A simple transverter for the 1.3GHz band
Rafal Orodzinski, SQ4AVS

Contents

I am pleased to say that some good articles arrived to fill this issue. I put in an article originally published in RadCom because it is a very interesting piece of hardware and a good "stocking filler" for Christmas.

If you have a project that you think would be suitable for the magazine please contact me and remember to subscribe for 2012.

Merry Christmas and a Happy New Year 73s - Andy

Rafal Orodzinski, SQ4AVS

A simple transverter for the 1.3GHz band

1.0
Introduction

The main design criteria of this system were the greatest possible simplicity of the system while maintaining good performance and low price. The transverter uses the minimum number of tuned elements (no transformers or helical filters after the mixer). The new solution greatly simplifies the construction and commissioning of the equipment. The system is modular for easy start-up and subsequent experiments.

2.0
The phase locked loop

Phase locked loop performance is achieved by using a high frequency standard, a good quality swept oscillator, the appropriate filter bandwidth and a low noise phase detector. A phase locked loop is usually built using a specialised microprocessor IC that often complicates the whole construction unnecessarily. The PLL circuit described here (the original circuit diagram has been split into three sections shown in Fig 1, Fig 2 and Fig 3) contains only a readily avail-

able integrated circuit and no microprocessor. The generator operates at a frequency of 1152MHz (1152MHz + 144MHz = 1296MHz) and produces 2 - 3mW output power. The oscillator is a BFR92 transistor, oscillation occurs because of positive feedback between the base and emitter. The frequency stabilising element is a transverse wave (TEM-type) ceramic resonator. This type of resonator can be obtained from the band filter or duplexers of a mobile phone with a 900MHz GSM band. It is important that the filter was built with a single resonator and not a single ceramic block

In my case it was necessary to shorten the resonator from 9mm to approximately 6mm to get the desired frequency. The resonator is shortened from the wire pin end. This operation requires some time and the easiest way to do it with the appropriate cutter. It is also possible to buy a resonator for the desired frequency. The frequency of the resonator should be about 20% higher than the frequency synthesiser (mounting capacity decrease the frequency). A tunable oscillator is implemented using an MA4STK250 diode, this provides a low phase noise (for voltages above 2V it's quality factor is 400). The tuning of the system depends on the amount of capacitive coupling of the resonator (C50), it should be as small as possible, however, it provides compensation for thermal effects. With the
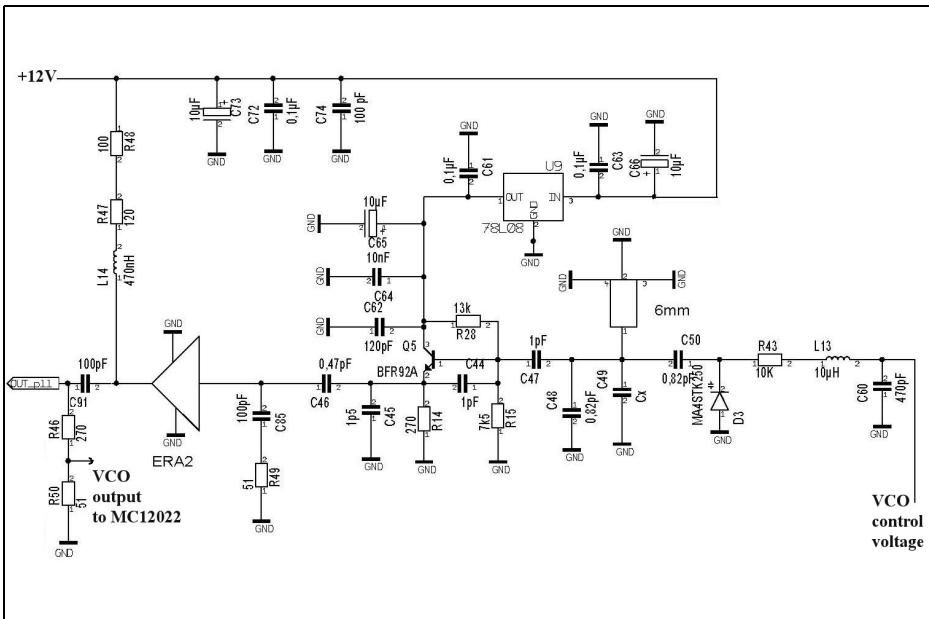
Fig 1: VCO section of the phase locked loop circuit.
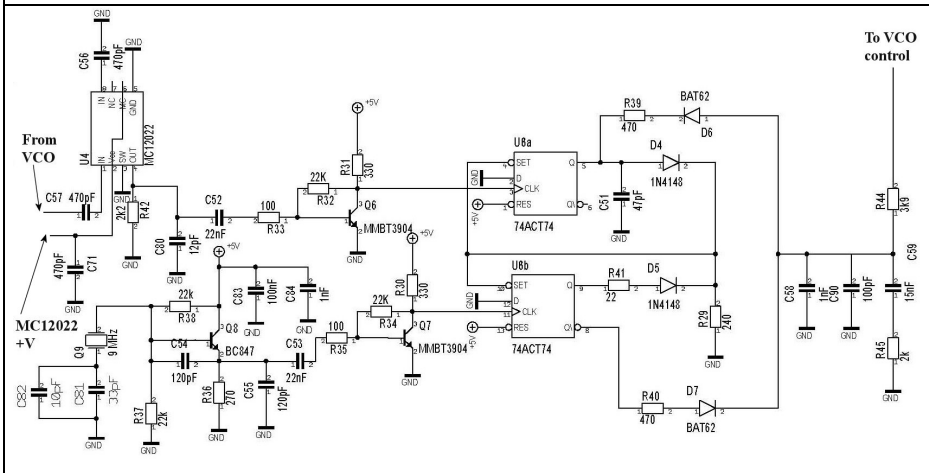
Fig 2: Divider and phase detector section of the phase locked loop circuit.

increase in voltage the output frequency of the oscillator increases. A monolithic microwave amplifier ERA2 provides isolation from the VCO and gives a reasonable output. A 120Ω and a 100Ω resistor in series are used to set the operating point for the ERA2 so that the power dissipation is not exceed for the resistors. The output level of the ERA2 amplifier is appropriate for driving the prescaler divider.
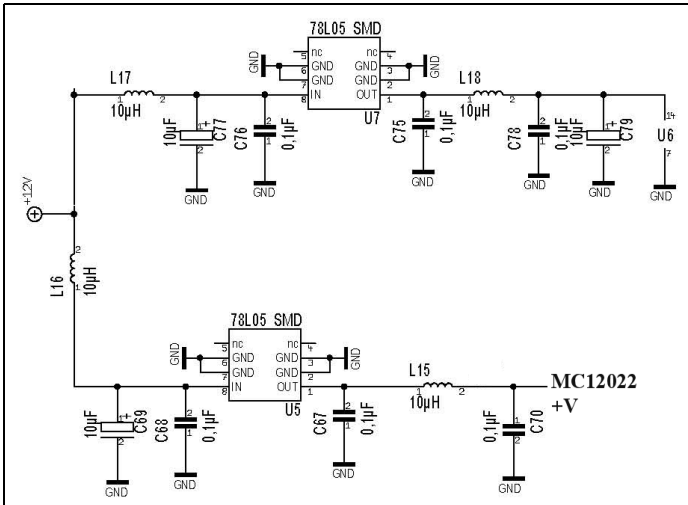
Fig 3: Power supply section of the phase locked loop circuit.

The VCO signal is divided by 128 using an MC12022 prescaler. At 1152MHz this gives an output frequency of 9MHz. An MB507 can also be used, however, different programming pins are required for division by 128. The divided VCO signal is compared in a phase-frequency comparator made from 74ACT74 logic flip-flops. The correct signal level to drive the 74ACT74 gates is provided by amplifiers using MMBT3904 transistors. They cannot be replaced the BC series transistors but with other suitable pulse types only. The reference signal is derived from a 9MHz quartz crystal oscillator. The 9MHz oscillator uses a BC series transistor, it is not applicable, even accidentally, to use a pulse transistor in this circuit because it would generate too much 3rd overtone (around 27MHz). The PLL filter bandwidth is 15kHz due to the noise properties of the VCO oscillator. The prescaler can be used with other division factors but the PLL filter components must be recalculated.

The PLL filter elements should be selected for each individual VCO (with resonators cut by hand each one is slightly different). In order to ensure a large attenuation of the reference frequency a 3rd order filter is used.

Disconnect the PLL output filter from the VCO and connect a 10kΩ potentiometer from ground to +5V with the slider connected to the VCO tuning input. Set the VCO tuning voltage to 3V and by selecting capacitor Cx obtain a frequency of 1152MHz (sometimes it is easier to slightly increase or decrease the return coupling capacitor C50). The oscillator is tuned by fractions of a pF. You can also use a good quality tubular trimmer, this solution is the simplest and least cumbersome in practice. The frequency of the system can be increased slightly by removing metallisation near the pin end of the resonator however it cannot be replaced. Removing too much copper (silver) makes the Q of the resonator worse. Next set two different voltages: approximately 2V and 4V and measure the frequency and use them in PLL filter calculations (a computer program can be used for this). To calculate the PLL filter components the free program AdiSimPLL [1] from Analog Devices can be used. Your voltage and frequency values can be entered into the AdiSimPLL program file to calculate the PLL loop filter components. Fig 4 shows the program settings. Solder in the calculated values of PLL filter components.

Use the potentiometer to increase the frequency of the VCO. The output volt-
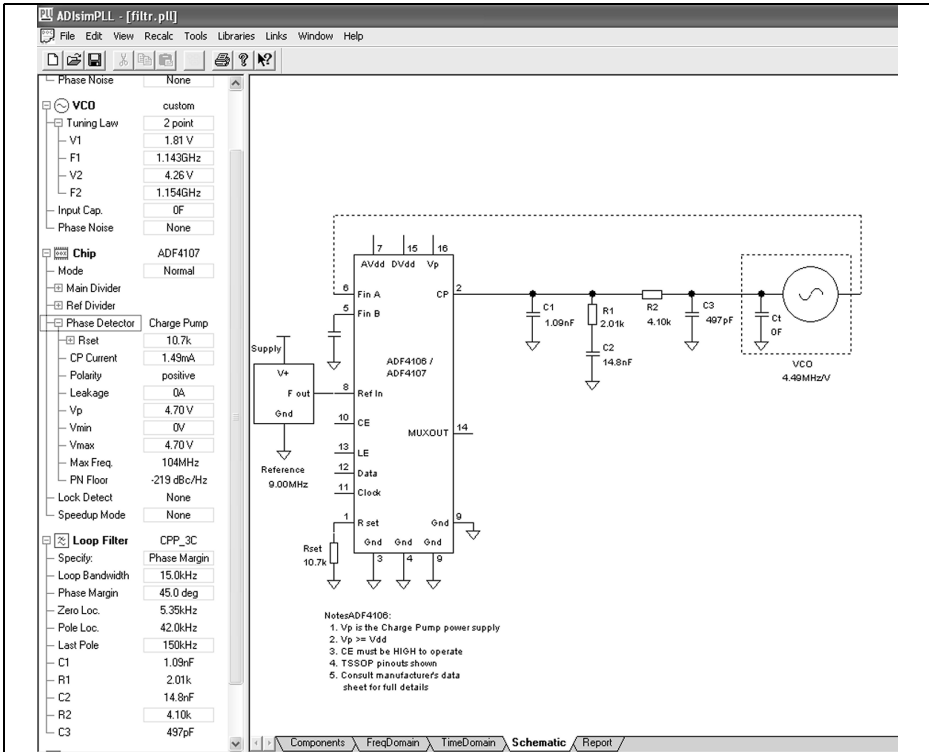
Fig 4: A screen shot of AdiSimPLL in action.

age after the low-pass filter should be about 0.3V when the output frequency of the prescaler is higher than the reference frequency, in our case, 9MHz. Then reduce the frequency of the VCO using the potentiometer, the voltage at the output of low-pass filter should increase to 4.7V when the output frequency of the prescaler is lower than the reference frequency, in our case, 9MHz. After the test, remove the potentiometer, reconnect the output of the PLL filter and the system should work flawlessly and stabilise the desired frequency. The output frequency of the prescaler should be equal to the phase detector reference frequency (9MHz). The exact frequency adjustment can be done with the 22pF capacitor (C82) in the crystal oscillator. The spectrum of the generator is shown in Figs 5 and 6, the -50dB level of noise

in Fig 5 corresponds to the noise used by the spectrum analyser (very narrow band filters and the analyser with a long scan time).

3.0

The transverter

The transverter is a typical design and generally does not require a broader description. It has separate transmitter and receiver circuits making it very easy to connect an external amplifier. Figs 7 and 8 show the circuit. The receiver input amplifier uses an ATF10136 transistor application according to the project developed by Reiner Bertelsmeier, DJ9BV.
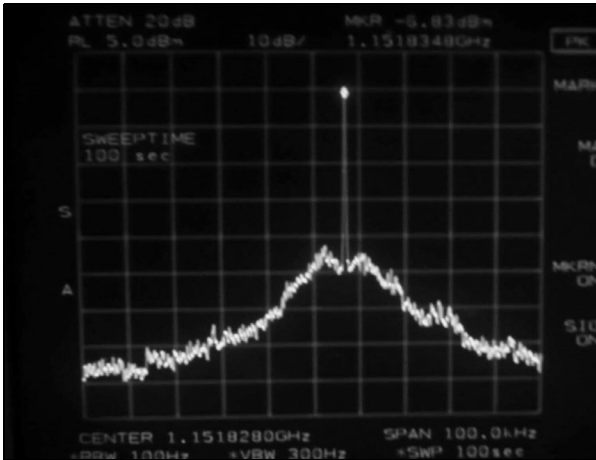
Fig 5: Phase locked loop output spectrum.

The advantages of this design are: low noise, good stability and good availability of the transistor used. The input capacitor should be 3.3pF ATC due to dielectric losses. The amplifier can be tuned for minimum noise with L4. The amplifier filter is tuned with a microwave trimmer, it improves the mirror signal attenuation (which is still around 50dB without this item). The ATF10136 transistor amplifier is followed by an INA 10386 MMIC amplifier.

The receiver amplifier circuit uses TEM bi-directional filters identical the resonators used in the PLL generator. These two resonators give an image frequency attenuation of approximately 50dB. The resonators have a different length from the resonator used in the PLL (cut to a different length). Filter tuning is done using a miniature trimers. The 22nH inductor in the filter circuit provides attenuation of higher frequencies. A filter without this coil would pass the odd multiples of the signal. Irek, SQ5MX measured the image frequency rejection given by the two TEM resonator filters, using a spectrum analyser, as 40dB (first version of transverter) in this version image suppression is 50dB. The resonators in the filter and PLL must be soldered to PCB ground along their entire length on both sides.
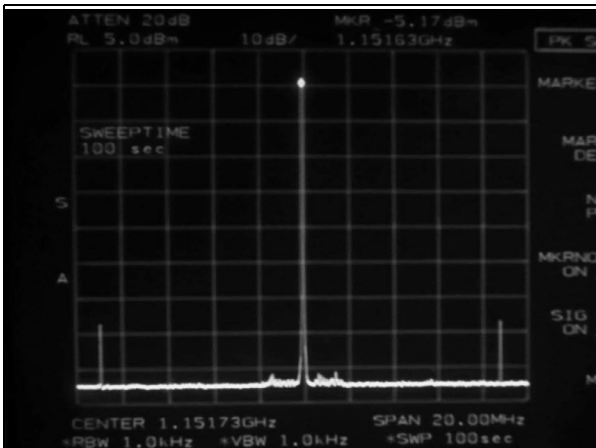
Fig 6: The phased locked loop spectrum showing that there are two symmetrical carriers at -67dB, which is a good value. This effect is a characteristic of the PLL.
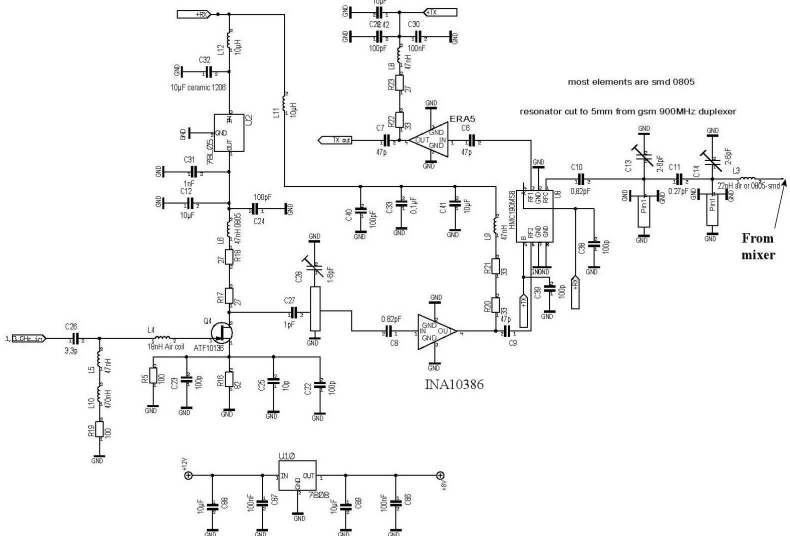
Fig 7: Filter, TX output and RX preamplifier section of the transverter circuit diagram.
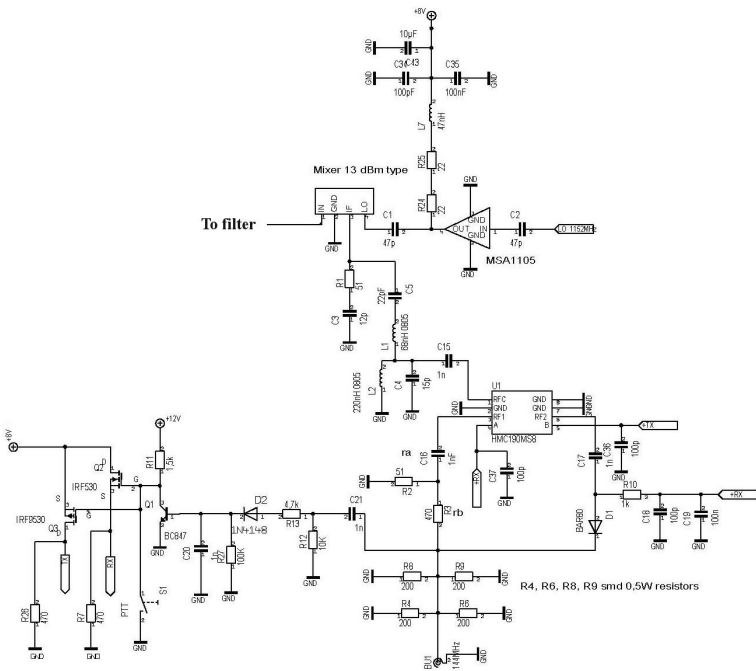
Fig 8: Mixer and switching section of the transverter circuit diagram.
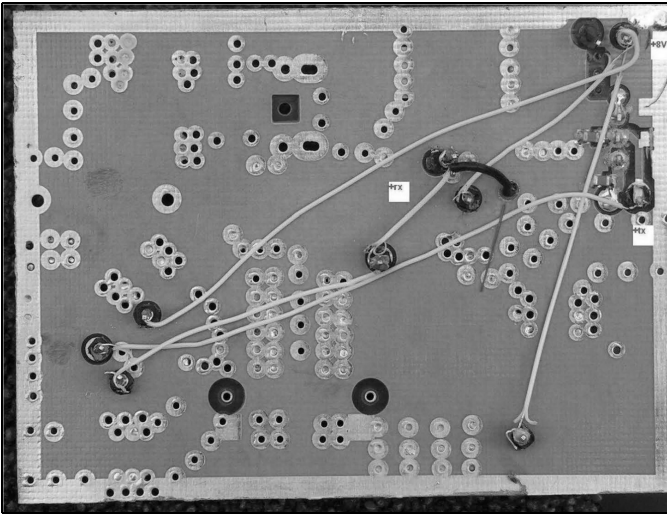
Fig 9: View of the bottom side of the transverter showing control voltage feeds.

The transmitter circuit uses an ERA5, it gives a gain of about 20dB.

Switching between transmit and receive is done using HMC190 semiconductor switches, their advantage is low cost and good dynamic performance. The system needs two voltages 0 and 8V in antiphase to control the inputs and the operation of the switches (that is if one of the inputs is 8V, the second must be 0V and vice versa). The system also uses an additional BAR80 diode switch that provides additional isolation during transmission to protect the HMC190 switch against large signal levels from the transmitter.

R3 reduces the power from the 2m transceiver, you must set the correct value to protect the transverter. The local oscillator signal is amplified to around 25mW by an MSA1105.

The transverter is constructed on a double sided FR4 PCB laminate with a thickness of 0.8mm. The control voltages are fed from the bottom of the PCB as shown in Fig 9.

