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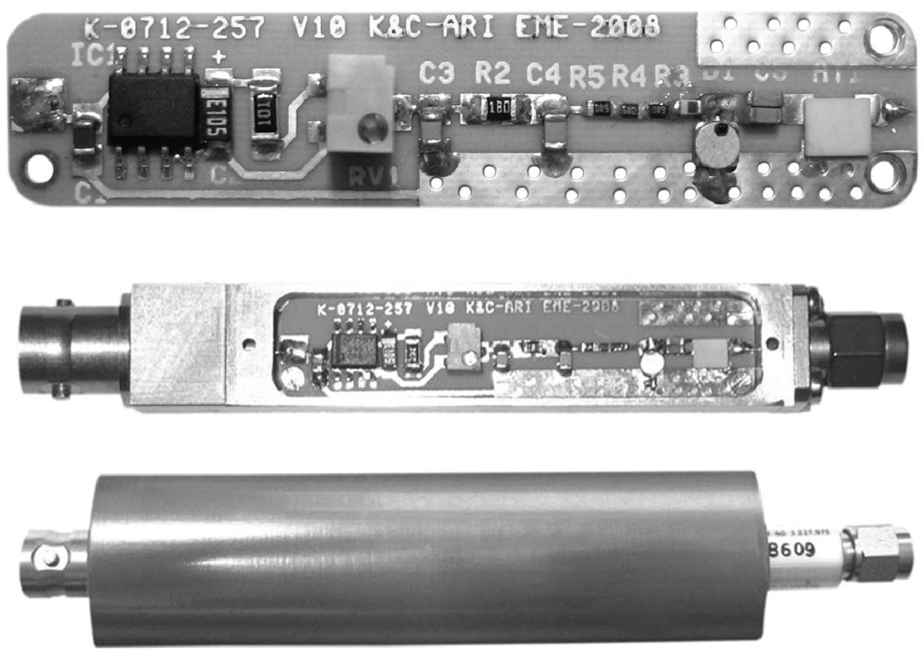

10MHz - 10GHz noise source diodes

Franco Rota I2FWH

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Another year of excellent articles but the supply of new articles is getting more and more difficult. If you have been working on an interesting project and think that it would make a good article in the magazine please contact me and I will help to get it ready for publication. Maybe something like the article that I wrote for issue 2/2008 about the 23cm PA that I built.

Don't forget to renew your subscription for 2009, there should have been a form with this issue of the magazine. If not you can always renew your subscription using the web site, by post to the address below or your normal agent. PayPal is a very popular way to subscribe, just send your subscription to the PayPal account vhfcomms@aol.com.

The second edition of The International Microwave Handbook that I edited is now available from the RSGB or ARRL, see the advert on page 209.

Merry Christmas and a happy New Year. 73s - Andy

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Stefan Przeliorz, SP9QZO

Retuning a GSM band PA from 900MHz to 1296MHz

In Poland and the Czech Republic there is very popular is PA using a BLV958. We can buy a module containing one of these transistors for €15 - €40. It is possible to retune this PA to deliver 50 - 60W on 1296MHz with a gain of 7 - 8dB and 40% efficiency. The supply voltage is up to 27V with a maximum current of 7A. Stefan SP9QZO devised this method, he has retuned a lot off these PAs using this method. It possible use a power combiner to obtain more power from several modules. With reference to Fig 1, the method is:

1 Remove the BGY916 hybrid amplifier.
2 Cut this track.
3 Remove capacitors. Add a piece cooper foil to connect both tracks. This is the input ground for the PA.
4 Cut this track, solder a high quality capacitor of 22 - 33pF across the cut.

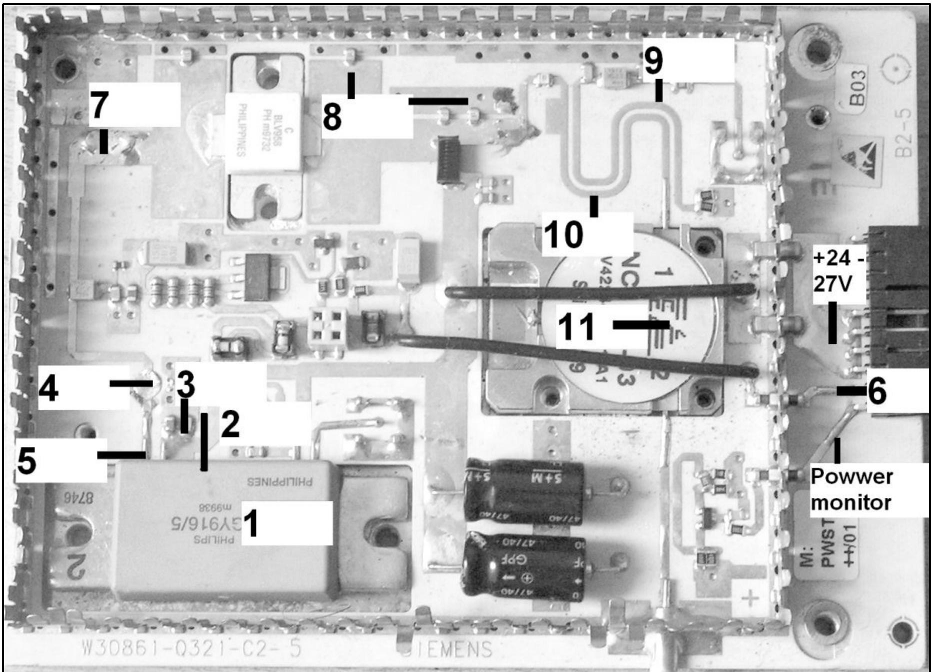


Fig 1: The un-modified PA module.

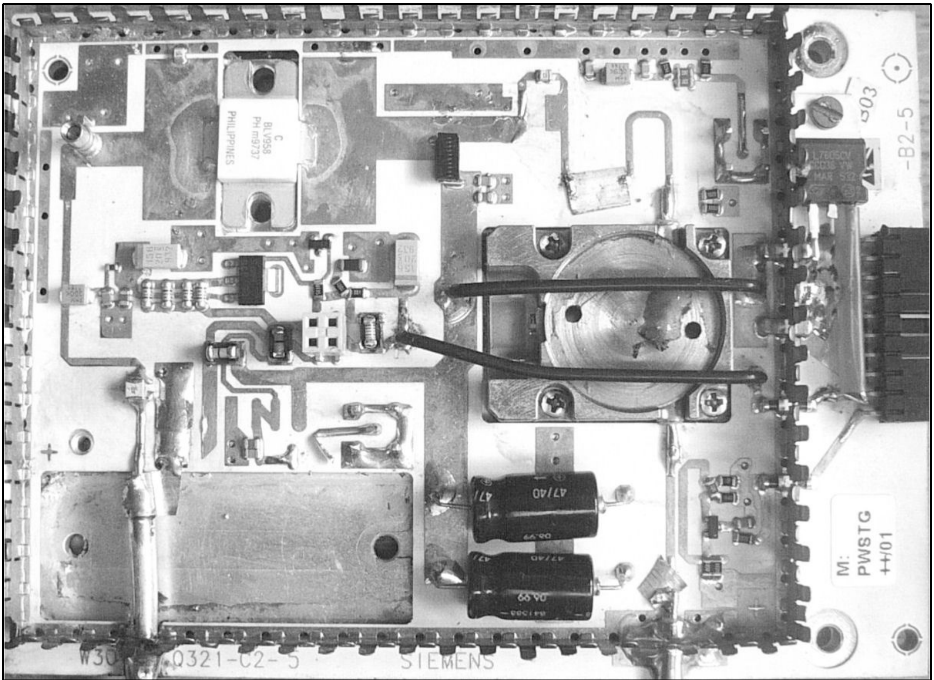


Fig 2: The modified PA module.

5 This is the PA input, connect PA drive using semirigid cable.

6 Connect a positive power supply to this point using a 7805 5V regulator. This sets the collector current to 250mA (class AB). The maximum voltage is 8V that gives class A.

7 Remove all capacitors (without trimmer)

8 Remove all capacitors (without trimmer)

9 Remove this track using a hot soldering iron. It is very simple to remove this track.

10 Add a piece of cooper foil (5mm x 12mm)

11 Remove the circulator and mount a piece of semirigid cable inside the circulator.

Fig 2 shows the modified PA module. With a low power input, tune the PA using both trimmers (typical). Sometimes it is good idea to remove the output trimmer in collector the collector of the BLV958 and tune the PA with a small capacitor (<1pF). Sometimes the output power increases, if so, add a 5mm x 5mm piece of foil on the output of the PA.

References

[1] Article by Stefan Przeliorz SP9QZO translated by Rafal Orodzinski SQ4AVS, sq4avs.googlepages.com

Matjaz Vidmar, S53MV

DDS RF signal generator

1. Theory

In this article a simple, modulated radio frequency signal source is presented. Both the RF signal and the modulating waveforms are generated in a digital way. The principle of operation of a Direct Digital Synthesizer (DDS) is shown in Fig 1.

During each clock cycle, a specified number (the parallel tuning word) is added to the content of an accumulator. The content of the accumulator therefore represents a saw tooth signal whose frequency is directly proportional to the number added to the accumulator. Frequency resolution is defined by the accuracy of the mathematical operation or in other words the number N of bits used in the computations. The saw tooth signal is converted to sine wave using a ROM lookup table. Finally the digital signal is converted to analogue format using a DAC.

The block diagram shown in Fig 1 can be implemented in many different ways. For audio frequencies up to a few MHz a software implementation on a DSP processor is a simple and efficient solution. Dedicated hardware is required for higher frequencies in the radio frequency range. Both solutions are used in this project, the former for the audio modulation and the latter for the RF carrier.

A digitally generated sine wave includes many spectral components as shown in Fig 2. This is taken from the AD9851

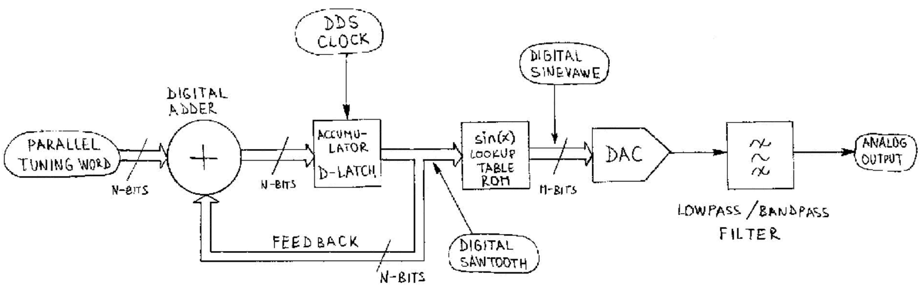


Fig 1: Principle of operation of a Direct Digital Synthesiser (DDS).

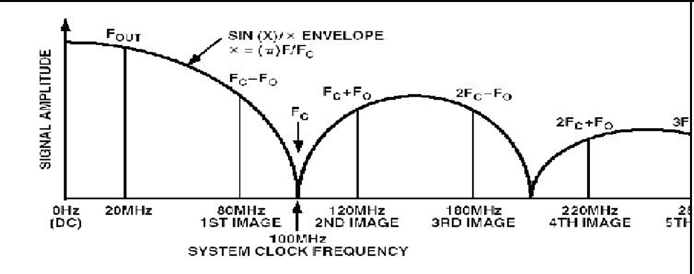


Fig 2: Spectral components of a DDS signal.

data sheet. The unwanted spectral components may be attenuated by using higher clock frequencies or additional filtering of the output signal. On the other hand, higher order image spectral lines can be practically used as additional signals available from the same DDS source.

The DDS RF signal generator presented in this article is based on two chips: an AD9851 generates the RF carrier as a 200MHz DDS, while a LPC2138 ARM microcontroller performs all remaining functions operating at a clock frequency of 60MHz. Inside the LPC2138 a DSP interrupt routine running at 100kHz generates all modulation signals up to 10kHz. Further the LPC2138 accepts commands from the keypad, drives a LCD module and programs the registers inside the AD9851 through a serial interface. Fig 3 shows a block diagram of the signal generator.

Frequency modulation is added to the digital carrier information sent to the AD9851. Since any aliasing products cannot be filtered, the modulation frequency has to be kept much lower than the DSP interrupt frequency (100kHz) in FM mode.

Amplitude modulation is applied in analogue way to the bias circuit of the AD9851 (carrier) DAC. An analogue lowpass filter (LM358) suppresses alias-

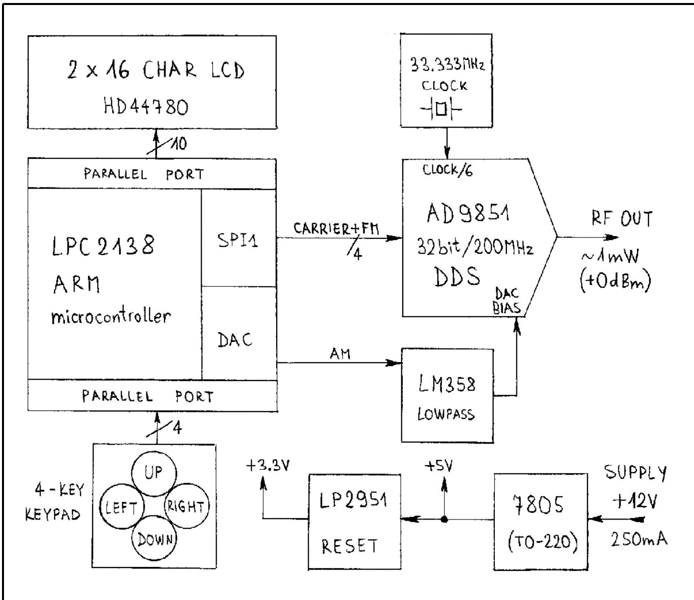


Fig 3: Block diagram of the signal generator.

