifiers are connected in series on the same PCB.

There is another topic to consider regarding stability. Those who have already developed circuits using the ABA 2563 will know that the influence of the plated through holes around pins 1, 2 and 5 is extremely critical. If their self-inductance is not kept as small as possible there are substantial problems. Fig 9 shows the
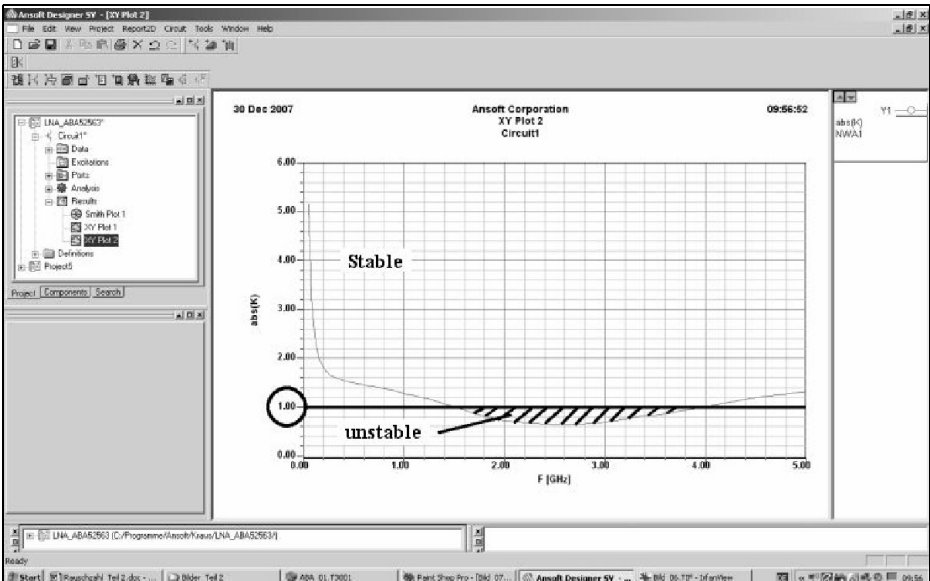

Fig 9: With 0.2nH added for the plated-through hole inductance the circuit becomes useless.
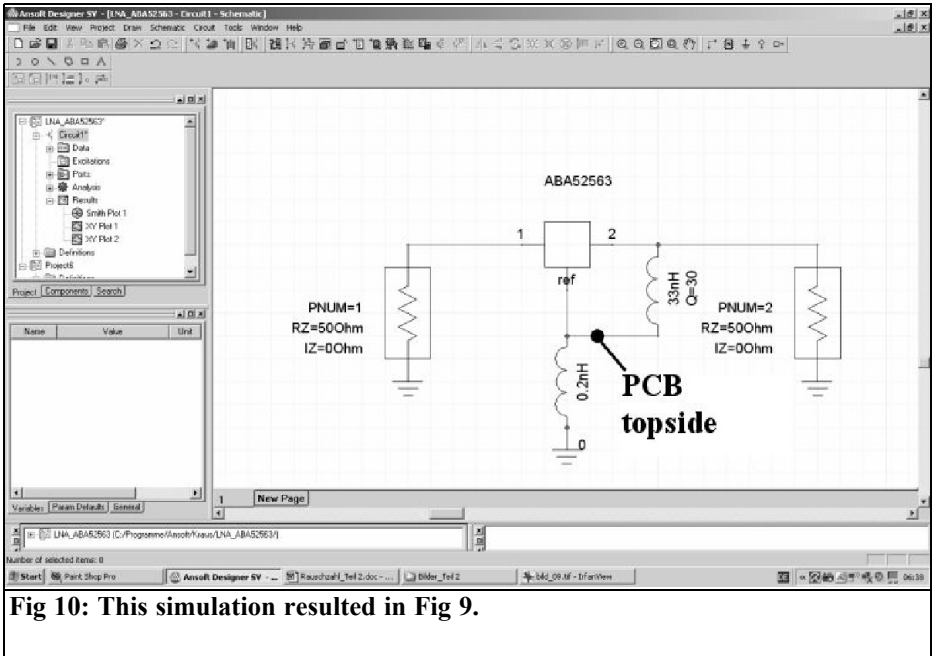
Fig 10: This simulation resulted in Fig 9.

simulation result for "k" with an inductance only 0.2nH between the earth references of the MMIC (pins 1, 2 and 5) and the ground surface. The simulation circuit is shown in Fig 10. It should be noted that the lower end of chip inductor on the output is connected to the top side of the PCB which is grounded (with 2 high-quality SMD ceramic capacitors

Fig 11: This is the technique to reduce the unwanted plated-through hole inductance.

joined in parallel) and therefore must be attached to pin 1, 2. The simulation result shows at the very small value of 0.2nH plated-through hole inductance the circuit could not be used because it would oscillate! Fig 11 shows that by enlarging the number of plated-through holes in the PCB layout it is possible to reduce the self-oscillating possibility below the danger level. So, when placing these holes as close as possible to the pins they are in parallel and the total inductance is reduced dramatically. Finally it is just as important to isolate the ground surface on the topside into an input and an output area. Both are connected by a sufficient number of plated-through holes to the continuous ground surface on the bottom side of the PCB.
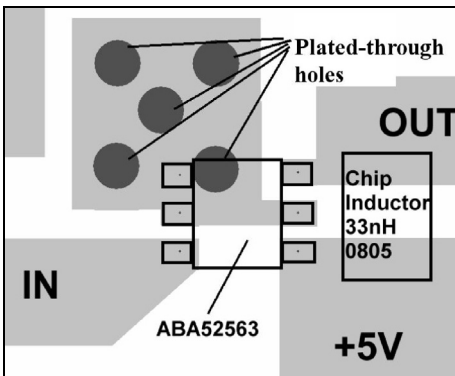
Fig 12 shows the circuit built into a milled aluminium enclosure, it is easy to locate the positions of the different components of the circuit diagram (please look carefully at the supply voltage which is fed carefully, using a coaxial cable to prevent any stray effects, to the
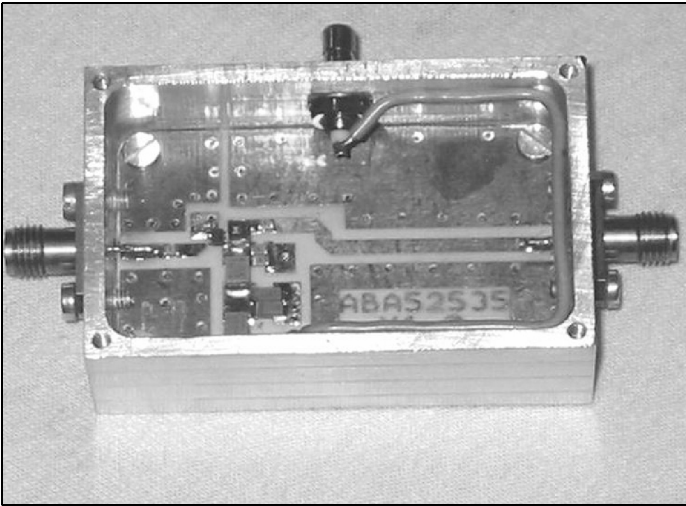
Fig 12: The circuit ready for operation, built into a milled aluminium enclosure. The SMB socket for the voltage supply is on the long side.

SMB socket on the long of the enclosure side). The results of measurements on the prototype are exciting; these are shown in Figs 13 to 16. It is very pleasing that with the network analyser (HP8410 with S parameter plug-in HP8746B for 0.5 to 12.4GHz and sweep oscillator HP8690 with HP8699 plug-in for 0.1 to 4GHz) the forecasts can be confirmed so beautifully. The reverse path was also taken: several cases with different plated-through hole inductances were simulated and showed on the basis of the measured S21 value that a value of 0.07nH gives the best agreement with the measure-ments.

In order to find the causes for the unexpected response at the high frequency end the simulation would have to be substantially extended and refined. Obvious periodic resonances are happening. Genuine micro strip lines at the input and output, including their effects, have not been considered. Nevertheless; the basic findings of the simulation and measurement agree also the circuit remains stable, therefore the actual development target i.e. the large amplifier, can continue.
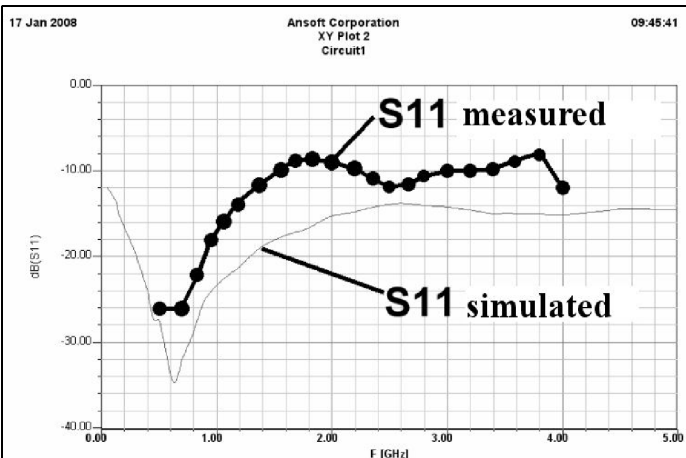

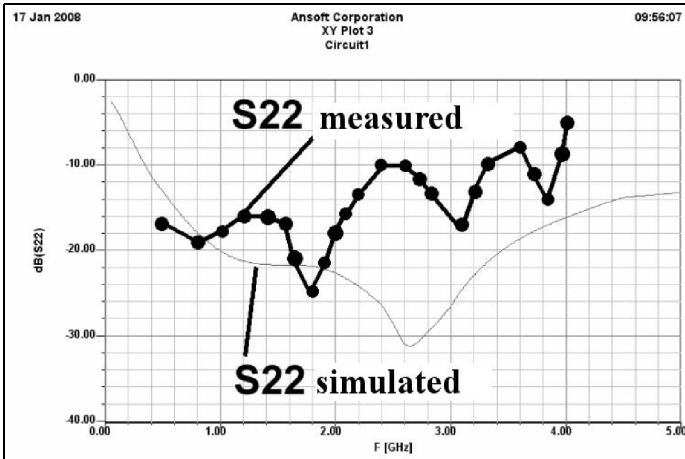**Fig 13: The measured and simulated result for reflection S11.**

Fig 14: The measured and simulated result for output S22.

4.2.3. The three stage amplifier

4.2.3.1. The way to the goal

Now it becomes exciting; how does series connection of three such circuits in a single aluminium enclosure behave? As a precaution the complete arrangement was simulated in order to guarantee that the interaction of each stage (expressed by its parameter S12) does not become unstable. Because if it were unstable there would not be much sense in further development and an alternative would be sought.

The PCB used to connect the individual stages using 50Ω microstrip is shown in Fig 17. The simulation circuit shown in Fig 18 gives the S parameter simulation shown in Fig 19 and the stability simulation shown in Fig 20. Everything looks good and it now depends on: the coupling factors in the enclosure, the mutual shielding of the individual stages, the correct arrangement of the ground surfaces and their plated-through holes as well as the decoupling of the three stages power supplies whether this large amplifier self-oscillates!

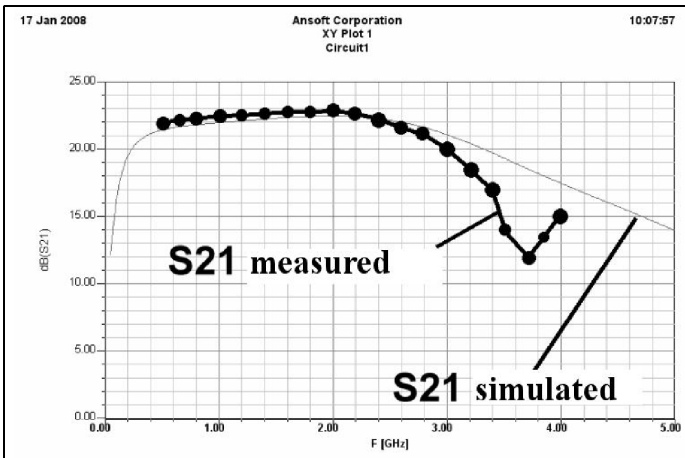The first prototype was made and screwed into the enclosure (Fig 21). A 50Ω resistor was connected to the SMA

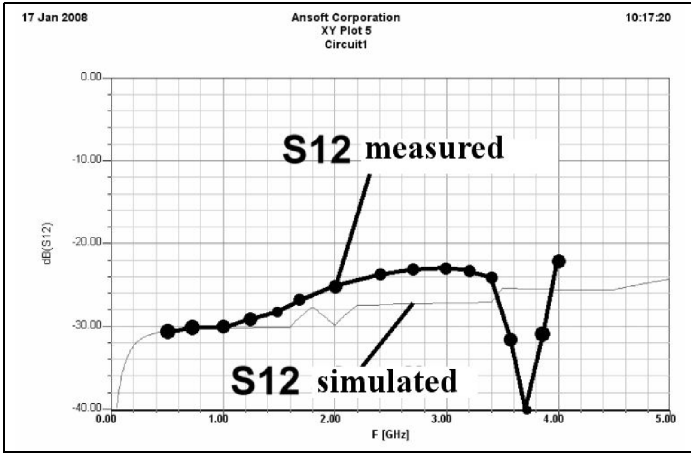**Fig 15: The measured and simulated result for S21.**

147

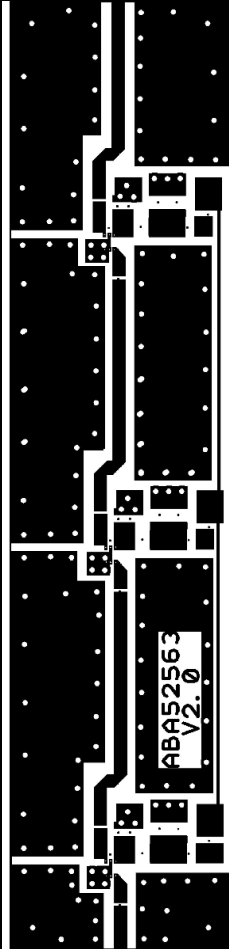Fig 16: The measured and simulated result for S12.

Fig 17: With a length of the new PCB of 130mm and width of 30mm makes the enclosure beautifully long!

input socket with a spectrum analyser connected to the output and then the amplifier examined under different conditions. Here are the results:

• If the aluminium cover of the enclosure is omitted, the circuit does not oscillate, but there was a strong peak in the output noise at approximately 1.5GHz .

• Presenting the blank aluminium cover was sufficient to bring the circuit to wild oscillation.

• If absorbing foam material was stuck onto the inside of the cover the amplifier looked nearly perfect; the oscillation was completely gone and the frequency response of the noise on the screen was nearly even. Three different materials were tried but the jet-black material used for the packing ESD sensitive semiconductors gave very good results.