PTT should be controlled by a sequencer, in my opinion RF vox should not be used even though it is often used. I recommend a sequencer such as the DB6NT
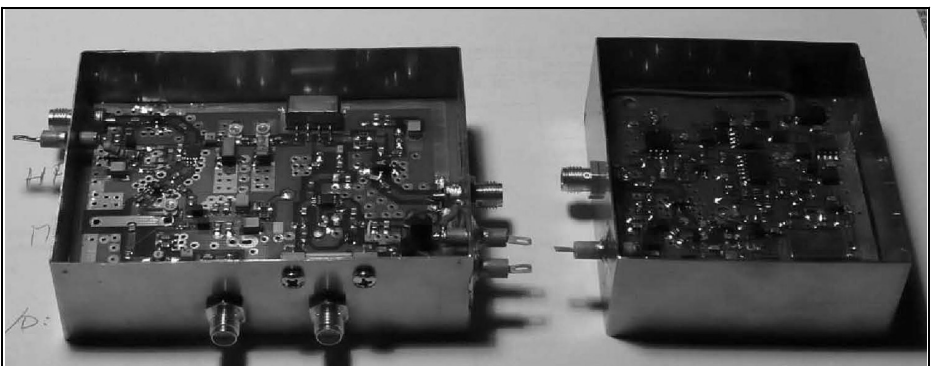
Fig 10: Picture of the completed transverter and phase locked oscillator.
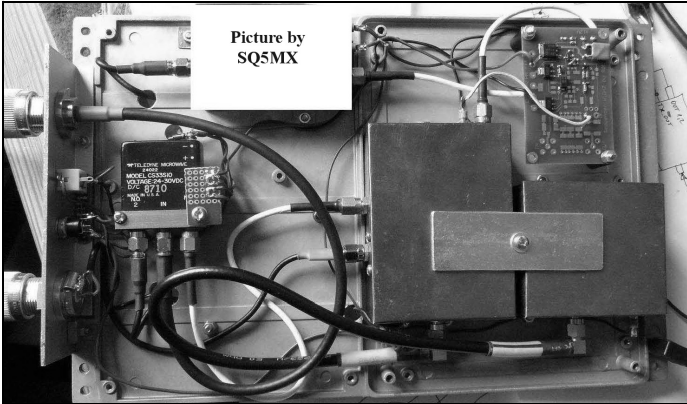
Fig 11: Photograph of the completed transverter in a housing.

offering [2]. A demonstration of communications via satellite using this transverter can be found on YouTube [3].

Finally, but by no means least, I want to thank Andy Barter, G8ATD, for translation of the text, Irek Szumski, SQ5MX for critical comments, performance measuremants and photos and Waldek Sznajder, 3Z6AEF for help correcting this text.

4.0

References

[1] Rafal Orodzinski, SQ4AVS presents all details of the project on his web site: http://sites.google.com/site/sq4avs/

[2] http://www.db6nt.de/

[3] The transverter can be seen in action on: http://www.youtube.com/watch?v=_xgzRrE1S70
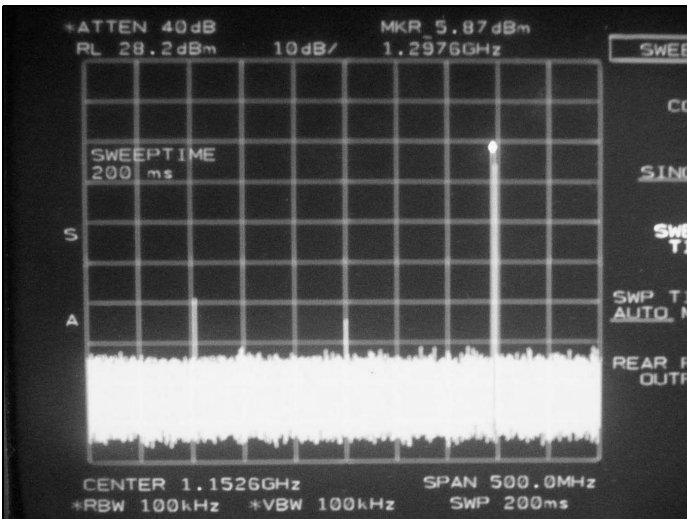
Fig 12: The 1296MHz output spectrum. On the right, the LO 1152 + 144 MHz, in the middle the local oscillator signal (1152MHz) and on the left the LO 1152 - 144MHz = 1008MHz.

Gunthard Kraus, DG8GB

Practical project: Development of a 50MHz to 2.5GHz wideband MMIC amplifier, Part 1

This article demonstrates the modern technique "from idea to the finished RF product" with all the important considerations, simulations and necessary experiences during the conversion to a working printed circuit board. A collection of instruments was used to test the characteristics of the amplifier so that all the vagaries at high frequencies would not be missed. The basis for this project by the author was a laboratory amplifier for the Baden-Wuerttemberg Dual High School in Friedrichshafen.

1.

Specification

1.1 How is it derived?

A product specification must be present before any work or thought can be put into a draft circuit. Product specifications are always the result of an analysis of the characteristics required for the component to carry out a specific task in a complete system or a special requirement e.g. as an important aid for measurements.

In this case, beside the educational requirement, the basic requirement was a preamplifier for noise figure measure-

ments with a spectrum analyser (also see [1] and [2]). The list of the requirements was:

- 3dB bandwidth: 50MHz to 2.5GHz

- The gain still present at 3GHz is to be examined

- Flatness: ±1dB

- Gain: 20dB minimum

- Input and output impedance: 50Ω

- Reflection attenuation: better than 10dB required

- Maximum power output (P1dB): +5dBm minimum

- Maximum permissible input level without compression: -20dBm

- Output 3rd order intermodulation point (OIP3): +15dBm minimum

- Noise Figure NF: under 5dB

- Absolutely stable (no oscillation, with any source or terminating resistance)

- Supply voltage: +5V (via an SMB socket); Sufficient wideband isolation of this supply voltage.

- Double sided printed circuit board with complete copper lower earth plane.

Features

- **Operating frequency: DC ~ 3.5 GHz**
- **21.5 dB gain**
- **VSWR < 2.0 throughout operating frequency**
- **9.8 dBm output P1dB**
- **3.3 dB noise figure**
- **Unconditionally stable**
- **Single 5V supply (Id = 35 mA)**
- **Lead-free option available**

Fig 1: The amplifier characteristics from the data sheet to be compared with the product specifications.

Fig 2: The amplifier characteristics taken from the data sheet.

- Material: Rogers R04003 (32MIL = 0.813mm thick) with 35μm copper cladding

- Printed circuit board size: 30mm x 50mm with 4 2.7mm mounting holes on a grid of 24mm x 44mm. Mounted in a milled aluminium housing with shielding cover.

- Connection of the topside of the printed circuit board to the lower earth plane by as many plated through holes (vias) as possible. They are implemented with hollow silvered rivets on the prototype.

- Continuous micro strip line with Z=50Ω (Guided Coplanar Waveguide) from the input to the output of the circuit. Only interrupted by DC isolation capacitors and the amplifier module.

- RF input and output by SMA sockets with two-hole or four-hole mounting flanges screwed from the outside onto the aluminium housing. Their inner connection rest directly on the associated micro strip lines of the printed circuit

board.

- The circuit to be designed compact enough so that an additional second identical stage could be fitted on the printed circuit board.

1.2 Explanation of the product requirement specifications

MMIC means Monolithic Microwave Integrated Circuit; this is a microwave IC with all components on a chip. The number of these components available is very large and ever more semiconductor manufacturers are entering this area. More and more chips are being manufactured with gallium arsenide or in germanium technology instead of silicon (these give less noise).

After examining the data sheet for the ABA-52563 from Agilent-Avago it was selected. It is offered at a reasonable price and delivery via the Internet. Delivery is usually the greatest problem with such components. Fig. 1 shows its most important characteristics. These are now examined in sequence and compared with the product requirement specifications.

The term "broadband amplifier from
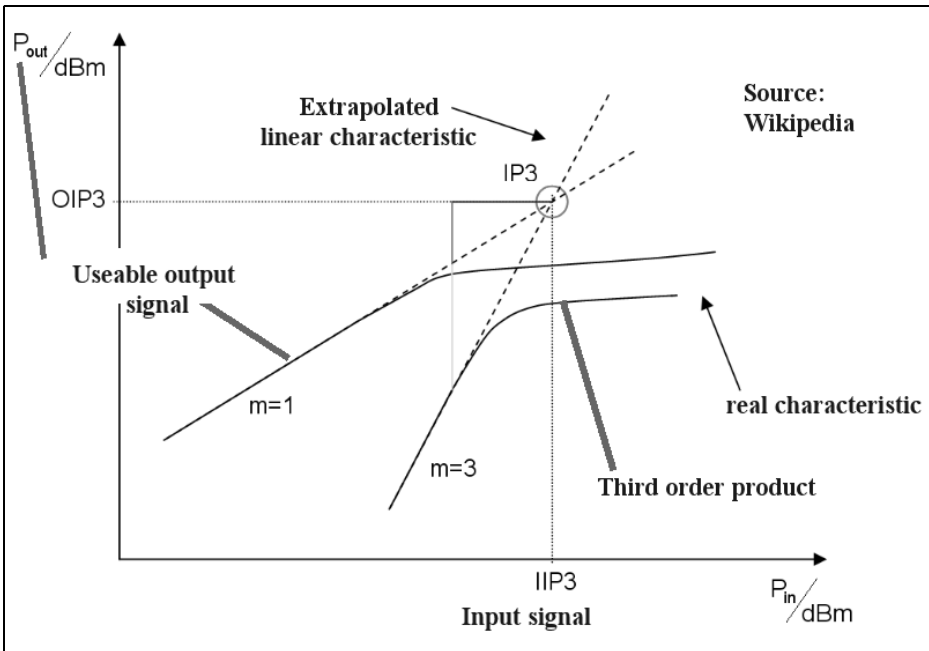
Fig 3: This shows the IP3 point with reference to the input signal.

50MHz to 2.5GHz" means that the gain at the 50MHz 3dB cutoff frequency should be the same as the gain at 2.5GHz. Between these it is to stay as constant as possible with at least +20dB gain with fluctuations up to 1dB allowed over the entire frequency range. Fig 2 shows that this component fulfils these demands.

The input and output impedance should deviate as little as possible from 50Ω over the entire frequency range. This is expressed by the requirement for as small a reflection factor "r" as possible (usually selected by the indication of the maximum permissible values of the S parameters S11 and S22). These represent the associated reflection factors, but expressed in dB - and this db value must be as large a negative number as possible, reaching the greatest reflection at a value of 1 = zero dB.

The VSWR (Voltage Standing Wave Ratio) is often used; it is associated with

$$|r| = \frac{VSWR - 1}{VSWR + 1}$$

the reflection factor as follows:

The maximum reflection factor for the MMIC port can be calculated from the value in the data sheet of VSWR < 2.0 throughout the frequency range:

"r "= (2 -1)/(2 + 1) = 1/3 = 0.33 = 33%

The maximum values for S11 and S22 are therefore:

$S11_{max} = S22_{max} = 20 \cdot \log(r) =$
$= 20 \cdot \log(0.33) = -9.6dB$

If the input power of an amplifier is increased it finally reaches limiting. Suddenly the power output stays constant and no longer constantly rises. From the data sheet the value for the 1dB compression point (P1dB) is where suddenly 1dB of output power is missing. For the ABA-52563 this is:

Fig 4: This shows that the noise figure in dB directly degrades the signal-to-noise ratio after amplification.

9.8dBm output P1dB

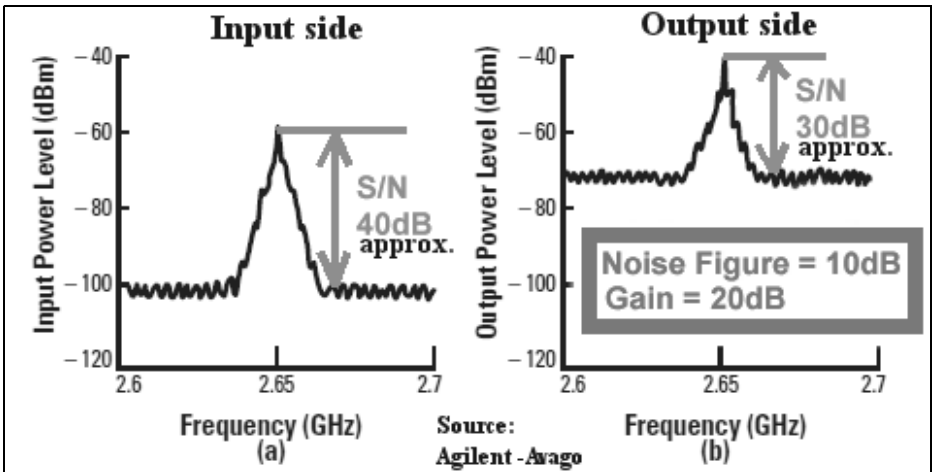The associated input level is calculated by subtracting the gain in dB from the output level. In the product requirement specifications +5dBm for P1dB is specified - thus an input level of more than:

+5dBm – 20dB gain = -15dBm

it is close to the requirement. The default of a maximum input level of –20dBm is realistic.

Before limiting is reached, the distortions at the output begin to rise. The third order products produced are particularly unpleasant, since they rise 3 times faster than the input signal.

The following applies:

If the input level is increased by 10dB, the distance between the usable output signal and the third order products decrease by 20dB.

Before these two characteristics intersect, the limiting discussed begins. If these two straight lines are extended to their intersection they come to the "Third Order Intermodulation Intercept Point" or IP3. It can be specified for the input or output side, but for the output side it is shown as OIP3. This is shown very clearly in Fig 3. The practical meaning of this diagram lies in the fact that the overload borders can be seen clear. With input levels shown below this overload the signal-to-intermodulation ratio or the upper limit for trouble-free operation can be read off. Additionally if the self-noise is indicated as the lower limit for the input signal the intermodulation free dynamic range is defined.

Noise Figure

Almost all components in communications degrade the signal-to-noise ratio of signals that are supplied to them for processing. This difference between the signal-to-noise ratios at input and output in dB (the degradation) is the Noise Figure also in dB. Fig 4 shows this for an amplifier with NF = 10dB and a gain of 20dB.

Stability factor "k"

To avoid any oscillation of an amplifier the stability factor "k" is defined. It can be calculated from the S parameters. The result is simple:

An absolutely stable circuit (under no circumstances inclined to auto-excita-
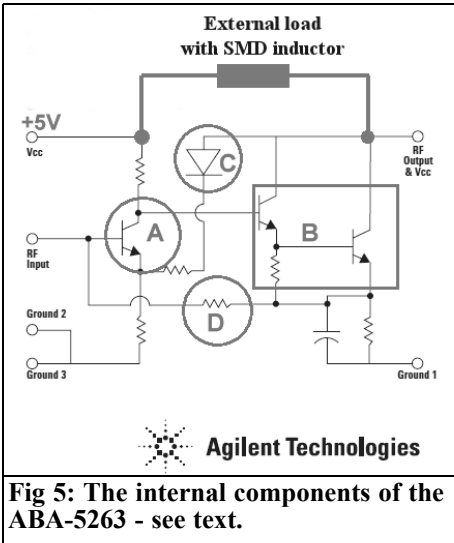
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Fig 5: The internal components of the ABA-5263 - see text.

tion) must exhibit a stability factor "k" that is larger than 1.

The calculation of "k" at microwaves is usually done with a CAD program that uses or determines the S parameters of the circuit in the simulation. Here the Ansoft Designer SV is used.

For the selected component that does not seem to be a problem because the data sheet says: "Unconditionally stable", but take care, and test it at the end of the simulation procedure.

2.

The circuit

2.1 The MMIC

Fig 5 shows a view inside the ABA-52563. It consists of an input amplifier stage (A) in grounded emitter that passes its output signal on to a second emitter follower stage (B). This second stage is arranged as a "Darlington circuit" to increase the input impedance and gain of the stage.

The diode "C" supplies negative feedback from the collector of the second stage to the emitter of the first stage to stabilise the characteristics indicated in the data sheet.

The resistor D connects the emitter of the right hand stage to the base of the left hand stage. Because the emitter of the right hand stage is grounded to RF with a capacitor, this represents a DC voltage negative feedback. Thus the operating points are stabilised in the overall circuit.

An external inductor forms the collector load of the right hand stage and ensures that the DC quiescent current can flow in the right hand stage.

Interestingly there are three points for grounding (Ground 1 - 3). The right hand stage with its higher gain levels must be grounded separately - otherwise there can be interactions with the first stage and the circuit can oscillate. The grounding point of the left hand stage is led out twice. The reason for two lead wires (gold bond wires with 0.02mm diameter) connected in parallel with the package legs means that the associated inductance is halved. It is in series with the emitter resistor of this stage and because inductive reactance increases with frequency the gain falls with increasing frequency. Thus it is important to reduce this inductance as far as possible for the higher frequencies.

2.2 The complete circuit

The circuit diagram can be seen in Fig 6; the following explanations apply:

- The input and output capacitors of the amplifier are necessary for DC isolation. These should be low-loss 470pf SMD capacitors (0805 size) made from NP0 material. These are light grey and violet in colour.

- The external SMD load inductor has a value of 100nH and size 0805 or 1206. It needs a Q of 30 to 40 at the high frequency. The self-resonance will be approximately 1 to 1.5GHz.
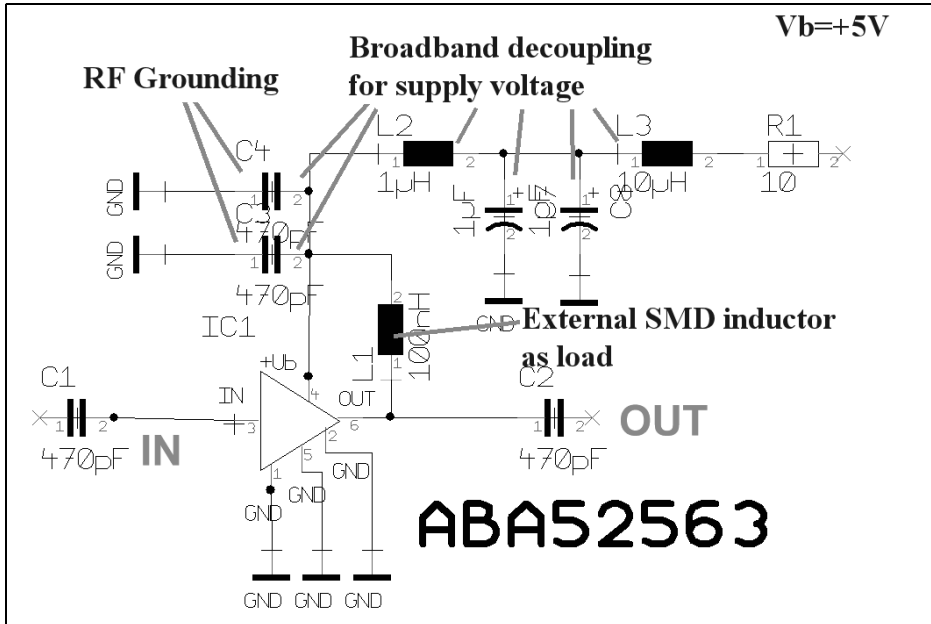
Fig 6: The circuit diagram of the amplifier designed to meet the specification - see text.

- Two low pass filters in series decouple the supply voltage.

- The first low pass filter (10µH/2µF) has a low cut off frequency. The filter capacitor is made from two parallel-connected tantalum electrolytic capacitors with low inductance and small series resistance. The parasitic elements (series inductance and series resistance) are halved by the parallel connection of the two capacitors.

- The second low pass filter (1µH/1nF) decouples the high frequencies. The two parallel-connected 470pF SMD capacitors are connect to the "+Vb" connection of the MMIC with the upper end of the 120nH inductor so that wideband alternating voltages are grounded. The previous considerations apply to the parallel connec-

tion of the two capacitors. Almost all SMD capacitors of this size have about the same series inductance that lies between 0.5 and 1nH because of their construction. By parallel connection of several capacitors the total inductance is reduced giving a clear advantage at the higher frequency range (the rising resistance is an annoyance with rising frequency).

- The preceding of 10Ω resistor is not only for decoupling but also as current limiting at switch on. The tantalum electrolytic capacitors cannot withstand starting current shocks of higher than 0.3 to 0.5A.

2.3 Simulation of the S parameters with Ansoft designer SV

2.3.1 Obtaining the S parameter file for the ABA-52563

That is not a problem, just work: it is at

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Table 1: S Parameters for the ABA-52563.

! ABA-52563 S Parameters								
# ghz s na r 50								
0.05	0.01	146.6	12.10	-2.6	0.03	0.3	0.15	-2.4
0.10	0.00	134.0	12.11	-4.8	0.03	-0.3	0.15	-5.1
0.20	0.00	-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

the end of the data sheet and must be captured by copying from the pdf file, inserting it into a new text file and reworking on the PC into the form shown in Table 1.

Explanations of Table 1:

- Lines starting with an exclamation mark (!) are comments and are ignored.

- The line beginning with a hash mark (#) indicates the form the parameters.

- ghz s ma r 50

  means that the following lines begin with the frequency in GHz, then the S parameters in the order:

  S11 S21 S12 S22

  Each parameter is indicated as a pair with "m" (magnitude) and "a" (angle).

  The reference "r 50" means that the system resistance Z is 50Ω

This table is copied into a simple text file and stored in a suitable file under the name ABA52563.S2P.