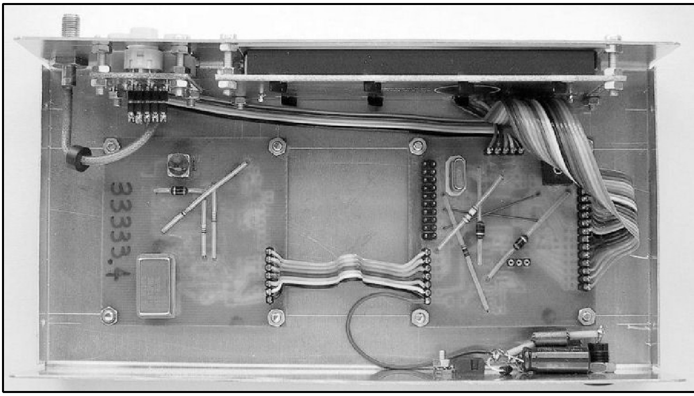


Fig 4: Picture of the two PCBs mounted in the aluminium enclosure.

ing products from the LPC2138 (modulation) DAC in AM mode.

No filtering is applied to the RF output so that all available spectral lines can be used. In order to measure communication radio receivers, an additional external adjustable attenuator of up to 120dB has to be added.

The complete DDS RF signal generator is built on two single-sided printed circuit boards, installed in a simple aluminium enclosure as shown in Figs 4 and 5.

The latter provides enough shielding for

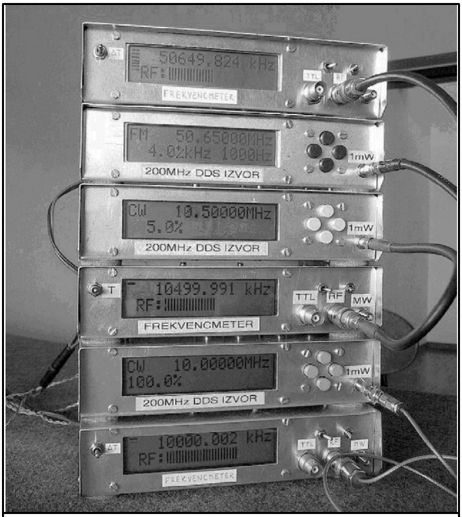


Fig 5: Three signal generator prototypes under test.

receiver measurements beyond -120dBm (0.2µV) using external attenuators. The power drain of the complete instrument is about 250mA at 12V.

The performance of the instrument is mainly limited by the AD9851 DDS chip. The latter is only specified for operation up to 180MHz, although reliable operation was obtained beyond 200MHz in all three prototypes built:

The resolution of the AD9851 DAC is only M=10bits, limiting the output signal-to-interference ratio to about 60dB. Therefore the described DDS RF signal generator has some limitations. For example its output signal is not clean enough to be used for adjacent channel interference measurements in communication receivers.

2. Design

The design is split into two main modules, each built on its own printed circuit board. The CPU board carries the LPC2138 ARM microcontroller and all associated circuits shown in Fig 6.

The LPC2138 ARM microcontroller offers comfortable 32-bit arithmetic and

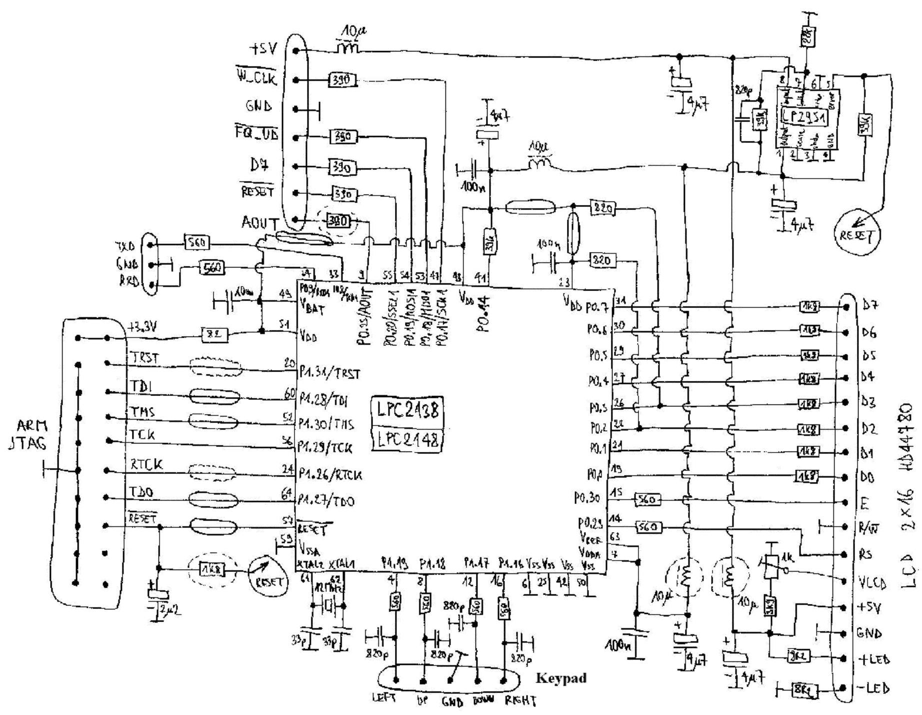


Fig 6: Circuit diagram of the CPU board.

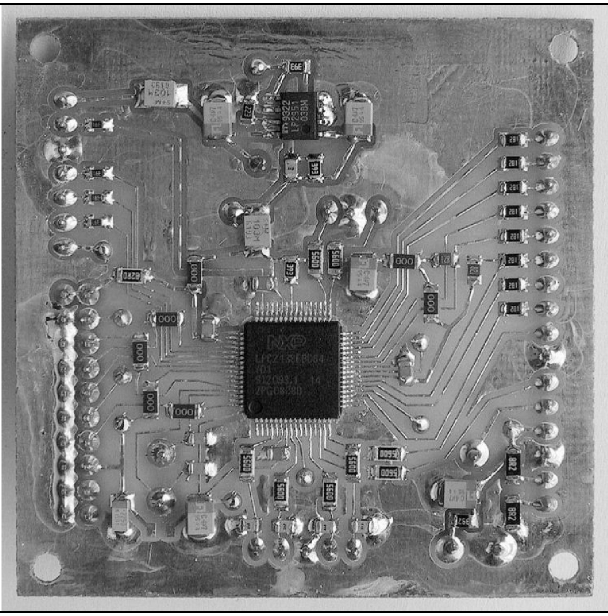


Fig 7: Photograph of the assembled CPU board.

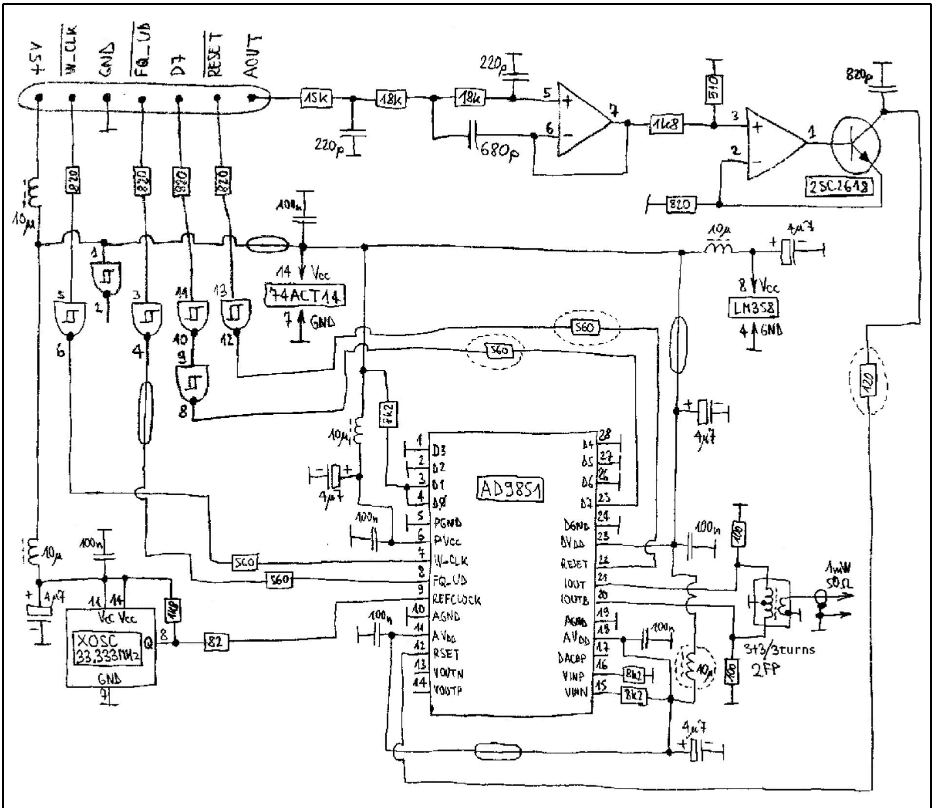


Fig 8: Circuit diagram of the DDS board.

fast instruction execution. Most instructions only take one clock cycle or less than 17ns at a 60MHz CPU clock frequency. This represents a significant advantage over obsolete 8-bit microcontrollers that are simply too slow for this application.

The CPU module includes a 12MHz clock crystal, multiplied internally to 60MHz inside the LPC2138, reset and power supply circuits. The module has several connectors: for the keypad (four simple switches to ground), for the LCD module, for the interface to the DDS module and power supply, for the JTAG interface used to program the microcontroller and a (spare) UART1 port. Most of the components are SMD and are installed on the solder side of the single sided PCB as shown in Fig 7.

A LP2951 regulator provides both the +3.3V supply voltage and the RESET signal for the LPC2138 microcontroller. The HD44780 controller built inside the LCD module accepts 3.3V logic signals although its operating voltage is +5V. Interference suppression resistors are used on all signal lines. Some pins of the LPC2138 require external pull-ups: P0.2, P0.3 and P0.14. Some JTAG programmers may require an additional pull-down resistor on the JTAG signal RTCK to enable the JTAG circuits inside the LPC2138.

The same circuit and software will also operate with the LPC2148 microcontrol-

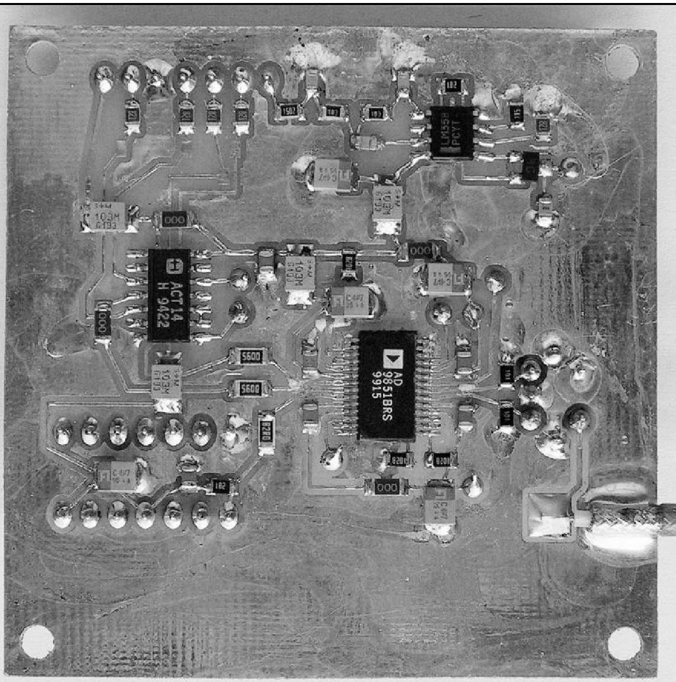


Fig 9: Photograph of the assembled DDS board.

ler. The LPC2148 has an additional USB port. The corresponding pins are left unused in this circuit to allow further expansions.

Besides the DDS chip AD9851 the second module includes a 33.333MHz clock oscillator, a lowpass filter (LM358) for amplitude modulation, a logic level conversion stage (74ACT14) for the serial data and comprehensive supply-filtering networks, the circuit diagram is shown in Fig 8

The clock frequency is multiplied by 6 to 200MHz by an internal PLL inside the AD9851. Other clock frequencies may also be used within the operating range of the AD9851. The software provides a simple menu to set the desired clock frequency and adjust accordingly all computations.

The AD9851 requires CMOS signal levels at all inputs. Since it is supplied at +5V in this project, a logic level conversion is required for the 3.3V signals coming from the LPC2138 ARM microcontroller. A 74ACT14 is used for this purpose. In addition the built in hysteresis inside the 74ACT14 cleans the pulses travelling along wires from the CPU module.

The first section of the LM358 performs as an anti-aliasing lowpass in AM modes. The second section of the same IC and the following 2SC2618 NPN transistor operate as a current source for the AD9851 DAC bias circuit. The 820pF capacitor on the output of the current source prevents instabilities and oscillations of the bias regulator inside the AD9851.

The DDS module includes a single connector for the interface to the CPU module and power supply. Most of the components are SMD and are installed on the solder side of the single sided PCB shown in Fig 9.

In order to obtain the highest signal-to-

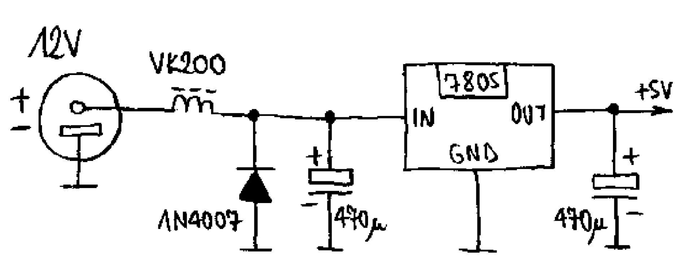


Fig 10: Circuit diagram of the power supply.