• Finally a professional microwave absorption material from Cummings was tried. This is designed for use up to 24GHz and was stuck on the inside of the cover (thanks to Dieter, DL6AGC, for providing a sample). The oscillation and the instability in the frequency response had almost disappeared. As a suspicious person, the 50Ω resistor was removed from
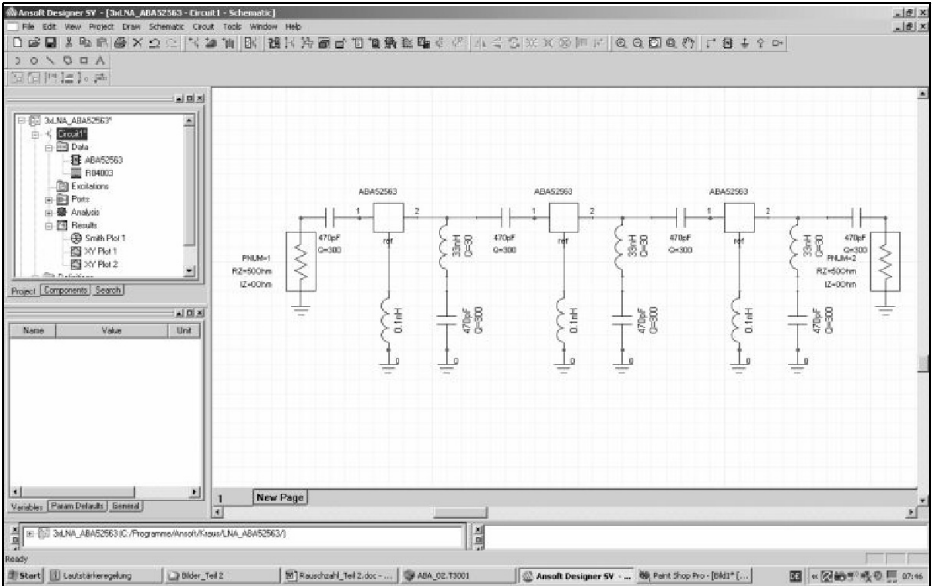
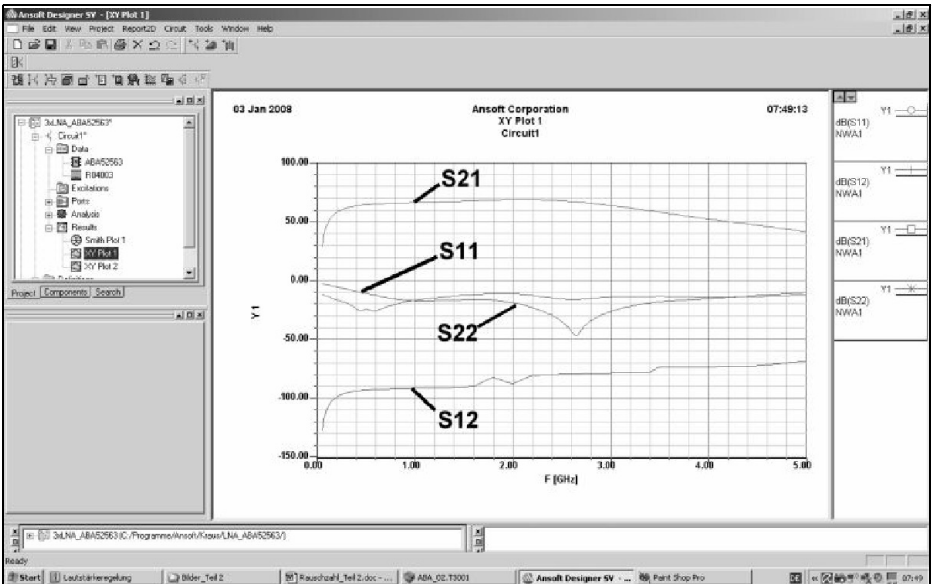**Fig 18: Simulation circuit for the three stage amplifier.**

Fig 19: The S parameters for the three stage version simulation are completely normally.
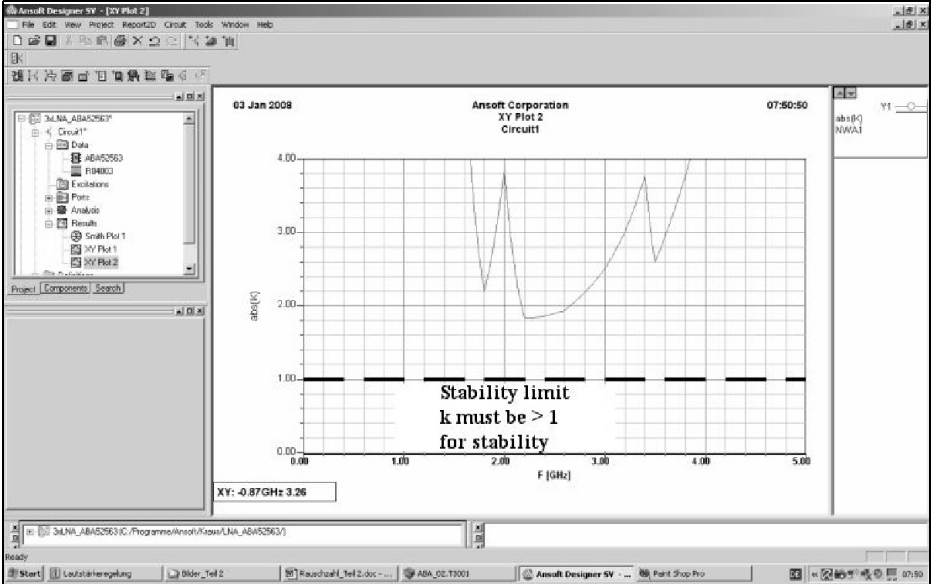
Fig 20: The simulation of stability for the three stage amplifier.

the input. A 1 – 2db rise in the noise at 1.5GHz resulted. In order to cure that completely a 1dB SMD attenuator can be permanently fitted to the input, however the self-noise figure of the amplifier increases by 1dB.

In order to verify the own measurements, a modern Agilent network analyser was used to measure the S parameters at a friendly company. Due to the high gain of over 60dB a 20dB SMD attenuator had to be attached to the input of the test specimen in order to avoid overloading the measuring input. Fig 22 shows the result, the curves for S21 and S22 supply the following information:

- The gain is expressed by the parameter S21 with the scale from +37 to 47dB on the left axis (20dB must be added because of the attenuator)
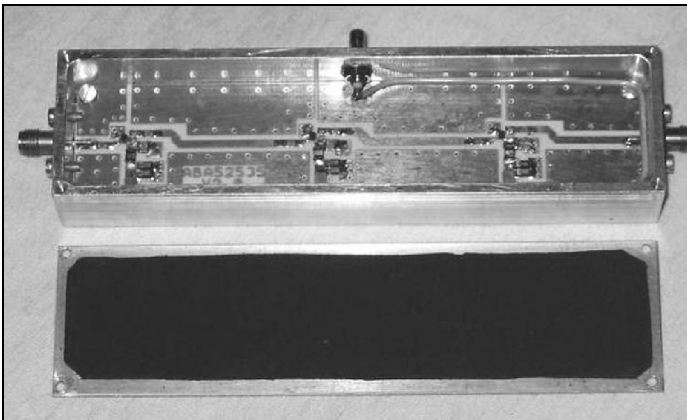

Fig 21: The three stage amplifier disappears in the enclosure if the cover pasted with absorption material is screwed on.
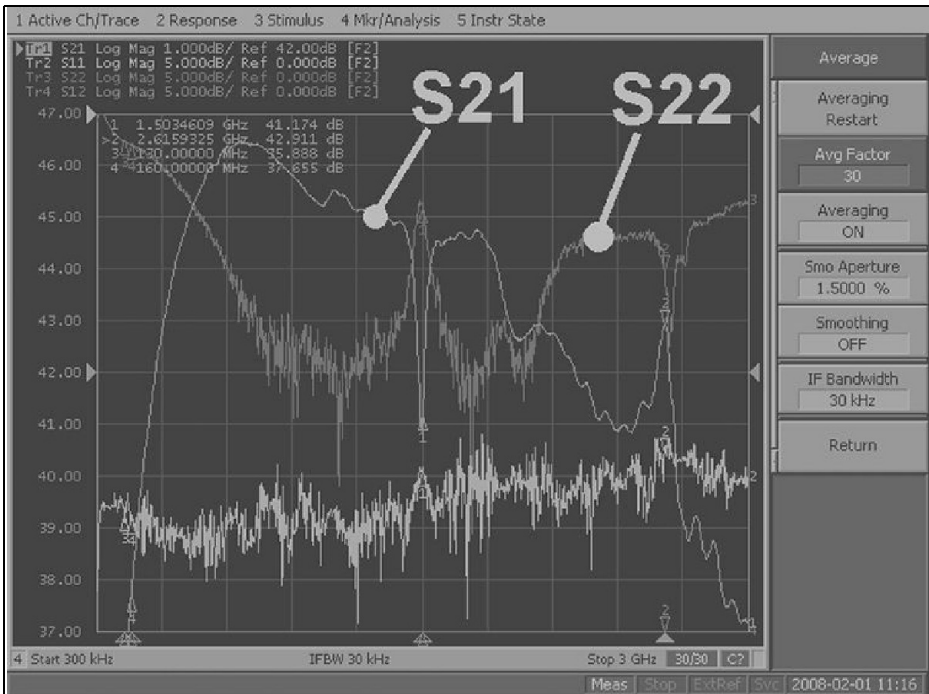
Fig 22: Measurements with an ultramodern network analyser! Detail: See text.

measures more than +60dB from approximately 200MHz to 2700MHz with a maximum value of +66.5dB at approximately 650MHz.

- At the sensitive oscillation point of 1500MHz a notch of approximately 3dB can be seen, this is nearly always an indication for a self-resonance, probably the 33nH choke.

- S22 (right axis) shows a blip at this frequency. At lower frequencies the reassuring value of -20dB is measured, this only worsens above 2100MHz up to -15dB and finally at 3GHz a maximum of -10dB.

The result was not perfect for the first prototype, but nevertheless for investigations and measurement of noise factor it will be quite usefully.

To use the amplifier first we have to find the suitable settings for the spectrum analyser. For optimum operation, accuracy and readability, manual scan proved most suitable when the analyser works as tuneable receiver. Now ensure that the self-noise of the analyser remains at least 10 to 20dB below the output noise of the preamplifier by using as low an amplification of the prescaler in the analyser as possible. At the same time keep an eye on the overload limit of the analysers mixer input. Select the 10Hz video filter and a scale factor of 2dB per division to get a fine bright point screen display that varies irregularly by approximately 0.2dB. Now you can relax for the first time and collect the following measurements:

- Removing and replacing the 50Ω resistor on the input changes the indicated output noise level by approximately 2dB.
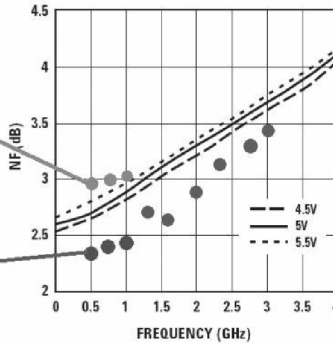
Fig 23: The hour of the truth: Measurements with professional Noise Figure meters.

- With the 50Ω resistor fitted, warming it by hand or violently with a soldering iron shows the forecast rise of the noise level very clearly since the resistor has now a temperature that is more than 290 Kelvin.

Thus the development phase is finished and genuine measurements can be made.

4.2.3.2. The single stage amplifier prototype as the Device Under Test (DUT)

The frequency range to 1.5GHz was selected because (owing to the help of the friend already mentioned) accurate measurements with modern professional instruments were available. Fig 23 shows these results of measurement compared with the data from the data sheet.

The three-stage prototype amplifier with more than 60dB gain was screwed onto the analyser, the 50Ω resistor fitted to the input, the centre line centred and only then the item under test with approximately 20dB gain was fitted. As expected the noise level rose by around approximately 20dB between 300MHz and 1.5GHz (according to the method from chapter 4.1. in part 1). Instead of the expected nF of approximately 2.5 to 3dB (see Fig 23), a nF of less than 0.5dB was measured, although the S21 of DUT was known from the network analyser meas-

urements for all selected measuring frequencies with an accuracy of approximately 0.2dB.

What had gone wrong? Interestingly the results of measurements from 2.5 to 3GHz converged slowly with the factory specifications according to Fig 23!

It is easy to be smart after the event, but the way to realisation was laborious despite the moral support of intensive discussions with competent people. Naturally the following rough calculations can be made:

The approximate middle noise output level in dBm of the chain of four prototype stages that feed the analyser input:

Pnoise = (spectral noise density per Hz) + (increase by larger bandwidth) + (overall gain of all amplifiers) + (middle noise Figure of the first stage)

That results in the following total level in dBm:

P = -174dBm/Hz + 10 x log (2.5GHz) + 84dB + 3dB = -174dBm + 94dB + 84dB + 3dB = +7dBm

The higher values of gain (of over 22dB per individual stage) between 500 and 1000MHz were not considered at all, and if one then looks at the data sheet:
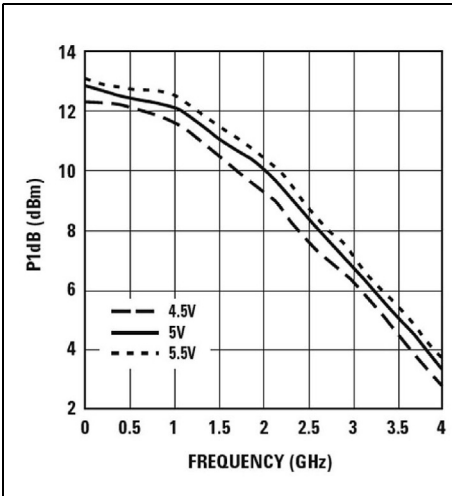
P1db = +9.8dBm

Fig 24: Here one can see, how the MMIC output goes onto its knees with rising frequency - therefore also the wrong results of measurement come from the test run!

It becomes clear: P1dB defines the level at which at the output amplitude (by limiting) is already reduced by 1dB and this affects the last amplifier stage. The larger voltage peaks naturally occurring in the noise suffer and they are cut off. Fig 24 supports this realisation from the data sheet of the ABA52563, which shows the reduction of the P1dB value with rising frequency.

Since this method is not valid the following work is required: A further PCB must be designed but this time with only 2 stages and a sufficient overall gain between 40 and 45dB. The following improvements will follow:

- The microstrip lines on the PCB are replaced with "Coplanar Waveguides". These resemble microstrip lines, but ground surface is much closer to the left and right of the line. This leads to substantially smaller scattering fields and improves stability. Designing the lines is not a problem because the line calculator in the Ansoft designer student version has all conceivable models and options ready, free of charge.

- A copper sheet is inserted between the two stages, soldered to the ground surface as a partition to reduce the crosstalk between the stages. The through hole that connects the two stages can be very small due to the Coplanar lines.

- Different values for the 33nH choke on the output of each stage are chosen e.g. 27nH and 33nH. In this way there is no longer the bandpass filter amplifier effect with the periodic resonances. A bandpass filter amplifier relies on resonant circuits that are all at the same frequency, this is avoided with this detuning.

- The absorption material stuck on the inside of the lid cover must not be forgotten. The copper partition should exert some pressure on the foam material so that a gap is avoided.

The three-stage amplifier prototype has not been thrown away. With a low-noise, narrow-band preamplifier and additional band limitation with filters (to the avoidance of the limiting effect described above) you can hear the grass grow on a suitable Patch or Yagi antenna.

5.

Summary

Once again much has been learned, not only about the phenomenon of the noise and noise measurement, prototype design of LNAs with a large frequency range and high gain and the fight with the self-oscillation. Also becoming acquainted with the spectrum analyser how to set the controls and switches in order to make the correct measurements. Naturally this gives food for thought.

In conclusion a small anecdote, in order to show that the phrase "nobody is perfect" will probably always apply:

After everything was clear with a good conclusion and the follow up procedure, it suddenly occurred to me that I had bought and stored four identical but defective Noise Figure meters at least 10 years ago from a HAM Radio Flea market. These were model 117A from the Magnetic company of Sweden constructed in 1970 including an associated solid state noise source for the frequency range to 1000MHz. After the question: "where did I put them" an intensive search followed and the answer was "in a heap with other electronics scrap iron in the garage" The next question was: "where is the service manual" once it was found this was followed by faultfinding. After appropriate expenditure of time now two of the four devices run again and third is waiting for two officially no longer available adjustable voltage regulators for the power supply that have been found using the Internet residing in the USA. The large meter in the fourth instrument is hopelessly defective and destroyed. So this set is only useful as a source of spare parts.

Excitingly after switching on and comparing with the measurements made by the friendly company in short; the two antique devices measured their 1000MHz NF values of the small and large preamplifiers that were only around 0.3dB over the data sheet curves (see Fig 23). What more does one want? Which one to use now with so many devices?

Another short remark from the author:

With my contribution to this topic I have released things like an avalanche, because the results of part of 1 were some discussions by email and by telephone. Fortunately all were friendly and enriching for my own knowledge, uncovering my own mistakes in reasoning or simply becoming acquainted with very nice and competent colleagues.