2.3.2 Simulation of circuit characteristics

2.3.2.1 The program

The Ansoft designer SV is meant for study purposes and is the free variant of the outstanding (and expensive) Microwave CAD program. It is interesting because it includes a complete linear S parameter simulation including innumerable components models. Unfortunately the manufacturer has taken this software off of the Internet, however the author has permission to make it available on his homepage (www.elektronikschule.de/~krausg) for prospective customers to download.
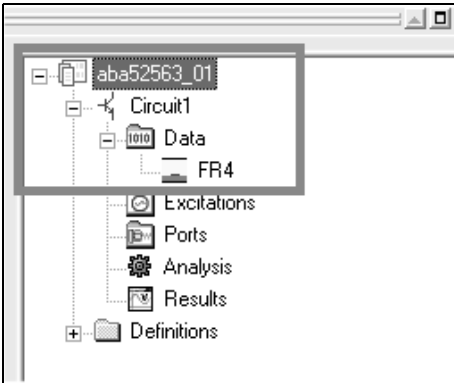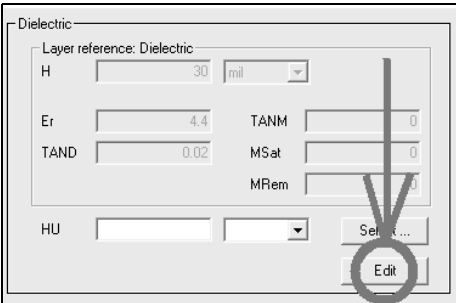
Fig 7: The material hides here, click on "FR4".

Fig 8: It can be changed to "Rogers RO4003" under "Edit".

2.3.2.2 Beginning of the project and input of the printed circuit board data

Start the program and click the "File" menu then "New" then "Project" and "Insert Circuit Diagram. The program responds with "Choose Layout Technology" and select from the Pop Up menu:

MS - FR4 (er = 4.4) 0.030 inch, 0.5 oz copper

and confirm with "Open". The screen will be completely covered with different windows and it is important that the project is saved under its own name (here use: "aba525623_01") in a suitable new file.

In the left upper window open the menu "aba52563_01" to see "Circuit" and "Data" (see Fig 7).

Double click on "FR4" in the project window to change the printed circuit board material to "Rogers RO4003" and select the line type "Grounded Coplanar Waveguide". A further menu emerges "Choose layout Technology", make the necessary entries (Fig 8). The data for the printed circuit board material is in the upper right half of this menu. To make changes click on "Edit" and enter the required values

The data required is:

- Substrate name: RO4003
- Type of substrate:
  Grounded Coplanar Waveguide
- PCB thickness: "H" = 32mil
- Dielectric constant: Er = 3.38
- Dissipation factor: TAND=0.002
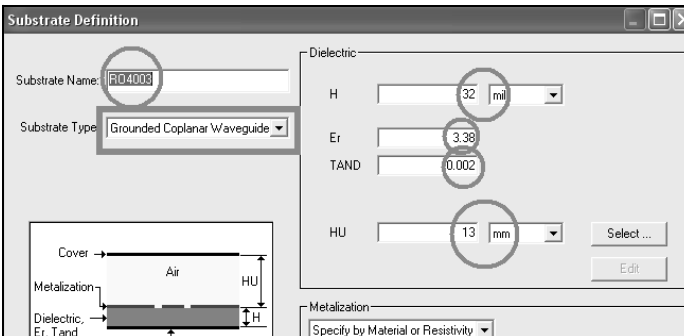- Distance between PCB and shielding cover: HU = 13mm

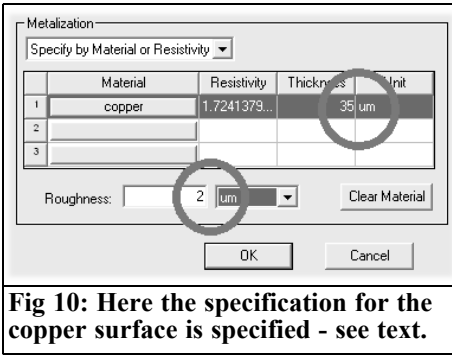**Fig 9: Some work is required to transfer the correct information - see text.**

Fig 10: Here the specification for the copper surface is specified - see text.

Clicking OK once confirms the data then examine the result to ensure that these characteristics were correctly stored (Fig 9).

Now clicking OK once opens this menu again (or double click again on "RO4003" in the project window), now click on "Edit" in the lower right corner. Enter the thickness of the copper layer as 35μm and the surface roughness with of

2μm (Fig 10) and save again with one OK click.

2.3.2.3. Draw the circuit

Use the circuit editor to draw the circuit see Fig 11:

1 Begin with the ABA52563 and get an "N-port" from the menu border

2 Call this N-port "ABA52563"

3 Now the program needs the path for the S parameter file of the component (Note: it is advisable to store this project file and the S parameter file in the same folder)

4 Mark the S parameter file of the ABA52563 with one mouse click and transfer it to the project file. Click OK twice and the N-port-symbol can be placed on the screen.
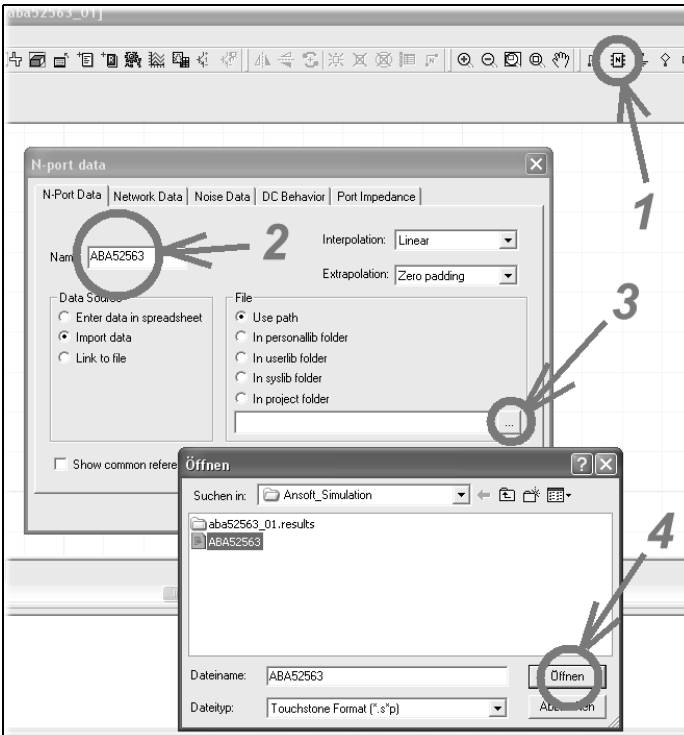
Right click on the N-port symbol and

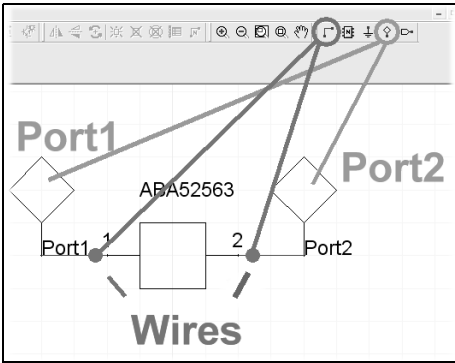**Fig 11: The procedure to use an N-port for the ABA-52563.**

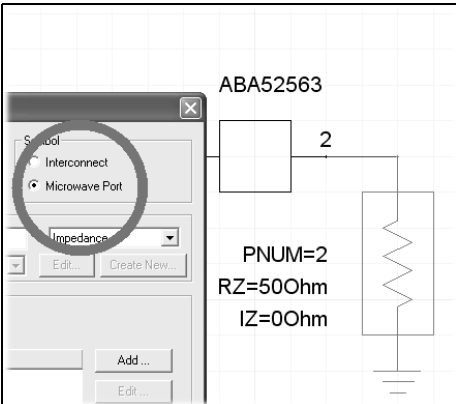**Fig 12: Add 2-port symbols at the input and output and connect with wires.**

Fig 13: Make the ports Microwave Ports.

select the "Properties" to show the record sheet "Property display". Click "Add" and "OK" and the MMIC designation appears next to the circuit symbol on the screen. Now 2 port symbols are required at the input and output of the component and connect them to the component with "Wires" (Fig 12).

Double left clicking on the port symbol opens its Property Menu (Fig 13). Change over to "Microwave Port" and check fro a system resistance of 50Ω. Use this procedure for the input and the output port.

Now it becomes somewhat more compli-

cated, because a 100nH SMD inductor is needed. First go to the Project Menu an choose "Components" and then "Circuit Element/Lumped/Inductors" go to "Chip Inductor/Q Factor". Turn the symbol through 90 degrees by typing "R" and attach it to the output of the port by clicking the left mouse button. Do not forget the ground character (the symbol is in the menu border between the port and N-port-symbol). Fig 14 shows this procedure.

Finally open the Property Menu of the inductor with a double click on the inductor symbol to enter the coil data; the
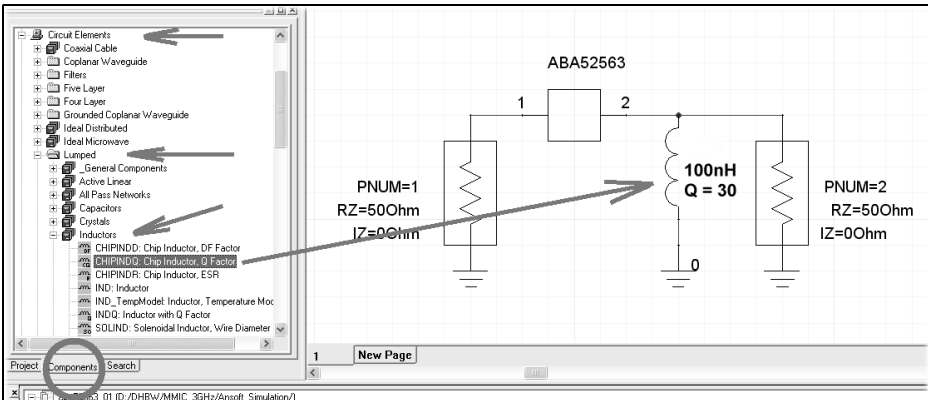

Fig 14: The procedure for adding the 100nH SMD inductor.

Fig 15: The inductor parameters.
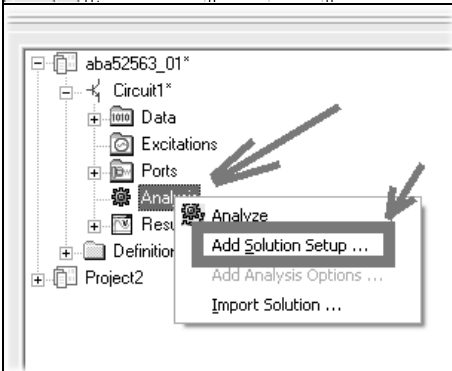
Fig 16: Before starting the simulation the sweep must be programmed.

Fig 17: Choose a linear sweep for the frequency.

Right mouse click on "Analysis" and select "Add Solution Setup". Fig 17 shows how the basic adjustments should look then confirm with OK. Complete the somewhat complex input procedure accurately according to Fig 18 in the following order:

1 Begin with "Add" in the menu and a further menu for the Sweep attributes opens.

four entries are shown in Fig 15.

2.3.2.4 Simulation of the S parameters
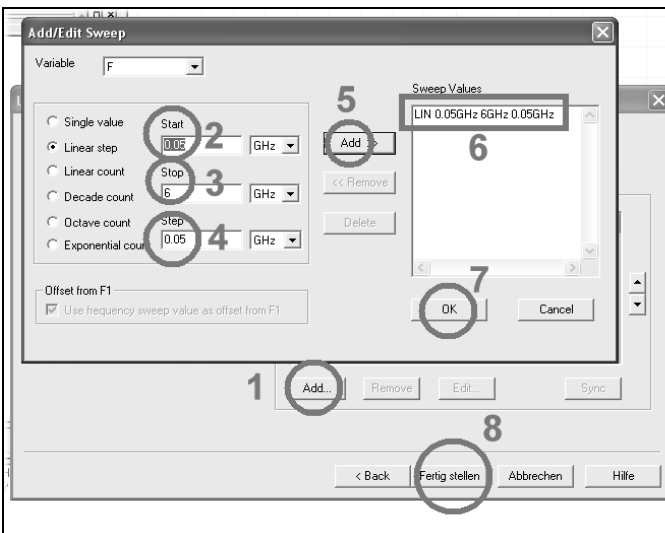
Move back to the "Project" see Fig 16.
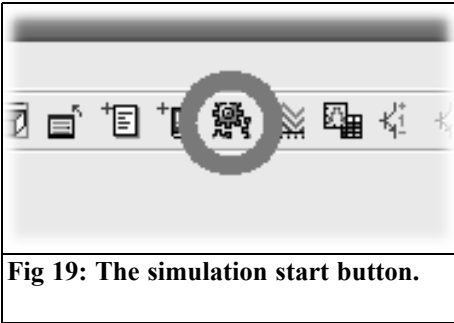

Fig 18: Programming the sweep parameters - see text.

Fig 19: The simulation start button.

2 to 4 a Sweep is programmed for 0.05GHz to 6GHz with an increment of 0.05GHz.

5 Press "Add" …

6 the Sweep programming should appear in the associated window and it can be examined.

7 After OK…

8 right click on "Finish" and everything is now prepared for the simulation.
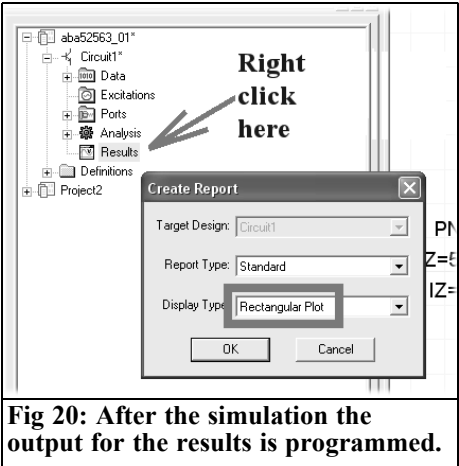
Fig 20: After the simulation the output for the results is programmed.

The simulation is started with the Simulations Button (the gear wheel) in the menu border (Fig 19).

After the simulation is completed the results are stored but the output for result must be programmed. Fig 20 shows the way to do this, select the submenu "Results", in order to produce a report. Check that "Rectangular Plot" type of
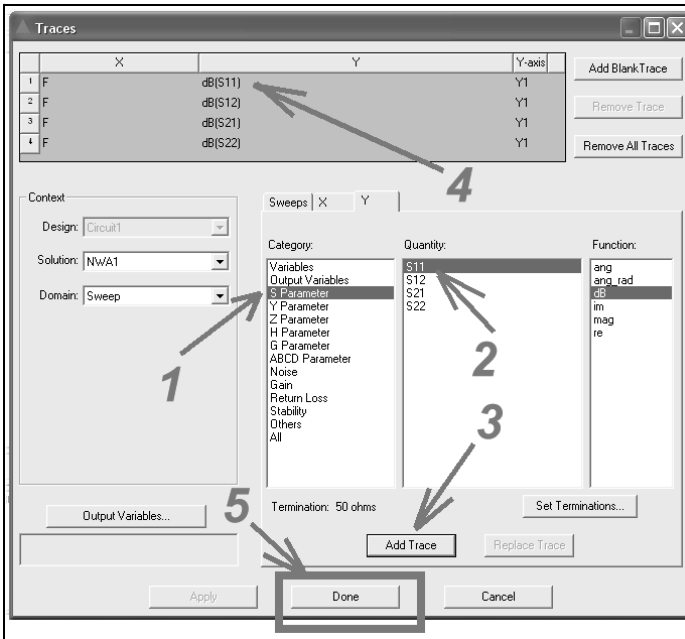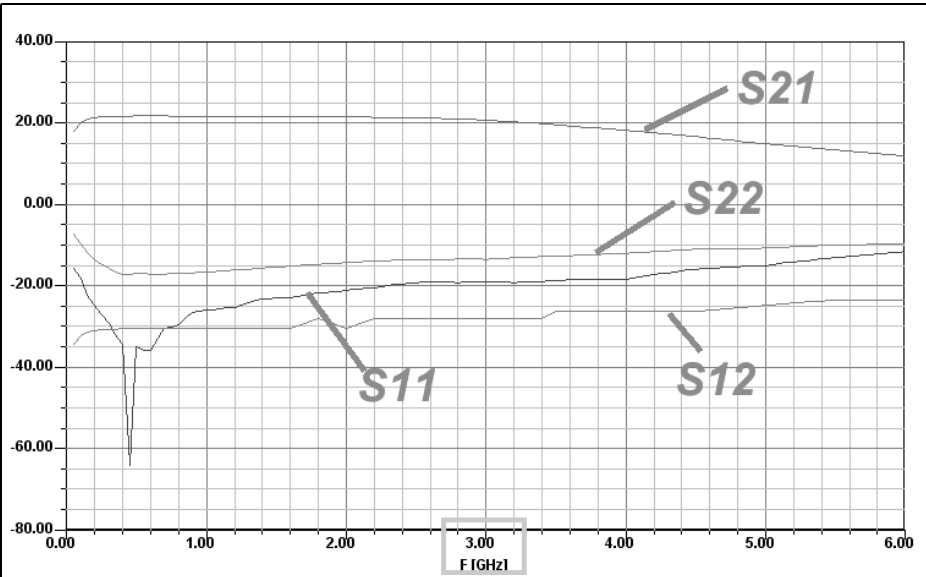

Fig 21: Selecting the S Parameters for output.

Fig 22: The result of the simulation.

display is selected: then click OK.

Now it is somewhat complex (Fig 21), because the steps (1) to (5) must carried out successively.

1 Check that the S parameters for presentation are selected.

2 to 4: Mark S11, then press "Add trace"; and check that it was transferred to the task list and appears there.

This procedure is repeated for the remaining S parameters S12/S21/S22.

5 After the last one is complete press "Done". That should produce the screen shown in Fig 22. It can now be seen if the requirements up to 2.5GHz are fulfilled.

The diagram has frequency f=3GHz marked and there the S21 value is good; so the goal has been achieved and it can be continued without further concerns.

To be continued.

X.

References for part 1

[1] Practical project: Noise factor measurement with older spectrum analysers: part 1, Gunthard Kraus, VHF Communication Magazine 1/2008 pp 2 – 11

[2] Practical project: Noise factor measurement with older spectrum analysers: part 2, Gunthard Kraus, VHF Communication Magazine 3/2008 pp140 – 154

Henning C. Weddig, DC5LV

The demodulator module

1.

Introduction

Besides the IF amplifier [1] and AGC module [2] already described in this magazine the demodulator described here is a further module to build a complete receiver chain. The demodulator is connected to the second output of the IF amplifier and used to demodulate the IF signal into an audio signal or talking in modern terms to translate the IF into baseband.

Although the IF amp and AGC module are part of the control loop the audio output of the demodulator is used to determine the "hold time" of the control voltage in AGC short and AGC long. Therefore the demodulated (audio) signal is fed into the AGC module. To verify the complete control loop chain the demodulator is also necessary and useful.

To build a complete receiver (single conversion or even dual conversion) the following further modules are necessary and partly developed/in development by the author: FM demodulator module, input filter to remove the image frequency, first mixer (and second mixer for a dual conversion receiver), IF filter bank (already designed), BFO and local oscillator. For the latter the DDS (with a AD9951) described in [3] can be used.

2.

Product requirement specifications

The demodulator fulfils the following tasks:
- Demodulation of SSB and CW signals with an external BFO (crystal oscillator or DDS)
- Demodulation of DSB AM signals
- Feeding in a demodulated FM audio signal
- Nominal input level -30dBm
- Overload margin > 10dB
- Input impedance: 50Ω
- High pass and low pass filtering of the demodulated signal with cutoff frequencies of approximately 100Hz and 5kHz
- Amplification of the AF signal to 1Vpp
- Distortion factor < 1% with nominal input level and 10dB above that level
- The AF output able to drive a low impedance load (<1kΩ)

3.