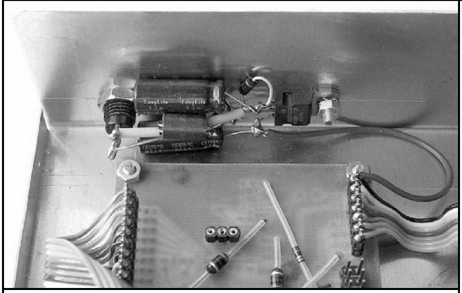


Fig 11: Power supply mounted on rear panel.

interference ratio, both true and complementary outputs of the AD9851 DAC are fed to a balun transformer. The RF output cable and in particular its braid is soldered directly to the PCB to avoid shielding problems!

Both generator modules require a +5V power supply. A 7805 regulator is a simple and efficient solution. Some additional components are required for interference suppression, the circuit diagram is shown in Fig 10. The 7805 regulator is bolted directly to the rear panel for heat sinking as shown in Fig 11.

Some older LCD modules may provide poor contrast or no display at all at a supply voltage of +5V. In this case a simple voltage doubler is required for the Vo or Vlcd voltage. The latter is referenced to the +5V supply and some modules may require it negative with respect to ground.

A simple low current voltage doubler may be built with the 7660 chip. A similar circuit can also be built with the

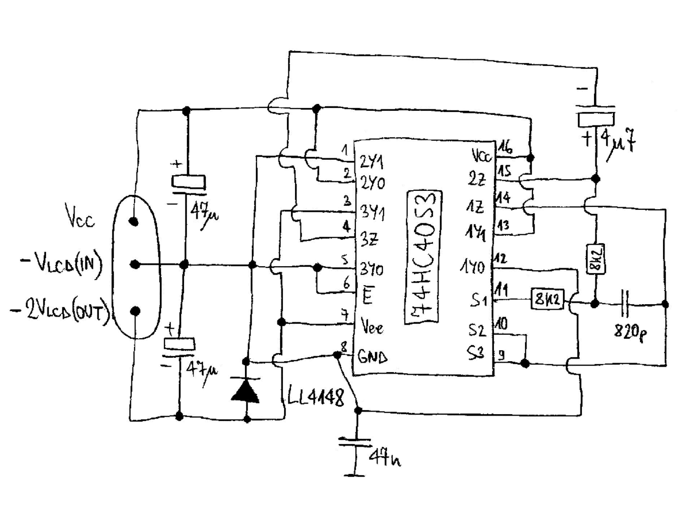


Fig 12: Circuit diagram of the low current voltage doubler.

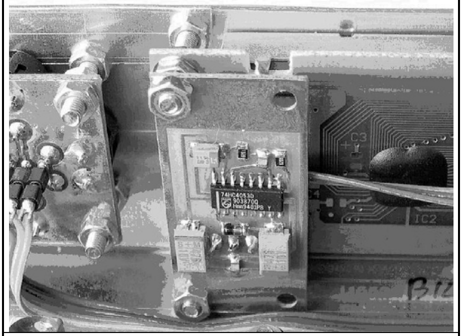


Fig 13: Voltage doubler mounted on the LCD module.

more popular 74HC4053 chip as shown in Fig 12.

The 74HC4053 requires just a few capacitors and resistors to operate as a low current (1mA) negative voltage doubler. The first switch of the 74HC4053 is used as a non-inverting amplifier for the oscillator operating at around 70kHz. The second switch acts both as an inverting amplifier for the oscillator and as an output driver. Finally the third switch operates as a synchronous rectifier for the negative output voltage. The voltage doubler is built on a small PCB bolted directly to the LCD module as shown in Fig 13.

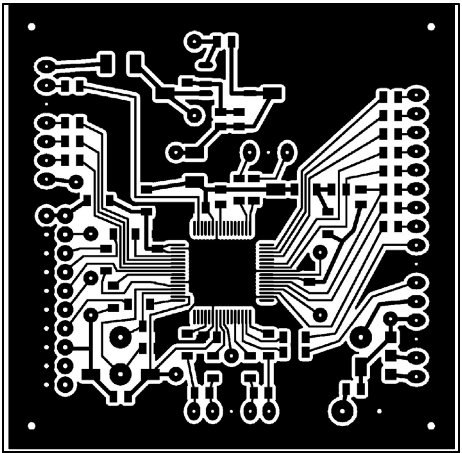


Fig 15: PCB layout for the CPU board.

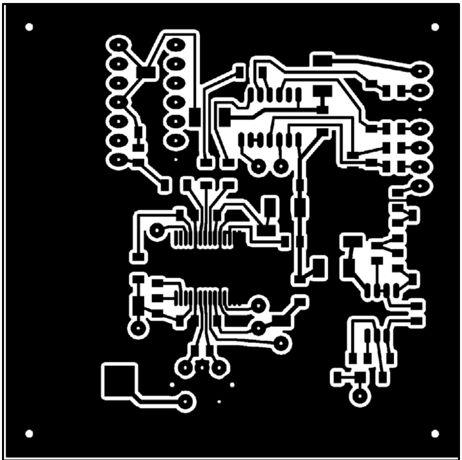


Fig 14: PCB layout for the DDS board.

3. Assembly

The DDS RF signal generator includes four single sided printed circuit boards: two larger 60 x 60mm boards for the CPU and DDS modules, a smaller 30 x 30mm board for the keypad and another smaller 43 x 20mm board for the LCD voltage doubler, if required. All four printed circuit boards are etched on an 1mm FR4 laminate. The recommended copper thickness is 17.5µm to avoid excessive under-etching. The PCBs are shown in Figs 14 – 17, the full scale

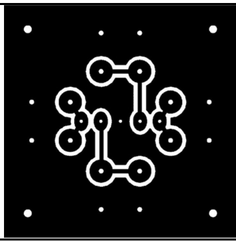


Fig 16: PCB layout for the keypad.

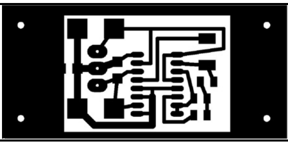


Fig 17: PCB layout for the voltage doubler.

tioned, additional flux added and the solder pads heated with the soldering iron. Do not add any solder at this time, just let the surface tension of the melted solder to do its work. As flux I am using stearine wax, available in the form of white powder. It

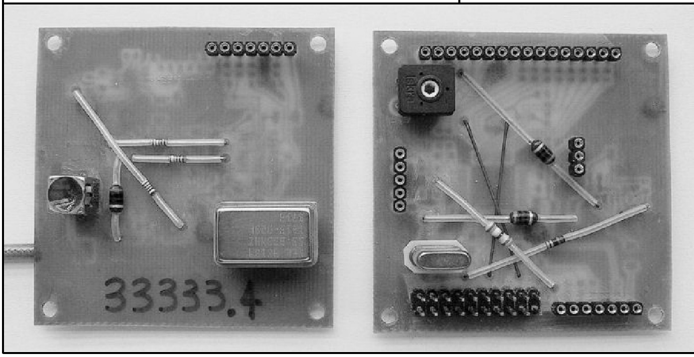


Fig 18: Photograph of the components mounted on the top of the PCBs.

drawings can be downloaded from [1].

There are just a few conventional components installed on the topside of both larger boards. These include connectors, a crystal, a trimmer, an oscillator, a balun transformer and some resistors and inductors used as bridges as shown in Fig 18.

Notes on assembly:

- **Soldering of narrow pitch SMDs,** in particular the LPC2138 (0.5mm) and the AD9851 (0.65mm). The PCB should be pre-tinned using lots of flux. Then the IC has to be positioned, … is recommended to check your work under a microscope and maybe add just trace amounts of solder by wetting the tip of the soldering iron. Any larger amount of solder will cause shorts between adjacent pins. After the soldering is complete, remove all flux. Warm, melted stearine wax can be removed with a piece of cloth. Any remaining traces of stearine wax can be removed with acetone.

- **Programming of 32-bit microcontrollers.** This is somewhat different from the dark ages of 8-bit micros. Commercial programming tools are

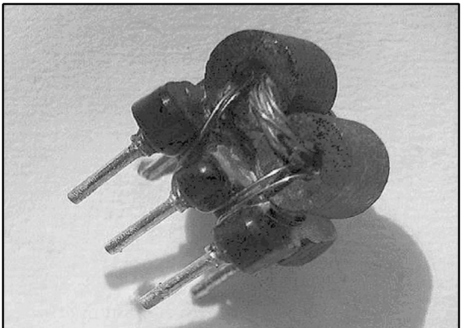


Fig 19: Picture of the balun.

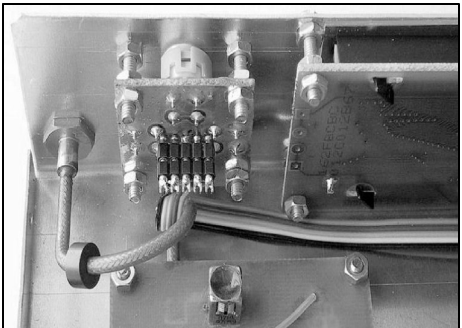


Fig 20: Picture of balun and output connection.

Fig 21: The display after power-up.

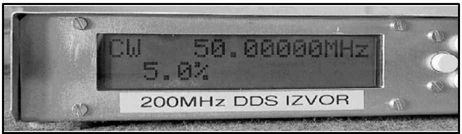


Fig 22: The display in CW mode.

very complicated and extremely expensive. Since there are no sockets for narrow pitch LQFPs, the micro has to be soldered into the target PCB and programmed there. All 32-bit micros are equipped with a JTAG port.

There is an excellent, free programming tool called H-JTAG that can be downloaded free from [2]. H-JTAG works through the parallel printer port of any PC and only requires a simple interface cable with a few resistors like my "Cigotag" [3].

Finally, the LPC2138 has some exceptions, its JTAG pins being shared with conventional I/O ports. In particular, the RTCK pin requires a pull-down resistor to switch the pins into JTAG mode. This resistor is provided in the "Cigotag", but may not be available in some expensive commercial USB programmers like "Keil". In the latter case just solder a 2.7kΩ resistor from RTCK to ground on the target micro board. I do not recommend using complicated commercial LPT port programmers like "Wiggler" or "Olimex", since the latter contain useless TTL buffers that may cause programming errors!

The balun transformer requires a careful selection of the magnetic core first. A simple selection rule is to use high permeability ferrite beads and select those that provide the highest impedance in the frequency range of interest. The transformer includes 3 + 3 turns for the primary winding and 3 turns for the secondary winding. The finished transformer is installed on the base of a 7 x 7mm IF transformer as shown in Fig 19.

Finally the whole transformer with its base is inserted in the original shielding can of the IF transformer and the latter is filled with glue. The prototypes exhibited a flat response from 200kHz to 200MHz. The finished transformer, RF output cable and connector and keypad are shown in Fig 20.

In order to improve shielding, a double braid cable is recommended together with an appropriate female SMA connector on the front panel. The ferrite bead on the cable also helps shielding. The keypad includes just four normally open pushbuttons and a connector.

The generator is installed in a box made of aluminium sheet. The bottom is made from 1mm thick aluminium sheet, the cover is made from 0.6mm thick aluminium sheet and the LCD is protected by a small piece of plexiglass. The useful internal width is 200mm, depth 100mm and height 45mm. The RF connector, keypad and LCD module are installed on the front panel. The power supply connector is installed on the rear panel.

Fig 23: The display in FM mode.

Fig 24: The display in AM mode.

Fig 25: The display in VOR mode.

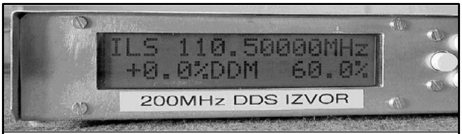


Fig 26: The display in ILS mode.

4.

Operation

Immediately after power-up, the generator displays the software version/date for about one second as shown in Fig 21.

About one second after power-up the display content is replaced by the mode and frequency in the top row and some more parameters in the bottom row. A blinking cursor shows the digit controlled by the UP and DOWN keys. LEFT and RIGTH keys select the digit. If there is no blinking cursor, then the UP and DOWN keys select the operating mode.

The frequency can be adjusted in 10Hz steps from zero up to 999.99999MHz. The software makes no reference to the actual DDS clock frequency at this stage. Therefore the frequency display just means that there will be some spectral component at the indicated value, but it does not tell whether it is the fundamental DDS product or a higher order image.

In the CW mode the generator produces an unmodulated carrier. Besides the carrier frequency its relative amplitude can be adjusted from zero up to 200.0% of the nominal output level as shown in Fig 22.

In the FM mode the generator produces a signal that is frequency modulated by an audio sine wave. The RF-signal level is set to the reference level (about 1mW at low frequencies). The peak deviation can be adjusted from zero to 999.9kHz and the audio modulation frequency can be adjusted from zero to 9999Hz in 1Hz steps as shown in Fig 23.

In the AM mode the generator produces a signal that is amplitude modulated by an audio sine wave. The carrier level is set to the reference level (about 1mW at low frequencies). The amplitude modulation depth can be adjusted from zero to 100.0% and the audio modulation frequency can be adjusted from zero to 9999Hz in 1Hz steps as shown in Fig 24.

In the VOR mode the generator produces a signal that is amplitude modulated by a standard VOR aviation-beacon signal. The latter includes the 30Hz bearing signal and the 9960Hz audio subcarrier that is frequency modulated by the 30Hz reference with a deviation of +/-480Hz. The carrier level is set to the reference level (about 1mW at low frequencies). The VOR radial (phase difference between 30Hz bearing and reference) can be adjusted from zero to 359.9 degrees and the amplitude modulation depth (sum of 30Hz+9960Hz) can be adjusted from zero to 100.0% as shown in Fig 25.

In the ILS mode the generator produces a signal that is amplitude modulated by a

Fig 27: The display in REF mode.

Fig 28: The display showing EEPROM saving.