Therefore I thank everyone for the lively reaction. Thanks also for all the additional information about things like the (fast aging) noise diodes that are probably still to be found in America over the Internet as a spare parts at bearable prices. And naturally for all practical assistance with measurements and discussions during the project.
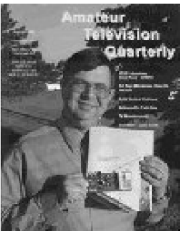
Sigurd Werner, DL9MFV

76GHz sextupler and amplification

This article describes two simple circuits to produce a useful 76GHz signal from a 12.7GHz source.

Module A consists of one MMIC (CHU3377 from UMS) giving a power output of 21mW.

In module B this chip functions as driver and the signal is amplified with an MMIC (APH633 from Velocium/Hittite) to give a power output of 70mW (18.5dBm).

A spectrum analysis of the output signal shows that output filtering (required for most applications) is not necessary.

1.

Introduction

In previous articles [1 - 3] frequency multipliers for 76GHz were described. All these circuits used the MMIC CHU3277 from UMS (United Monolithic Semiconductors) in combination with a frequency quadrupling chip (CHX2092a). Output powers were approximately 50, 90 and 160mW, depending upon the number of MMICs used.

For some time UMS have produced a special sextupler chip with an integrated amplifier. The CHU3377 processes an input signal of, for example 12.7GHz First it is trebeled, then it is amplified and finally doubled and amplified again. The data sheet [4] promises a power output of approximately 15dBm (measured on the chip). Since there are many radio amateur frequency allocations in the range from 95 to 12.8GHz, a functioning transmitter for 76GHz could be built with extremely few components.

This gives a continued examination into the benefit of a downstream MMIC amplifier.

2.

Mechanical and electrical construction

The housings were milled from brass and electroplated with gold (pH 8.2), thay had dimensions 26mm x 20mm x 15mm (module A) and 34.1mm x 24.6mm x 15mm (module B). The principle of construction was described in the earlier articles and is shown in Fig 1 - 4. The input signal of 12672GHz is fed by an SMA socket and the output is coupled into a modified E band waveguide
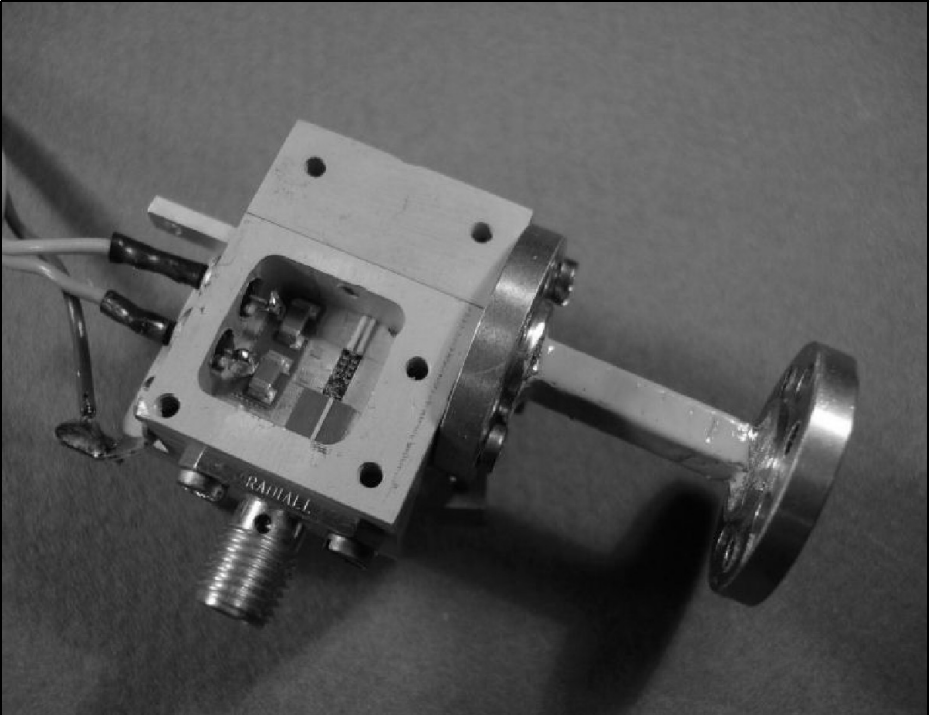
Fig 1: A picture of module A in a milled enclosure. The points of interest are the two enclosures, the feed to the waveguide, the flanged waveguide, the SMA socket for the feed from 12.7GHz and the two DC feeds. The total volume of the module corresponds to half a matchbox.

(1.5mm x 3.0mm) in its H plane [2]. The waveguide flange on the housing side is a special design with a smooth flange (available via [7]) and a commercial flange with its typical λ/4 choke at the other end. All strip lines are made from 5 mil (0.124 mm) Al_2O_3 substrate, so that the gap between the ceramic and the MMICs (height of 0.1mm) is controlled.. The connections to the chip substrstes is made using bond technology as described in [1]. The connection to the DC supply is made via single layer ceramic capacitors.

Module A

Fig 1 and 2 show those extremely "simple" construction. It has two straight strip lines and an MMIC.

Module B:

Here the CHU3377 chip is used as driver with its output connected to an APH633 amplifier MMIC from Velocium/Hittite (Fig 3 and 4).

3.

Results

Module A

The following control voltages were determined: Minus 4.6V (approximately 15mA) and plus 4.7V (220mA; the voltage at the chip was 4.6V). The input power at 12.672GHz was regulated to
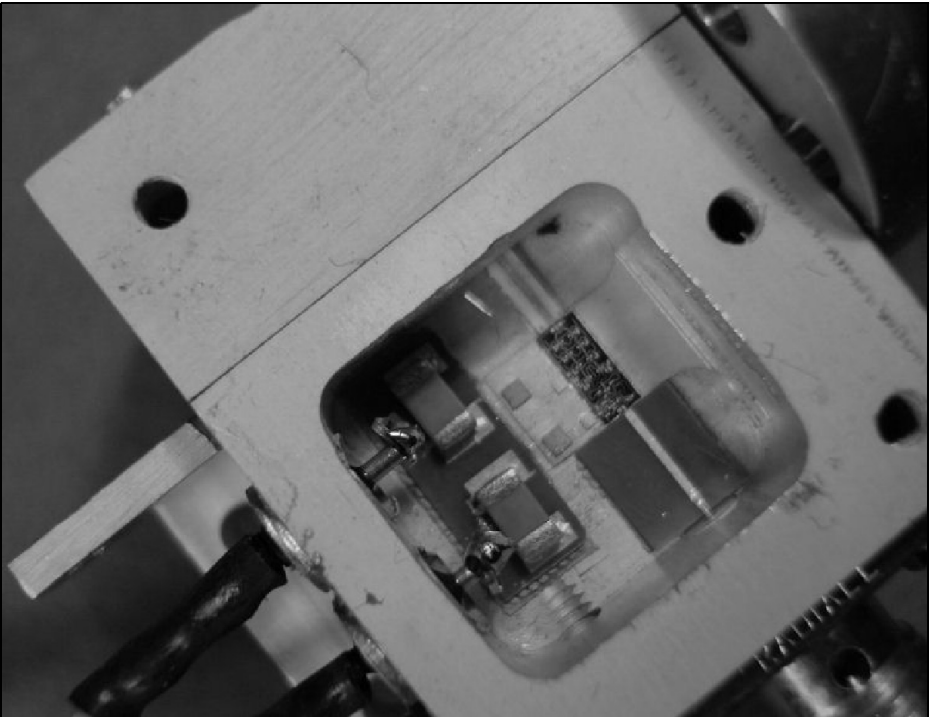
Fig 2: A detailed picture of module A. The CHU3377 MMICs CHU3377 can be seen clearly. The linking substrate into the waveguide is Al_2O_3 (bright), the other substrate for the 12.7GHz feed is made from AlN (dark). The DC pad with the ceramic capacitors is on the left.

approximately 8dBm. The signal strength at 76.032GHz was measured as 21.2mW (13.3dBm).

Module B

The CHU3377 driver was connected to minus 4.5V (15mA) and plus 4.5V (205mA). The gates of the APH633 amplifier chip were connected to the ground potential (0.0V) wilth the drain supply at plus 4.3V (235mA). 3.2dBm was sufficient drive for the amplifier to give a power output at 76.032GHz of 71.1mW (18.5dBm).

The RF transition from the MMIC to the strip line and the following energy transfer into the waveguide must taken into account at these frequencies; even though this is as good as it can be, a loss

of approximately 1.5dB must be added. If this is considered, the maximum values listed in the data sheets [4, 6] are reached. For example the APH633 a maximum of 20dBm (measured on the chip) [6]; 20dBm - 1.5dB = 18.5dBm (70.8 mW).

The signals from the modules were examined in a frequency range from 72 to 80GHz for spurious emissions and mixing products using an HP 70000 system. Two signals at 73.3 and 78.8GHz were found that were 20dB below the main signal. Thus filtering of the power output is unnecessary, at least for most applications.

The modules presented here are suitable for signal processing of 12.7GHz, for beacons or as transmitter components. In
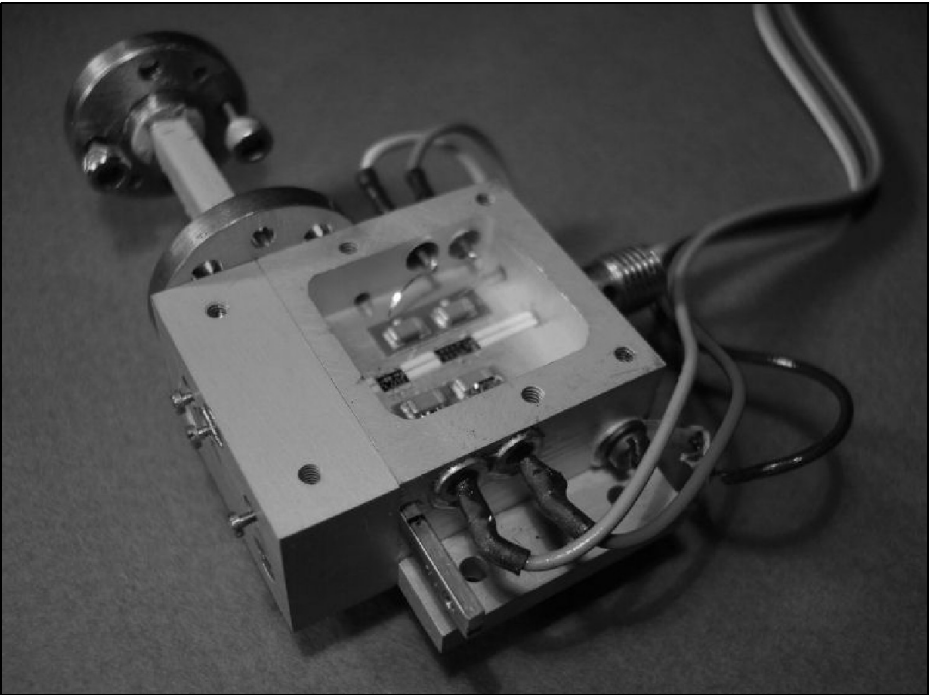
Fig 3: View of the opened module B. A 5mm thick cover is fitted to complete the connection to the waveguide flange. The RF circuit in the centre is flanked by the two DC pads.

particular the single chip model (module A) is hard to beat particularly since it produces a good 20mW output.

4.

Conclusion and thanks

The power output of the driver chip (CHU3377) is sufficient for two amplifier MMICs. Therefore a further module will be built with a Wilkinson divider delivering the driver output to two APH633 amplifiers. Using the proven parallel feed into the waveguide it should produce well over 100 mW. I will report on it soon.

Based on the described multiplier modules (sixfold and eightfold) attractive concepts for a further 76GHz Transverter [5] are to be presented shortly. The efficiency of the components used gives a quantum leap forward compared to diode multipliers.
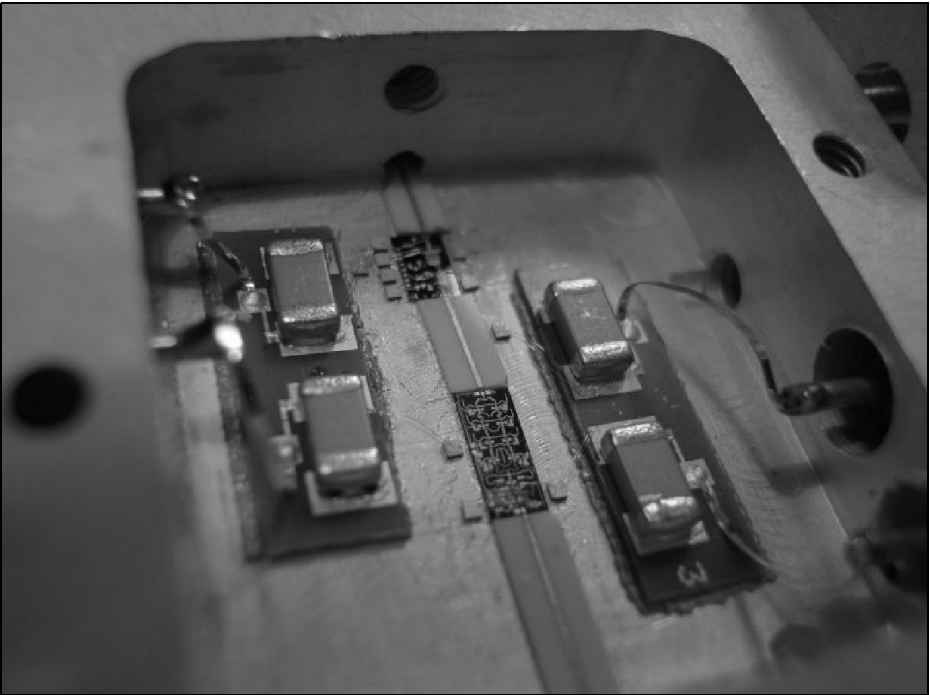
Fig 4: Cloesup view of the main components in module B. All substrates are manufactured from 5mil Al$_2$O$_3$. The probe into the waveguide was tapered in order to keep the hole as small as possible. The upper (smaller) chip is the amplifier (APH633), the lower chip is the MMIC (CHU3377) used as a sextupler and a driver.

5.

References

[1] New transmitter with higher performance for 76GHz, Sigurd Werner, DL9MFV, VHF Communications Magazine 4/2003, pp 194 - 198

[2] Combining power at 76GHz: Three possible solutions discussed, Sigurd Werner, DL9MFV, VHF Communications Magazine 1/2004, pp 13 - 19

[3] Frequency doubler for 76GHz with 130/160mW output, Sigurd Werner, DL9MFV, VHF Communications Magazine 3/2004, pp 130 - 138

[4] Data sheet from United Monolithic Semiconductors, SAS. Ref.DSCHU3377082, 23 Mar 07

[5] A simple concept for an efficient 76GHz transverter, Sigurd Werner, DL9MFV, VHF Communications Magazine 2/2003, pp 77 - 83

[6] Data sheet from Hittite for HMCAPH633, ATMs HEMT MMIC medium power Amplifier, 71 - 76GHz

[7] VHF Communications Magazine web site www.vhfcomm.co.uk

Primoz Lemut S53KS

GPS-LCD, an add-on to the GPS Disciplined Oscillator

1.

Introduction

While assembling the GPS Disciplined Oscillator (GPSDO) that Zeljko S52ZB published in the second issue of VHF Communications in 2007, I realised I would like to use it as a standalone unit, without connection to a PC. Nowadays a PC sits in almost every corner but I didn't like the idea I would have to start it up just to check that GPS reception is OK. To overcome this I decided I would build a simple microcontroller circuit that would do roughly the same as PC programs; process GPS receiver NMEA data from the serial port and display its content. The circuit is named GPS-LCD.

The basic requirement for the unit was to display signal strength for 12 satellites, type of GPS fix (none, 2D, 3D) and number of satellites actually used in position calculation (fix). With all this information one can quickly determine whether signal quality is good enough for GPSDO to lock to the receiver output signal. If there were enough room on the LCD, displaying date and UTC time would be beneficial. Some GPS receivers like Navman Jupiter-T model TU60-D120-041 require initialisation at startup others (Rockwell Jupiter model TU30-D140-221) do not. To initialise the receiver RS232 output format needs to be changed from Motorola to binary and

further to NMEA. To perform initialisation two strings need to be sent to the receiver.