The block diagram

The block diagram of the demodulator module is shown in Fig 1. The output signal of the IF amplifier is fed to a
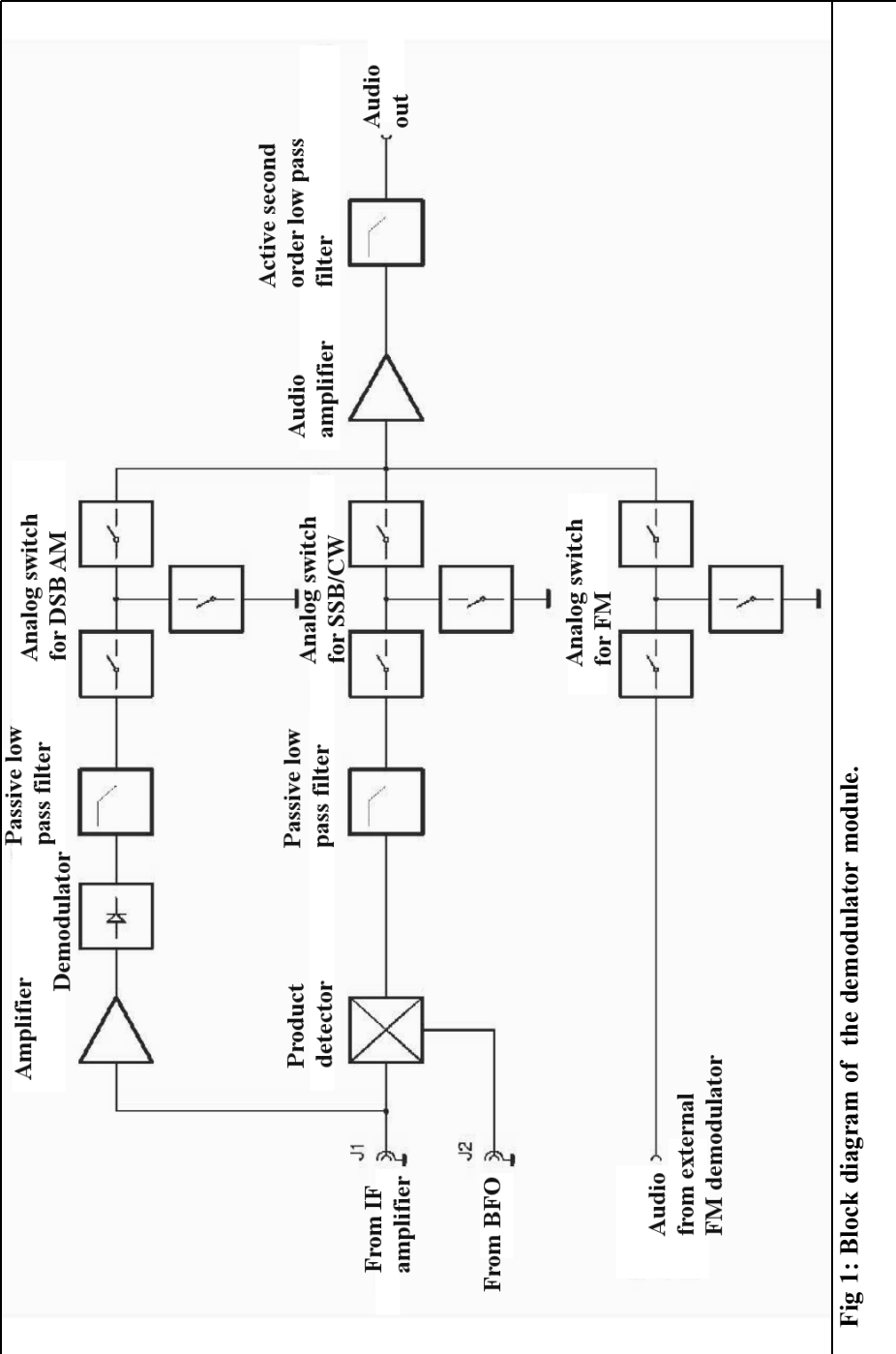
Fig 1: Block diagram of the demodulator module.
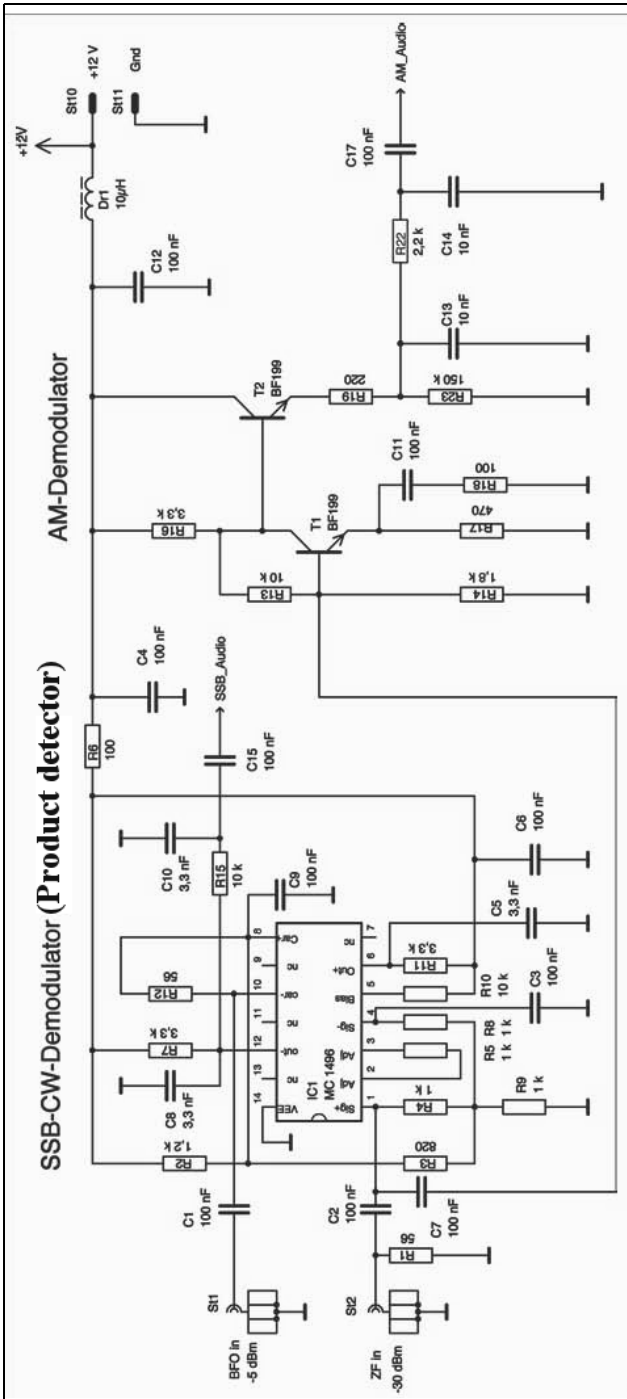
Fig 2: Circuit diagram of the SSB and AM demodulators.

mixer (product detector) to demodulate suppressed carrier (SSB) signals or unmodulated carrier (CW) signals. The second input of the product detector is connected to a BFO module.

To be able to demodulate amplitude-modulated signals with full carrier (DSB AM), an envelope detector is used.

The respective AF output signals are fed to an RC low pass filter at the AF change over switch. The AF signal from an externally FM demodulator can be also fed to this switch. An amplifier stage with an active second order low pass filter follows. A high pass filter is implemented by the value of the coupling capacitors between the stages and at the output.

4.

Description of the circuit

4.1 Overview

The circuit diagram of the demodulator module is divided for better legibility into three partial diagrams; see Figs. 2, 3 and 4.

The circuit of the SSB/CW demodulator

Fig 3: Circuit diagram of the audio switching.

Fig 4: Circuit diagram of the audio amplifier and low pass filter.

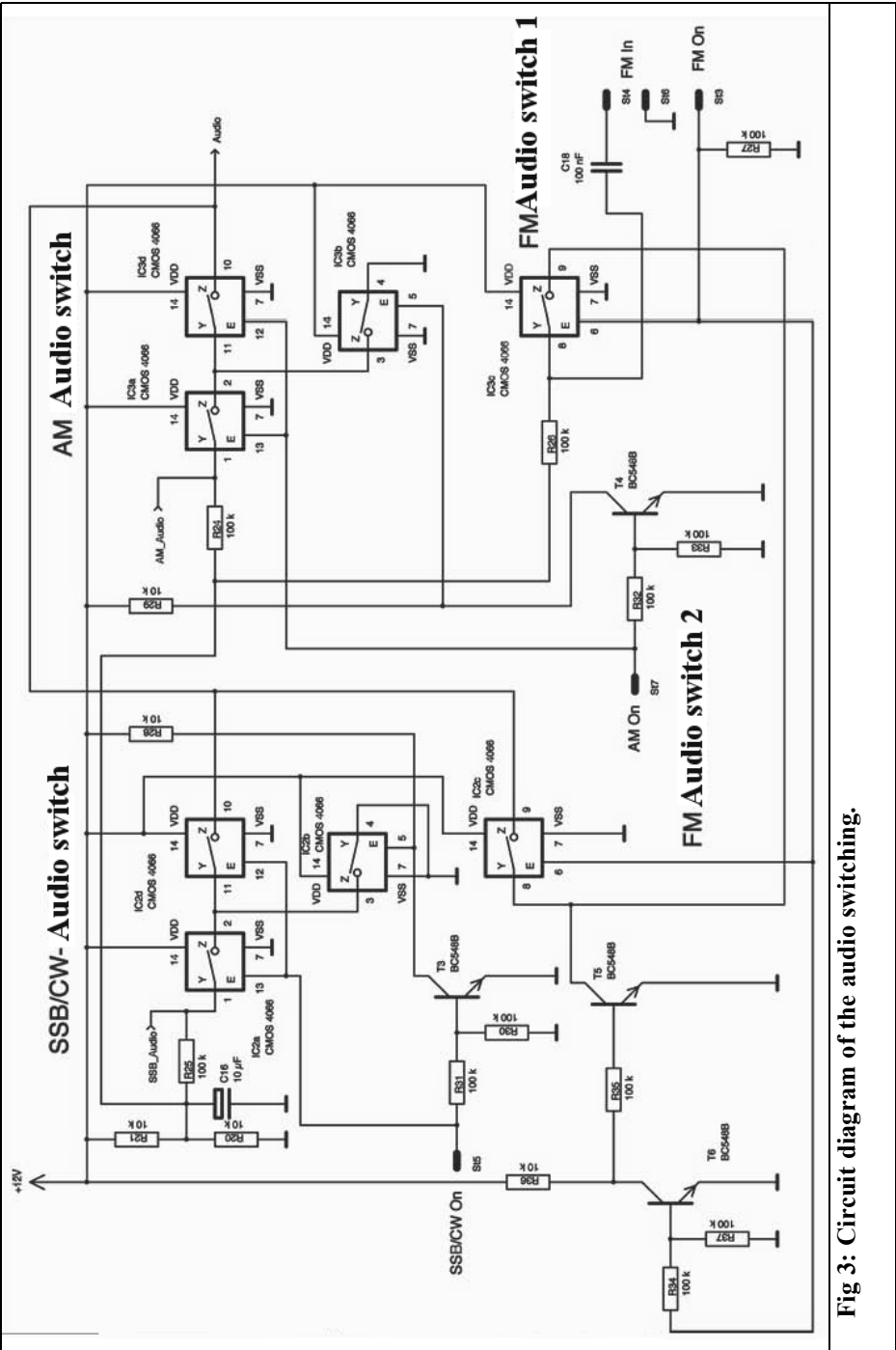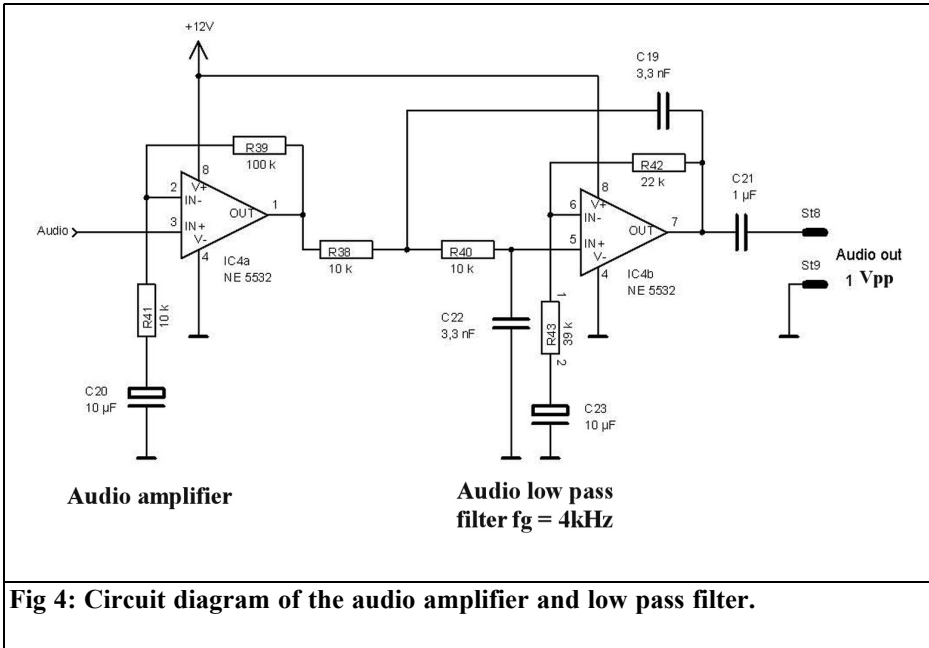and AM demodulator is shown in Fig 2, the circuit of the audio change over switch is shown in Fig 3 and the circuit of the audio amplifier with active low pass filter is shown in Fig 4.

4.2 SSB/CW demodulator

4.2.1 Choice of the demodulator ICs

From [1] the nominal output power of the IF amplifier within the gain control range is -30dBm.

In many publications the product detector NE 612 or NE 602/SA602 are used. Looking at the data sheet [4], it shows that the input level should not exceed -40dBm (2.24mV at 50Ω). With input levels 10dB higher than -40dBm compression starts resulting in a worsening of intermodulation products, which are only 35dB down in relation to each wanted signal at this higher input level. This IM spec does not comply to the specification of the demodulator module given in chapter 2 of the article.

The familiar MC1496 can stand substan-

tially higher input levels [5] and despite its great age it is still available, even in an SMD package. The MC1496 is therefore favoured as a product detector. The larger packages (DIL 14, DIL8 or SOIC 14/SOIC 8) are not favoured as much as the SMD package because they require more wiring.

Both ICs mentioned are called Gilbert cell mixers after its inventor, Barry Gilbert. The Gilbert cell consists of two emitter coupled sum-and-difference amplifier stages, whose outputs (summed current) are connected with the opposite phase. Also the corresponding base connections of the two sum-and-difference amplifier stages are connected.

The collectors of a third sum-and-difference amplifier stage feed the connected emitters of both sum-and-difference amplifier stages. The collector currents are the emitter currents of the upper difference amplifiers. Therefore the collector and emitter currents of these stages and the respective input voltages are linearly dependent. This crosswise interconnec-

tion of the collectors of the two upper sum-and-difference amplifiers makes a four-quadrant device. With input signals larger than 25mVpp between the bases of the upper sum-and-difference amplifier stages the linear range of the sum-and-difference amplifier is exceeded, these stages then work as switches.

4.2.2 Demodulator circuit

The external wiring for the MC1496 is shown in Fig 26 of the data sheet [5]. The IF signal is fed to the bases of the lower sum-and-difference amplifier on pin 1 via a coupling capacitor. The second input (pin 4) is decoupled to ground by the capacitor C3.

The input impedance of the IC is high in relation to system impedance of 50Ω, R1 presents the correct impedance for the IF module.

The BFO signal (-5dBm) is fed via C1 to pin 10 of IC1. The second input (pin 8) is decoupled to ground by the capacitor C9. Pin 10 and 8 are the bases of the two upper sum-and-difference amplifier stages of the Gilbert cell mixer. The resistor R12 between pin 8 and 10 brings both bases of the IC to the same DC voltage potential and provide a 50Ω impedance match for the BFO signal.

According to [5] the IC can be used at different operating points depending on the amplitude of the carrier signal. With a carrier amplitude <25mVpp the demodulated output signal depends on both the amplitude of the carrier signal and the amplitude of the IF signal. With carrier amplitudes above this the dependence on the carrier amplitude is void. The second mode of operation was selected for the demodulator, since a dependence of the demodulated output signal on the amplitude of the BFO is unwanted.

From Fig 22 in [5] the carrier suppression is largest if the output voltage of the BFO approximately 150mV, 3.5dBm (f = 10MHz). For use as a product detector the carrier suppression is not crucial

because the low pass filters suppress any RF remaining in the demodulated signal at the output.

The amplitude of the demodulated signal can be adjusted with R5, the value of 1kΩ is shown in the data sheet as the default. With this circuit and an input level of -30dBm the output signal in the prototype had an amplitude of 66mVpp.

4.2.3 RC low pass filters at the output of the demodulator IC

The two outputs of IC1, pin 6 and 12 are connected by R10 and R7 to the supply voltage. C5 and C8 decouple any remaining RF. C5 and C8 with R7 form a first order low pass filter at audio frequencies. This statement is surprising at first; consider that the capacitors not only represent a capacitive shunt for high frequency but also the load impedance for the IC collectors with increasing audio frequency. This lowers the gain giving a low pass function.

The cutoff frequency of the RC element R7/C8 is:

$$f_{gr} = \frac{1}{2\pi * R7 * C8} = \frac{1}{6.28 * 3.3\,k\Omega * 3.3\,nF}$$

$$= \frac{1}{6.28 * 3.3 * 10^3 * 3.3 * 10^{-9}} \frac{1}{s}$$

$$= \frac{10^6}{68.423}\,Hz = 14.6\,kHz$$

R7 is in parallel with the output impedance at pin 12 of IC1, from the data sheet [4] this is 40kΩ and can therefore be ignored.

The demodulated signal from pin 12 is low pass filtered by the RC filter R15/C10 with a source impedance of 0Ω and infinitely high load impedance gives a cutoff frequency of:

$$f_{gr} = \frac{1}{2\pi * R15 * C10} = \frac{1}{6.28 * 10\,k\Omega * 3.3\,nF}$$
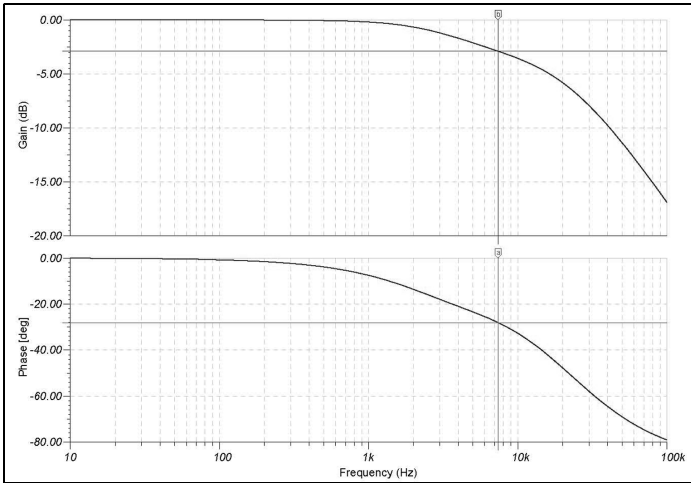
$$= \frac{1}{6.28 * 10 * 10^3 * 3.3 * 10^{-9}} \frac{1}{s}$$

Fig 5: Simulation of the output of the low pass filter R5/C7 loaded with R15/C10.

$$= \frac{10^6}{207.345} Hz = 4.8\,kHz$$

Note: since R15 is large, approximately three times the value of R7, the cutoff frequency of the RC components R15/C10 must be lower by a factor 3, as the calculation confirm.

Because the RC components R7/C8 are loaded by R15/C10, the cutoff frequency of the amplitude response is 7.8kHz. This is shown in Fig 5 by a simulation using "TINA". (Note that the 3dB cutoff frequency is approximately half the calcu-

lated value of 14.6kHz in the "TINA" simulation that shows 7.4kHz. The phase shift at the 3dB cutoff frequency is -28° compared to the unloaded RC low pass filter. This is because there are two RC low pass filters, not one, decoupled by a buffer stage. The two different cutoff frequencies are connected into series.

For an unloaded first order low pass filter the phase shift of the input signal is -45° at the cutoff frequency. The loading on the low pass filters above lowers the phase shift at the 3dB cutoff frequency to -28°.
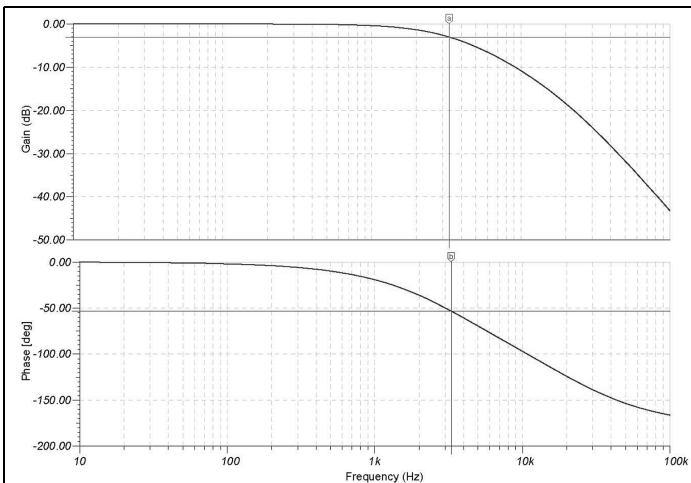

Fig 6: Simulation of the low pass filter R5/C7 and R15/C10.

If the series connection of both low pass filters is simulated, Fig 6 shows that the results are different. The cutoff frequency is 3.3kHz and the phase shift is -53° (Note that the 3dB cutoff frequency is 3.3kHz in the "TINA" simulation and the phase shift is -51° that is higher than the first order RC low pass filter. This is because there are two RC low pass filters, not one, decoupled by a buffer stage. The two different cutoff frequencies are connected into series.

R2, R3 and R9 form a voltage divider to feed the bases of the sum-and-difference amplifier stages in the IC via decoupling resistors.

4.3 AM demodulator

4.3. The circuit

A simple envelope detector is used to receive AM modulated broadcast transmissions.

The input signal of -30dBm is too low for demodulation using diodes see [11]. Therefore the signal is amplified with T1 in grounded emitter configuration. The voltage at the collector is set to approximately half the supply voltage by R13/R14. In the prototype the voltage was measured at 7.37V and the emitter voltage was 0.37V giving an emitter current of 780µA.