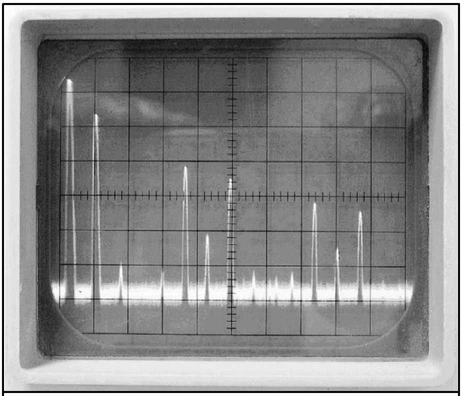


Fig 29: Typical output spectrum.

standard ILS (localiser or glideslope) aviation beacon signal. The latter includes 90Hz and 150Hz sine waves. The carrier level is set to the reference level (about 1mW at low frequencies). The ILS difference in depth of modulation (DDM) between the 90Hz and the 150Hz sine waves can be adjusted from zero to the sum of both modulation depths and the latter can be adjusted from zero to 99.9% as shown in Fig 26.

In the REF mode the DDS clock frequency can be adjusted while the generator operates in the CW mode. This allows a frequency counter or receiver to remain connected to the output and measure the frequency actually generated by the DDS. The CW level and frequency are shown in the bottom row but can not be adjusted in the menu shown in Fig 27.

Without any filtering of the RF DDS output, a conventional frequency counter will only show the correct frequency of the lowest (fundamental) spectral line of the DDS when the latter is less than about one third of the DDS clock frequency (less than about 65MHz) in CW mode. In other modes the modulation may interfere with the operation of the counter and this limit may be much lower.

All operating parameters of the DDS RF signal generator can be conveniently stored permanently into the EEPROM inside the LPC2138 microcontroller. There is just a single command to write the parameters into the EEPROM by depressing both LEFT and RIGHT pushbuttons simultaneously. The new parameters are written into the EEPROM only if a difference is found between the actual and stored parameters. In the latter case the display indicates the successful update of the EEPROM for about one second as shown in Fig 28.

All parameters are copied from EEPROM to RAM at power-up RESET. When the DDS RF signal generator is powered up for the first time, the values of all parameters are meaningless and correspond to empty EEPROM locations. Therefore the first operation is to set all parameters in all different menus and most important the DDS clock frequency in the REF menu to some meaningful values and store them in the EEPROM.

The typical output spectrum of the DDS RF signal generator set to 35MHz is shown in Fig 29 (spectrum analyser set to 50MHz/div horizontal resolution and 10dB/div vertical resolution).

The highest peak on the left is the DC component, followed by the fundamental DDS signal at 35MHz. Many higher-order images are also visible. The image at 435MHz is less than 30dB below the fundamental signal; therefore this signal generator can be used for testing receivers in the 70cm band and even beyond. Clock crosstalk from the DDS chip at 200MHz and 400MHz is also well visible. The smallest peaks are produced by second-order nonlinearities elsewhere in the analogue circuits.

Of course, the same output spectrum is obtained if the signal generator is tuned to 35MHz, 165MHz, 235MHz, 365MHz, 435MHz, 565MHz, 635MHz, 765MHz, 835MHz or 965MHz at a nominal DDS clock frequency of 200MHz. In the case of a different DDS clock frequency, the software makes all necessary computations so that there appears a spectral line at the desired frequency. One just has to keep in mind that there are many more lines elsewhere in the output signal spectrum. Further, any signals close to the DDS clock or its harmonics are rather weak due to the sin(x)/x nulls.

When measuring the sensitivity of receivers and transceivers, it makes sense to keep a fixed 20dB SMA attenuator directly on the output connector at all times. This attenuator both reduces further shielding requirements and protects the DDS chip in the case of an unwanted transmitter activation. In any case it is recommended to use double braid cables like the RG223 or its teflon equivalents for all connections to and from attenuators while measuring receivers.

Additional menus could be developed to generate DTMF, CTCSS or other common signalling tones. The DDS serial programming represents a major limitation. The latter takes about 5 microseconds either by using the SPI1 interface or by direct software implementation in the LPC2138. Therefore the DSP interrupt rate cannot be much higher than 100kHz in FM mode.

Recent DDS chips operate at clock frequencies up to 1GHz with much improved 14bit DACs on their outputs, producing cleaner output signals. All these chips have a similar serial interface to the CPU. Therefore the same LPC2138 microcontroller module could be used to drive new DDS chips, unfortunately including the speed limitations in FM mode.

5. References

[1] All information can be downloaded from http://www.s5tech/s53mv/dds/dds.zip

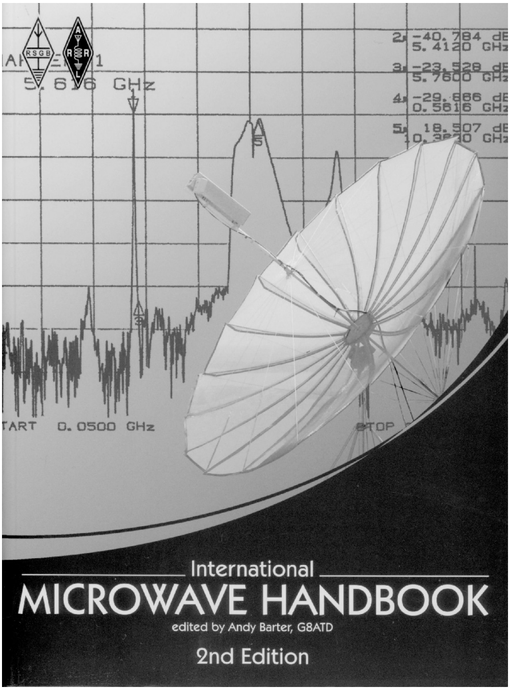
[2] http://www.hjtag.com/

[3] http://www.s5tech.net/s53mv/equipment/cigotag.html

[4] NXP (Philips) web: http://www.nxp.com/

[5] Analog Devices web: http://www.analog.com/

The International Microwave Handbook 2nd Edition



The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area, along with excellent designs from around the world to fire the imagination. Material has been drawn from many sources including the RSGB journal RadCom and the ARRL publications QST & QEX. Alongside this material a truly international range of sources have been used including items from Germany, Denmark, New Zealand, Slovenian and many more.

The earlier chapters of the book provide invaluable reference material required by all interested in this exciting area of experimentation. Techniques and devices are covered in depth, leading the reader to understand better the wide range of equipment and techniques now available to the microwave experimenter. This book contains a wide selection of designs using the latest technology that can reasonably be used by radio amateurs and ranges from ones that can be reproduced by most radio amateurs to those that require a high degree of skill to make.

With the explosion in consumer electronics using microwave frequencies the opportunity to experiment has never been greater and this book is simply the.

The handbook is available from The RSGB or ARRL

Wolfgang Schneider, DJ8ES

Universal GPS clock

Many things these days in amateur radio require the exact time, particularly modern digital modes like WSJT, for Meteorscatter, or JT65B, for EME, require a second accurate change over between transmit and receive modes. This accurate time can be made available at minimum cost using a modern GPS receiver and a simple decoding circuit based on a microcontroller.

1.
General information

An exact time is required for many of today's digital transmission modes. Meteorscatter and WSJT rely on switching from transmit to receive at the beginning of each minute or half minute depending on your location. The same applies to EME (JT65B), to say nothing of such time critical applications as Coherent CW (CCW) or the like.

Using the GPS (Global Positioning System) satellite signals from a suitable receiver and a decoder to extract the required data, the exact time and current date can be displayed as shown in Fig 1.

The correct local time (e.g. British Summer Time or Central European Summer Time) for different large cities or countries can be shown worldwide by changing the time zone. The number of satellites being received can be displayed as well as an item of special interest to radio amateurs, the QTH Locator. An example display is shown in Figs 2 and 3.

2.
What is GPS?

GPS (Global Positioning System) is a satellite based navigation system designed for the American military based on 24 individual satellites. These circle around the earth twice a day at approximately 20000km and make it possible for GPS receivers to determine their current position worldwide. Additionally further standard information is made available. Everyone can use this satellite service for private or economical use; it is free.

The accuracy of the position depends on many of factors. One is determined by the American military that reserves the right to reduce the accuracy or to close the satellites to private reception. Since the beginning of 2000 the accuracy has not been reduced to a deviation of 95% of the measured values e.g. 7m radius. Therefore it is possible for a GPS receiver that receives the signals of several of these satellites at the same time to

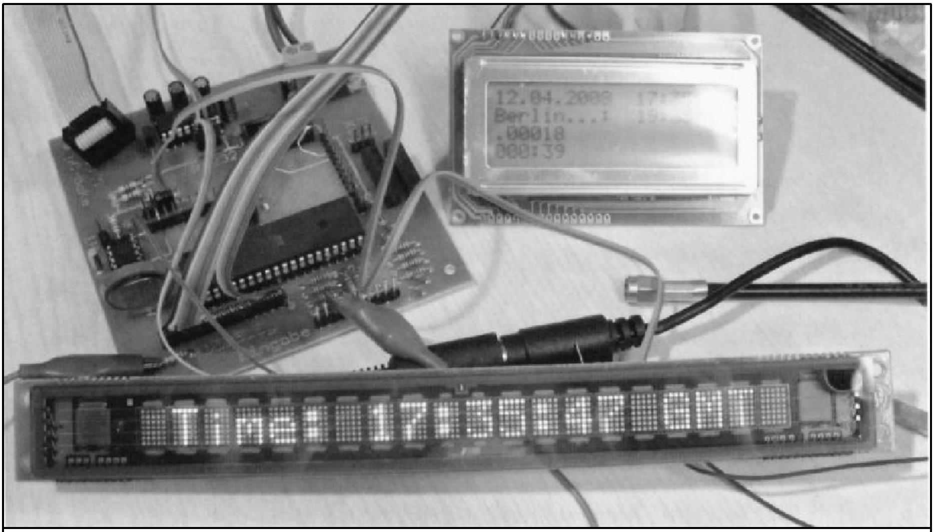


Fig 1: The universal GPS clock with LCD display and Vacuum Fluorescent display.

provide the current location in form of longitude, latitude and height. Also the data records contain highly exact time and date information.

GPS receivers that cost between €50 and €100 can be connected to a PC or laptop and with suitable software can display the information from the GPS system. The receiver and display device are normally separated, with the receiver close to the antenna so that it has a view of as much of the sky as possible. Even small impairments to the free view of the satellites by trees or houses can affect the reception. That is because of the relatively high frequency range of the satellites at 1.575GHz and their relatively small transmitting power.

Fig 2: The display showing the time, date, number of satellites and locator.

Fig 3: Display showing date, time, city and local time.

The data from a GPS receiver is generally transmitted over a serial link at 4800 Baud at regular intervals usually of one second.

The protocol for data records were standardised in 1980 by a consortium of firms in NMEA (National Marine Electronics Association). Since then most GPS receivers use this protocol. The data is transmitted in the ASCII format with all printable characters as well as CR (Carriage Return) and LF (Line Feed). The data is transmitted in the form of sentences. Each of these sentences begins with the identifier "$", two characters for transmitter identification (GP) and three characters for the type of sentence. Sentence identification (e.g. RMC or GGA) is followed by a set of data that is separated with commas. Finally the sentence has an optional check sum and terminated with a CR/LF.

Each sentence can contain up to 82 characters including of the "$" and the CR/LF. If a data field in a sentence is not available then it is omitted simply by using the commas for separation of a blank field. The receiver can simply count the commas to assign the data correctly.

The following two data records in NMEA format show examples of GPS data:

The RMC data record contains:

$GPRMC, 191410, A, 4735,5634, N, 00739.3538, E, 0,0, 0,0, 181102.0,4, E, A*19

- Time in UTC
- Receiver data of ok or warning
- Degree of latitude N (northern latitude), S (southern latitude)
- Degree of longitude E (eastern length), W (western length)
- Speed
- Course
- Date….

The GGA data record contains:

$GPGGA, 191410.4735,5634, N, 00739.3538, E, 1.04.4,4, 351,5, M, 48,0, M, *45

- Time in UTC
- Degree of latitude N (northern latitude), S (southern latitude)
- Degree of longitude E (eastern length), W (western length)

Fig 4: The HOLUX GR-213 GPS receiver.

- Quality of the measurement
- Number of satellites received

3.

The GPS receiver GR-213

The HOLUX GR-213 GPS receiver (Fig 4) uses the most up to date technology and can evaluate even weak GPS signals correctly. It is currently available for €60.

There is an orange LED status indicator on the GPS receiver that indicates:

LED on = searching for GPS signal

LED flashes = signal recognised

A mini DIN socket is used to connect the receiver to a suitable decoder, the pin connections are shown in Fig. 6. As well as the serial data the power supply (5V/80mA) is connected is this socket. Either TTL level or standard RS232 (±15V) signals can be selected. The data transmission rate is preset to 4800Bit/s in the standard format 8-bit, No parity and a stop bit. The factory installed data records are for the GGA, RMC and GSV NMEA protocol valid for GPS.

After first switching on (cold switch on) it takes approximately 40 seconds for the data plus date and time to be sent in the first 1-second pulse. This time reduces to less than one second for following activations (warm start). There is further information on the manufacturers homepage [2].

4.

Circuit description of the GPS clock

The main item of the GPS clock circuit is the ATMEL ATMega32 (IC1) microcontroller (Fig 5). This is in a 40 pin DIL package offering plenty of inputs and outputs for connection to different peripherals. There is plenty of room for the system software in the 32k ROM. Not much of this was used for the GPS clock, at least in the present state of development.