2.

Hardware

In its simplest form this circuit requires 7 signal lines to the character LCD, 2 serial lines and a crystal. Such a circuit can also be very quickly assembled on a perforated FR4 PCB for prototyping using a microcontroller in the DIL package. I simply used a PCB with Atmega8 microcontroller, created by Darko S57UUD. That PCB was originally used in a frequency counter within a VHF SSB transceiver. The size of the PCB is the same as that of the most common 2 x 16 characters LCD; 80 x 37mm. The LCD and the controller PCB are assembled in a sandwich shown in Fig 6. There are several other components on Darko's PCB that I didn't install. If there is enough interest for this circuit, Darko S57UUD is willing to provide a small series of PCBs.

3.

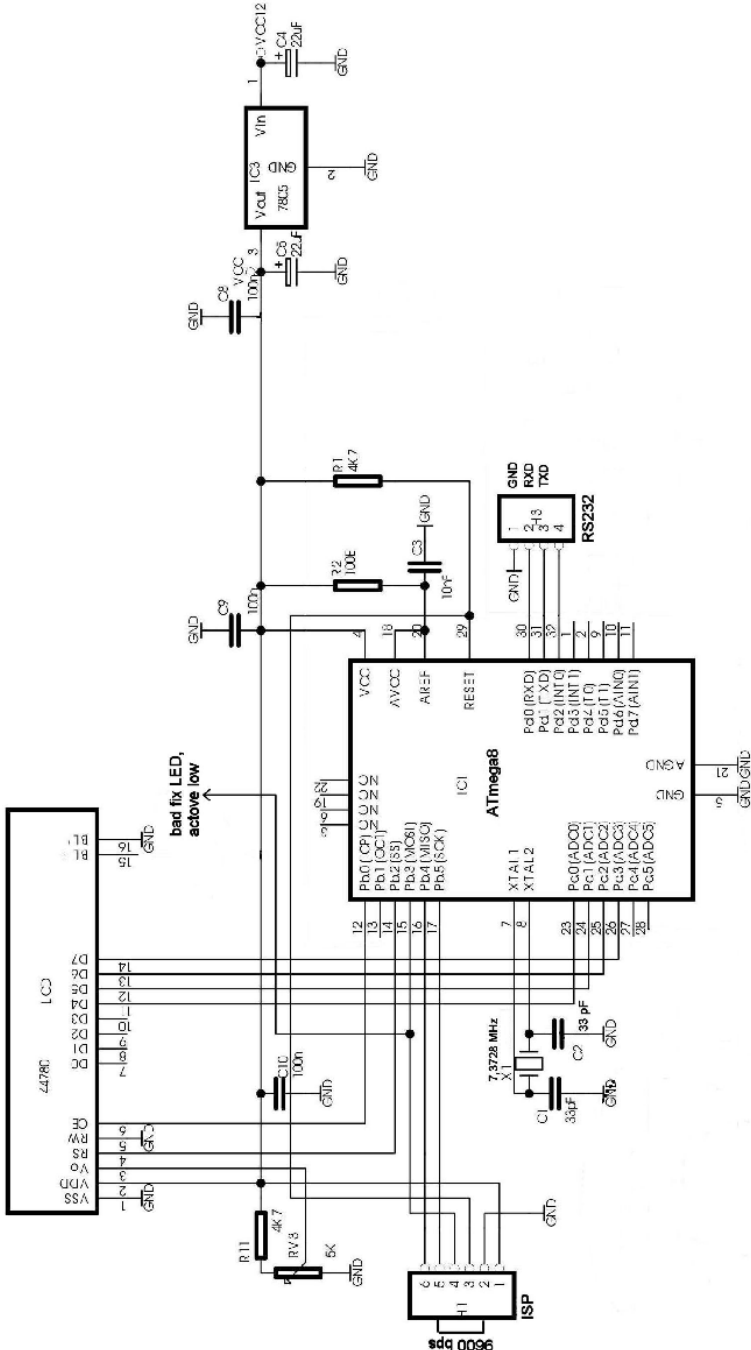## Software

The software was entirely written in the

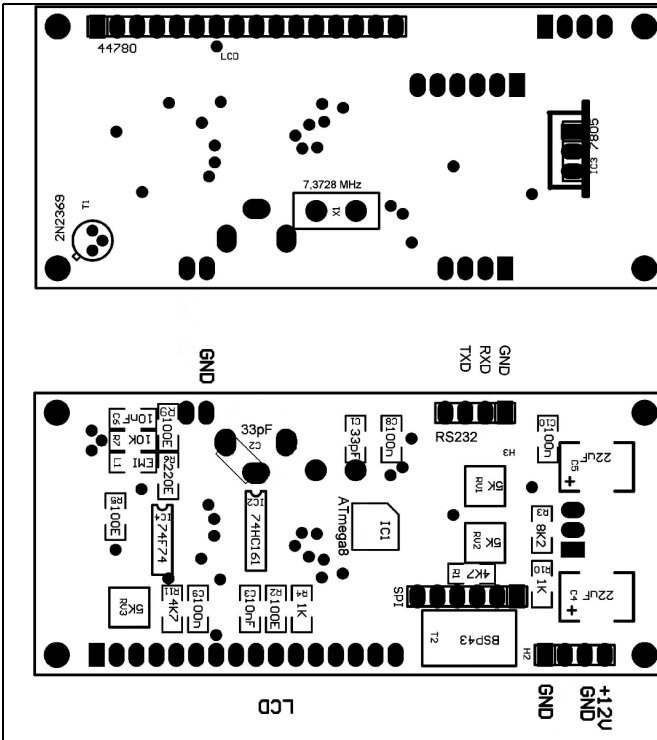**Fig 1: GPS-LCD schematics (courtesy of S57UUD).**

Fig 2: GPS-LCD component placing (courtesy of S57UUD).

C language, there was no need for high-speed operation and assembly language coding. After a function for parsing the GPS data had been written, everything else was very straightforward. When a line (sentence) of GPS data (GPGSV, GPGSA or GPRMC) is properly received, useful data is extracted from it. Some attention was paid to timeouts for total loss of GPS sentences. Also in such
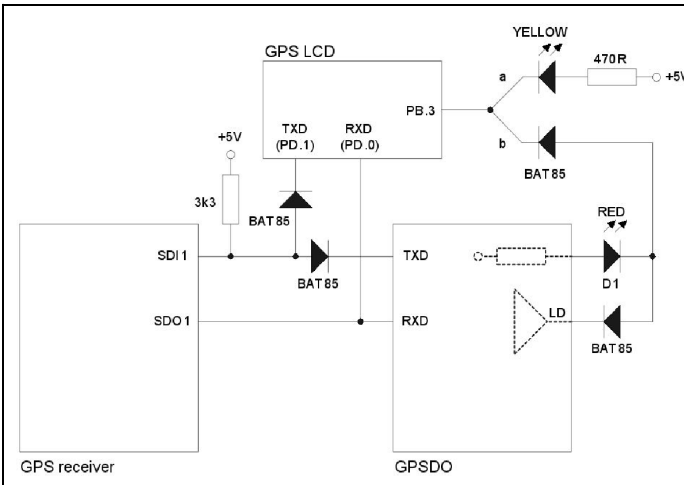

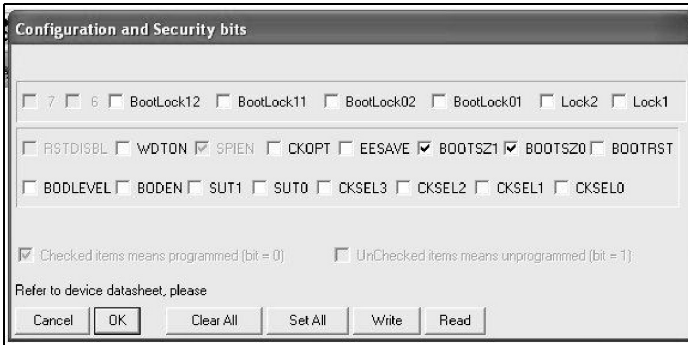Fig 3: Integration of GPS-LCD into GPSDO.

Fig 4: ATmega8 fuses setting in PonyProg.

case all status information must be regularly refreshed.

When the welcome message is displayed during the unit start-up, two initialisation strings are sent to the GPS receiver. Both strings are stored in the EEPROM and can be at maximum 63 bytes long. The first is stored at location 0x00 and second at 0x40. The first byte of the message location is not data but the actual length of the string. By placing the strings into the EEPROM the user is able to change them without re-compiling the program. The HEX code for programming the microcontroller is available free of charge from the author.

4.

Integration with the GPSDO

In the direction from the microcontroller to the GPS receiver a simple diode circuit with pull-up resistor operates as an AND gate. This circuit allows both, the microcontroller and the PC, to control the receiver. From the GPS receiver the signal is directly fed to the RS232 level translator and GPS-LCD circuit. There is a LED connected to the PB.3 microcontroller output. It blinks with short pulses at 0.5 Hz rate when there is no 3D fix available. This is helpful when sometimes TCVCXO locks to 10kHz GPS output signal that is not synchronised to the GPS due to poor reception conditions. You can use either a separate LED (path "a" in Fig 3) or the lock detector LED (path "b") from the GPSDO. For the later two additional diodes are required.

The FLASH and EEPROM memory within the Atmega8 microcontroller can be programmed either using an external parallel programmer and TQFP adapter or in-circuit (ISP) using an interface with a cable. In both cases you have to take care that fuse bits are correctly set. For
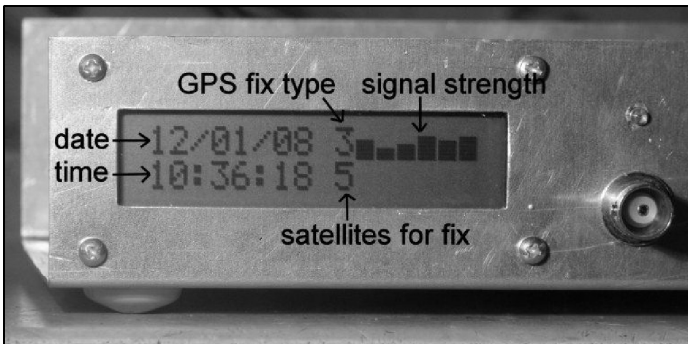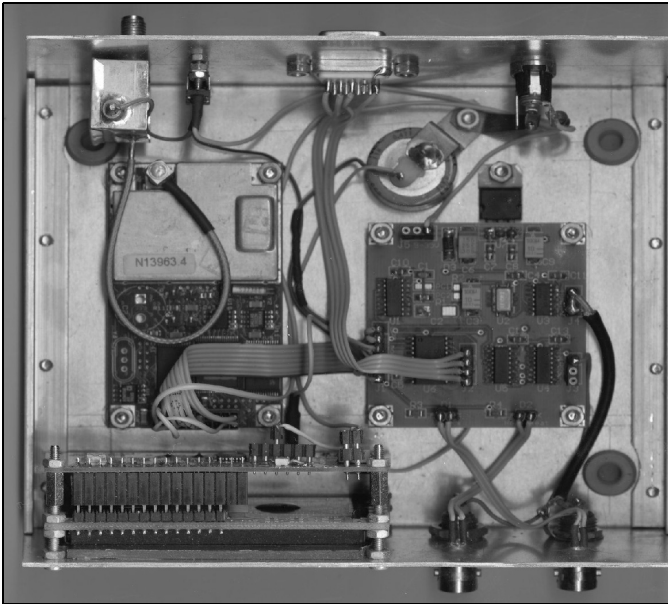

Fig 5: Meaning of LCD data.

Fig 6: Completed unit.

the freely available PonyProg tool that I am using the settings are shown in Fig 4.

In the component layout (Fig 2) all the components from the S57UUD frequency counter are visible. For the GPS-LCD there are only few component to be fitted. Following the schematics and reference designators you'll be able to correctly assemble the board. Instead of the variable capacitor C2 a fixed one (SMD or through-hole) of 33pF is installed.

As shown in Fig 6 I added a few things to the GPSDO unit. One is an external bias tee for the active antenna supply. The switch on the rear of the unit is used for selection of 5V/12V supply. As the receiver doesn't have any current limit I realised it would be much easier to repair an external bias tee rather than the one built into the receiver. The second thing added is a goldcap that assures few hours of GPS almanac data retention and thus quicker start-up time.

GPS-LCD is supplied from +12V because there is a voltage regulator on-board. Current consumption (without backlight) is less than 50mA and if IC3

(7805) is omitted the board can be supplied directly from the GPSDO +5V line.

5.

Operation of the GPS-LCD

When the unit powers up, a welcome message "GPS disciplined 10MHz reference" is displayed for four seconds. After this normal operation starts. Until the GPS receiver receives its first valid data, "--/--/--" and "--:-:-" are displayed for date and time, respectively. GPS fix type (0 to 3) digit (Fig 5) is blinking 5 minutes after 3D fix is available. I selected this timeout because GPSDO typically needs at least 5 minutes for initial lock. If 3D fix is lost, the digit starts blinking again. There is additional digit in front of the second line of signal strength bars. It is displaying how many satellites are actually used in the current GPS fix. This number is typically lower than the number of all visible satellites, for which signal strengths are displayed.

If you are using a receiver that outputs data at 9600 bps, you need to short pins 2 and 5 on the ISP connector. For 4800 bps just leave them open.

6.

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Aristoteles Tsiamitros, DD5FT
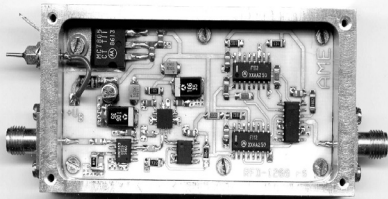
Attenuators

The main task of pure resistive attenuators is to reduce high signal levels independent of frequency. Half L sections can also perform wideband matching between different characteristic impedances. The equations for calculating T and PI attenuators for different characteristic impedances at the input and output are described by examples.

The LTspice/SwCad3 simulator from linear Technology enables the current, tolerance of components and changes in temperature to be examined for attenuators.

1.

Introduction

The subject of this article is the design of asymmetrical T and PI attenuators. Asymmetrical attenuators have different characteristic impedances at the input and output. The design equations can however be used for symmetrical attenuators, these have the same characteristic impedances at both sides.

The characteristic impedance at the input is denoted in the following equations by Z_1 and at the output by Z_2. The attenua-

tion is denoted by A in dB and the load resistance by R_L. For matching $R_L = Z_2$. The objective is it to determine three resistors for the attenuator. Because the surplus power is not reflected but absorbed, the power rating of the resistors must be considered.

The design equations for asymmetrical attenuators are summarised in Fig 4 for the T attenuator and in Fig 6 for the PI attenuator. The linear attenuation factor is given by:

$$\alpha = P_1/P_2 = 10^{A/10}$$

It is the relationship of the input to output power. Thus $\alpha > 1$ must always apply. For $\alpha = 1$ the attenuation A = 0dB.

Both sets of equations for symmetrical attenuators and their derivation can be found in literature (e.g. in [1] P. 603). For asymmetrical attenuators the equations can be found in [2] and [3] but the derivation is searched for in vain in the literature. The only source of these derivations that is known at this time is the article in [4]. A further method for the derivation of the equations in Fig 4 and 6 is shown in the appendix. Using the theorem of reflection factor the computation becomes very clear. This theorem should be familiar to high frequency technician from line theory (e.g. see [5], P. 68). It means that the reflection factor at the input of a lossy line is smaller by the line loss than that at the output. This
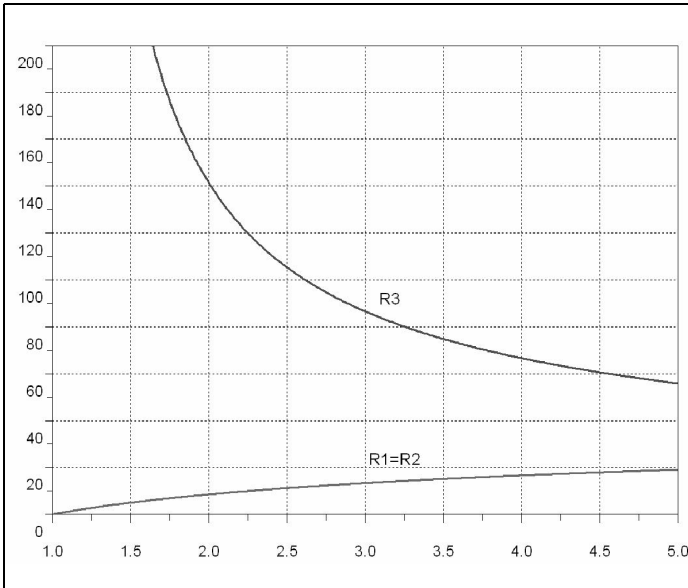
Fig 1: Plot for a symetrical 50Ω T attenuator showing $R_3\Omega$ and $R_1 = R_2\Omega$.

can be applied to any four terminal network including an attenuator. The reflection factor α at the input of the attenuator is smaller by the attenuation factor , than at the output.