T2 operates as a diode; the bias current is adjusted with the high value resistor R23 for a minimum distortion factor of the demodulated signal. This depends on the amplitude of the RF input signal. In the prototype a value of 180kΩ was used, with an emitter voltage of 6.37V this gave an emitter current of 35µA for T2. T2 can also be considered as an emitter follower. Because of the capacitive load on the output (C13), the input impedance can become negative. To prevent this situation a series resistor in the base of T2 prevents this situation. The effect of variations in transistors on the distortion factor was not examined. The capacitor C13 suppressed the remaining RF and

works as a low pass filter for very high audio frequencies.

The gain of the transistor stage T1 was set so that the demodulated AM signal (m = 30%) is the same as a demodulated SSB signal with an RF input of -30dBm.

4.3.2 Shaping the audio signal

The demodulated signal is low pass filtered with the RC filter R22/C14. With a source impedance of 0Ω and an infinitely high load impedance the cutoff frequency is:

$$f_{gr} = \frac{1}{2\pi * R22 * C14} = \frac{1}{6.28 * 2.2\,k\Omega * 10\,nF}$$

$$= \frac{1}{6.28 * 2.2 * 10^3 * 10 * 10^{-9}} \frac{1}{s} = \frac{10^6}{138}\,Hz = 7.3\,kHz$$

As for the SSB demodulator the cutoff frequencies of the total low pass filter deviate from the cutoff frequency of a single RC. The formulae in [7] can be used to measure the internal resistance of the driving source (output impedance of T2):

$$r_a = \frac{r_{BE}}{\beta}; \ r_{Be} = \frac{U_T}{I_{B0}}; \rightarrow r_a = \frac{\frac{U_T}{I_{B0}}}{\beta}$$

From the data sheet [6] the current gain is β=50 with a collector current IC=100µA. The thermal voltage Vt at ambient temperature is 26mV. With a collector current of only 35µA and a current gain of 50 the base current is large at 700nA. Thus the dynamic output resistance r_a is:

$$r_a = \frac{\frac{U_T}{I_{B0}}}{\beta} = \frac{\frac{26\,mV}{0.7\,\mu A}}{50} = \frac{37.14\frac{10^{-3}\,V}{10^{-6}\,A}}{50}$$

$$= 0.7 * 10^3\,\Omega = 700\,\Omega$$

The cutoff frequency of the low pass filter R19/C13, ignoring R19, is:

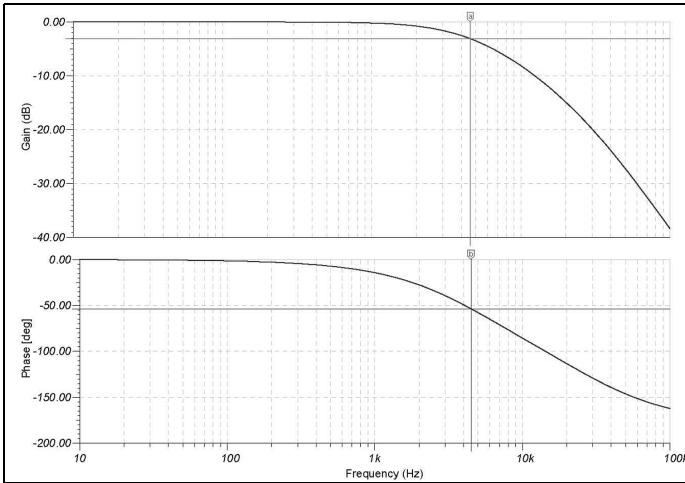$$f_{gr} = \frac{1}{6.28 * 0.92 * 10^3 * 10 * 10^{-9}} \frac{1}{s}$$

Fig 7: Simulation of the low pass filter of the AM demodulator.

$$= \frac{10^6}{62.83} Hz = 17.3\,kHz$$

The cutoff frequency of the second RC low pass filter with a source impedance of 0Ω:

$$f_{gr} = \frac{1}{6.28 * 2.2 * 10^3 * 10 * 10^{-9}} \frac{1}{s}$$

$$= \frac{10^6}{138.23} Hz = 7.3\,kHz$$

Assuming the circuit a source impedance of 0Ω and both low pass filters are buffer by a stage with very high input impedance and low output impedance the cutoff frequency of the two degrades to 4.58kHz see Fig 7 (the 3dB cutoff frequency according to TINA is 4.8kHz and the phase shift is -51° somewhat higher than a single low pass filter. This is because there are two RC low pass filters, not one, decoupled by a buffer stage. The two different cutoff frequencies are connected into series.

4.4 Audio switching

Two 4066 CMOS ICs were chosen for switching the audio signals; each IC contains four analog switches. A "T" configuration was selected for higher attenuation when the analog switches are switched off. In addition, two switches are connected in series and between them there is a third switch giving a low impedance path to ground when the other switches are turned off. The analog switches are connected to half supply voltage with the voltage divider R20/R21 and the decoupling resistors R24 to R26. C16 is a short circuit for the audio signals.

The two remaining analog switches give the option for an external audio signal to be connected, e.g. from an FM demodulator.

The series analog switches are operated by +12V on St5 or St7. The grounding switches are activated with the inverters formed by transistors T4 and T5.

4.5 Audio amplifier and active low pass filter

Not all operational amplifiers can drive low impedance loads (1kΩ). For high quality audio applications the single op amp NE5534 is often used. The NE5532 contains two single op amps in one housing. This IC is used for the AF amplifier and the second order active low pass filter.

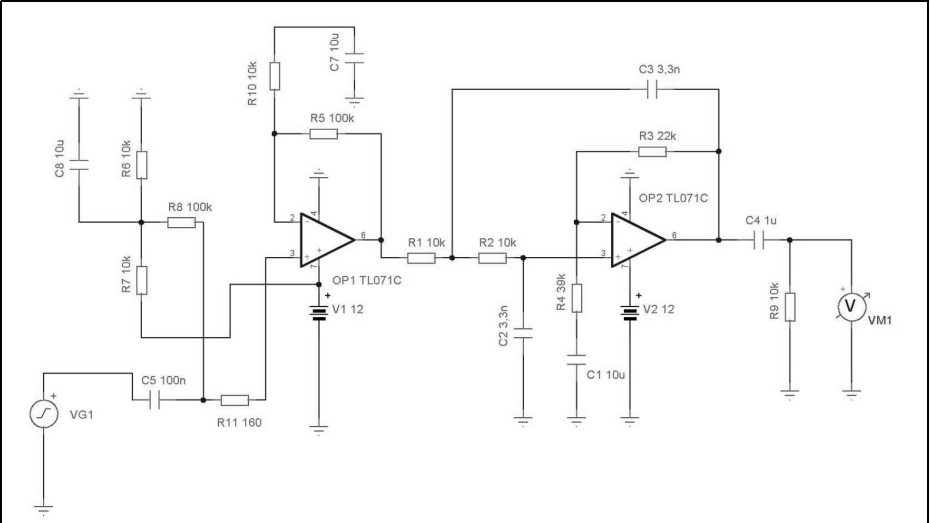The active low pass filter uses the "Sallen-Key" design, the design and

Fig 8: Simulation of the second order active low pass filter.

component values are shown in [9] an [10].

The ratio of R42 to R43 determines the attenuation factor (d) given by:

d = 2 − R42/R43

The gain of the filter from [9] is:

1 + R42/R43 = 1 +d i.e. 1+ 22/39 = 1.56

With the attenuation factor (d) determined a Bessel function approximation was selected to avoid overshoot. The cutoff frequency of the filter is determined by the time constant R38 * C19:

$$f_g = \frac{1}{2\pi * R38 * C19}$$

$$= \frac{1}{6.28 * 10 * 10^3 * 3.3 * 10^{-9}} \frac{V}{A} \frac{A \sec}{V}$$

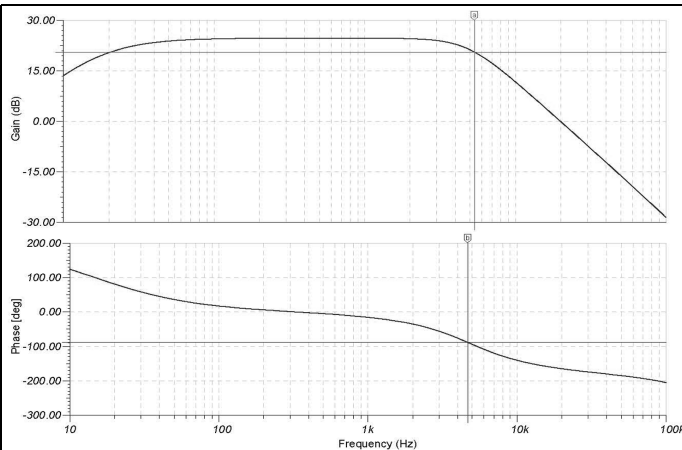$$= \frac{10^6}{207} \frac{1}{\sec} = 4825 \, Hz$$

Fig 9: Simulation of the audio amplifier and the second order low pass filter. The frequency response is from 20Hz to 4.8kHz.
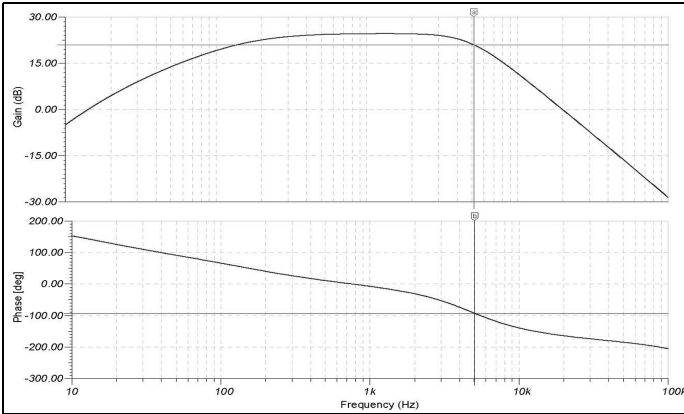
Fig 10: Simulation of the audio amplifier and the second order low pass filter with a 1kΩ load. The frequency response is now 150Hz to 4.8kHz.

The capacitors C15 (CW/SSB) or C17 (AM) form a high pass filter with C21.

The simulation circuit for a load resistance of 10kΩ is shown in Fig 8. Unfortunately TINA does not have an NE5532 in its library so a TL071 was used.

The simulation results in Fig 9 show a gain of 24dB at 1kHz, the cutoff frequencies are 24.8Hz and 4.82kHz.

The fall off of gain above the cutoff frequency is 12dB per octave because of the active second order filter.

If the load resistance is reduced to 1kΩ, the lower cutoff frequency increases to approximately 150Hz sees Fig 10. This is an indication that for low impedance loads the coupling capacitor C21 is too small.

5.

PCB design and alignment

The circuit is developed on a 145mm x 35mm double-sided printed circuit board. Fig 11 and 12 show the track and component sides and Fig 13 the compo-
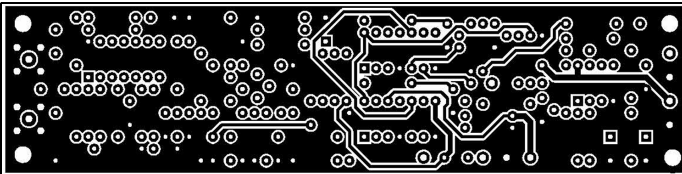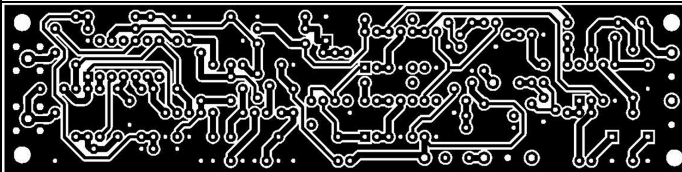

Fig 11: Componet side of demodulator PCB.

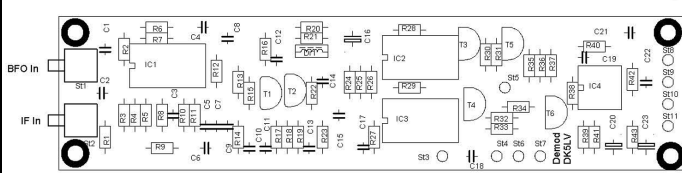**Fig 12: Track side of demodulator PCB**

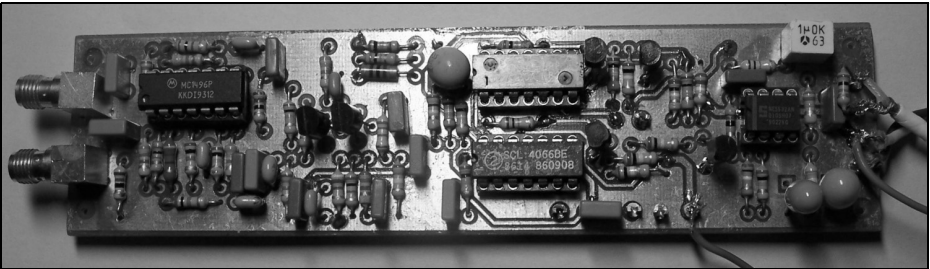**Fig 13: Component layout for the demodulator.**

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Fig 14: Picture of the prototype demodulator.

nent layout. Fig 14 shows a photograph of the prototype. The parts list of the demodulator module is shown in Table 1.

Following a visual check the circuit should work when switched on. The operating point voltages of the transistors should be measured as a functional check. Also the voltage on the pins of IC1 can be measured. The following voltage levels were to be measured on the prototype: pin 1, 4 3.75V, pin 2, 3 3.07V, pin 5 1.23V, pin 6, 12 8.16V, pin 8, 10 6.87V. The DC voltage on the input of the analog switches (pin 1 of IC2, and pin 1 of IC3) also at the inverting and non-inverting inputs of the two operation amplifiers in IC4 should be approximately half supply voltage. The power input of the prototype was 16mA.

6.

Measurements

6.1 Frequency responses

6.1.1 Frequency response of the SSB demodulator

A 1kΩ load was placed on the output of the low pass filter R15/C10 and St8/St9 the output of the module. Fig 15 shows the results of measuring the frequency responses at the output of the SSB demodulator IC IC1 (pin 12).

The measuring instruments used were two signal generators (Marconi 2024, Marconi 2018) and an oscilloscope (Hameg 1005). The frequency response was plotted using a spreadsheet analysis.

Some kinks in the curves were caused by inaccuracies in reading the amplitudes from the oscilloscope. The cutoff fre-
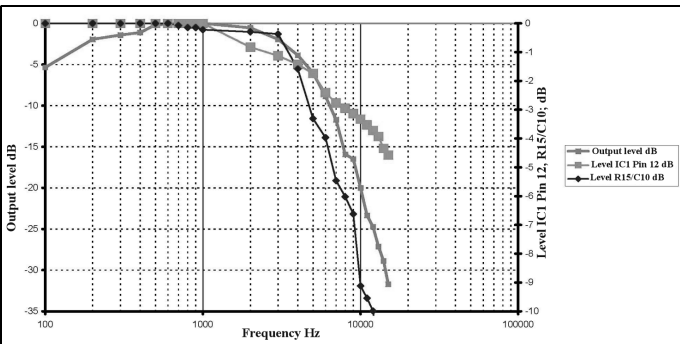

Fig 15: Measurement of the frequency response of the SSB/CW demodulator with nominal input level. 1kHz is the reference level of 0dB. The two axes have different divisions.

Table 1: Parts list for the demodulator.

Ref	Value	Type	Manufacturer	Ref	Value	Type
C1	100nF	MKS	Wima	R25	100k	
C10	3.3nF	FKC	Wima	R26	100k	
C11	100nF	Z5U		R27	100k	
C12	100nF	Z5U		R28	10k	
C13	10nF	MKS	Wima	R29	10k	
C14	10nF	MKS	Wima	R3	820	
C15	100nF	MKS	Wima	R30	100k	
C16	10µF			R31	100k	
C17	100nF	MKS	Wima	R32	100k	
C18	100nF	MKS	Wima	R33	100k	
C19	3.3nF	FKC	Wima	R34	100k	
C2	100nF	MKS	Wima	R35	100k	
C20	10µF			R36	10k	
C21	1µF	MKS	Wima	R37	100k	
C22	3.3nF	FKC	Wima	R38	10k	
C23	10µF			R39	100k	
C3	100nF	Z5U		R4	1k	
C4	100nF	Z5U		R40	10k	
C5	3.3nF	FKC	Wima	R41	10k	
C6	100nF	Z5U		R42	22k	
C7	100nF	MKS	Wima	R43	39k	
C8	3.3nF	FKC	Wima	R5	1k	
C9	100nF	Z5U		R6	100	
Dr1	10µH			R7	3.3k	
IC1	MC 1496 DIL		Motorola	R8	1k	
IC2	CMOS 4066			R9	1k	
IC3	CMOS 4066			St1	SMA	Rt angled connector
IC4	NE 5532		NXP	St10	1.3m	Solder pin
R1	56			St11	1.3m	Solder pin
R10	10k			St2	SMA	Rt angled connector
R11	3.3k			St3	1.3m	Solder pin
R12	56			St4	1.3m	Solder pin
R13	10k			St5	1.3m	Solder pin
R14	1.8k			St6	1.3m	Solder pin
R15	10k			St7	1.3m	Solder pin
R16	3.3k			St8	1.3m	Solder pin
R17	470			St9	1.3m	Solder pin
R18	100			T1	BF199	TO92
R19	220			T2	BF199	TO92
R2	1.2k			T3	BC548B	TO92
R20	10k			T4	BC548B	TO92
R21	10k			T5	BC548B	TO92
R22	2.2k			T6	BC548B	TO92
R23	150k					
R24	100k					

quency of the low pass filter R7/C8 at IC1, Pin12 agrees well with the simulation (Fig 5). The cutoff frequency at the junction R15/C10 is however higher than the simulation predicted (Fig 6; 3.3kHz) at 4.8kHz.

Above 4.8kHz the second order active low pass filter causes a further attenuation rise of 12 dB per octave so that altogether a fourth order low pass filter results.

As the simulation already showed, the size of the coupling capacitor C21 is responsible for the lower cutoff frequency.

6.1.2 Frequency response of the AM demodulator

A 1kΩ load was placed on the output of the module St8/St9. Fig 16 shows the results of measuring the frequency responses at the output of the AM demodu-
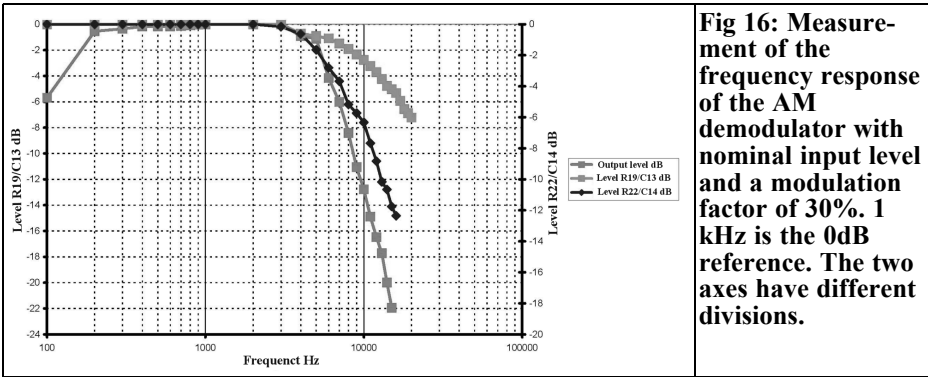
Fig 16: Measurement of the frequency response of the AM demodulator with nominal input level and a modulation factor of 30%. 1 kHz is the 0dB reference. The two axes have different divisions.

lator (R23/C13).

The cutoff frequency measured at the output of the AM low pass filter (R22/C14) was 5kHz, according to simulation (Fig 7) this should be 4.5kHz, so the simulation and measurement agree quite well.

6.2 Distortion factors - SINAD

6.2.1 Distortion factor - SINAD of the SSB demodulator

With nominal levels the SINAD is 46.6dB. The conversion SINAD to distortion factor can made with the formula:

$$k = 10^{-\frac{SINAD}{20}} * 100\%$$

$$\Rightarrow k = 10^{-\frac{46.6}{20}} * 100\% = 0.46\%$$

The SINAD is evaluated correctly in the above formula from the distortion factor and the noise but only with infinitesimal small noise. The "Stabilock" test set indicates a background noise of 0.4mV, (without the demodulator 0.1mV), the information signal is 1Vpp = 353mV.

Related to the background noise of 0.4mVs the signal-to-noise ratio of the demodulator module in SSB mode is:

$$\frac{S + N}{N} = 20 * \log\frac{353\,mV}{0.4\,mV} = 58.9\,dB$$

Because the SINAD is worse by only

12.3dB, i.e. the amplitude of the noise approximately ¼ the amplitude signal, the noise component cannot be neglected in the conversion of SINAD from distortion factor. The distortion factor is thus lower than the conversion shows.

6.2.2 Distortion factor – SINAD of the AM demodulator

The measuring instrument used was a generator Marconi 2024 signal generator; the SINAD was measured with a Stabilock 4040 test set. Measurements showed that the distortion factor of the Marconi 2024 is much lower than the Marconi 2018. According to the data sheet of the 2024 [11] the distortion factor is < 1.5% with a modulation factor of m = 30% and < 2.5% with m = 80%.