The unusual clock frequency of 9.216MHz was selected to generate the serial interface clock rates, the more usual clock rates are e.g. 8, 12 or 16MHz. The actual clock speed is of little impor-

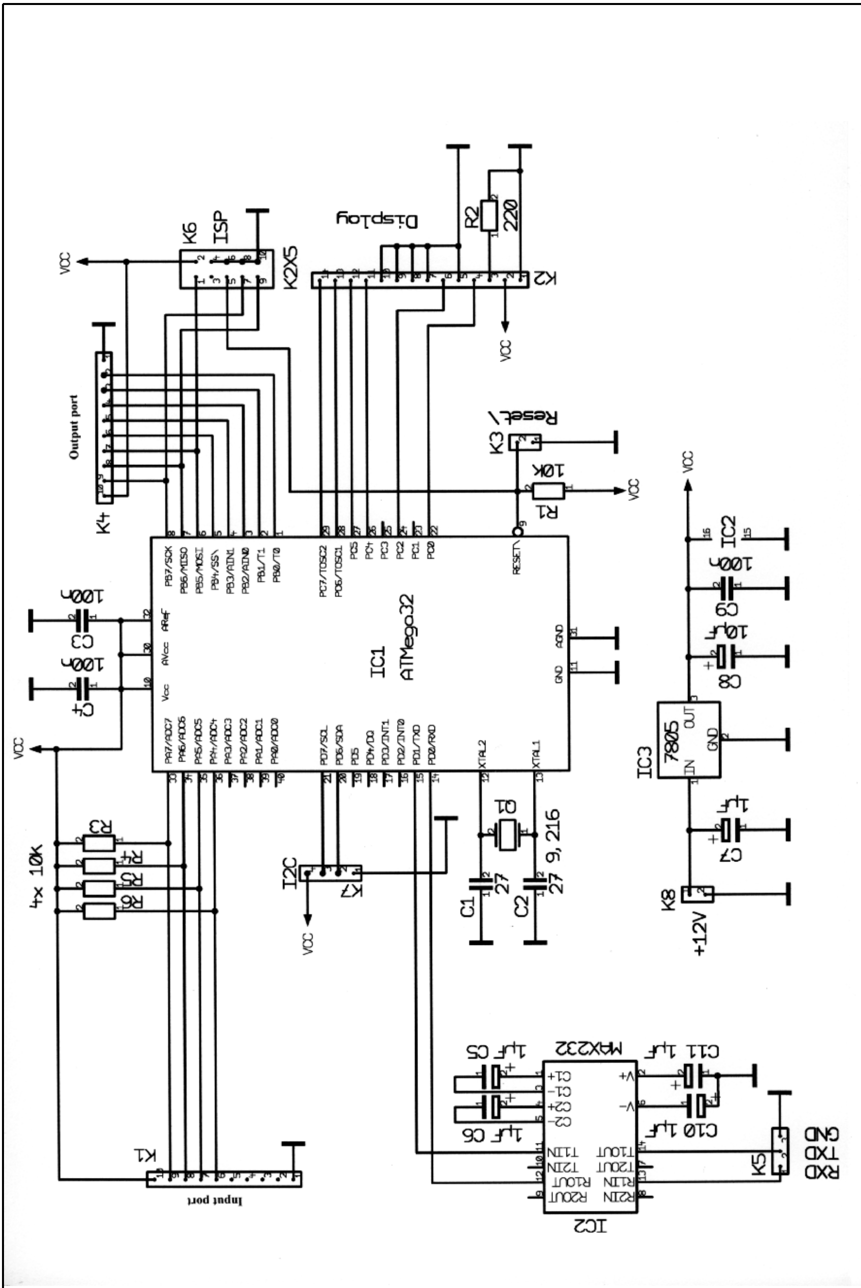
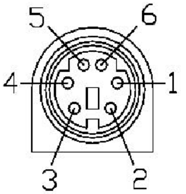


Fig 5: Circuit diagram of the universal GPS clock.



Pin	Signal and description
1	TX_RS232 (RS232 level)
2	+5 VDC
3	TX_TTL (CMOS TTL level)
4	Ground
5	RX_TTL (CMOS TTL level)or DGPS input(RS232 level)
6	RX_RS232 (RS232 level)

Fig 6: Pin connections for the mini DIN connector on the GPS receiver.

tance for the GPS clock application. A MAX232A (IC2) is used read the data from the serial interface because it has the required level change required for standard connections. The serial port is connected to the contact strip K6.

Four input ports (PA4 – PA7) are used for the GPS clock:

PA7: Time mode/position display

PA6: Summer time/winter time

PA5: Time zone +1

PA4: Time zone -1

The current data (date, time, Locator, number of satellites) is displayed on 2 lines by 20-character LCD display. This display is connected by the patch cord K2 (14 pin) directly to the microcontroller port C.

As shown in Fig 1 it is possible to use a second large external Vacuum Fluorescent (VF). This was connected using K4 (4 pin) used for to control a 7-segment display over a I²C bus. This requires amendments to the software.

5.

Program description

The software for the GPS clock is programmed in BASCOM (version 1.11.8) on an ATMega32. The microcontrller is clocked at 9.216MHz. This unusual frequency is an advantage for the accuracy

Table 1: Parts list for the universal GPS clock.

IC1	Microcontroller ATMega32, ATMEL
IC2	MAX232A, level translator
IC3	7805, 5V voltage regulators
Q1	Crystal 9.216MHz
Resistors:	
R2	220Ω, 1/4W, RM 10mm
R1, R3	10kΩ , 1/4W, RM 10mm
R4 - R6	10kΩ, 1/4W, RM 10mm
Ceramic capacitors, RM 2.5mm:	
C1, C2	27pF
C3, C4, C9	100nF
Electrolytic capacitors 25V, RM 2.54mm:	
C5 - C7	1µF
C10, C11	1µF
Connector pins, single-row, straight, RM 2.54mm:	
K1, K4	10-pin
K2	14-pin
K3, K8	2-pin
K5	3-pin
K6	5-pin
K7	4-pin
1	PCB DJ8ES 076 (µC beacon)

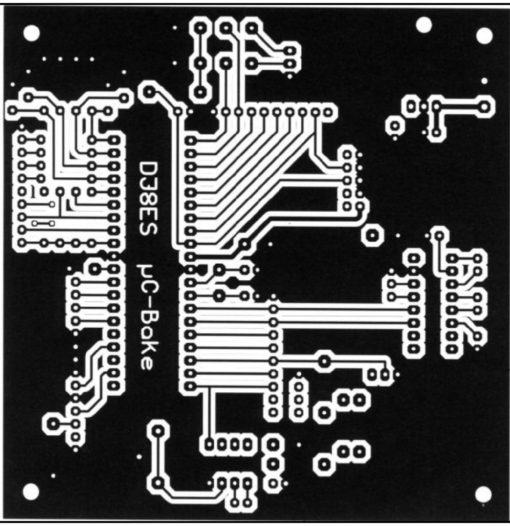


Fig 7: PCB layout for the universal GPS clock.

of the serial interface.

In a first step the interfaces are defined: Port A for input of the control information, port C for the LCD display. Port B is not needed in the present program version. The serial port (RS232) and optionally the I²C bus are on the appropriate connector pins of port D (see Fig 5).

After the initialisation phase the GPS receiver is configured. The only data records that are required are RMC and GGA. Other records contain no relevant information.

The GPS receiver supplies all data in serial form over the RS232 interface, this transmission is interrupt controlled.

There are two display modes, the current times in UTC and local time (e.g. MEZ) with associated time zone (e.g. Berlin for Germany) or UTC and the Locators with number of up-to-date satellites received.

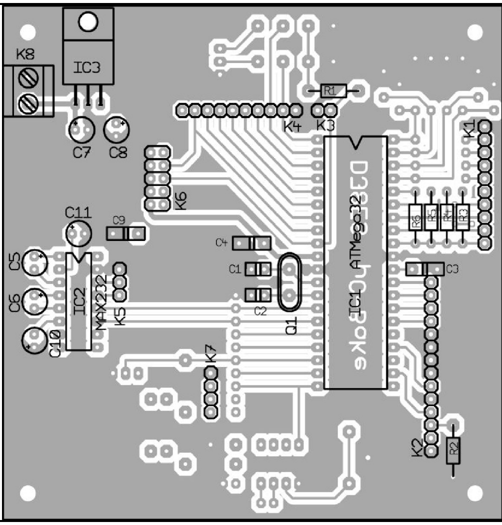


Fig 8: Component layout for the universal GPS clock.

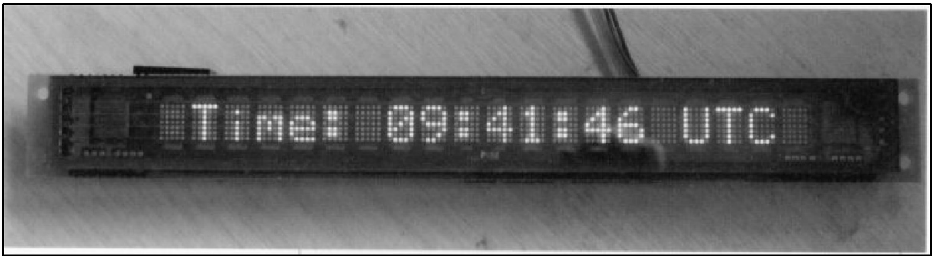


Fig 9: The optional vacuum fluorescent display.

These are selected with a 2-pole rocker switch. There is also a selection between display of the summer or winter time and the possibility for the choice of the individual time zone (12 zones).

The display can optionally be on a VF display or a wide 7-Segment display using the I^{2}C bus. This option is already included in the software. Data formats and contents require individual programming.

The GPS receiver always sends data at the beginning a minute. Because of the time for data communication and the time for processing in the microcontroller the display is approximately 300ms delayed. This should not represent a problem. Display using the exact 1-second pulse can give a precise display with appropriate software modification.

6.

Construction

The circuit for the GPS clock is built on a double-sided epoxy PCB, DJ8ES 076, the original the PCB for the beacon controller, (Fig. 7). The 100mm x 100mm PCB is plated through in several places. Only very few components are fitted, no particular problems should be experienced.

The assembly takes place in an informal order as shown in Fig. 8. Only the ATmega32 (IC1) microcontroller, the MAX 232 (IC2) RS232 interface and the LCD display should be added after testing the +5V supply at the output of the 7805 (IC3) voltage regulator. Therefore sockets are useful for assembling of the ICs.

7.

Start-up

The GPS receiver must be switch on for start-up because the software examines the data stream for validity and only leaves the wait loop when this is true. The VF display used in the prototype is optional.

Note: After first switching on (cold start) it takes approximately 40 seconds for the data to be sent for the first time.

The individual components in the overall system are supplied with a 5V supply voltage. Approximately 80mA are required for the microcontroller, including a 2 line LCD display. Approximately 25mA are required for the GPS receiver (without lighting) and the optional VF display requires approximately 500mA.

The GPS receiver is supplied over the RS232 cable, it can use TTL signal levels but standard RS232 signal levels are less prone to interference over long connecting cables.

Fig 10: The completed universal GPS clock.

There are GPS receivers with an additional PPS (Pulse Per Second) signal. This precision second signal (clock) can be used for synchronization e.g. frequency stable oscillators, counters, etc. The GR-213 GPS receiver used here does not offer this possibility.

8. Optional display units

When building the prototype it was realised that a large VF display was possible. The FC20X1SA from NEC has such a display with 20 characters in a single row. These displays are used in such things as POS (Point Of Sale) terminals so they are available fairly cheaply. The connections are made using a serial bus with TTL or RS 232 levels They have good readability even in daylight over a distance of several metres. Unfortunately they consume relative high power of 500mA at 5V. This must be considered when designing the power supply.

Alternatively a solution using large 7-segment displays is possible, however this requires a lot of wiring. A design using 56.9mm high displays controlled over an I^2C bus with 2 x SAA1064 drivers was considered but the cost of the printed circuit board meant that it was not manufactured. An interface for the microcontroller board was considered.

9. References

[1] Klaus H. Hirschelmann, DJ7OO: Simple GPS decoders with LCD displays, http://www.kh-gps.de/

[2] Holux: data sheet for the GPS receiver GR-213, http://www.holux-gps-vertrieb.de/

[3] Claus Kühnel: Programming the AVR RISC microcontrollers with BASCOM AVR Edition 2, ISBN 3-907857-04-6, http://www.ckuehnel.ch/

[4] Mailte Pöggel: Control of Vacuum Fluorescent display with the PC, http://www.maltepoeggel.de/

[5] NEC Electron DEVICE: Data sheet FC20X1SA, http://www.maltepoeggel.de/

[6] Matjaz Vidmar, S53MV: A DIY receiver for GPS and GLONASS satellites part 1, VHF Communications Magazine 1/1994 pp 35 – 44, A DIY receiver for GPS and GLONASS satellites part 2, VHF Communications Magazine 2/1994 pp 66 – 77, A DIY receiver for GPS and GLONASS satellites part 3, VHF Communications Magazine 1/1994 pp 151 – 165, A DIY receiver for GPS and GLONASS satellites part 3 Quadrifilar backfire helix antennas, VHF Communications Magazine 4/1994 pp 197 – 200

Aristoteles Tsiamitros, DD5FT

Example active low pass filters

This article describes an example of active low pass filters with an order of n > 2. These use of cascade filter technology i.e. the filter constructed from partial filters in series. First and second order filters were designed. The partial filters use operational amplifiers. The mathematical formulae to determine the transfer functions of the partial filters are simplified using the program SCILAB [3]. The filters designed are examined with the LTspice simulator [4] using models of the operation amplifiers.

1. Introduction

There are many different ways to design active filters; the cascade technique is the most popular. The reasons that this technique is popular are: the small number of operation amplifiers required, the relatively low sensitivity to component tolerances and the simple alignment of the individual filter stages.

It is based on the fact that the transfer function of a filter can be split into a product of first and second order sub-functions. These sub-functions are then realised by partial filters. The product of the sub-functions is the series connection of the partial filters. The filter circuits must be decoupled from each other; this is achieved with the high input and low output resistance of the operation amplifiers.

The circuit components can be calculated by comparing the general form of the sub-functions with the split of the transfer function and the circuit specific transfer functions.

First the split of the transfer function of the filter is defined. In the next section an example of a Tschebyscheff filter using the cascade technique is described step by step. All other low pass filters can be designed in the same way.