Independently of whether a T or PI network is realised, three cases can be discussed on the basis of the values of Z_1, Z_2 and A.

Case 1) The characteristic impedances Z_1 and Z_2 are equivalent, A is arbitrary.

This is the most common case. The range of values of attenuation A varies from A = 0dB up to (theoretically) any value. The attenuator is symmetrical, i.e. the parallel resistors are the same for the T attenuator or the longitudinal resistors are the same for the PI attenuator. The resistor values for a T attenuator are plotted against attenuation in Fig 1.

Case 2) The characteristic impedances Z_1 and Z_2 are different, $A > A_{min}$.

Fig 2 shows the resistor values plotted against the attenuation factor α for a T

attenuator with $Z_1 = 50\Omega$ and $Z_2 = 75\Omega$.

It can seen from this graph that for $\alpha = 1$ to $\alpha_{min} = 3.732$ the resistor R_1 is negative. Similarly a resistor would be negative for the PI attenuator below the minimum attenuation factor. Attenuation values below the minimum attenuation cannot be realised. Asymmetrical T or PI attenuators can only be realised above the minimum attenuation factor.

Case 3) The characteristic impedances Z_1 and Z_2 are different, $A=A_{min}$.

For the minimum attenuation the longitudinal resistor on the low impedance side of a T attenuator becomes 0. For the PI attenuator the leak resistor becomes approximately infinitely on the high impedance side of the circuit. In both cases this is a half L section (minimum loss pad) consisting of two resistors. Such half L sections are used e.g. in order to match 75Ω devices to 50Ω.

In the following the above cases are described with examples of T and PI networks.
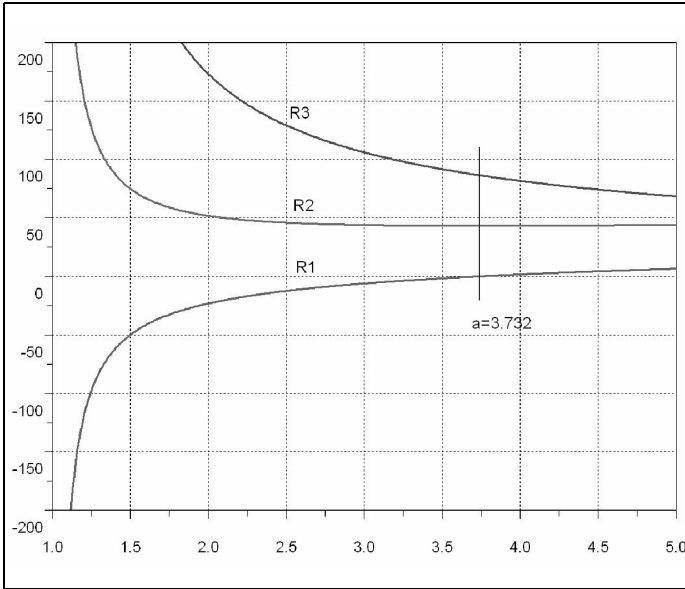
Fig 2: R1, R2 and R3Ω plotted agains attenuation factor α for a symmetrical T attenuator with 50Ω input and 75Ω output impedance. α must be greater than 3.732, otherwise R₁ is negative.

2.

T attenuators

The circuit diagram of a T attenuator is shown in Fig 3. The design equations are summarised in Fig 4. The calculations can be carried out with a pocket calculator, a spreadsheet program or with LTspice [7].

The cases shown in the previous section are described by the following three examples.

Example 1:

Symmetrical T attenuator, A = arbitrary

For $Z_1 = Z_2 = 50\Omega$ a T attenuator is to be realised with an attenuation A = 6.02dB. First, for the given attenuation in dB the attenuation factor α is calculated.

$$\alpha = 10^{A/10} = 10^{6.02/10} = 4$$

A symmetrical attenuator is realisable for any attenuation. The given characteristic impedances and α are used in the equa-

tions from Fig 4

$$R_3 = \frac{2\sqrt{\alpha}Z_1 Z_2}{\alpha - 1} = \frac{2 \cdot \sqrt{4 \cdot 50\Omega \cdot 50\Omega}}{4 - 1} = 66{,}67\Omega$$

$$R_1 = \frac{\alpha + 1}{\alpha - 1}Z_1 - R_3 = \frac{4 + 1}{4 - 1} \cdot 50\Omega - 66{,}67\Omega = 16{,}67\Omega$$

$$R_2 = \frac{\alpha + 1}{\alpha - 1}Z_2 - R_3 = \frac{4 + 1}{4 - 1} \cdot 50\Omega - 66{,}67\Omega = 16{,}67\Omega$$

Example 2:

Asymmetrical T attenuator, A > A_min

In this example a T attenuator between $Z_1 = 40\Omega$ and $Z_2 = 80\Omega$ is to be designed. The attenuation is A = 12dB.

For different characteristic impedances the first check is to determine if the circuit is realisable. The required attenuation must be larger than the minimum attenuation. The minimum attenuation is given in Fig 4:

$$\alpha_{min} = \left(\sqrt{\frac{Z_g}{Z_k}} + \sqrt{\frac{Z_g}{Z_k} - 1} \right)^2 = \left(\sqrt{\frac{80}{40}} + \sqrt{\frac{80}{40} - 1} \right)^2 = 5{,}828$$
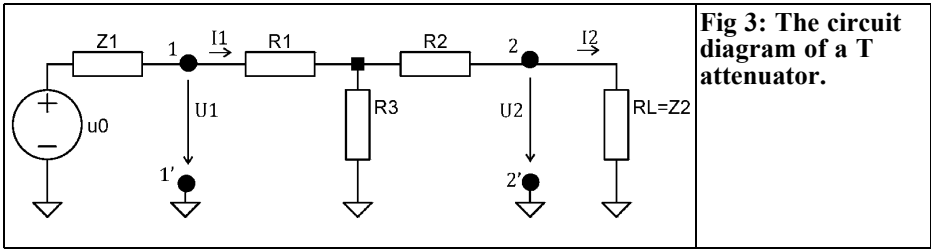
Fig 3: The circuit diagram of a T attenuator.

or expressed in dB:

A_{min} = 10 lg (5.828) = 7.656dB.

The required attenuation (12dB) is larger than the minimum attenuation (7.65dB) and is realisable. Now the attenuation factor α for 12dB is calculated:

$\alpha = 10^{A/10} = 10^{12/10} = 15.85$.

Using this in the equations from Fig 4 gives::

R_3 = 30.33Ω, R_1 = 15.06Ω, R_2 = 60.44Ω

Example 3:

Half L section, A = A_{min}

The T attenuator is between the characteristic impedances Z_1 = 50Ω and Z_2 = 75Ω. It is to be used to match two devices with different characteristic impedances over a wideband and to insert as small an additional attenuation as possible.

$$R_3 = \frac{2\sqrt{\alpha Z_1 Z_2}}{\alpha - 1}$$

$$R_1 = \frac{\alpha + 1}{\alpha - 1}Z_1 - R_3$$

$$R_2 = \frac{\alpha + 1}{\alpha - 1}Z_2 - R_3$$

$$\alpha_{min} = \left(\sqrt{\frac{Z_g}{Z_k}} + \sqrt{\frac{Z_g}{Z_k} - 1} \right)^2$$

$$A_{min} = 10\ lg(\alpha_{min})$$

Fig 4: Design equations for a T attenuator. In the equation for the minimum absorption factor α_{min} the denominator (Z_g) is the larger and (Z_k) the smaller of the two characteristic impedances Z_1 and Z_2.

The attenuator is always realisable for the minimum attenuation:

$$\alpha_{min} = \left(\sqrt{\frac{Z_g}{Z_k}} + \sqrt{\frac{Z_g}{Z_k} - 1} \right)^2 = \left(\sqrt{\frac{75}{50}} + \sqrt{\frac{75}{50} - 1} \right)^2 = 3,732$$

Calculating the resistors gives:

R_3 = 86.60Ω, R_1 = 0Ω, R_2 = 43,30Ω

The resistor on the low impedance side of the circuit, R_1, is as expects 0. The attenuation in dB for this "minimum loss PAD" is:

A_{min} = 10 lg (3.732) = 5.72dB

3.

PI attenuator

The circuit diagram of a PI attenuator is shown in Fig 5. The design equations are summarised in Fig 6. The definition cases used for the T attenuator apply to the PI attenuator. For the symmetrical PI attenuator the leak resistors are the same, i.e. R_a = R_c. The minimum attenuation is calculated as shown in the appendix and uses the same formula as for the T attenuator.

Example 1:

Symmetrical PI attenuator:

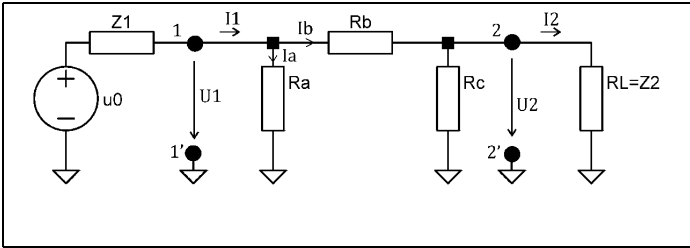A PI attenuator is to be realised for Z1 = Z2 = 50Ω and an attenuation A = 20dB.

Fig 5: The circuit diagram of a PI attenuator.

First the attenuation factor is calculated from the given attenuation in dB:

$$\alpha = 10^{A/10} = 10^{20/10} = 100$$

Using this in the design equations gives:

$$R_b = \frac{1}{2}(100-1)\sqrt{\frac{50\Omega \cdot 50\Omega}{100}} = 247{,}50$$

$$R_a = R_c = \frac{1}{\dfrac{1}{50\Omega}\dfrac{100+1}{100-1} - \dfrac{1}{247{,}5\Omega}} = 61{,}11\Omega$$

Example 2:

Asymmetrical PI attenuator, A > A$_{min}$

The values from example 2 for the T

Fig 6: Design equations for a PI attenuator. In the equation for the minimum absorption factor α$_{min}$ the denominator (Z$_g$) is the larger and (Z$_k$) the smaller of the two characteristic impedances Z$_1$ and Z$_2$.

attenuator will be used: $Z_1 = 40\Omega$, $Z_2 = 80\Omega$ and A = 12dB.

The test for feasibility would have to take place in the same way as for the T attenuator. Since the minimum attenuation has the same value as for the T attenuator, the PI attenuator is realisable. With $\alpha = {}^{10A/10} = 10^{12/10} = 15.85$ This is used in the equations in Fig 6 giving:

$R_a = 52.94\Omega$, $R_b = 105.49\Omega$,

$R_c = 212.55\Omega$

Example 3:

Symmetrical PI attenuator:

If the minimum attenuation is required, then it follows that the half L section will be exact the same as the half L section for the T attenuator. In the PI network the leak resistor on the high impedance side becomes infinitely large.

4.

Simulation with LTspice/SwCad3

I wrote an article on LTspice/swCADiii in this magazine [7]. The program (called LTspice for brevity) can be downloaded free of charge from the homepage of linear Technology [6].

4.1. Power absorption in the attenuator

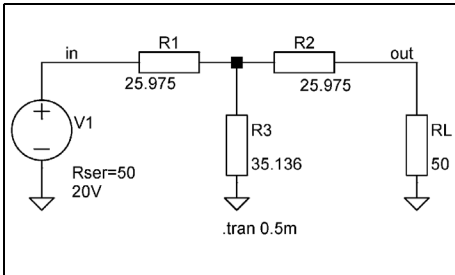An attenuator designed for an attenuation A = 10dB and 50Ω at the input an output

Fig 7: T attenuator for $Z_1 = Z_2 = 50\Omega$, A = 10dB.

is shown in Fig 7. The objective is it to determine the power dissipated by the individual resistors.

LTspice can simulate the attenuator with a source connected, pressing the ALT key with the mouse pointer over one of the components changes the mouse pointer into a thermometer. The power is shown on the Waveform Viewer, power dissipation is shown as positive and output as negative. If the circuit is fed with a DC source then the rms power is shown. With an AC source the instantaneous power is shown. Averaging can be carried out with the Waveform Viewer by pressing CTRL + the left mouse button over the legend e.g. over the text "V (out) *I (RL)".

For the example shown in Fig 7 with a DC source the available power is:

$$P_{max} = \frac{U_0^2}{4 \cdot Rser} = \frac{(20V)^2}{4 \cdot 50\Omega} = 2W$$

The power at R_L is 1/10 of the available power, i.e. 200mW. This can be seen in Fig 8, which shows the power dissipations or the individual resistors. The sum of these powers amounts to 2W, the largest power dissipation is in the resistor that is next the source. To read a value, click the right mouse button over the legend e.g. V(N001)*I(R3), this opens the Expression Editor dialogue. The cursor is placed on the curve and the power e.g. P_3, can be seen in the window.

4.2. Component tolerances (Monte Carlo analysis)

The Monte Carlo analysis allows the component tolerances of the circuit to be analysed. The function mc(x, y) is available in LTspice that produces random numbers with a uniform distribution between x*(1-y) and x*(1+y) (See the online help on Dot Commands, [.PARAM] that can be applied to a filter circuit and the examples in the LTspice file MonteCarlo.asc of the installation list).

A family of curves is produced for the parameter that can be measured (voltage, current, power etc.) in several runs. The number of runs can be controlled with
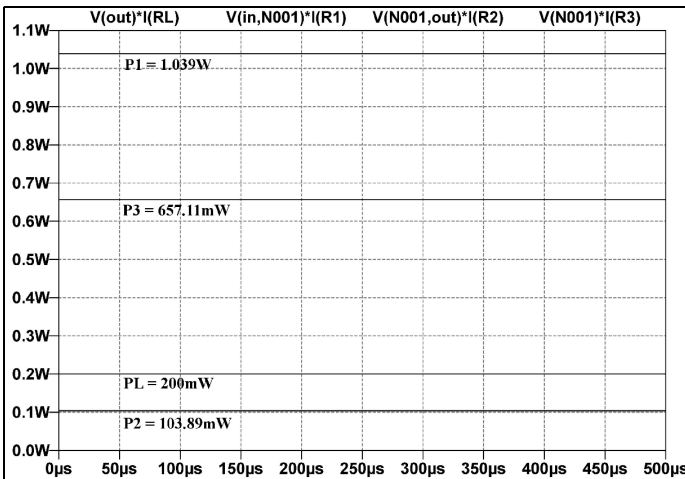

Fig 8: Power disipation of R_1, R_2, R_3 and R_L.

.step param N 0 50 1

R1
{mc(27,0.05)}

R2
{mc(27,0.05)}

V1
AC 1
Rser=50

R3
{mc(36,0.05)}

RL
50

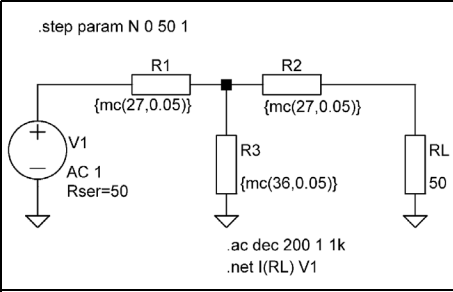.ac dec 200 1 1k
.net I(RL) V1

Fig 9: The the effect of the component tolerances on the attenuation is determined using Monte Carlo analysis.

the instruction "step" (see on-line help). With each new run the function mc(x, y) produces new values used for the simulation.