Because the Stabilock measures the distortion using a fixed notch filter set to 1kHz only the distortion was measured at an audio frequency of 1 kHz and not at other audio frequencies within the audio band

Because the RC components R19/R23/C13 and R22/C14 cause the averaging of the high frequency signal the distortion factor can rise with the demodulation of lower audio frequencies due to unfavorably time constants.

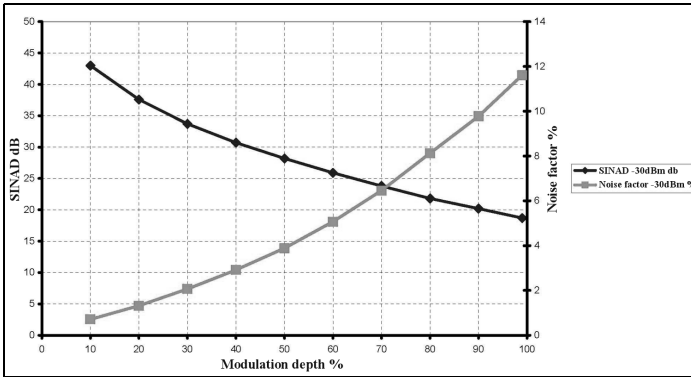With a nominal input level and a modulation factor of 30% the SINAD was 33.7dB.

Fig 17: Measurement of the SINAD with nominal input level and variable modulation factor from 10 to 99%. The distortion factor is calculated from this for AM operation.

The background noise of 0.4mV in AM mode gives the same signal-to-noise ratio as in results for the SSB/CW mode i.e. 58.9dB.

During the first measurement the modulation factor was changed from 10 to 99% with a constant RF carrier level of -30dBm. The results of that measurement are shown in Fig 17. The SINAD with the lowest modulation factor (10%) is highest (-43dB) and with 99% it is lowest (-18dB). The distortion factor of the signal generator for a modulation factor of 99% is not specified so the result with that modulation factor needs to recognize that fact. It seems that the distortion factor of the signal generator contributes significant to the result.

The second measurement was done with a constant modulation factor (30%) and a variable carrier level see Fig 18. It is surprising that the best SINAD and thus lowest distortion factor is reached with the input level of -20dBm. Therefore a further series of measurements were carried out with the modulation factor varied from 10% to 99% and the input level constant at -20dBm. These measurements and measurements with an input level of -30dBm are shown in Fig 19.

It can be seen clearly that with a 10dB higher input level the distortion factor is better. In order to achieve same level conditions for SSB/CW and AM (m = 30%), the gain of the T1 stage could be increased by reducing the value of R18 but the demodulated audio signal would have to be reduced after the low pass filter R22/C14 by using a high impedance voltage divider.
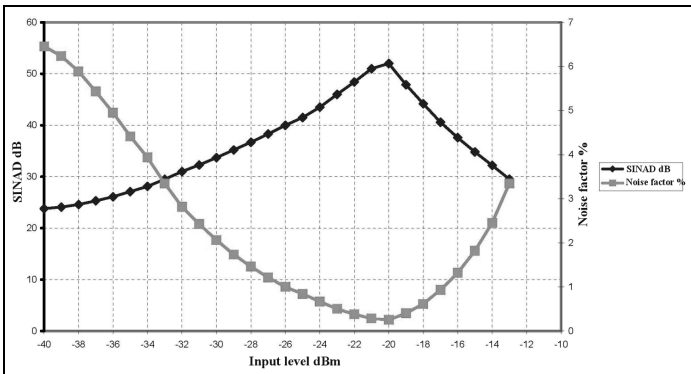

Fig 18: Measurement of the SINAD with input levels of -40dBm to -10dBm and a modulation factor of 30%. The distortion factor is calculated from this for AM operation.
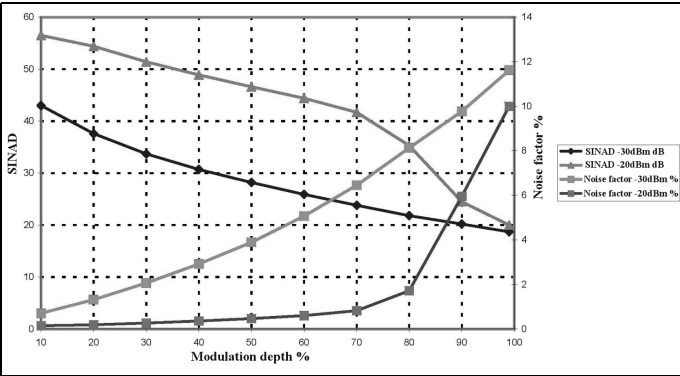
Fig 19: Measurement of the SINAD with input levels of -30dBm and -20dBm with modulation factors from 10% to 99%. The distortion factor is calculated from this for AM operation.

6.3 Determination of the 1dB compression point

6.3.1 1dB compression point of the SSB/CW demodulator

For this measurement the RF input level (Marconi 2018, f = 9.000MHz) was increased from -40dBm in 1dB steps to -6dBm and the relative output level measured using the oscilloscope in the Stabilock 4040. A Marconi 2024 signal generator, set to 8.999MHz with a level of -5dBm, was use as a BFO. A 0dB reference was set at the nominal input level of -30dBm. The result of this measurement is show in Fig 20. The 1dB compression point was reached with an input of -8dBm that is approximately 12dB higher than specified in the product requirement specifications (overload margin >10dB related to the nominal level of -30dBm).

6.3.2 1dB compression point of the AM demodulator

For this measurement the RF input level (Marconi 2018, f = 9.000MHz, AM modulation, m = 30%) was increased from -40dBm in 1dB steps to -6dBm and the relative output level measured using the oscilloscope in the Stabilock 4040. A 0dB reference was set at the nominal input level of -30dBm. The result of this measurement is shown in Fig 21. The 1dB compression point was reached with an input of -11dBm that is very much higher than specified in the product requirement specifications.

From the diagram it is remarkable that the measured curve deviates from an ideal straight line with an input level of < -32dBm.

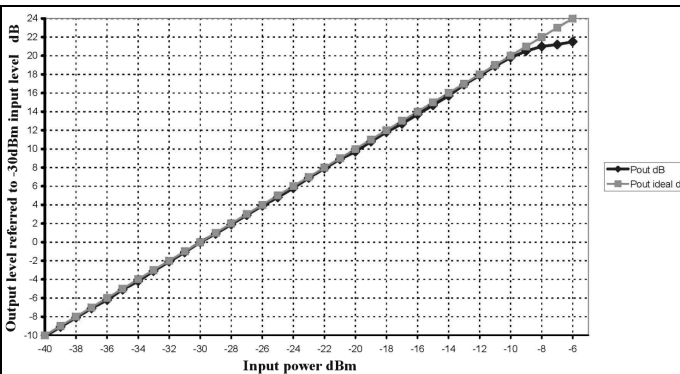The remainder of the curve above -12dBm shows a large overload. The

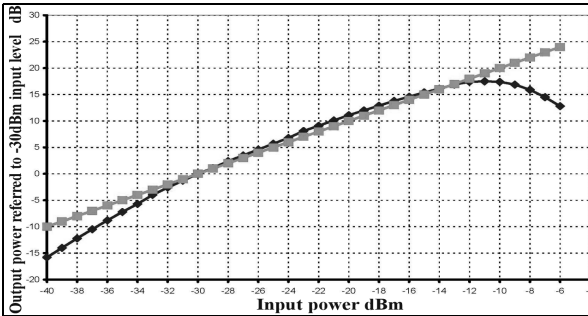Fig 20: Linearity of the SSB/CW demodulator as a function of the RF input level.

Fig 21: Linearity of the AM demodulator as function of the RF input level, modulation factor = 30%, and an ideal straight line.

product requirement specifications needs an overload margin of 10dB, related to the nominal level of -30dBm this is exceeded by 8dB.

With this result, it appears at first sight, that the demodulator could have an RF signal input higher than -30dBm for a lower distortion factor; see chapter 6.2.2. However the higher output voltage of the AM demodulator feeding the AF amplifier would cause the overload margin to be reduced. A voltage divider between the output of the AM demodulator and the AF amplifier can correct this problem.

7.

Literature

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[2] The AGC module,, Henning Christof Weddig, DK5LV, VHF Communications Magazine 1/2008 pp 26 – 30 and 35 – 42. This article was continued in issue 4/2009 pp 242 - 250

[3] DDS using the AD9951, Henning C. Weddig, DK5LV, VHF Communications Magazine 2/2009 pp 104 - 117

[4] Data sheet NE 602, NXP Semiconductor

[5] Data sheet MC 1496, ON Semiconductors

[6] Lecture script, Prof. Dr. Ing. Heinz Schmidt-Walter, Bipolar transistor: Emitter follower, Power Source, Sum-and-Difference Amplifier, 17.1 Emitter Followers (collector connection): http://schmidt-walter.eit.hda.de/ el/skript_pdf/el_17.pdf

[7] Data sheet BF199, Motorola http://www.datasheetcatalog.net/de/ datasheets_pdf/B/F/1/9/BF199.shtml

[8] The active filter cook book, Dan Lancaster, TeWi publishing house see: http://www.amazon.co.uk/s/ ref=nb_sb_noss/ 277-0194207-7071731?url= search-alias%3Daps& field-keywords=don+lancaster

[9] Sallen-Key Filter http://www.vias.org/mikroelektronik/ filter_sallen_key.html

[10] Analysis of the Sallen-Key Structure http://focus.ti.com/lit/an/sloa024b/ sloa024b.pdf

[11] AM Demodulators using Silicon Semiconductors, D.E. Schmitzer, DJ4BG, VHF Communications Magazine 3/1971 pp 190 - 193

[12] AM/FM SIGNAL GENERATOR 2023 and 2024, Operating Manual, Document Part No. 46892/225, Issue 9, 18 May 1998 http://www.aeroflex.com/ats/products/ prodfiles/opsmanuals/ 2023_2024OperatingManual.pdf

Bernd Kaa, DG4RBF

Soldering advice for 0.5mm pitch SMD ICs

In an age when miniaturisation is progressing very fast it is understandable that there are more and more 0.5mm pitch ICs. At first sight soldering these ICs seems extremely difficult because the pins are very close to each other (Fig 1). With this advice it should be possible for someone with some soldering experience to solder these SMD components by hand.

1.

What tools are required?

- A normal soldering iron (approximately 25W) with a 1mm broad chisel bit. Or better an adjustable soldering station (approximately 60 - 80W) with replaceable bits.

- A strong magnifying glass or a microscope
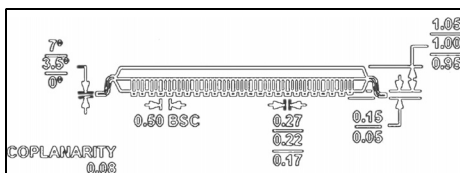
Fig 1: A typical IC package in 0.5mm pin spacing format.

- Special flux

The three important aids: Magnifying glass, soldering iron and flux:

It is most important to have a good optical magnifier for the work area because you can work on things that you see well. The simplest solution is a strong magnifying glass (Fig 2). I soldered my first 0.5mm pitch IC with such a simple folding magnifying glass.

A hand magnifier or strong reading glasses (3 - 4 diopters) can be used. A bench magnifying glass (Fig 3) or magnifying eyeglasses (Fig 4) also give good

Fig 2: An old folding magnifying glass.

Fig 3: A desk top magnifier.

Fig 4: Magnifying eyglasses.

magnification. A magnification of at least 6 times is needed but a magnification from 10 – 20 is better.

If small components are soldered regularly it is more comfortable to use a stereomicroscope with at least 5 - 6cm work distance. Searching for an inexpensive stereomicroscope with sufficiently work distance and acceptable magnification I found the BMS S-10 (Fig 5) [available from Amazon [2] in the UK]. The manufacturers specification states that it has a work distance of 80mm and 20 times magnification with an option for lighting.

The next important tool is a suitable soldering iron; do not use an SMD needle point soldering iron (Fig 6) they are too unstable. The thin design means that they are not very strong and they are usually not powerful enough.

A normal soldering iron of approximately 25W with a 0.8 – 1mm wide chisel bit is suitable and small enough. The ideal is an adjustable soldering station with a 0.8mm wide chisel bit (Fig 7). An example of a 0.8mm broad bit is shown in Fig 8. The very thin bit shown in Fig 9 is also suitable because it has a sturdy shank that guarantees a fast heat supply. A very thin point is not necessary when soldering this type of component; the somewhat broader bit has advantages during handling and guarantees a fast

Fig 5: The BMS S-120 steromicroscope.
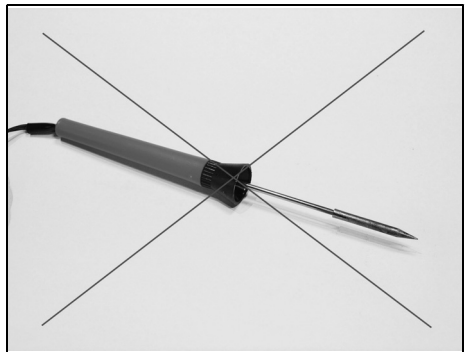
Fig 6: The SMD needle point soldering iron. Do not use.

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Fig 7: A WSP80 soldering iron from a Weller WS 81 soldering station. This is very suitable.

heat supply.

FL88 FLUX (Fig 10) liquid flux works well and comes in a handy dispenser. There are also gel fluxes available in syringes that were developed specially for fine pitch soldering (e.g. EDSYN FL22); both fluxes are available from [1] or [2].

2.

Working procedures

- First the soldering iron bit must be tinned. Make sure that all pads are tinned evenly (Fig 11 and 12).

- Put a generous amount of flux on the tinned pads.

Fig 9: A very thin soldering iron bit.

Fig 8: A 0.8mm LTH bit from a Weller WS 81 soldering station. This is very suitable.

- Place the IC on the solder pad in the exact position and solder two diagonally opposite pins. Before soldering the other pins check that the IC is positioned correctly because it can be corrected easily at this stage.

- Apply more flux so that the pins of the IC are moistened as well.

- To solder the other pins they are pressed downwards with the hot soldering iron. The existing solder on the pins is sufficient to solder the individual pins. The liquefied solder is pulled by the flux between the solder pads and the IC pin. This gives a clean soldered joint and prevents short-circuits between the individual pins. Even if two pins are pressed down together with a wide bit, no solder joint develops between them if sufficiently flux is used.

Because no additional solder is added

Fig 10: A flux pen.
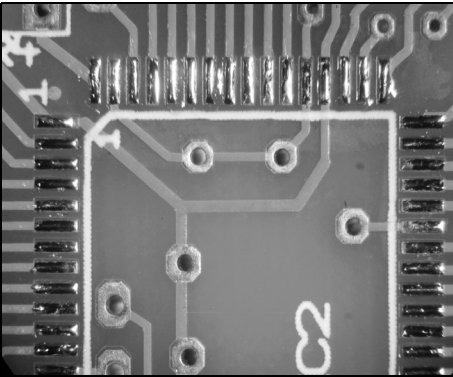
Fig 11: A PCB with the pads tinned.
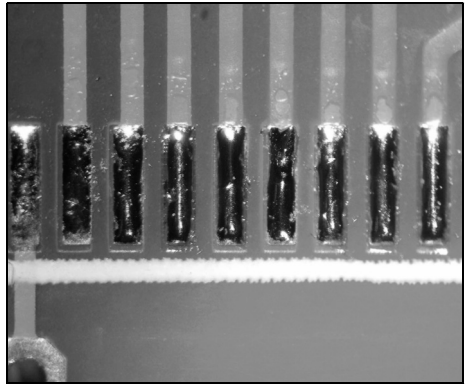
Fig 12: A PCB with the pads tinned and covered with flux.

during this process (this would lead to short-circuits) you have one hand free. This free hand can be used to hold a magnifying glass around the soldering iron to correctly position it and observe the soldering process.

This method has worked well for me. After soldering all pins the work can be inspected with a magnifying glass. Figs 13 to 16 show examples of 0.5mm or 0.65mm pitch ICs soldered using this procedure. Even SMD ICs without pins can be still used with this method if they have lead connection up to the edge of the IC. Figs 14 to 16 show that this technique works even on a even corroded PCB. A professionally manufactured

PCB with a solder mask is the simplest to use.

It is important to take suitable steps to prevent static when soldering. The soldering station and the operator must be earthed. The soldering iron should be earthed to a suitable point such as a water pipeline or heating element. The operator can be earthed using an anti-statics bracelet.

Remember

The most important things are a good optical magnifier, plenty of flux and not to have a fear of very small ICs.
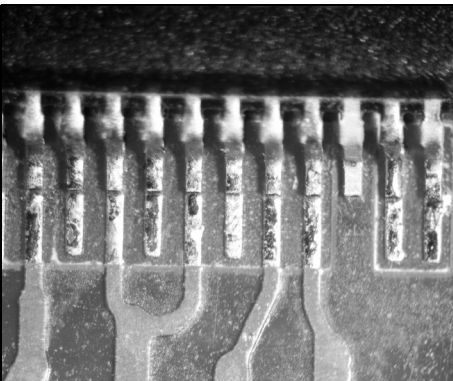
Fig 13: View of a 0.5mm IC solder using this technique.
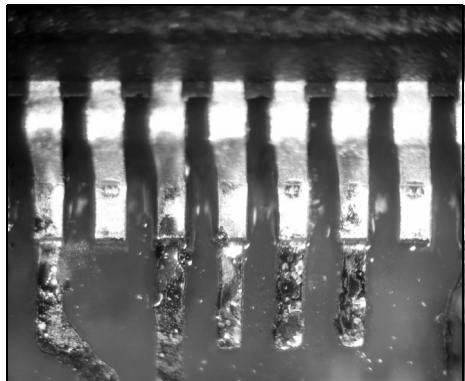
Fig 14: View of a 0.65mm IC solder using this technique.

Fig 15: View of a 0.5mm IC solder using this technique on a poor PCB.
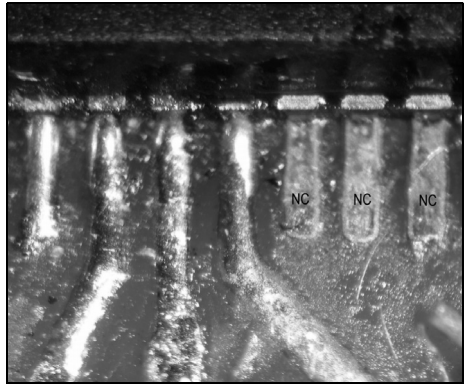
Fig 16: View of a 0.5mm IC solder using this technique on a poor PCB.

3.

Source of supply

The items mentioned in this article can be obtained from the following:

[1] Realm ELT electronics: www.reichelt.de

[2] The BMS S-10 stereomicroscope is available in the UK from Amazon: www.amazon.co.uk

[3] FL88 flux is available in the UK from Farnell: www.farnell.com

André Jamet, F9HX

Become familiar with ceramic capacitors for good use from DC to microwaves

1.

Introduction

Many materials are, or were, used as the dielectric for capacitors: air, mica, paper impregnated with oil, wax or PCB, polypropylene, polystyrene, PTFE (Teflon®), glass, Mylar® and various ceramics. The so-called electrochemical capacitors aluminium or tantalum are a class apart. Ceramics capacitor replace more and more other dielectrics and even compete at the higher values. They are in the form of leaded components: tubes, round or rectangular plates and SMD for surface mounting.

To obtain high capacity in the minimum size the 'multi-layer' technique is widely used. As shown in Fig 1 for an SMD component, several layers are superimposed, up to several thousand. This is the MLCC technique: "Multi Layer Ceramic Capacitor".

2.

Ceramics

Without going into unnecessary detail of their chemical composition for our purposes, it is necessary to know some properties of the ceramics used. Indeed, certain characteristics of capacitors like capacitance, reactance, impedance,
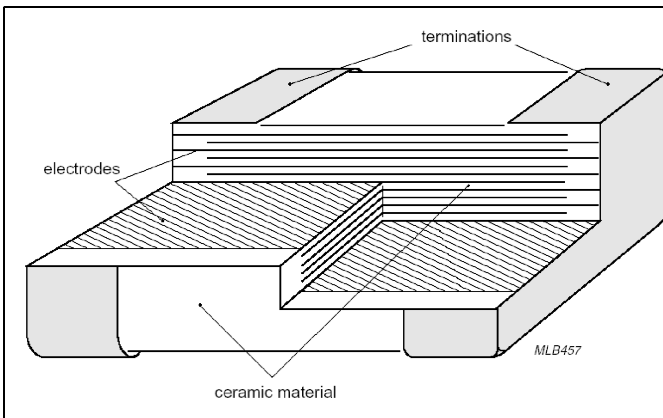
Fig 1: Cross-section of a multilayer SMD capacitor.

Table 1: Parameters of dielectric materials.

Material	Permittivity εr	Temperature coefficient $(10^{-4}/K)$	Losses $\tan\delta$ (10^{-4})
Teflon ®	2.1	-2.7	6 at 1MHz 9 at 5GHz
Mica	6 to 7	-0.2 +1	≤2 at 1MHz
COG SMD [1]	6 to 30	<10pf <±1.5 ≥10pF <±0.3	-40°C:6 0°C:6 120°C:7.2 at 1MHz
Mylar ®	3.2	+3	40 at 1MHz
Polystyrene	2.5	-1.2	1 at 100MHz 3 at 3GHz

resistance, dielectric loss, resonant frequency, temperature, aging, etc. are specific for a type of ceramic.