Circuits for first and second order of partial filters are shown in the appendix.

2. Splitting the transfer function

The transfer function of a low pass filter of nth order is given by the following equation where s = σ + jω a complex frequency.

$$H(s) = \frac{U_{OUT}(s)}{U_{IN}(s)} = K \cdot \frac{b_0}{s^n + b_{n-1}s^{n-1} + b_{n-2}s^{n-2} + \ldots b_1 s + b_0}$$

[1]

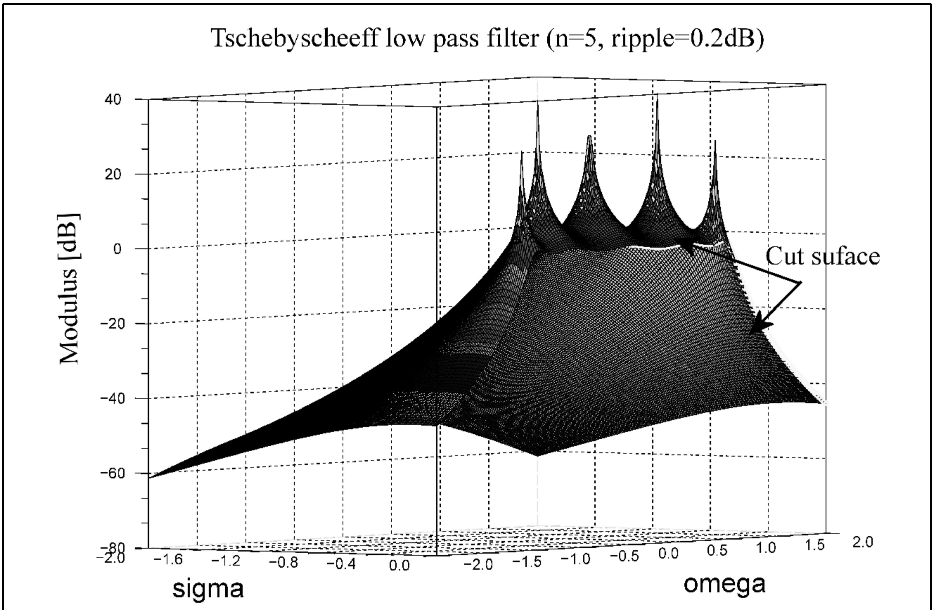


Fig 1: The surface H(s) for a 5th order Tschebyscheff filter.

In a complex number σ is for the real part and ω is for the imaginary part, a value of s can be assigned to each point of the curve H(s) and represents a surface over the complex numbers. The surface H(s) for a 5th order Tschebyscheff filter is shown in Fig 1. There are five salient points in this representation, where the transfer function goes approximately to infinitely. Those are the places, where the denominator of H(s) becomes zero; these are the poles of the transfer function. The position of these points in the complex plane is shown in the zero pole diagram Fig 2. The position of the poles determines the ripples in the pass band, the

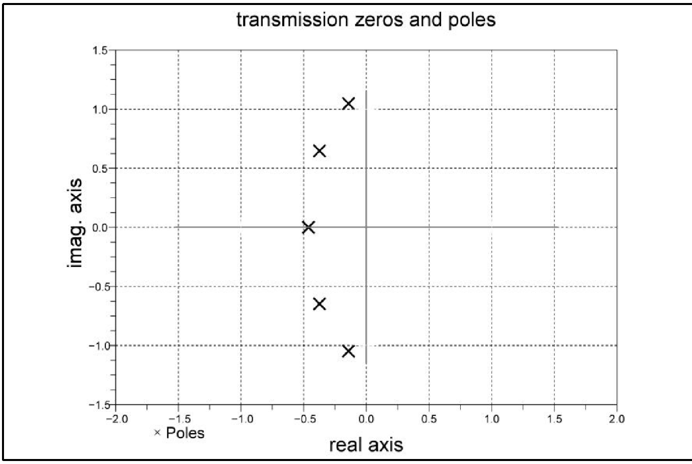


Fig 2: Positions of the poles for a Tschebyscheff low pass filter (n = 5 ad - 0.2dB)

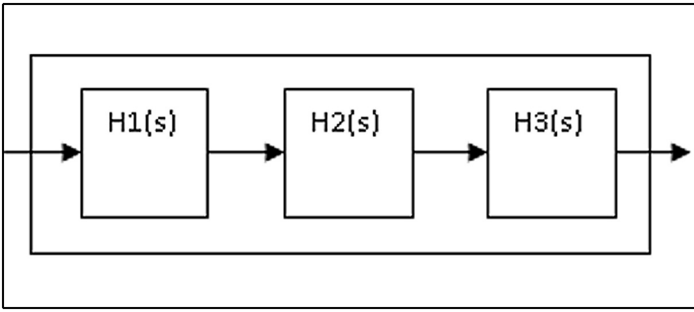


Fig 3: Block diagram of cascade filters for a 5 pole low pass filter. The transfer function is $H(s) = H_1(s) * H_2(s) * H_3(s)$.

selectivity and the transient behaviour of the filters.

If the surface H(s) in Fig 1 is cut along the omega plane, the cut curve gives H(j), the modulus of the transfer function for the case when σ = 0. The surface for this was yellow on the original diagram but because Fig 1 is a greyscale the cut surface has been indicated. It corresponds to ω ≥ 0 in the amplitude response shown in Fig 9.

The poles are the roots of the denominator polynomial. Odd order filters have a pole on the real σ axis (Fig 2), the remaining poles are complex conjugate pairs. Even order filters have exclusively complex conjugate poles.

For cascade synthesis of an active low pass filters one of the complex conjugate poles is used. For odd order filters the real pole results in a term for a first order filter. The result for odd order filters is shown is equation [2]. For even order filters the first order term is void.

$$H(s) = \frac{K_o b_0}{s+b_0} \cdot \frac{K_1 b_1}{s^2 + a_1 s + b_1} \cdot \frac{K_2 b_2}{s^2 + a_2 s + b_2} \cdots \frac{K_1 b_1}{s^2 + a_1 s + b_1}$$

[2]

The split of a fifth order low pass filter is into a single order filter $H_1(s)$ and two second order filters $H_2(s)$ and $H_3(s)$ as outlined in the block diagram Fig 3. The transfer function of the whole filter and the sub-functions are calculated in section 5 using the program SCILAB (Fig 6 and 7).

3.

Calculation of the components

After splitting the transfer function, active RC filter circuits were chosen for the first and second order partial filters. There is a large selection of second order filters, in this article a double negative feedback circuit was chosen (see appendix and [1]).

Next the specific transfer functions for the selected circuits are compared with those of the partial filter. The general transfer function for a single order filter is:

$$H(s) = K\frac{b}{s+b}$$

with s = - b, the real pole of the filter.

The example in section 5 is:

$$H(s) = \frac{0.46}{s+0.46}$$

The transfer function of an RC circuit for the partial filter is:

$$H_{RC}(s) = \frac{\frac{1}{RC}}{s + \frac{1}{RC}}$$

Comparing of the two transfer functions it follows:

$$RC = \frac{1}{0.46}$$

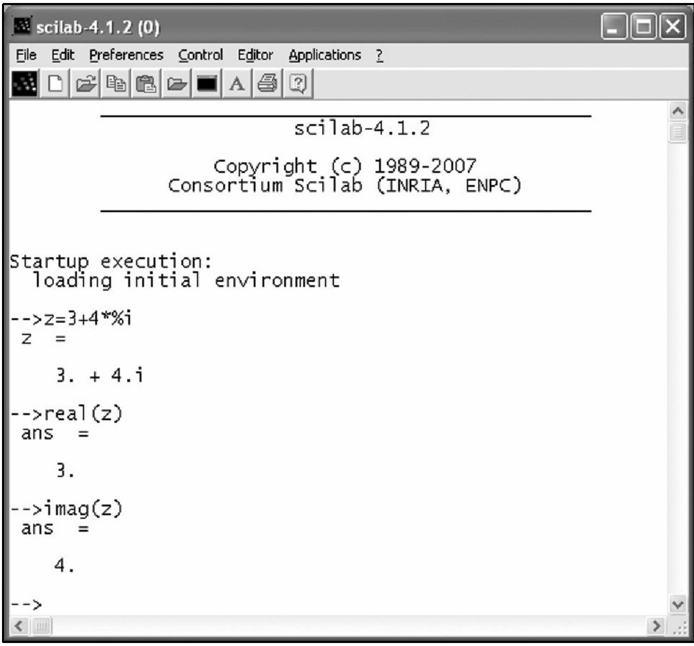


Fig 4: The opening window of the SCILAB software.

There are two unknowns (R and C) but only one can be computed so the resistor is set to a standardised resistance R = 1, therefore:

$$C = \frac{1}{0.46}$$

Finally R and C must be evaluated for the desired frequency and resistor value denormalised.

A similar approach is used for the transfer function of the second order filter. The general form is:

$$H_2(s) = K \frac{b}{s^2 + as + b}$$

The circuit specific transfer function is set up next. The circuit values depend on the factors K, A. Since insufficient design equations are available some components must be described as parameters. The remaining components are calculated by comparing of the appropriate factors.

4.

Programs used

Two programs are used for this article. The program SWCADIII [4] from Linear Technology (referred to as LTspice) is used as the last step in the development of the active filters, it concerns the frequency response of the complete filter circuit to evaluate the frequency response of the operation amplifiers used. LTspice will be familiar readers of VHF Communications Magazine from earlier publications.

A second program is needed for the filter synthesis; this is inevitably complex because it deals with the poles and zeros of the transfer function. The program SCILAB is used, it can be downloaded free of charge. There is plenty of reference material about the program and downloadable help on the homepage for SCILAB [3]. After installation a window showing the desktop for scilab-4.1.2 is

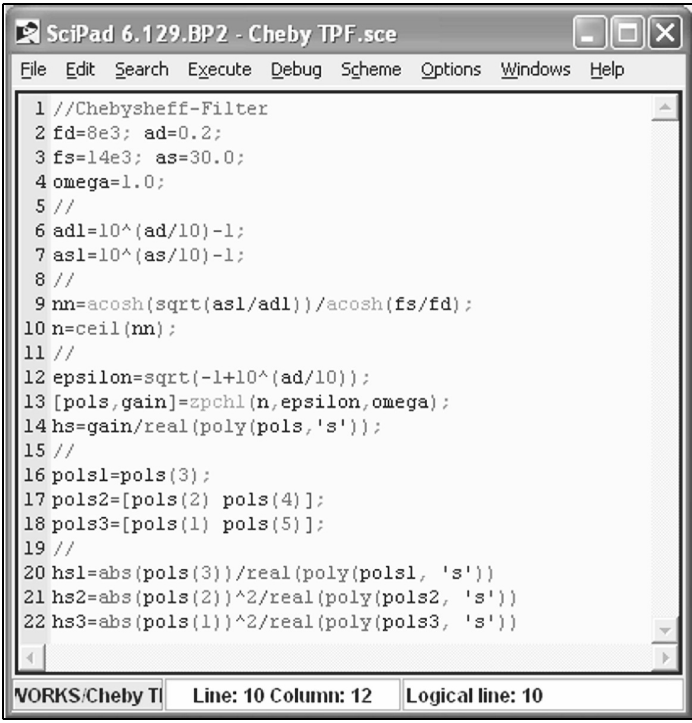


Fig 5: An SCILAB script for calculation of the pole positions and transfer function of a low pass filter.

displayed (Fig 4). Simple calculations can be carried out by direct input into the input line but it is better to prepare a script using "open Scipad". Such a script is shown in Fig 5. It contains the mathematical formulas for calculating the filter order, the poles and zeros, as well as the transfer functions of the partial filters. The calculation is started with CTRL+L.

There is plenty of help for the user on the SCILAB homepage that will quickly familiarise the user with the basic functions. In particular to the bibliography on the homepage.

5.

Example of a Tschebyscheff low pass filter

In this section the cascade synthesis of a Tschebyscheff filter is described in detail. The steps in detail are:

- Specification of the low pass filter, determination of the approximation method
- Computation of the order of the filter
- Computation of the pole positions
- Summary of the conjugate complex pole positions and if necessary the real pole position of the partial filters
- Selection of active circuits to realise of the partial filters
- Comparison of the transfer functions of the partial filters with the circuit specific transfer functions and design of the standardised componments
- De-standardisation

5.1. Specification

The starting point of the development is naturally specification of the low pass filter. First the filter characteristic (e.g. Butterworth, Tschebyscheff, Bessel, etc.) is specified. Several factors contribute to

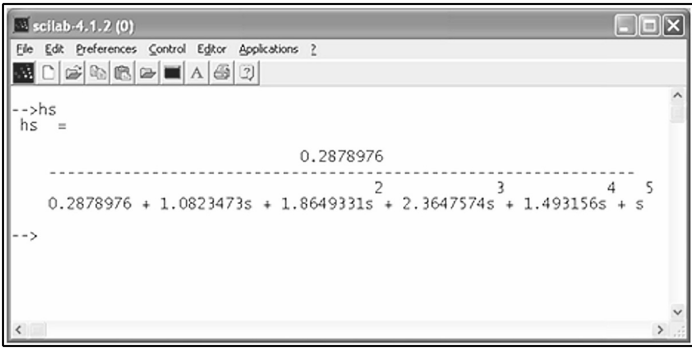


Fig 6: Transfer function for a Tschebyscheff filter with n = 5 and ad = 0.2dB.

this e.g. desired selectivity of the filter, requirement to this selection, the envelope delay, etc.