In Fig 9 the parameter N is stepped from 0 to 50. For each new value the resistors "mc ($R_{nominal}$, 0,05)" have a new value between $R_{nominal} \times (1 - 0.05)\Omega$ and $R_{nominal} \times (1 + 0.05)\Omega$ assigned and the network parameters computed. The computation of the network parameters is determined by the instructions .AC and .net.

The transmission function S21 for N = 0 to 50 is shown Fig 10. The deviation from the nominal value of 10dB due to the 5% tolerance of the resistors is approximately 0.4dB.

4.3. Temperature effects (DC-Sweep)

Resistors change their value as a function of the temperature. The resistance value at a temperature ϑ_W amounts to:

$$R_W = R_K \times [1 + TC_1 \times (\vartheta_W - \vartheta_K) + TC_2 \times (\vartheta_W - \vartheta_K)^2]$$

The temperature factors TC_1 and TC_2 are material constants. They can have positive or negative values (PTC or NTC resistors). The second temperature factor TC_2 is usually very small and only considered for larger temperature differences. The nominal temperature ϑ_K at which the components have their nominal values R_K, is defined in LTspice with $t_{nom}=27°C$ that can be changed using the "option" parameter, see Fig 11.

The objective of this section is it to plot the attenuation over a temperature range (Fig 12). The "DC sweep" function is used for this analysis. The actual purpose of this function is to determine the DC voltage or a direct current for one or several sources. A possible application of this is to plot a transistor family of characteristics. This function can be used to determine the values as a function of temperature.
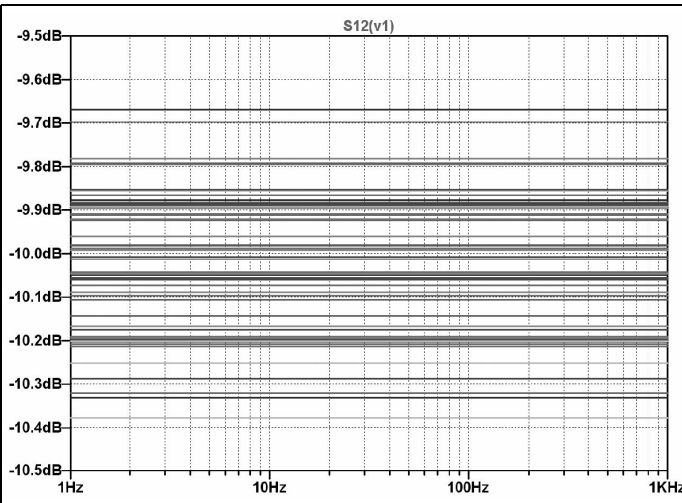

Fig 10: Attenuation dependence on the tolerance of the components, the nominal value is 10dB.
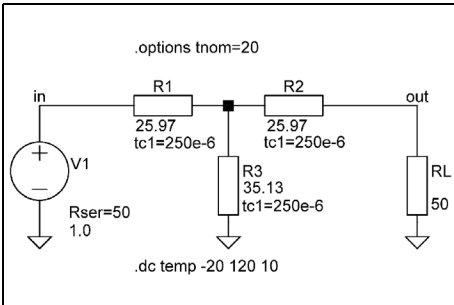
Fig 11: Circuit for determining the temperature dependence. The option "tnom" defines the nominal temperature for the resistors.

For the 10dB attenuator shown in Fig 11 the source must be a DC source, U = 1V with an internal resistance equal to the characteristic impedance of the attenuator. The first temperature factor in this example is:

$$TC_1 = 250 \times 10^{-6} \text{ K}^{-1}$$

The second temperature factor $TC_2 = 0$. To the assign of the temperature factor to the resistors, right click on the component while pressing the CTRL key, this opens the Component Attribute Editor. Enter "tc1=250e-6 in the line "Spice-Line" and close the dialogue. Repeat this procedure for all the other resistors.

The attributes for the simulation are set in the "Edit Simulation Command" dialogue using the menu "Simulate > Edit Simulation Cmd". Select "DC sweep" under "1st source" and give the field "1st Source to sweep" the name "temp". For this example the "linear" temperature range from -20°C to 120°C in steps to 10°C was selected. Entering ".op" or with "S" on the keyboard opens the dialogue "Edit Text on the Schematic" for the simulation instructions shown in Fig 11.

If "DC-Sweep" is selected the network analysis cannot be run automatically as in the previous section, but the desired size must be entered manually. Activate the Waveform Viewer window after the simulation and open the menu "Plot Settings > Add traces". The dialogue "Add Traces to plot" opens where the expression 20*log10(V(in)/V(out) should be placed in the field "Expression (s) to add". The result is shown in Fig 12.

At the temperature of 20°C, which was set in Fig 11 as the nominal temperature the attenuator has the nominal attenuation of 10dB.

The formula entered applies to the case for same the characteristic impedances at the input and output of the attenuator. For
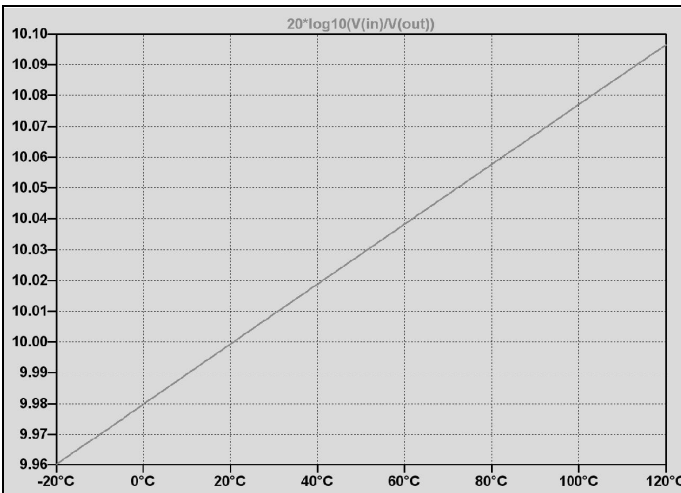

Fig 12: Attenuation as a function of the temperature. At 20°C the attenuation is exactly 10dB.

173

different characteristic impedances the input must be corrected according to equation (2) in the appendix. It becomes $20*\log10((V(in)/V(out))*\sqrt{Z_2/Z_1})$, with Z_2 the characteristic impedance at the output and Z_1 the characteristic impedance at the input.

5.

Summary

The design equations for T and PI attenuators have been described by examples. The practical importance of power dissipation of the individual resistors as well as tolerance and temperature effects were determined using LTspice.

6.

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http://www.rfcafe.com/references/calculators/attenuator_calc.htm

7.

Appendix

7.1. General information

The following equations apply to both the T, and PI attenuators as shown in Fig 3 and Fig 5.

The logarithmic attenuation value A is expressed by a positive dB number. The linear attenuation factor α is defined as the quotient of the power at the input terminals (1-1') and the power at the output terminals (2-2').

$$\alpha = \frac{P_1}{P_2} = \frac{\frac{U_1^2}{Z_1}}{\frac{U_2^2}{Z_2}} = \frac{U_1^2}{U_2^2}\frac{Z_2}{Z_1} \qquad (1)$$

From this the power ratio follows:

$$\ddot{u} = \sqrt{\alpha} = \frac{U_1}{U_2}\sqrt{\frac{Z_2}{Z_1}} \qquad (2)$$

and the output voltage:

$$U_2 = \sqrt{\frac{Z_2}{\alpha Z_1}} \cdot U_1 \qquad (3)$$

Since dual matching is assumed, the input impedance is Z_{in} at the terminals 1-1' equal to Z_1 and the input current is:

$$I_1 = \frac{U_1}{Z_1} \qquad (4)$$

The output current using U_2 from (3):

$$I_2 = \frac{U_2}{Z_2} = \frac{U_1}{\sqrt{\alpha Z_1 Z_2}} \qquad (5)$$

The reflection factor at the output of the four terminal network with load resistance R_L is:

$$\rho_2 = \frac{R_L - Z_2}{R_L + Z_2} \qquad (6)$$

The reflection factor at the input terminals 1-1' is smaller than at the output by the attenuation factor of the four terminal network. This is similar to the reflection factor at the input of a transmission line:

$$\rho_{in} = \frac{Z_{in} - Z_1}{Z_{in} + Z_1} = \rho_2 \frac{1}{\alpha} \qquad (7)$$

7.2. T attenuator

Resistor R_1:

The reflection factor at the output of the four terminal network with an open circuit output i.e. $R_L = \infty$, from (6):

$$\rho 2 = 1$$

The reflection factor at the input from (7):

$$\rho_{in} = \frac{Z_{in} - Z_1}{Z_{in} + Z_1} = \frac{1}{\alpha}$$

Re-arranging this equation for Z_{in} results in:

$$Z_{in} = \frac{\alpha + 1}{\alpha - 1} Z_1 \qquad (8)$$

Z_{in} is the resistance of the four terminal network at terminals 1-1'. With the open circuit four terminal network is:

$$Z_{in} = R_1 + R_3$$

Using this in (8) and bringing R_3 to the right side gives the design equation for R_1 shown in Fig 4.

Resistor R_2:

If the four terminal network in used in the reverse direction, i.e. feed at the output with open circuit input, then this is

the reflection factor at terminals 2-2':

$$\rho_2 = \frac{Z_{out} - Z_2}{Z_{out} + Z_2} = \frac{1}{\alpha}$$

Re-arranging this equation for Z_{out} results in:

$$Z_{out} = \frac{\alpha + 1}{\alpha - 1} Z_2 \qquad (9)$$

On the other hand Z_{out} is the resistance of the four terminal network at terminals 2-2', if terminals 1-1' are open circuit.

$$Z_{out} = R_2 + R_3$$

Using Z_{out} in (9), and bringing R_3 to the right side gives the design equation for R_2 shown in Fig 4.

Resistor R_3:

If Kirchhoff's law is applied to the input circuit of the T attenuator this results in:

$$I_1 R_1 + I_3 R_3 = U_1$$

With the currents I_1 and I_2 in (4) and (5) the current is I_3:

$$I_3 = I_1 - I_2 = U_1 \left(\frac{1}{Z_1} - \frac{1}{\sqrt{\alpha Z_1 Z_2}} \right)$$

If I_1, I_3 and R_1 are used in the network equation it results in:

$$\frac{U_1}{Z_1} \left(\frac{\alpha + 1}{\alpha - 1} Z_1 - R_3 \right) + U_1 \left(\frac{1}{Z_1} - \frac{1}{\sqrt{\alpha Z_1 Z_2}} \right) R_3 = U_1$$

After simplifying for U_1 the equation gives the design equation for R_3 shown in Fig 4.

Minimum attenuation

The minimum attenuation occurs when the longitudinal resistor on the input side of the circuit is zero. If Z_2 is $> Z_1$, the resistance becomes $R_1 = 0$. Computing α can be achieved using the substitution ü $= \sqrt{\alpha}$ resulting in a quadratic equation. The attenuation in dB is:

$$A_{min} = 10 \lg (\alpha) = 20 \lg (ü).$$

$$R_1 = \frac{\alpha + 1}{\alpha - 1} Z_1 - R_3 = 0$$

Using the equation for R_3:

$$\frac{\alpha + 1}{\alpha - 1} Z_1 - \frac{2\sqrt{\alpha Z_1 Z_2}}{\alpha - 1} = 0$$

After simplifying the denominator ($\alpha - 1$) as well as simplifying for Z_1, the substitution is introduced to give:

$$\ddot{u}^2 + 1 - 2\ddot{u}\sqrt{\frac{Z_2}{Z_1}} = 0$$

Simplifying using $m = \sqrt{\frac{Z_2}{Z_1}}$ gives.

$$\ddot{u}^2 + 1 - 2\ddot{u}\sqrt{\frac{Z_2}{Z_1}} = 0$$

The solution of this quadratic equation is:

$$\ddot{u} = m + \sqrt{m^2 - 1} = \sqrt{\frac{Z_2}{Z_1}} + \sqrt{\frac{Z_2}{Z_1} - 1}$$

The two square roots contain the two characteristic impedances. It is assumed that $Z_2 > Z_1$.

7.3. PI attenuator

Resistor R_a

The resistor R_a is determined from the relationship of the reflection factors at input and output. In addition the output (2-2') is short-circuited (Fig 5). The reflection factor from (6) with RL = 0

$$\rho_2 = \frac{R_L - Z_2}{R_L + Z_2} = -1$$

The reflection factor at the input is:

$$\rho_{in} = \frac{Z_{in} - Z_1}{Z_{in} + Z_1} = \rho_2 \frac{1}{\alpha} = -\frac{1}{\alpha}$$

Re-arranging this equation for Z_{in}:

$$Z_{in} = Z_1 \frac{\alpha - 1}{\alpha + 1}$$

The resistance Z_{in} at the terminals 1-1' is obtained from the parallel connection of R_a and R_b:

$$Z_{in} = \frac{1}{\frac{1}{R_a} + \frac{1}{R_b}} = Z_1 \frac{\alpha - 1}{\alpha + 1}$$

Re-arranging this equation for R_a, gives the design equation shown in Fig 6.

Resistance R_b

The resistance R_b can be determined from Kirchhoff's first law applied to the input current:

$$I_1 = I_a + I_b$$

For the four terminal network:

$$I_1 = \frac{U_1}{Z_1} \qquad I_a = \frac{U_1}{R_a}$$

Using (3) gives:

$$I_b = \frac{U_1 - U_2}{R_b} = \frac{U_1}{R_b}\left(1 - \sqrt{\frac{Z_2}{\alpha Z_1}}\right)$$

Using these currents in the current equation above and re-arranging for U_1 gives:

$$\frac{1}{Z_1} = \frac{1}{R_a} + \frac{1}{R_b} - \frac{1}{R_b}\sqrt{\frac{Z_2}{\alpha Z_1}} = 0$$

The term: $\dfrac{1}{R_a} + \dfrac{1}{R_b} = \dfrac{1}{Z_1 \dfrac{\alpha - 1}{\alpha + 1}}$

is from that equation, and allows the calculation of R_a. Re-arranging for R_b results in the equation given in Fig 6.

Resistor R_c

To determine R_c the four terminal network is turned round and the reflection factors calculated for the new input and output. The terminals 1-1' are short-circuited. The reflection factor at the terminals 2-2' is:

$$\rho_2 = \frac{Z_{out} - Z_2}{Z_{out} + Z_2} = -\frac{1}{\alpha}$$

Re-arranging for Z_{out} results in:

$$Z_{out} = Z_2 \frac{\alpha - 1}{\alpha + 1}$$

The resistance Z_{out} at the terminals 2-2' is obtained from the parallel connection of R_c and R_b:

$$\frac{1}{\dfrac{1}{R_c} + \dfrac{1}{R_b}} = Z_2 \frac{\alpha - 1}{\alpha + 1}$$

Re-arranging this equation for R_c gives the equation in Fig 6.

Minimum attenuation

To design the attenuator for the minimum attenuation one of the two resistors R_a or R_c becomes infinite or one of the conductances $1/R_a$ or $1/R_c$ becomes zero. For $Z_2 > Z_1$

e.g. $1/R_c = 0$.

$$\frac{1}{Z_2} \frac{\alpha + 1}{\alpha - 1} - \frac{1}{R_b} = 0$$

Re-arranging the equation using R_b gives:

$$\frac{1}{Z_2} \frac{\alpha + 1}{\alpha - 1} - \frac{1}{\dfrac{1}{2}(\alpha - 1)\sqrt{\dfrac{Z_1 Z_2}{\alpha}}} = 0$$

The term α-1 in the denominator can be simplified with the substitutions $ü = \sqrt{\alpha}$

and $\quad m = \sqrt{\dfrac{Z_2}{Z_1}}$

after re-arranging the same quadratic equation as for the T attenuator is produced:

$ü^2 + 1 - 2üm = 0$.

The minimum absorption for T and PI attenuators is identical. An L half section with identical resistor values results in both cases.

Wolfgang Schneider, DJ8ES

Beacon controller using an ATmega32 and Bascom

The question of what controller to use for a new beacon project has arisen. Previously discrete circuitry would have been used but now a micro controller offers many advantages. However the problem of programming such a device can be an unsolvable problem for a radio technician

1.