Here is an overview of the characteristics of the types of ceramics used and the capacitors that they make.

2.1 COG

This type I (electroceramic) of dielectric is called COG (NP0) because it has a low variation of permittivity εr (dielectric constant) with temperature. It is of the order of a few dozen stronger than that of the so-called "noble" dielectrics, mica, Teflon® or Mylar®. The losses are comparable, very low aging and application of DC or AC voltage does not change the characteristics. See the Table 1 giving published values for the noble dielectrics used in electronics.

Table 1 also gives the values for the material used for COG SMD capacitors. Manufacture of capacitors strongly influences the results that may not fully reflect the intrinsic properties of the dielectric used (mechanical actions, mixture of different materials, impregnation, and coating).

COG dielectric capacitors can replace those with so-called noble dielectric. This is the case for tuned circuits, filters, timers, links between stages and other

circuits that require very good stability of capacity (temperature, voltage, and aging) and low loss.

Unfortunately it is currently not possible to manufacture capacitors above 39nF for wired capacitors and 3.3nF for the SMD. It is necessary to parallel connect capacitors to get larger capacity while maintaining the same properties.

There are also in COG capacitors manufactured "for microwave". These use the same dielectric used for the ordinary capacitors, the characteristics are the same except that the second parallel resonant frequency is specified.

2.2 X7R

This type II dielectric (ferroelectric) provides a permittivity of the order of 1200 with significant losses: $\tan\delta \approx 0.03$ [1,2,3,4]. Unfortunately, this is obtained with large influence by temperature, the applied voltage constancy and aging. However, the influence of frequency is low up to 12kHz. Losses decrease with temperature. The use of X7R capacitors should be limited to decoupling and noncritical coupling. It is possible to obtain 1µF in a 1206 SMD package.

When a circuit is designed it is essential to oversize the capacitors to take account of these predictable changes. See the
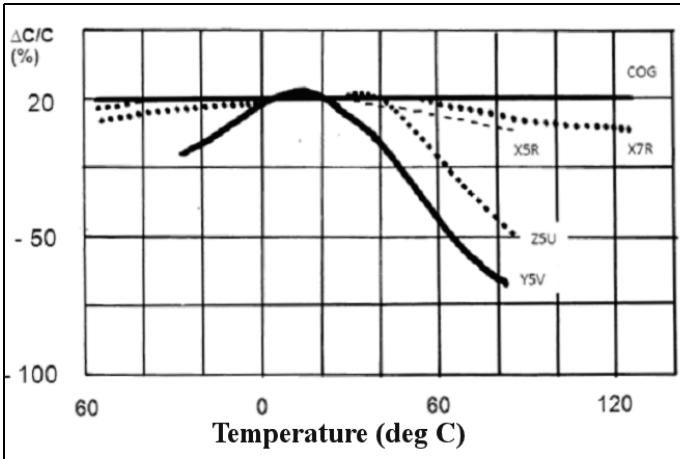
Fig 2: Capacitance variation vs. temperature of various ceramic materials.

curves in Figs 2, 3, 4 and 5.

2.3 Z5V and Y5V

Type III (mixed, conductive insulation), their permittivity is higher and can be up to 8000 [1,2,3,4]. This means that very high capacity can be obtained in SMD form that can compete with tantalum capacitors. It is possible to obtain more than 10μF in 1206. But the above defects of the X7R type are increased further, their use must be considered well to avoid any problems. Permittivity varies greatly with temperature, the applied voltage and aging. Filtering can be insufficient during the life of a circuit using this type of capacitor, especially in extreme temperatures. Tanδ is in the order of 5.10^{-2}. See the curves in Figs 2, 3, 4 and 5.

2.4 Porcelain

It is familiar as a special case for the microwave enthusiast. They commonly use the ATC 100a 1pF with different dimensions to the usual SMD components. The permittivity is in the order of 15. Losses are very low at high frequencies that are significant both upon a receiver (noise factor) or a transmitter (power transmission). There are models with capacity of a fraction of a picofarad
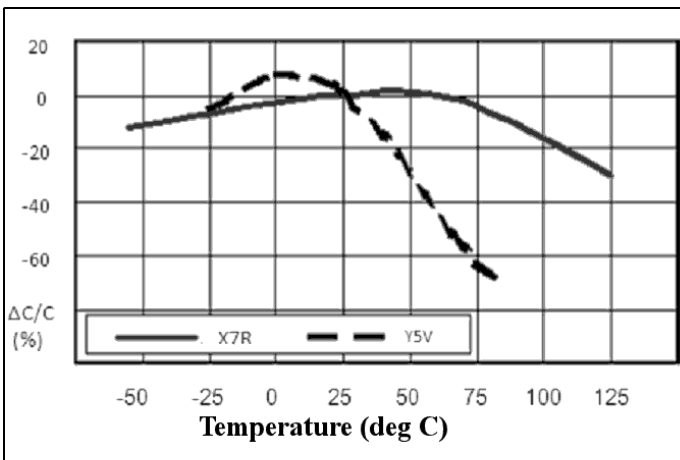
Fig 3: Capacitance variation vs. temperature for a 10μF 10V capacitor.
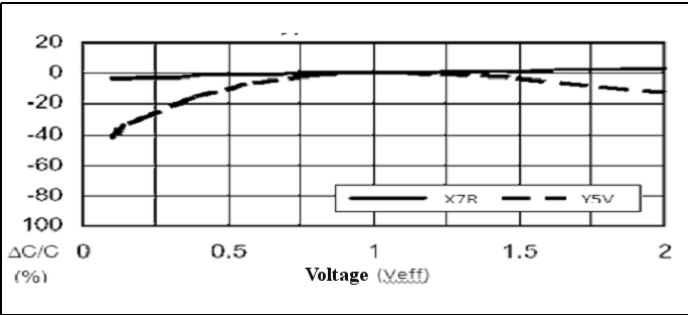
Fig 4: Capacitance variation vs. AC voltage for a 10µF 10V capacitor.

to a nanofarad and voltages (powers) that can reach high levels.

Losses are: $\tan\delta = 7.10^{-5}$ at 100MHz, 10^{-4} at 300MHz and $2.5.10^{-3}$ at 1GHz

Temperature coefficient is: $+90 \pm 20$ppm from -55 to +125°C (more than COG!).

Because these capacitors are used as transmission elements for microwaves, this is not a problem. However at VHF or UHF they are used in tuned circuits, this should be taken into account.

As with all SMD components this small beautiful porcelain capacitor must be soldered flat on the printed circuit board for a minimum of parasitic inductance and the visibility of markings [3].

The familiar "small" 1pF capacitor is ten times the price of a COG. It is used at 10GHz that is approximately its resonance frequency.

2.4 Other difficulties with high permittivity

You should know that this type of dielectric causes effects similar to those of piezoelectric ceramics used in microphones and loudspeakers. The capacitors "sing" if a musical frequency is applied to them, and they produce a voltage if they are subjected to shocks or vibrations (transformer, fan). Avoid using them in Hi-Fi, preamplifiers at low level and VCOs.

Their low resistance series (ESR) is an asset in DC smoothing. But, in some cases, they have caused headaches to designers when they wanted to replace aluminium or tantalum capacitors. The later are very "placid" while ceramics are more "nervous"! The current absorbed at power-up can be very high.

In addition, applying a voltage step to
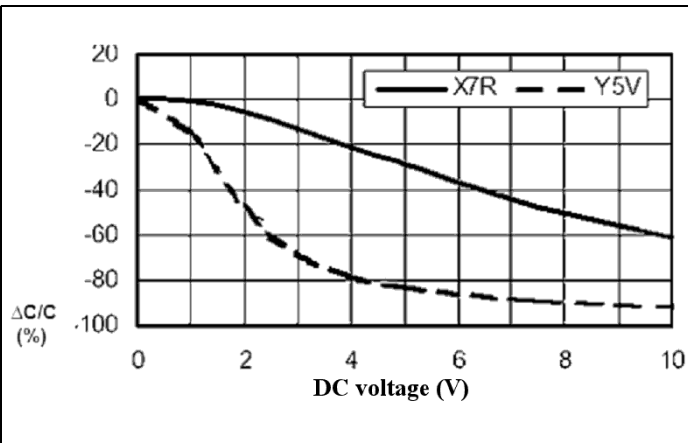
Fig 5: Capacitance variation vs. DC voltage for a 10 µF 10V capacitor.
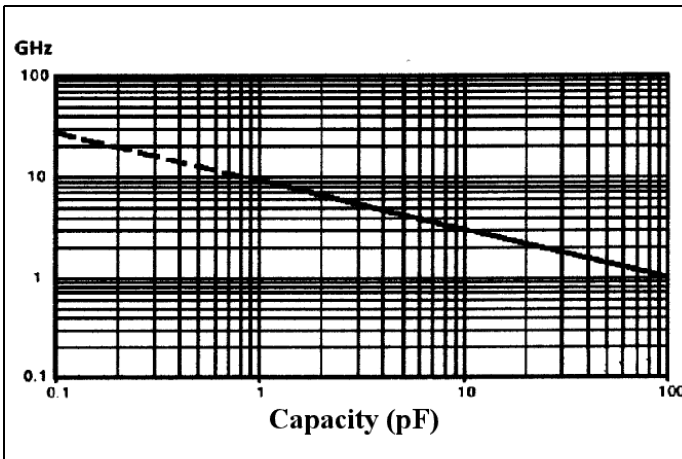
Fig 6: Series resonance vs. capacitance for an ATC100A capacitor.

these capacitors causes a large current that stores energy in the inductance of the power leads. A large voltage spike is created when the stored energy is transferred from this inductance into the ceramic capacitor. These voltage spikes can easily be twice the amplitude of the input voltage step. It must therefore be taken into account in this type of design and eventually a damping resistor is needed in series with the capacitor [8].

3.

Conclusion

For home made devices, it is necessary to know the components used well to achieve the desired results, especially when its material is used in difficult conditions. Portable operating in winter or at a sunny place in summer may cause much frustration, even with amateur equipment sold by professionals.

4.

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Sam Jewell, G4DDK

Signal Hound: Compact 4.4GHz spectrum analyser and matching tracking generator

1.

Introduction

A spectrum analyser is a very desirable item of test equipment. It enables you to check the spectral cleanliness of your signal, find instability problems and can act as a selective measurement receiver and a great deal more. A new professional spectrum analyser can cost many tens of Thousands of pounds. Even preowned spectrum analysers can be very costly. Then if something goes wrong they can be expensive to repair. But all that is changing. Several companies now offer high performance spectrum analysers, based on software defined radio (SDR) technologies, within the budget of many amateurs.

Fig 1: The SA44B spectrum analyser (front) and TG44 tracking generator.

The Signal Hound SA44B (Fig 1) costs less than many HF transceivers. It is made by the American company Test equipment Plus (TEP) [11 and is a high performance spectrum analyser covering from near DC to 4.4GHz with a performance that should be more than adequate for most radio amateurs.

The SA44B can also be used as a measurement receiver with CW, SSB, AM and FM demodulation capability across the entire 4.4GHz frequency range. The claimed amplitude accuracy is better than ±1.5dB from 0dBm down to the displayed noise level (which depends on frequency and resolution bandwidth) and ± 2dB from 0 to +10dBm. Using the internal reference the frequency accuracy is better than 1ppm (part per million).

2.

Prerequisites

The Signal Hound is not a standalone item. A laptop or desk top computer with two or more USB 2.0 interfaces is required to use the Signal Hound SA44B and TG44.

The Signal Hound is powered from the computer over the USB interface. The same interface is used to pass data to and

Fig 2: Front of SA44B analyser, The front of the TG44 tracking generator is similar.

Fig 3: SA44B rear panel. The TG44 is similar

from the instrument. All data processing is done in the computer; you'll need a Windows system with at least a 2GHz processor and 1GB of RAM.

The SA44B spectrum analyser and TG44 tracking generator are housed in identical extruded aluminium cases. At just 20cm long and weighing 290g they should be easily accommodated on most work benches.

3.

Brief overview of the analyser system

I was requested not to open up the case of the review analyser so, of necessity, this section relies heavily on the manufacturer's description of the Signal Hound hardware.

From the basic block diagram in the manual, it appears to use a conventional superhet RF front end with a digital sampling back end that converts the filtered intermediate frequency signal to an I/Q (in-phase and quadrature) bit stream. The bit stream passes over the USB connection to the host computer for signal processing and display on the computer screen. Few details of the signal processing are given in the manual.

The spectrum to be displayed enters the SA44B through an SMA socket on the front panel (Fig 2) and then to an attenuator with three selectable steps, from 0 to 15dB. The attenuator is directly connected to the input in order to retain response down to DC. This could make the attenuator vulnerable to damage if any DC appears at the input. For this reason an external DC block, with good frequency response to at least 4.4GHz, should he used whenever the low frequency response is not needed. The DC block is not supplied with the SA44B, but is available as an accessory.

The attenuator is followed by a wideband preamplifier that can be switched into the signal path when extra sensitivity is required. The preamplifier cannot be used below 500kHz.

After the preamplifier the input spectrum passes to one of two mixers. In its simplest operating mode, the local oscillator that feeds the two mixers is stepped, under software control, over the required frequency range in 200kHz steps (or less) in order to produce a series of bands at 200kHz intervals. These are filtered to 250kHz wide individual bands by the 10.7MHz IF filter. This arrangement produces a whole series of responses from the input spectrum. At least one of the responses will be the band of image frequencies from the input spectrum. The software is left to process this complex spectrum to identify and remove not only the image frequencies but also any other internally generated spurious signals. It does this extremely effectively.

Table 1: Measured parameters.

Parameter	Manufacturer's claim	Measured	Comments
Frequency range	1Hz to 4GHz	<10kHz to 4.4GHz	
Marker frequency accuracy	±1ppm	0.25ppm*	Internal TCXO at 21°C * measured with 100kHz span at 2.4GHz
Marker amplitude accuracy 100MHz			
0dBm		-0.6dB	
-30dBm		-0.4dB	
-60dBm		-0.5dB	
1GHz			
0dBm		-0.35dB	
-30dBm		-0.3dB	
-60dBm		-1.0dB	
2.4GHz			
0dBm		-0.1dB	
-30dBm		-0.5dB	
-60dB		-1.5dB	
1dB gain compression			
100MHz	+16dBm typical	+18dBm	15dB attenuation, preamp off
1GHz	+19dBm typical	+20dBm	
Displayed average noise level (DANL)			Marker offset to avoid low level spurious output at 100MHz and 1GHz
100MHz, preamp off	-148dBm	-158dBm	Normalised to 1Hz from
100MHz, preamp on	-161dBm	-170dBm	measurement at 100Hz
1GHz, preamp off	-144dBm	-152dBm	resolution bandwidth
1GHz, preamp on	-158dBm	-165dBm	
Residual spurious at the input connector		Att. 15dB, preamp on	Att. 0dB, preamp off
1Hz to 500kHz	-70dBm	Not measured	Not measured
500kHz to 1GHz	-57dBm	-57dBm	-20dBm
1GHz to 2.3GHz	-47dBm	-55dBm	-20dBm
2.3GHz to 2.6GHz	-40dBm	-55dBm	-28dBm
2.6GHz to 3.0GHz	-27dBm	-60dBm	-27dBm
3.0GHz to 4.4GHz	-35dBm	-48dBm	-27dBm

A disadvantage of this technique is that covering a span of 4.4GHz in 200kHz steps can take over 30 seconds. Narrower spans take considerably less time. A 5MHz bandwidth filter is used during wide frequency sweeps. Narrow bandwidth filtering down to as low as 0.1Hz takes place in the host computer using FFT (fast Fourier transforms).

For anyone familiar with conventional spectrum analysers the Signal Hound SA44B may require a period of adjustment. It took me about an hour of playing with the SA44B to familiarise myself

with it enough to start to get the best out of it. It can be time consuming to use the SA44B to adjust circuits as the screen display is not in real time. There can be a noticeable delay between making adjustments and seeing the result on the screen. In these circumstances it is best to make circuit adjustments using narrow spans and then to open out the span to see the results at higher or lower frequencies.

An external 10MHz reference input, via a BNC socket on the rear of the SA44B (Fig 3), can improve the frequency accuracy of the instrument as well as its phase
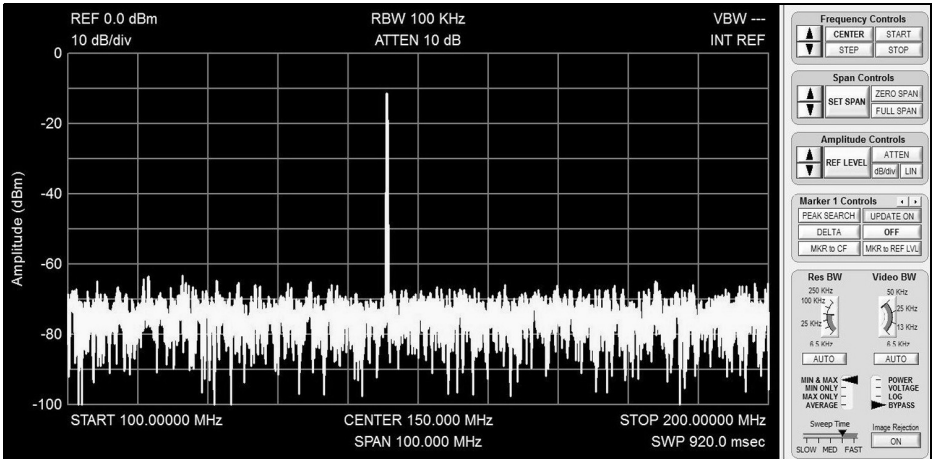
Fig 4: 144MHz output from Kenwood TH-6F.

noise performance. A second BNC connector on the rear of the SA44B provides a trigger input/output connection. This is useful for synchronising sweeps to some external event. It will accept both 3.3V and 5V (TTL) levels. The TG44 tracking generator sweep is triggered from this same connection.

Located centrally on the rear panel is a type B USB 2.0 connector to allow the supplied USB lead to connect the Signal Hound to the host computer. A bi-colour LED on the front panel shows green when the Signal Hound is powered and flashes amber when it is receiving data from the host computer.

4.

Graphical user interface (GUI)

A software defined radio relics heavily on the quality of the controlling application software. I'm pleased to say that the Signal Hound GUI seems to have been well thought out, is easy to use and works well. The version current at the time of this review was 2.07. This version provides support for the tracking

generator as well as a number of useful utilities such as harmonic measurement, channel power and phase noise. These are in addition to the usual spectrum analyser facilities of carrier level and spectral response.

Fig 4 shows the captured output spectrum from my Kenwood TH-6F handheld (suitably attenuated) when transmitting an unmodulated carrier on the 144MHz band. The GUI provides a large screen area for the spectrum display in a 10 by 10 matrix. The usual spectrum analyser control 'buttons' are located down the right side of the screen. The 145.500MHz signal is the vertical line near the middle of the frequency scale.

Frequency, Span and Amplitude buttons control the main functions of the analyser. In this case the top of the screen has been set to a 0dBm (1mW/50Ω) reference level.

The vertical screen display is set at 10dB/division and the horizontal screen is set to show from 100MHz on the left to 200MHz on the right. Sweep time is 920msec. The displayed average noise level (DANL) is approximately –75dBm with the 144MHz carrier at –12dBm. No markers have been selected; the resolu-
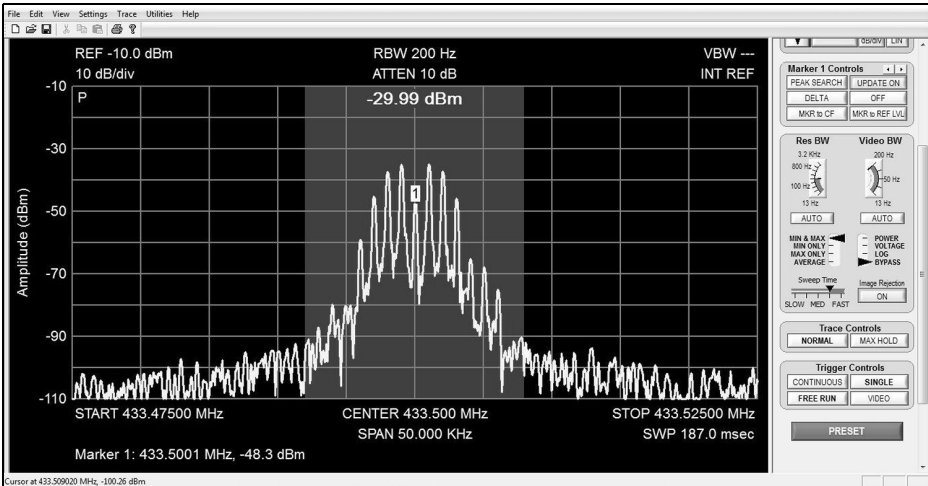
Fig 5: 433MHz FM modulated output from the Kenwood TH-6F.

tion bandwidth (set automatically by the computer) is 50kHz.

Image reject is switched on. When switched off there are several spurious responses in addition to the main image 21.4MHz below the wanted display signal. This indicates over 70dB of spurious display rejection at this frequency. The same or better performance is seen with a carrier at 433MHz.