The filter in this example has the following data:

- Tschebyscheff characteristic
- Cutoff frequency f_D = 8kHz
- Maximum attenuation in the pass band a_D = 0.2dB
- Notch frequency f_s = 14kHz
- Minimum stop band attenuation a_s = 30dB
- Amplification in the pass band K = 1 (0dB)
- Reference resistance R_0 = 10kΩ

5.2. Filter order

From the specification the filter order can be determined.

$$n = \frac{a \cdot \cosh\left(\sqrt{\frac{10^{as/10}-1}{10^{ad/10}-1}}\right)}{a \cdot \cosh\left(\frac{fs}{fd}\right)}$$

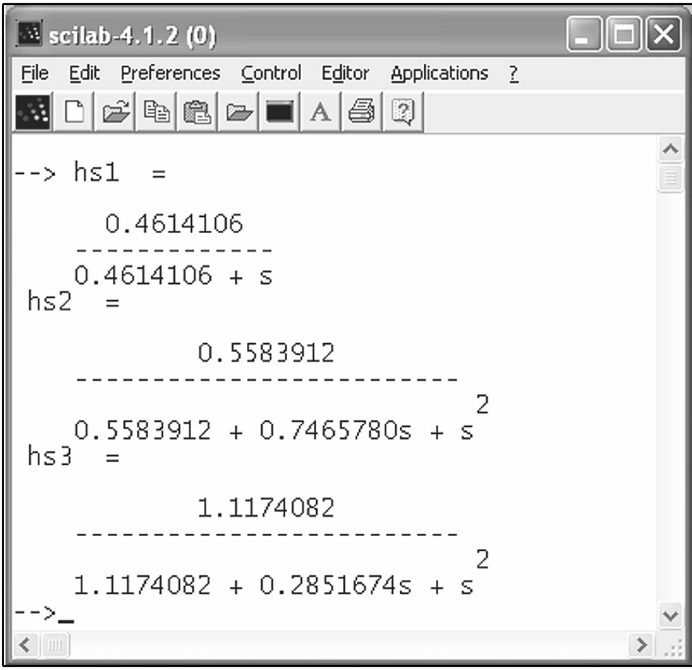


Fig 7: Transfer functions for the low pass filter.

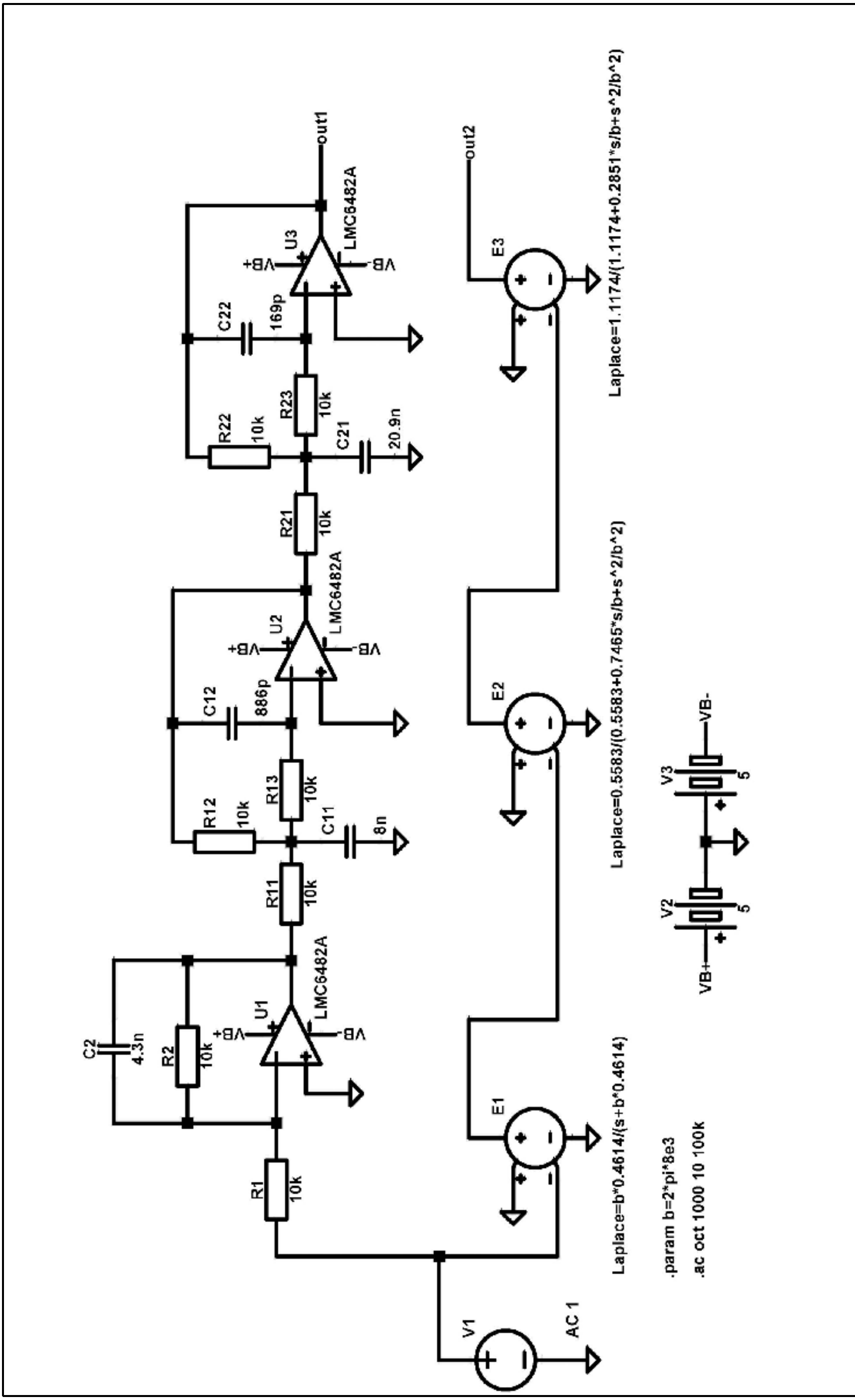


Fig 8: Circuit diagram of the low pass filter.

As mentioned the calculations are carried out using the program SCILAB. The script is shown is Fig 5. In line 9 of the script the accurate value of n = 4.89 is computed and rounded up in line 10 to the next whole number n = 5. The calculation is terminated with semicolons at the end of the lines 9 and 10. If the variable names are entered into the input line of SCILAB (after the execution of the script with CTRL+L) and pressing Enter, then SCILAB calculates the values of the variables.

5.3. Transfer functions of the partial filters

The pole positions of the Tschebyscheff partial filters are calculated in line 13 using the function "[pols, gain]=zpch1 (n, epsilon, omega)". The function needs the filter order n and the ripple:

$$\varepsilon = \sqrt{10^{aD/10} - 1}$$

as input parameters as well as the cutoff frequency . The ripple is calculated in the previous line 12. The function places the pole positions and the gain of the transfer function in the variables "pole" and "gain".

The function in the following line 14 "hs=gain/real (poly (pole, ,s`))" is the transfer function H(s) of the total filter as a polynomial as in equation [1].

The function "real" removes the imaginary part. H(s) is not need; it is only calculated for the sake of clarity (Fig 6).

The input line of SCILAB is "-->pols", to computed the pole positions. SCILAB uses the notation "2.5i" for the imaginary part and not "j2.5" as used for other mathematical representation.

The pole positions in the order they are calculated by SCILAB are:

p1 = -0.142583 + j1,047415

p2 = -0.373289 + j0,647338

p3 = -0,461410

p4 = -0.373289 - j0,647338

p5 = -0.142583 - j1,047415.

The pole position "pole (3)" is the first with an imaginary part. Ii can be seen that this has am imaginary part with a power of 10^{-17}. The function "clean (pole (3))" will remove the imaginary part.

The form of sub-functions for low pass filters is very easy to see. Conjugate complex pole positions are combined in pairs of second order functions. In the example there is p_1 and p_s, as well as p_2 and p_4. The real pole position p_3 is a transfer function for a first order.

In the lines 16, 17 and 18 of the script the pole positions are summarised for (p_1, p_s), (p_2, p_4) and (p_3). The lines 20, 21 and 22 compute the sub-functions and output them (Fig 7).

5.4. Partial circuits

After the transfer functions of the partial filters are determined the circuits to realise them are chosen. The first order filter is realised with the inverting amplifier circuit shown in Fig 11 and the second order filter by the circuit in Fig 12. The design equations are shown in the appendix.

Design of the partial circuits:

Transfer function

$$H_1(s) = \frac{0.4614}{s + 0.4614}$$

K = 1, b =0.4164, $R_1 = R_2 = 10k\Omega$

$$C_2 = \frac{1}{0.4614 \cdot 2\pi \cdot 8kHz \cdot 10k\Omega} = 4.3nF$$

Transfer function

$$H_2(s) = \frac{0.5583}{s^2 + 0.7465s + 0.5583}$$

$R_1 = R_2 = R_3 = 10k\Omega$

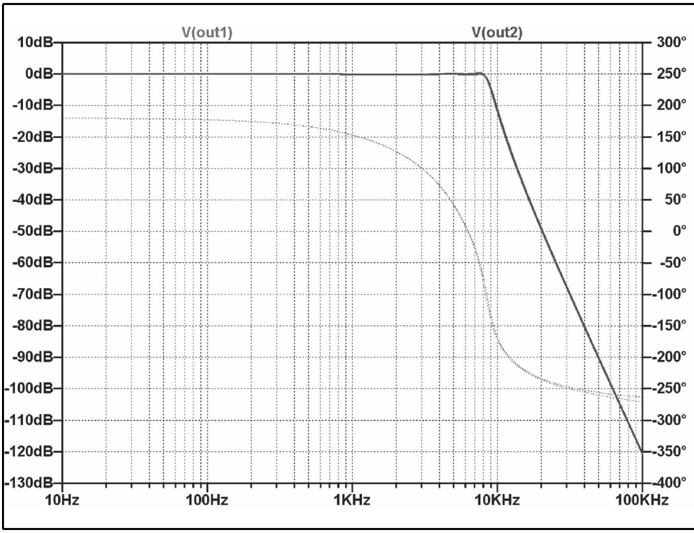


Fig 9: Graph showing the characteristics of the ideal filter and the real filter.

$$C_{11} = \frac{2K+1}{aK} \cdot \frac{1}{2\pi f_0 R_0} = \frac{3}{0.7465} \cdot \frac{1}{2\pi \cdot 8kHz \cdot 10k\Omega} = 8.0nF$$

$$C_{12} = \frac{a}{b(2K+1)} \cdot \frac{1}{2\pi f_0 R_0} = \frac{0.7465}{0.5583 \cdot 3} \cdot \frac{1}{2\pi \cdot 8kHz \cdot 10k\Omega} = 886pF$$

$$Q_2 = \frac{\sqrt{0.5583}}{0.7565} = 0.987$$

Transfer function

$$H_3(s) = \frac{1.1174}{s^2 + 0.2851s + 1.1174}$$

$R_1 = R_2 = R_3 = 10k\Omega$

$$C_{21} = \frac{2K+1}{aK} \cdot \frac{1}{2\pi f_0 R_0} = \frac{3}{0.2851} \cdot \frac{1}{2\pi \cdot 8kHz \cdot 10k\Omega} = 20.9nF$$

$$C_{22} = \frac{a}{b(2K+1)} \cdot \frac{1}{2\pi f_0 R_0} = \frac{0.2854}{1.1174 \cdot 3} \div \frac{1}{2\pi \cdot 8kHz \cdot 10k\Omega} = 169pF$$

$$Q_2 = \frac{\sqrt{1.1174}}{0.2851} = 3.70$$

6.

Simulation with LTspice

The completed filter circuit is shown in Fig 8. National Semiconductor operational amplifiers were used, unfortunately these are not in of the LTspice library and must be added. This is done quickly with LTspice. Spice models for different amplifiers can be found on the National Semiconductors web site [6]. The file LMC6482A.MOD can be found in the National Semiconductors component list. Copy this file into the list "…\swcadiii\lib\sub\" and call it LMC6482A.SUB.

Now open the symbol editor with LT-spice and "File>New symbol". With "File->Open" load from the listing "\lib\sym\opamps \" a suitable symbol, e.g. "opamp2.asy" and store it under "LMC6482A.asy ".

With "Edit->Attributes->Edit attributes" (or CTRL+A) open the dialogue "Symbol Attributes Editor" and make the following inputs:

SpiceModel	LMC6482A
Value	LMC6482A
Value2	LMC6482A
ModelFile	LMC6482A.SUB (if the file was renamed).

Store the changes the close and re-open LTspice so that the new component is included. Thus all the necessary components to be entered are available for the circuit in Fig 8.

The second order filters have been swapped into rising quality order.

In LTspice the frequency response can be specified as voltage controlled voltage supply (VCVS=Voltage Control Voltage Source) by a transfer function. In Fig 8 each filter stage was simulated with its own source so that both the transfer function of the total circuit and the partial circuits could be compared. The comparison in Fig 9 shows a high agreement between an ideal filter and the operation amplifier circuit.

7. Summary

In this article the cascade synthesis of multi-pole, active low pass filters was described in detail with an example. Using the program SCILAB it was possible to accomplish the otherwise very complex mathematical operations in a simple way. The complete circuit was examined with the LTspice simulator and in particular this includes the frequency response of the operational amplifiers used.

8. Literature

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9. Appendix

In the following circuits the values are standardised values. To de-normalise all resistor are multiplied by the reference resistance, the capacitors and the cutoff frequency are divided by the reference resistance.

$$R = R' \times R_0$$

$$C = \frac{C'}{2\pi f_0 R_0}$$

9.1. Realisation of first order filters

The simplest first order filter is an RC network. The amplifier in Fig 10 decouples the RC network of the following stage and if necessary provides gain.

A second option is shown in the circuit in Fig 11 using an inverting operation amplifier.