General information

A minimum solution for a beacon is to control the CW keying and repeater hardware. This can be either only the call sign or different text strings. The necessary character sequences are uniquely programmed and sent nonstop at the desired speed. But a micro controller can do much more! Using the plentiful input and output ports many auxiliary functions can be simply realised. The hardware presented here covers the following:

- time control of the identification
- up to 4 different texts
- Monitoring of important parameters like operating voltage, power output and temperature
- Output of telemetry over the serial interface

Fig 1: Display during the starting sequence with date and time.
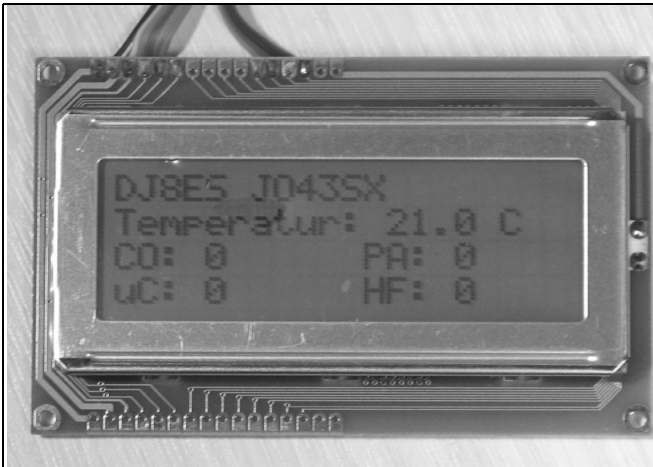
Fig 2: Display showing the operating temperature and all operating voltages.

- Remote control of various beacon auxiliary functions

To make this all possible several additional components are required around the micro controller. These are: a real time clock, a temperature sensor and naturally an LCD display to show the status locally. Sufficient input/output ports are already integrated into the micro controller. In addition there are more channels available such as an analogue to digital converter (ADC) and counters.

Micro controllers used to be programmed in machine code or assembler that is specific to the type of micro controller. Today various compilers and interpreters are available for most high-level languages. This makes the life much simpler and makes such projects feasible even for a radio technician.

2.

What is BASCOM?

BASCOM is a complete basic development environment for the AVR controller from ATMEL. It offers an efficient, user friendly, development environment. With the free demonstration version available on the Internet, everyone can produce code up to a size of 4 KB that uses up all of the memory of an AT90S2313. BASCOM AVR contains many complete functions that make it simple to use external components such as LCD displays. The demonstration version does not have any restrictions in relation to the full version except the program size. The full version can be found on the MCS Electronics web site [3].

3.

Description of circuit

The main component in the circuit of the beacon controller, shown in Fig 3, is an ATMega 32 micro controller (IC1) from ATMEL. This component is in a 40-pin DIL package that offers plenty of input/output ports for connection to peripherals for control functions and remote control. The 32k ROM in this controller has room for extensive system software. However this is not used at present for the beacon controller. The unusual clock frequency of 9.216MHz was selected rather than 8, 12 or 16MHz for the bit rates of the serial interface. The operating speed, related to the clock frequency, is
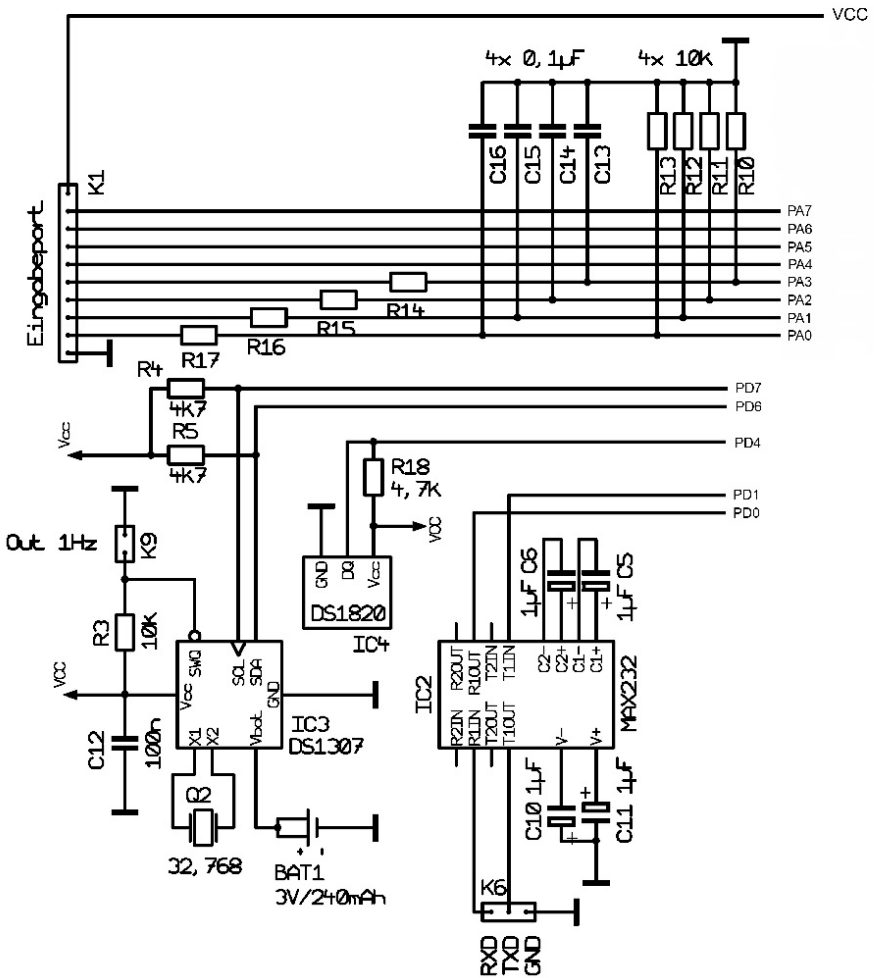
Fig 3: Circuit diagram of the beacon controller.
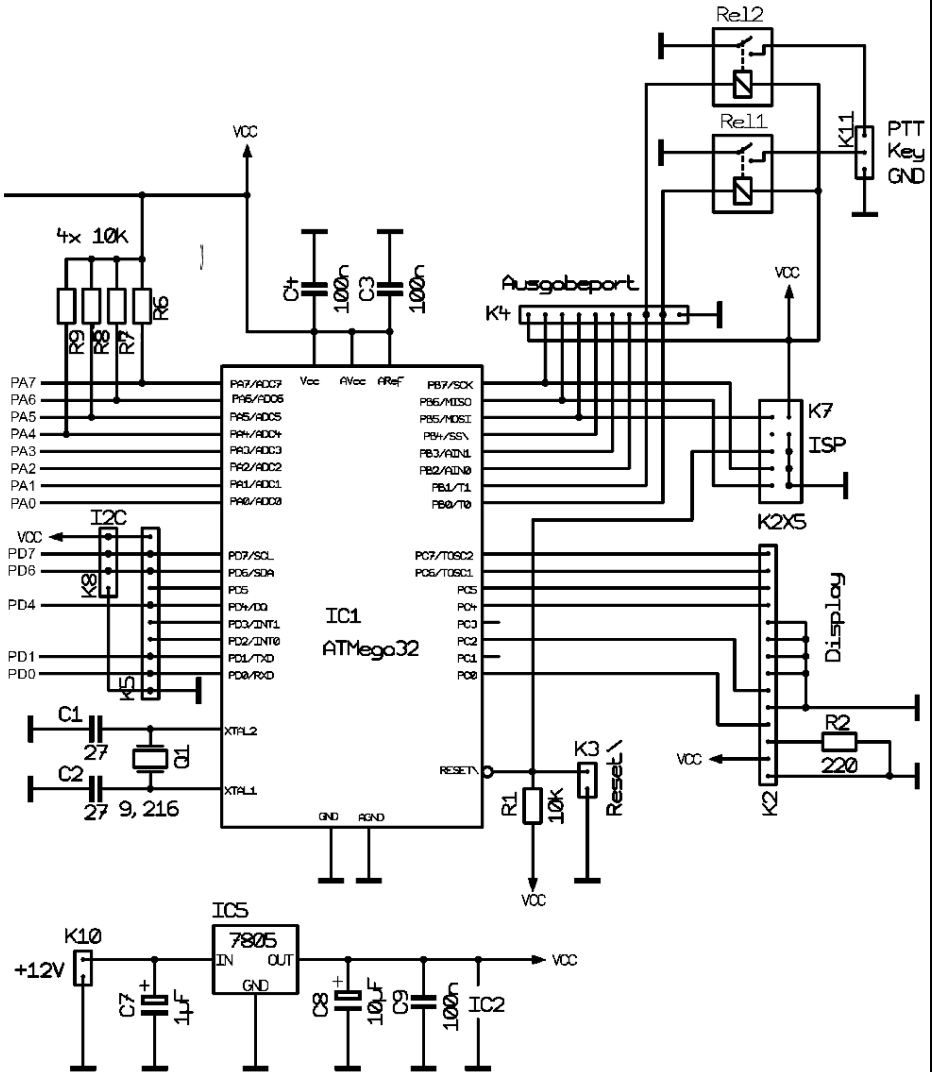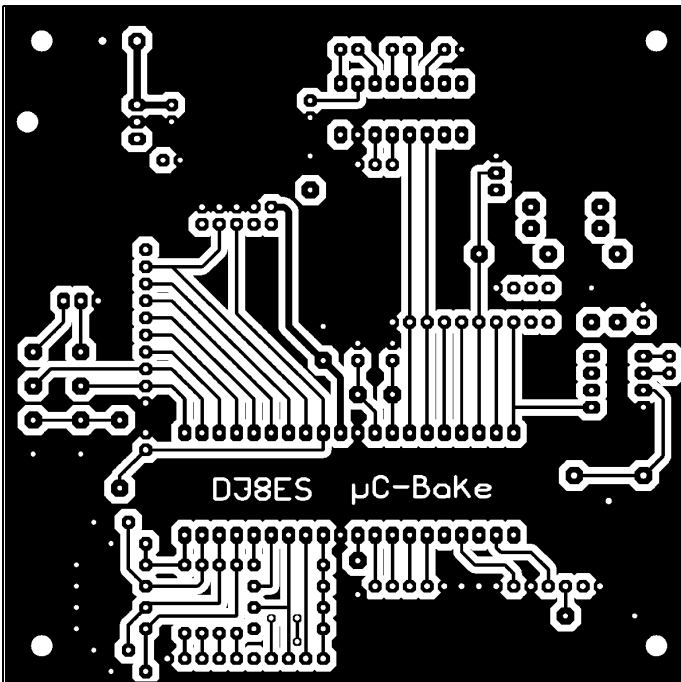
Fig 4: Bottom side PCB layout for the beacon controller.

of little importance for this application of beacon controller.

The serial port for the RS232 interface is connected to the contact strip K6. A PC can be attached directly to this port. Possible applications for this are telemetry or connection to a GPS receiver to obtain the exact time.

The beacon controller board has 4 free input ports (PA4 - PA7). This could be used to select different CW identification text. In addition other functions could be implemented in software for special beacon projects. In the project described here the 4 input ports (PA0 – PA3) are used to measure the power output of the beacon and monitor the operating voltages for the oscillator, output stage and the controller:

- PA0: 0 to 5V (Output power of the beacon)
- PA1: 0 to 15V (Output stage voltage)
- PA2: 0 to 15V (Oscillator voltage)
- PA3: 0 to 15V (Controller voltage)

The measuring range of the inputs is standardised to 0V to 5V using resistors. The system software takes care of the computation of the correct voltages. The current measured values can be displayed on the 4 lines by 20-character LCD display. This display is connected by the patch cord K2 (14-pole) directly to the micro controller on port C.

An important auxiliary component of the beacon controller is the real-time clock type DS1307 (IC4). This is in an 8-pin package offering all of the functionality required. It has an integrated clock, a calendar and 64Byte of static RAM with 56Byte free for the user (the clock needs 8 bytes) but is still very inexpensive. With an external 3V battery (1 x CR2032) connected to pin 3 this IC runs continuously at least 30 years. It continues to supply the component if the external operating voltage is switched off. In standby mode the power input is a maximum of 500nA. It is addressed over an I^2C bus or a TWI (twin Wire) bus. Pin
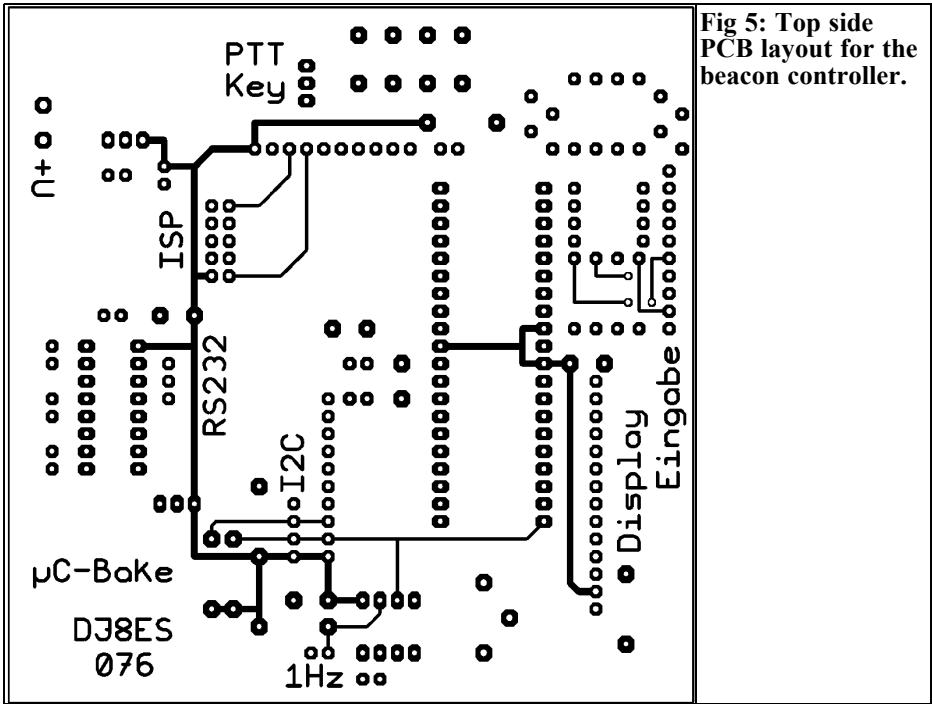
Fig 5: Top side PCB layout for the beacon controller.

5 (SDA = serially DATA) and pin 6 (SCL = serially Clock) are used to connect to the micro controller. The only other component required is a miniature 32.768kHz crystal to provide the accurate timing.

A temperature sensor type DS1820 (IC5) connected to the micro controller. This contains a very exact temperature sensor as well as the processing electronics that make it possible to attach this sensor directly to a PC or micro controller. This considerably simplifies temperature measurement. Similar temperature sensors use a temperature dependent of resistor or thermocouple, today in digital circuits digital sensors are used. These have the advantage that they are a similar size and not only contain the digital to analogue conversion but can be directly of controlled by the micro controller. Line resistances or noise that disturbs analogue sensors plays no more part with digital signals. The DS1820 has a 1-Wire interface requiring only one data line making it useful over large distances. More than that, several such sensors can be connected to one data line since each sensor has an individual serial number used to address it. Beside the data line the sensor must be connected to earth and the operating voltage between 3V and 5.5V. The measuring range of the sensors is from -55°C to 125°C with an accuracy over the entire temperature range of at least 2°C, between -10°C and 85°C it is only to 0.5°C.