Of course an analyser needs to be able to 'zoom' right into a signal in order to be able to analyse modulation. In order to obtain a clear and unambiguous display this means using a resolution bandwidth filter that is much less than that of the occupied bandwidth of the modulated signal. Fig 5 shows a 433.5MHz narrow band FM signal. This is a screen capture showing the output from my Kenwood TH-6F handheld, modulated by a single tone. The resolution bandwidth of 200Hz clearly shows the individual sidebands of the FM signal.

Along the top of the display in Fig 5 you can see a row of buttons including one marked 'Utilities'. When you click on this a drop-down menu allows access to a number of useful functions. The first of these is 'Audio listen'. This opens a small window from which you can select frequency, IF bandwidth, mode and de-emphasis. It is first necessary to display the signal you want to listen to by placing the marker on that signal. Now you can apply the selected demodulation parameters. The computer speaker allows you to listen to the selected signal. A little lower down the utilities menu you will see 'Measuring Receiver'. This opens a window showing the frequency and modulation characteristics of the selected signal. Many other utilities are included but space precludes reporting on them all.

5.

Measuring spectrum analyser

Since a spectrum analyser is likely to be relied upon for a number of amateur radio related measurements, some idea of its accuracy was required.

I felt it was worth checking a number of the manufacturer's claims against professional test equipment that originally had
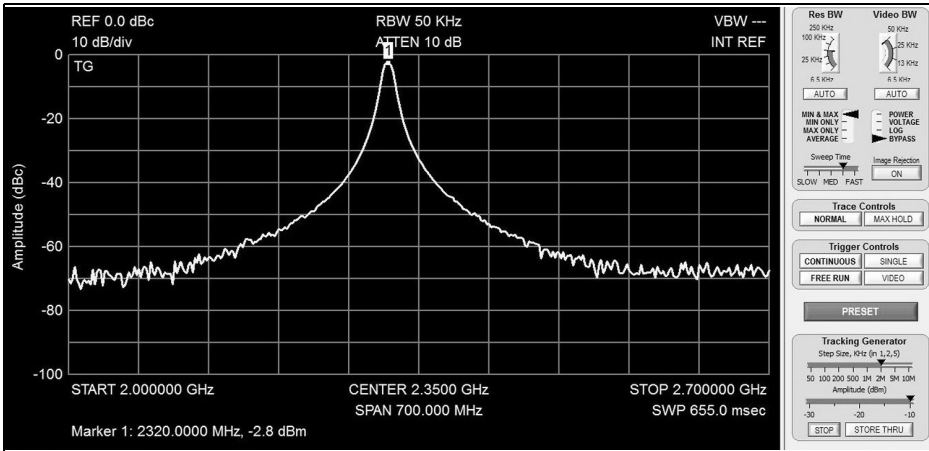
Fig 6: 2.3GHz filter response plotted with the help of the TG44 tracking generator.

cost many times the price of the Signal Hound. In Table l the first column identifies parameters, column 2 shows the manufacturer's specification and column three contains my measurements. The fourth column has my comments. Although my equipment is not formally calibrated, it is regularly compared with calibrated equipment. Even so, some discrepancies may be noticed, which may be due to differences in measurement technique or interpretation of specifications.

6.

TG44 tracking generator

The TG44 is housed in the same robust case as the SA44B and has a similar single SMA connector on the front panel for RF output, two BNC connectors on the rear panel for interconnection to the spectrum analyser and a type B USB 2.0 USB connector for connection to the host computer. A BNC to BNC lead, USB cable and SMA male to male adapter are supplied with the TG44.

The TG44 can be used with the SA44B spectrum analyser to create a simple but effective scalar network analyser system that can be used to measure RF gain, frequency response, gain compression and insertion loss. Used with a suitable directional coupler or bridge, it can also be used to measure return loss. The reviewed software does not allow for normalisation in this mode, but this may be added in a future release.

In addition to use as a tracking generator, the TG44 can also be used as a CW (unmodulated) signal generator will output from 10Hz to 4.4GHz. This is an enormous frequency range and makes the TG44 worth having in the shack for this purpose alone. Note that the harmonic output of the generator is quite high, which may make it unsuitable for some measurements. An external low pass filler or two could easily solve the harmonic problem for many applications.

7.

In use

Thc TG44 was used in conjunction with the SA44B to measure the response of a two pole interdigital filter tuned to

247

2.320MHz. This filter forms part of my 2.3GHz band system.

Control of the TG44 is from the spectrum analyser GUI. The controls for sweep and level are located at the lower right hand side of the GUI screen. In order to perform a sweep of a filter, such as shown in Fig 6, the sweep frequency limits should be set as appropriate. For a filter the RF level should not be important, so use the maximum available from the TG44 in order to achieve the largest on-screen dynamic range.

First, connect the tracking generator directly to the spectrum analyser using the RF cable and connector adapters you will be using to connect to the filter.

There should be a flat, straight line on the display. However, it will probably be some way below the top-of-screen reference line. Click the 'STORE THRU' button on the GUI screen. The through loss will now be subtracted and the display will move to the top of the screen. Connect the filter or other device under test (DUT) between the test leads. The frequency response of the DUT will appear on screen. Click the marker peak search and the on-screen marker will appear at the peak of the DUT response. If it is a filter then the filter insertion loss will be shown at the bottom of the screen. Using marker delta or two markers, the filter bandwidth or filter out-of-band response can be measured. Wide sweep widths require the TG44 10MHz output to be connected to the SA44B 10MHz reference input.

8.

Conclusion

My measurements on the spectrum analyser revealed surprisingly good results. Amplitude accuracy was particularly good and certainly more than adequate for the to majority of amateur radio

purposes.

Frequency accuracy, even using the internal TCX0, was much better than expected, but could he improved upon further by using a suitable external 10MHz reference, preferably one will low phase noise. This makes the SA44B very useful as a selective frequency measuring device, particularly when several signals are present or waveforms are very distorted. A frequency counter alone can easily be fooled into measuring and indicating the wrong frequency under these conditions.

The discrepancies in the DANL measurements in the measurements table may be due to differences in where the visually averaged noise waveform was taken to lie.

I wonder if TEP might not have updated their manual from a previous version of the SA44B since the spurious outputs at the input connector seemed unduly pessimistic. I added a fourth data column to the table entry to show that without the attenuator set to 15dB and the preamp switched out, the spurious levels were closer to those in the SA44B manual. The measured figures are very good for this type of receiver where there is no RF band-pass filter to help with LO rejection.

I mentioned near the beginning of this review that it might take a little getting used to compared to a conventional spectrum analyser. The reason for this is that wide band sweeps can take many seconds to complete. In some situations this can make this type of at analyser difficult to use for adjustment purposes, since you might have to wait a long time to see the results of each adjustment made. For narrow sweeps this should not be a problem.

The TG44 tracking generator proved extremely useful but the GUI display froze a few times when making adjustments to the swept frequency range. This necessitated re-booting the software. However,

an updated TG44 was due for imminent release and this should eliminate these (firmware related?) problems.

I would have liked to have seen a facility for saving screen shots. Printing is already taken care of, but if you want to save a screen shot, such as the display of the TH-6F FM modulation, it is necessary to use CNTL Prt SC and then import the bit map into a suitable editing program. If there is a facility to do this already, I didn't find it!

For the tests I used a Sony Vaio laptop with Intel CORE i3 processor with 2.3GHz clock and 6GB RAM running on Windows 7 Professional.

Overall, I was very pleased with I the performance of the SA44B. If I didn't already possess a good spectrum analyser I could be very tempted to buy one of these. The TG44 is perfectly useable and has excellent frequency coverage.

My thanks to Test Equipment Plus for the loan of the review samples. The Signal Hound SA44B spectrum analyser costs €799 + 19% VAT and the TG44 tracking generator is €520 + 19% VAT. The European distributor is Dutch company Foltronics [2]

9.

References

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[2] Foltronics: www.signalhound.eu
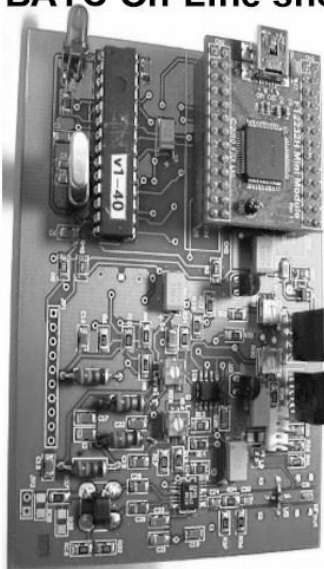
André Jamet, F9HX

Temperature measurement of electronic components, applied to a 10GHz PA

In this article I suggest possible ways to carry out temperature measurements on electronic components in situ.

1.

Introduction

The temperature of electronic components affects their characteristics. First of all, in the short term, for example the diode voltage drop, the collector current of a transistor, the value of resistors or capacitors, etc. In the long term, ageing appears. The Arrhenius equation - to be considered carefully according to the type of component - indicates that the lifetime of a component decreases to half whenever its temperature increases by 10°C. During this life, features evolve and may make it to perform poorly.

It is therefore important to operate a component within its boundaries of temperature from the design parameters. To confirm calculations, measurements in situ are useful or necessary.

2.

What is temperature?

In physics, temperature defines itself in several ways: such as increasing the degree of thermal agitation of particles, by the balance of heat transfer between multiple systems or entropy. Several scales are used according to countries and the use. Water vaporisation or boiling is given for a pressure of 1 atmosphere = 101.325 Pascal. Table 1 shows some commonly used temperature scales

Temperature is measured with a thermometer.

3.

Thermometers

Some characteristics of a body vary with its temperature. It is therefore possible to measure one of them and convert to degrees.

Table 1: Commonly use temperature scales.				
Scale	Degrees Celsius	Degrees Centigrade	Degrees Fahrenheit	Degrees Kelvin
Absolute zero	-273.15		-459.67	0
Ice fusion	0	0	32	273.15
Water vaporisation	99.975	100	212	373.125

Almost all bodies increase in volume when temperature increases. This is used in an alcohol thermometer. The line elongates in a capillary tube. After calibration, or use of the manufacturing parameters, it can have a graduated scale. The displayed value is then the temperature of the bulb containing the reserve of alcohol. Similarly, a metal spiral elongates with temperature and index indicates the temperature on a tick mark.

NTC resistors and diodes used in conjunction with an electrical measurement of the variation of their characteristics are often used to make thermometers. Their active element is housed in a probe with dimensions appropriate to its intended use.

Thermocouples formed using two different metals produce a junction voltage as a function of the temperature.

The method of measurement of infrared with laser light is very interesting and offers many opportunities.

There are also temperature sensitive labels that change colour at defined temperatures.

4.

The difficulties of temperature measurement

As for electric measurement, it is necessary that the means of measurement does not disrupt the value measured. A voltmeter must have a resistance much higher than that of the source to be measured. A thermometer must not alter the temperature of the measured body.

Almost all thermometers measure their own temperature (the part that is used to measure). This temperature is that of the local medium if there is thermal equilibrium between the thermometer and the environment. Most measurements only

concern the surface of a body unless the thermometer is housed in the measured body.

This explains the difficulty of measuring the temperature of electronic components in situ.

The last three methods referred to in paragraph 3 are best suited to the measurement of the temperature of electronic components: thermocouples may have very small dimensions, in the order of a millimetre. The sensitive labels can be fixed to the component if it is large enough and attached on its surface.

5.

Thermocouples

When two different metals are connected by two junctions subjected to different temperatures, the Seebeck effect produces a potential difference proportional to the temperature between the junctions. The thermocouple method measures a temperature difference. To measure the unknown temperature of a junction, the other junction must be maintained at a known temperature. The size of the potential difference requires sensitivity because it is only microvolts. Measuring instruments with a compensation device giving a virtual cold source simplifies the use of thermocouples. Several pairs of metals are suitable as thermocouples.

There is difference in the voltage produced by the same type of thermocouple. For precise measurements, it is essential to calibrate the thermocouple after ageing of each specimen used.

A thermocouple made from copper/constantan is possible but great accuracy is required to make the junction.

It is simpler to use a ready made K type thermocouple. It consists of the chromel (alloy of nickel and chromium) and the
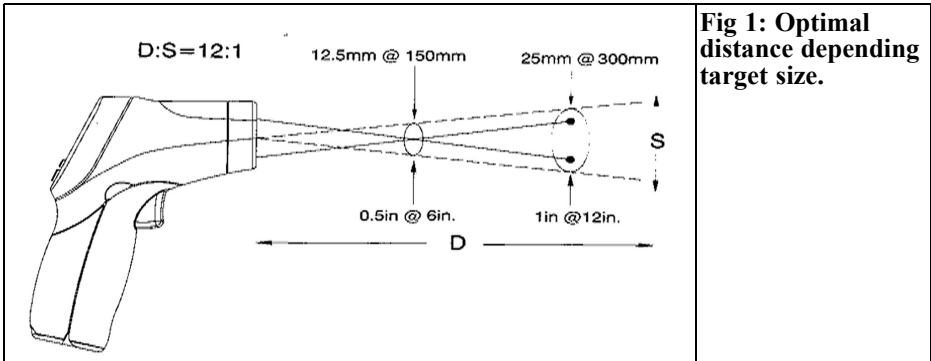
Fig 1: Optimal distance depending target size.

alumel (alloy of nickel and aluminium). The voltage produced is of the order of 40μV/°C and requires the correct measuring instrument designed for its use.

There are very small K type thermocouples, as small as a 1mm sphere. Despite this small size the wire connection adds to this and causes a loss of measured body heat. Measurement of transistors and the body of a resistor can be distorted. The temperature of larger objects can be done with precision, such as the inside of a transformer winding, a heat sink or the inside of an OCXO.

6.

The infrared thermometer

This method of temperature measurement is based on the radiation in the infrared spectrum of a body due to its own temperature. The radiation is transferred with conduction and convection, which are means of heat transfer. This method measures light energy with an IR detector to convert the radiation to an electrical signal. The measured temperature is that of the target surface.

The overriding interest of this measurement method is that it takes place without contact with the object to be measured. There are "guns" to cover the measurement target. The most advanced have a

dual-beam laser for very precise aiming. Temperatures can range from –50 to + 650°C

The viewfinder to object distance is critical for measurement to be taken at exactly the desired point. The use of two lasers makes it very easy to do because the optimum is obtained when the two laser beams converge in this point (Fig 1).

A very important factor is the value of the emissivity of the surface to be measured. It depends very much on the nature of the material and the state of the surface that must be clean and dust free.

The emissivity factors for common materials in electronics is as follows (not certified values by all authors working on this subject):

- matt painting 0.93
- plastics brilliant 0.5 ?
- plastics matt 0.93
- brilliant paint 0.8
- polished copper 0.07
- oxidized copper 0.78
- polished aluminium 0.07
- black oxidized aluminium 0.8
- black smoke 0.96

Infrared imaging comes into this category of temperature measurement. A camera can record and show the entire surface of a printed circuit in operation. Using a set of colours we see each
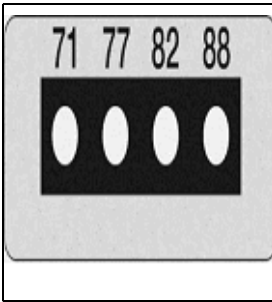
Fig 2: Example of a temperature sensitive label.

component showing its temperature. What a valuable tool in the design, maintenance, and troubleshooting, even production quality control. It allows small defects to be highlighted such as bad welds and bad contacts. The cost of this method seems out of reach to the amateur who doesn't have a laboratory or a production unit.

7.

Temperature sensitive labels

This method is very accurate and easy to implement. Sensitive labels have the property to irreversibly change colour when the assigned temperature is reached. The most useful for our application comes in the form of a strip with 4 small temperatures areas (Fig 2).

The product characteristics are:

- Strip dimensions: 11 x 4mm
- Response time: 1 second
- Change of colour: from light grey to black
- Impermeable to oil, water and steam
- Accuracy: ±1°C <100°C ±1% >100°C
- Available temperature range:
- Reference 1: 40 42 44 46 49 54 60 65 71 77 82 88 °C
- Reference 2: 93 99 104 110 116 121 127 132 138 143 149 154 °C

Table 2: Temperature measurement results.

Method	Thermocouple	IR Laser	Label
°C	64.6	≈60	>65 <74

- Reference 3: 160 166 1714 177 182 188 199 199 204 210 216 224 °C

Use: clean, flat surface over the size of the strip

The cost is quite high and a label that has "turned" is unusable except for temperatures above those already affected.

8.

A practical example

The 15W 10GHz PA is a good example to measure the surface temperature of the critical components [1,2]. The final transistor is attached by two screws onto a milled aluminium case fitted to a heat sink with forced air convection. This transistor dissipates a static 30W and the temperature of the bottom of the aluminium case was measured with a K type thermocouple as 47°C for 20°C of ambient. The calculation of the thermal resistance leads to a 118°C junction temperature.

The upper surface of this transistor has been measured successively by three methods studied above.

- A 1mm type K thermocouple connected to a device for converting μV/°C with a resolution of 0.1°C and a precision of 1°C. It was placed against the top of the transistor case and held in place by a drop of glue next to the contact point.
- A dual-beam infrared thermometer with a resolution of 0.1°C and an accuracy of ±1% and ±1°C. The surface of the transistor was cov-
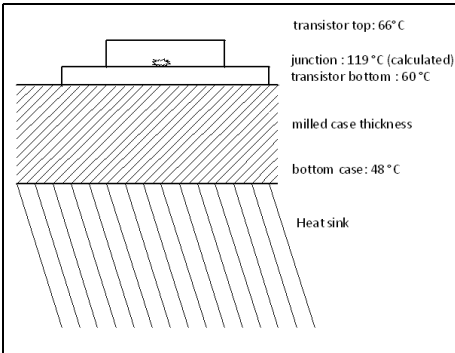
Fig 3: Temperatures measured and calculated for the final transistor.

ered with a small pencil soaked with matt black paint.

• The sensitive label was stuck with its glue on the surface of the cleaned transistor. Two strips were successively glued to the value that is between two values of labels.

Table 2 gives the values obtained.

The two measurements made by physical contact are very consistent. However, that obtained with the laser gives a lower value. It was conducted by using an emissivity $\varepsilon = 0.93$ being the value for matt black. It can be assumed that the paint layer plays the desired role of low emissivity, but it introduces a temperature gradient between the hot surface and the atmosphere.

This method is subject to errors because the emissivity factor varies significantly according to the state of surface: 0.07 for polished copper, 0.8 for oxidized copper. Therefore, it is difficult to assign the correct emissivity for the target.

Fig 3 shows the measured temperatures and those calculated for the junction of the final transistor based on the values in this section and [1,2].

9.

Conclusion

Knowledge of the temperature of electronic components is valuable to highlight weak points that can lead to more or less serious failures.

10.

References

[1] PA…d'OMsss! (5.7 and 10 GHz), F6BVA, CJ Proceeding 2010

[2] My future 15W 10GHz PA will not run too hot!, André Jamet, F9HX, VHF Communications Magazine, 2/2011 pp 119 - 124

Gunthard Kraus, DG8GB and Andy Barter, G8ATD

Internet Treasure Trove

Windfreak Synthesiser

Two reader of VHF Communications (José Geraldo Chiquito and Ted Bottomley) have contacted Andy to tell him about this great synthesiser The Windfreak Synthesizer is a low cost 137MHz to 4.4GHz software tuneable, PLL synthesized RF signal generator, controlled and powered by a PC running Windows XP or Windows 7 via its USB port. It costs €249.

Address:
http://www.windfreaktech.com/
synthusb.html

Continental Compliance Ltd

This one is for the weak signal enthusiasts with noise sources and noise measurement products. Plus there is an impressive list of links for all those standards that you are wondering about.

Address:
http://www.rfdesignuk.com/

Linear Technology

A familiar name that we know from the free Spice software LTspice. But the product range is unbelievably extensively e.g. active mixers in the GHz range with high IP3. It is worthwhile looking at the documents that are available under "Technical Documentation".

Address:
http://www.linear.com/designtools/

Microwave journal

It is good that this magazine is available and it is on-line. For many years it has been the place that presents new components and good articles. In addition the on-line archives means that you no longer need a mountain of paper.

Address:
http://www.mwjournal.com/
default.asp

Centellax

A company that makes equipment at the forefront of technology and supplies test equipments for electrical and optical data communication up to 100GBit/s. It is worth looking at the application notes offered.

Address:
http://www.centellax.com/support/
appnotes.shtml

What Circuits

If there were a question mark behind the homepage name then the contents of this homepage would be clear. It is an enormous collection of practical circuit examples for almost all areas of electronics. Very interesting and sometimes somewhat amusing.

Address:
http://www.whatcircuits.com/
Category/rf-circuits/

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This DVD contains magazines from 1/2000 to 4/2001 in scanned image format and from 1/2002 to 4/2009 in text and image format. This DVD is £35.00
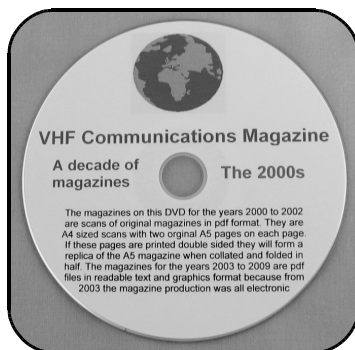
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