A MAX 232A (IC2) is used for the serial interface that contains the necessary level adjustment. A modem or PC with an appropriate terminal program can be connected directly. Handshake signals are not used.
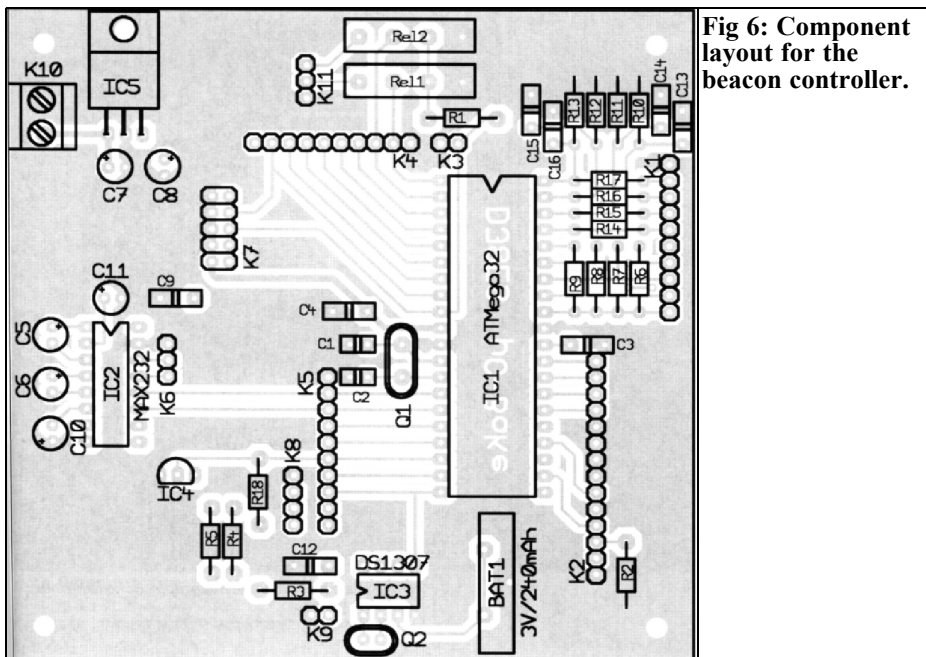
Fig 6: Component layout for the beacon controller.

4.

Program operation (software)

After switch on or a reset of the micro controller the program jumps into the starting sequence. The date, time and an information message are displayed (Fig 1). The data is updated with a 1 second pulse.

When the seconds go through zero or the PTT port is activated the desired text is sent in CW (e.g. DJ8ES JO43SX). A speed of 60WPM is programmed into the software but can be changed easily. Different text messages can be selected using the I/O ports. The current software limits the text to 20 characters. In each case a short break of 5 seconds is added at the beginning and at the end of the CW text. After 5 seconds the controller measures the operating temperature using IC4 the temperature sensor DS1820 and then the input ports (PA0, PA1, PA2 and PA3) are tested. The results of the meas-

urements are displayed for 15 seconds on the LCD display.

Finally the software switches to the starting sequence. If the PTT port remains active for a minute the beacon output is set to a continuous level.

The current version of the software is programmed in a straight line. Various possibilities exist for individual adjustment.

5.

Construction

The circuit beacon controller is shown in Fig 3 and is built on a 100mm x 100mm double sided epoxy PCB that is plated through in several places. The PCB layout is shown in Figs 4 and 5 with the component layout in Fig 6.

The assembly takes place in an informal order. Only the micro controller AT-
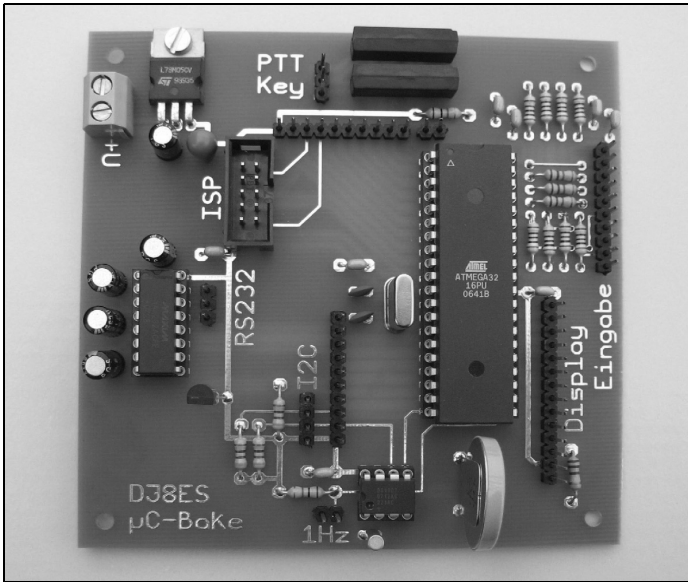
Fig 7: Completed PCB for the beacon controller without LCD display.

mega32 (IC1), two further integrated circuits (IC2, IC4) and the LCD display should be fitted after measuring the internal +5V supply at the output of the 7805 voltage regulator (IC3). Caution is required for the installation of the two relays. The protection diodes should be fitted as shown in Fig 8. Unfortunately the marking for these is missing from the PCB.

6.

Start-up

Before switching on a visual inspection should be carried out. The external supply voltage can vary between 8V and 15V DC. Usually for amateur radio use

Table 1: Parts list for beacon controller.

IC1	Micro controller ATMega32, ATMEL	Electrolytic capacitors 25V RM 2.54mm:	
IC2	MAX232A	C5...C7	1µF
IC3	DS1307	C8	10µF
IC4	DS1820	C10, C11	1µF
IC5	7805	Ceramic capacitors RM 2.54mm:	
1 x	LCD Display 4 x 20 characters	C1, C2	27pF
Q1	9.216MHz crystal	C3, C4, C9	100nF
Q2	32,768kHz miniature crystal	C12...C16	100nF, RM 5mm
Relay	2 Reed relay 12V, with Diode	Resistors 1/4W RM 10mm:	
BAT1	Lithium battery 3V / 240mAh	R1	10kΩ
K10	Screw terminal 2 pole, RM 5mm	R2	220Ω
Dual row contact strip RM 2.54mm:		R3	10kΩ
K7	2 x 5 pole	R4, R5	4.7kΩ
Single row contact strip RM 2.54mm:		R6..R13	10kΩ
K1, K4, K5	10 pole	R14	0Ω (see text)
K3, K9	2 pole	R15...R17	22kΩ
K6, K11	3 pole	R18	4.7kΩ
K2	14 pole		
K8	4 pole		
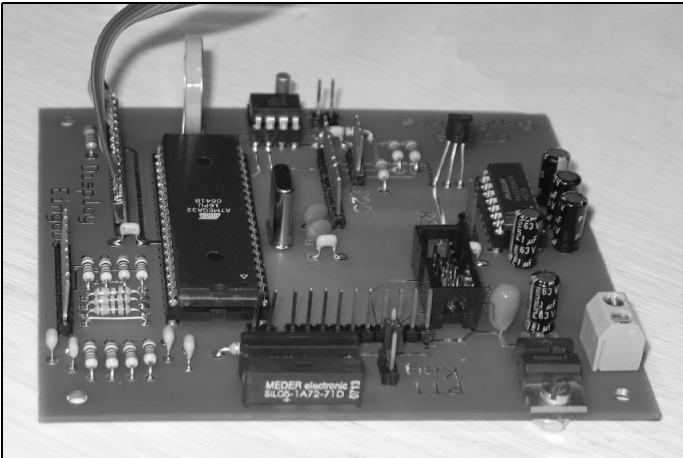
Fig 8: Completed PCB for the beacon controller showing the relays fitted correctly.

this will be 13.8V. The supply current should be 50mA with the LCD display connected.

Importantly, the real-time clock (IC3) must be programmed for the current date and the time. After this it will run with the external backup battery (1 x CR2032) for over 30 years. A useful addition, regarding running accuracy, is to synchronise the clock with a GPS receiver using the serial port.

The software for the beacon controller is in a file "Bakensteuerung_V1-0. bas" and additionally requires the file for the real-time clock "Subroutine_RTC.bas". These can be downloaded from the VHF Communications Magazine web site.

The micro controller can be programmed by ISP (In-System-Programming), this is briefly described in the development environment of BASCOM. However this is not possible with the demonstration version, the full version is needed.

7.

References

[1] Data sheet ATmega32, ATMEL Corporation, http://www.atmel.com/

[2] Morse code Beacon, C. Potma, PA3CKR, http://www.qsl.net/pa3ckr/

[3] MCS Electronics web site: http://www.mcselec.com/

John Fielding, ZS5JF

John's Mechanical Gem No. 3

Semi-rigid coaxial cable is widely used in µwave systems. Of the many types in use the UT-141 type is the most common. UT-141 is available in 25Ω, 50Ω and 75Ω impedance from several manufacturers.

For those who do not know, the UT prefix stands for "Uniform Tubes Inc"; the American Company who invented the product and the 141 is the outer diameter in inches. UT-141 is 0.141 inch in diameter (3.6mm). The common sizes above and below the 141 are 085, which is 0.085 inch in diameter (2.16mm) and 250, which is 0.25 inch in diameter (6.35mm). These types are UT-085 and UT-250. UT-141 is more commonly known today by its RG-402 designation.

UT-141 is commonly used with SMA connectors and the coaxial cable centre conductor is the exact size needed to interface with the female version of the SMA connector. Many SMA male plugs do not come with a centre pin, unlike a BNC or N type (Fig 1). The centre conductor is chamfered with a special

Fig 1: Commercially manufactured UT-141 cable and male connector.

tool to form the pin. For the 50Ω version the centre conductor is 0.0362 inch diameter (0.91mm).

To prepare a piece of UT-141 to accept a male connector therefore requires the user to do several things.

Firstly, we need to remove a section of the outer copper jacket and then trim back the PTFE insulator and finally chamfer the centre to form the centre pin. You can buy special tools for these applications but they are expensive.

A cheaper method

To remove the outer copper jacket is not as simple as it seems. The UT-141 coaxial cable is made in the same way that seaside rock is made. It starts with a much larger diameter and this is progressively reduced in diameter by drawing the piece through dies. This reduces the diameter and the piece increases in length. The proper tool is a miniature version of a pipe cutter with a revolving hardened steel blade, which is forced into contact with the copper jacket. As the tool is rotated it cuts a helical groove the depth of the copper jacket thickness and it is set to the required amount by a stop collar fastened on the outer jacket behind the tool. When the tool butts up against the stop collar the movement along the tube is stopped and the cut is square to the outer jacket. However, such a tool costs in the region of $750.

To remove the outer jacket we need a sharp knife; a craft knife or penknife serves us well. A nick or groove is formed around the tube and then the end is gripped with a pair of pliers and the end bent backwards and forwards to

Fig 2: Scoring the outer jacket with a knife.

cause the copper jacket to snap off.

The best way to do this is to press down on the tube with the sharp knife and roll the tube on a flat surface so the groove is applied all the way around the tube (Fig 2).

Alternatively, a sharp pair of side cutters will do the same thing if the cutters are lightly clamped onto the tube and the side cutter is rotated through 360° several times. When the side cutter blades are almost through the thin wall tube waggling the end backwards and forwards with a pair of pliers causes the tube to break away.

Do not let the knife blade go all the way through the copper jacket, or the PTFE insulator will be weakened and the copper jacket and the PTFE insulator will be

Fig 3: The piece of outer jacket removed.

removed as one piece. Gripping the piece tightly and pulling will make the piece of copper jacket pull away leaving the exposed PTFE insulator.

Having removed the copper jacket the PTFE is treated in the same way except that we need to cut down to the centre conductor. Gripping the PTFE with a pair of pliers we twist the insulator off the centre conductor (Fig 3).

The exact dimensions we obtain from the connector manufacturer data sheet. In some connectors the PTFE is cut off so that it protrudes about 0.5mm proud of the copper jacket, others have the PTFE flush to the jacket, so we need to study the data sheet for the connector.

To finish the centre pin requires a fine file; a toolmakers flat warding file is the best tool (Fig 4). Rotate the piece of UT-141 and apply the file to the end to form a taper for about 1mm with an angle of 45°. Again a special tool is sold for this application, known as a pointing tool, but for the small number of cable assemblies needed by amateurs the expense isn't warranted.

Finally the outer jacket is abraded to remove any oxidisation with a piece of steel wool and the connector body soldered in place, remembering to slide the nut on first if it is a two-part connector!

Fig 4: UT-141 cable prepared for fitting the SMA male connector. The warding file and PTFE insulator are shown.

Gunthard Kraus, DG8GB

Internet Treasure Trove

Skyworks Solution

This web site has many items to browse or download. There are detailed documents and application notes on components and modules for microwave engineering. To access these quickly choose the sub menu "Technical library".

Address:
http://www.skyworksinc.com/
products_literature.asp

IMEC

The introduction to this site says: IMEC is Europe's leading independent research center in the field of micro and nanoelectronics, nanotechnology, enabling design methods and technologies for ICT systems. This site has much of interest including information on Software Defined Radio (SDR) and a Direct Conversion Receiver for 5GHz.

Address:
http://www.imec.be/wireless/zero-if/

Effelsberg Radio Telescope

For anyone who wants to know about the concepts used for extremely low-noise receivers between 400MHz and 100GHz, take a look at this homepage. The information is sorted by frequency range with one document per receiver and the associated antennas. Radio telescopes work at the physical limit of receiver technology and it is interesting to study how they move closer and closer to these limits.

Address: http://www.mpifr-bonn.mpg.de/
div/effelsberg/receivers/receiver.html

National Semiconductors school of analogue design

That is a great site for almost all areas of analogue techniques including: operation amplifiers, mixer, DC to DC converters… Please look at the on-line courses, in particular the wireless topics.

Address: http://www.national.com/AU/

CMOS Self Mixing Free front end for Direct Conversion Applications

This publication from University of Hong Kong has good presentation and accurate information.

Address: http://www.ee.ust.hk/~eezzf/
download/frontend_iscas01.pdf

A simple printed broadband antenna

This is an informative and clearly written article on microstrip antennas with good suggestions on this topic.

Address: http://ceta.mit.edu/pier/pier60/07.0509263.Gao.S.pdf

Terabeam HXI

Sounds like Science Fiction, it is however a company who specialise in millimetre wave systems and components. Beside documents on their products there are interesting applications. For example: Frequency Multiplier Application Notes and Gunn Oscillator Application Notes.

Address: http://www.terabeam-hxi.com/prodserve/notes.php

Universitiy of Berkeley California

It is always interesting to search the pages of universities. This homepage has many interesting topics.

Address:
http://kabuki.eecs.berkeley.edu/papers.html#rf

Embedded Antennas get the signal

This is a strange title but it is a very interesting article. It describes the theoretical background and the production technology of this new technique.

Address: http://www.edn.com/contents/images/236435.pdf

RF Avenue

This site is a joy to visit, it has an unbelievable array of topics for developers. The heading "RF and Microwave Resources" says it all.

Address:
http://www.newwaveinstruments.com/resources/rf_microwave_resources/digital_wireless_communication_directory.htm

Philips RF manual

It is a good idea to put a manual on the design basics for RF circuits on the Internet. This is an excellent document containing lots of useful information.

Address: http://www.nxp.com/acrobat_download/other/discretes/philips_rf_manual_3rd_edition.pdf

QTH Viewer for Google Earth

Darek Milka, SP6NVK has produced a QTH Locator Square overlay for Google Earth. Using the link below select "QTH Locators viewer" to download the .kmz file. When this is used for the first time with Google Earth it produces an overlay of the main QTH locator squares over the globe. Under "My Places" there are three check boxes to control the overlay. A tick in the subdivisions check box will generate subdivision squares.

Address: http://www.hamatlas.eu/Gg/ KML_en.htm

Ecoflex coax cable

For those in the UK the low loss coax cables from SSB electronics can be purchased from Diode Electronics. There is a range of cables in the Aircell, Ecoflex and Aircom ranges. Ecoflex 15 offers 9.8dB/100m loss at 1GHz with a maximum frequency of 6GHz, Aircom plus offers 13.4dB/100m at 1GHz with a maximum frequency of 12GHz.

Address: http://www.diodecomms.co.uk/

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

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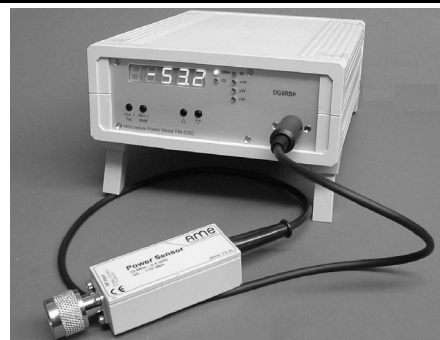
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