The transfer function of the circuit in Fig 10 is:

$$H(s) = K \cdot \frac{b_0}{s + b_0}$$

with:

$$K = 1 + \frac{R_b}{R_a}$$

and:

$$b = \frac{1}{R'_1 C'_1}$$

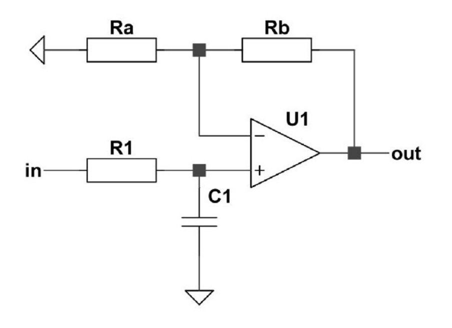

Fig 10: First order filter using a non-inverting amplifier.

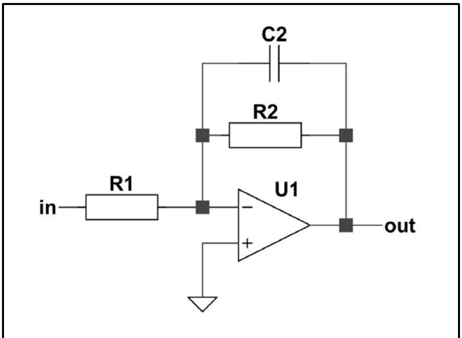

Fig 11: First order filter using an inverting amplifier.

Design

Default: $R'_a = 1$ and $C'_1 = 1$

Computed: $R'_b = K - 1$ and $R'_1 = 1/b$

The transfer function of the circuit in Fig 11 is:

$$H(s) = K \cdot \frac{b_0}{s + b_0}$$

with:

$$K = \frac{R_2}{R_1}$$

and:

$$b_0 = \frac{1}{R'_2 C'_2}$$

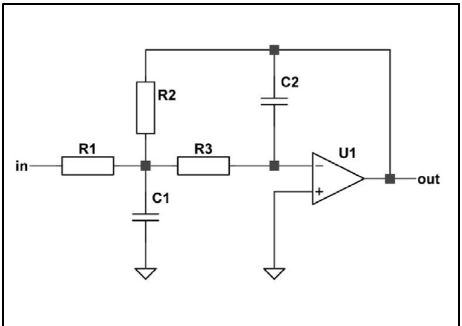

Fig 12: Second order filter.

Design

Default: $R'_1 = 1$

Computation: $R'_2 = K$ and:

$$C'_2 = \frac{1}{K_b}$$

9.2. Realisation of second order filters

The transfer function of the circuit in Fig 12 is:

$$H_{RC}(s) = \frac{R'_2}{R'_1} \cdot \frac{\frac{1}{R'_2 R'_3 C'_1 C'_2}}{s^2 + \frac{1}{C'_1}\left(\frac{1}{R'_1} + \frac{1}{R'_2} + \frac{1}{R'_3}\right)s + \frac{1}{R'_2 R'_3 C'_1 C'_2}}$$

The general transfer function for a second order filter is

$$H(s) = -K \frac{b}{s^2 + as + b}$$

The comparison results in:

$$a = \frac{1}{C'_1}\left(\frac{1}{R'_1} + \frac{1}{R'_2} + \frac{1}{R'_3}\right)$$

$$b = \frac{1}{R'_2 R'_3 C'_1 C'_2}$$

$$K = \frac{R'_2}{R'_1}$$

Design

It is given $R'_1 = R'_3 = 1$. Thus from the third equation $R'_2 = K$.

Now all resistors from which first equation are

$$C'_1 = \frac{2K+1}{aK}$$

from the second equation it can be calculated

$$C'_2 = \frac{a}{b(2K+1)}$$

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Current digital radio standards similar to FM voice transmission, Part 2

Continued from 2/2008

In the first part of this article the general conditions of digital radio standards were described, in a similar way this article describes the DECT standard. A comparison is made to the similar FM voice transmission standards. The achievable range as well the ability for trouble free reception and reception of data signals at low signal level is considered under comparable conditions.

2.7. DECT

DECT (Digital Enhanced Cordless Telecommunications) [19], [24], [4] was specified as an ETSI standard in 1992. It is a micro-cellular digital portable radio network for high user densities, to be used predominant in company buildings for data and language transfer. A combination of TDMA and FDMA are used for the DECT standard. The available frequency spectrum from 1880MHz to 1900MHz is divided into 10 channels with 1728kHz spacing. The centre frequency f_c is calculated as follows:

$f_c = f_0 - c \bullet 1728\text{kHz}$ with $C = 0, 1, \ldots, 9$

and $f_0 = 1897.344\text{MHz}$ {1}

The centre frequency for an active state can vary by a maximum of ±50kHz. The modulation is Gaussian minimum Shift Keying (GMSK) with a range time product, $B \cdot T = 0.5$. GMSK is a special type of the MSK (Minimum Shift Keying). Instead of switching violently between the two frequencies, the switching edges are flattened using a Gauss filter leading to the fact that the signal range is reduced.

The transmission capacity of each channel is divided into 10ms long frames containing 11520 bits. This framework gives a data transmission rate of 1152kbit/s. Each 10ms long frame is divided into 24 time slots. The first 12 time slots are intended for downlink, the next 12 for the uplink (Fig 6). For a duplex connection pairs of time slots are formed following each other within 5ms. Each of the 24 time slots has a length of 480 bits. The first 32 bits are reserved for the preamble and synchronisation. For a connection 64 bits of control information follow:

C (signalling for higher layers)

P (Paging information)

Q (system information, downlink only)

M (MAC Layer information)

N (handshake, identity information)

as well as 320 bits of useful information and a 4 bit parity check. A protection gap of 60 bits forms the remainder.

DECT uses a building block principle, the system component parts are: the air interface (Physical Layer), the access layer (Medium Access Control layer,

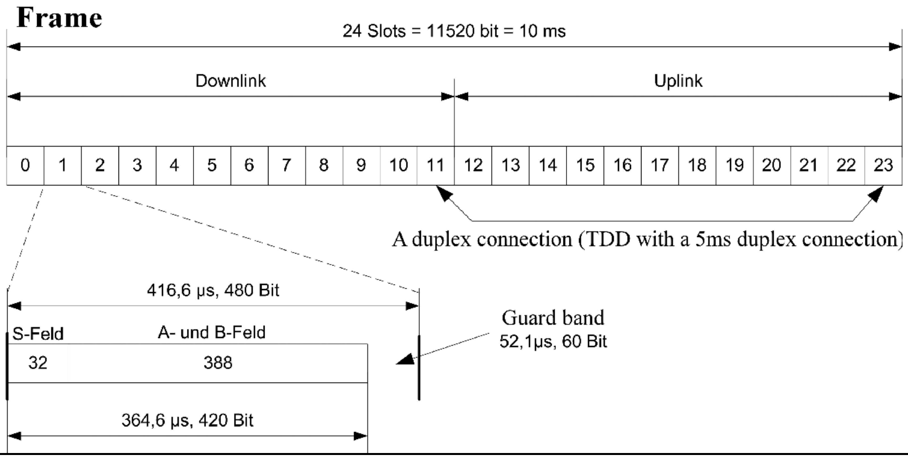

Fig 6: Frame structure for a DECT system.

MAC Layer), the link layer (Data Link Control Layer), the network or switching layer (Network Layer) and the administration of the lower layers (Lower Layer Management Entity).

Changing the radio cell without any interruption passes on a conversation for a moving mobile user. The maximum distance between the base station and mobile equipment in a free range is 300m and in buildings is up to 50m. This range can be increased using relay stations or mobile relay stations. For cost reasons DECT was developed for user speeds up to 20km/h.

The bit duration is 0.868µs. Without error correction time differences of under 10% of the bit duration are permissible, DECT can work with a delay spread up to 200 – 300nS, which corresponds to a transmission distance difference of approximately 100m being suitable for difficult environments Note: Stations with big reflections or problems with reflecting metal walls suitable cell planning and sector antennas can be used if necessary.

DECT offers the following service possibilities:

- Quality speech transmission in fixed network
- Duplex data communication rates with a total bit rate of 38.8kBit/s. In the half duplex mode the total bit rate of up to 931.2kBit/s is possible if all 24 time slots used. Still higher data rates are possible by multiplexing several carriers. Thus many cordless data applications are possible such as Internet, ISDN, Fax or TV.
- Applications for DECT covers PBX (Private Branch Exchange, private PABX), Telepoint and wireless local networks as well as speech telephones.

The DECT system can handle up to 1000 users in a Local Area (LA) [7]. For larger numbers of users, different cell areas can be planned that can be administered internally by DECT. A Dynamic Channel Selection procedure (DCS) is used to manage the high speech and data traffic loads and their uneven load on a cell with temporary and very variable load peaks. Thus in principle the entire frequency spectrum with all 120 channels are available to the mobile station and a suitable channel is selected in each cell. The 120 channels result from the 10 carrier frequencies and the 12 time slots for the downlink or uplink.

In cellular portable radio systems (GSM,

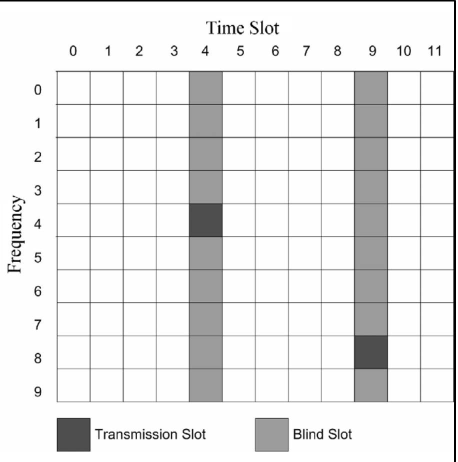

Fig 7: Blind slots for a transceiver with two connections and a permanently installed station.

UMTS,…) channel assignment takes place via a Fixed Channel Allocation (FCA). This is a very precise cell planning system but short load changes on an individual cell are not dealt with very well. With DECT there is no frequency planning, just one planning for the locations is necessary called DCS. The system can lower the blocking probability independently of changing loads adjust itself better than FCA networks. The dynamic assignment takes place according to certain rules so that there are no large disturbances. One rule is the interference adaptive channel assignment; dispatching decisions are made based on online measurements of the interference level. There are two types central and decentralised systems, DECT uses decentralised systems. The highest capacity gains can be obtained with centralised systems because all cell values are available to the control system and all dispatching decisions can be optimised based on the disturbances in adjoining cells. However the signalling expenditure between cells and control unit is higher. With decentralised systems, like with DECT, each permanently cell installed decides independently of other cells, so that the danger exists that channel dispatching will clash with another cell. A high threshold value for the C/I (Carrier to Interference ratio) reduces this danger, however this affects the capacity gains. The advantage is that there is no frequency planning and the system can cope with changes in network loads. Using interference adaptive channel assignment procedures it is possible to extend a DECT system by simply adding further permanently installed stations and increase the capacity.

The actual maximum number of channels of a DECT cell depends on the existing number of transceivers. If it has only one transceiver, then only one mobile station can be served per time slot, since a transceiver cannot work with two different frequencies at the same time. Thus only 12 channels can be used. If a channel on a terminal is occupied by a mobile, then the remaining channels in the same time slot are marked as blind slots (Fig 7). The mobile station looks for the possible channels within each framework in order to attain information about their quality. If the terminal occupies a channel for transmission then it cannot supervise the channels of the other frequencies in this time slot. Because of economies in the design it is not usual to be able to examine the time slots for quality before and after use. The blocked channels are no longer available (Fig 8).

To make the data transmission rate of individual connections variable several time slots can be assigned. This dispatching can take place for the uplink and downlink asymmetrically.

The digital coding used for DECT system is Adaptive Differential Pulse Code Modulation (ADPCM) using 32kBit/s.

In micro-cellular systems with small cell sizes and moving mobile stations the number of cell changes are counted. Because of the substantial signalling overhead it is best to keep the number of handovers as low as possible. The control system uses a decentralised handover algorithm steered by the mobile station

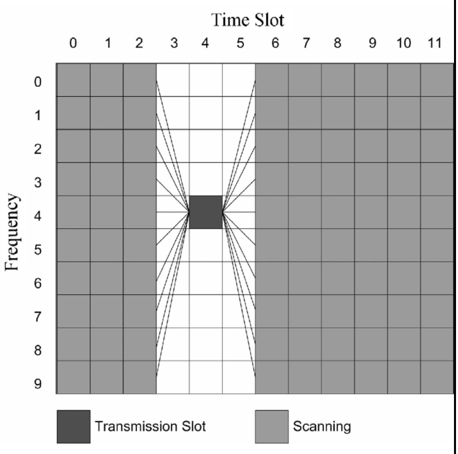

Fig 8: Un-usable channels because of the switching duration of the mobile station channel measurement.

that decide whether and when a handover is necessary. A seamless handover is possible because the old channel is only left if the new channel is already available. The user does not notice anything unlike the non-seamless handover schemes.

In order to protect DECT against unauthorised interference and abuse rules were formulated that ensure a high level of system security:

- Identification of a participant
- Avoidance of unauthorised use of the mobile station
- Identification of a permanently installed station
- Avoidance of unauthorised use of permanently installed stations
- Illegal listening to user or signalling data

The following safety precautions, similar to those used in public portable radio systems, are part of the standard system: A pin enquiry can be implemented when the mobile station is switched on. A mobile station has an international subscriber identification that is clearly defined in a certified range. Each mobile station is examined before the connection is establishment in order to prevent unauthorised network access. In addition an authorising key is stored in the mobile station and in the base station that is compared by an identification procedure. Thus a mobile station is protected against being manipulated by a permanently installed station to access the network. To protect other users from illegal listening to the user or signalling data the radio interface uses coded transmission [24].

3.

Comparison and evaluation

A direct comparison of the wireless transmission systems from chapter 2 in comparison to the similar FM voice transmission can made on the basis of the performance of standard receiver sensitivities. This is a measure of the quality of the demodulated signal from the smallest possible RF signal; it also determines the best transmission range. For speech transmission, based on similar types of modulation, the signal quality is expressed as signal/signal-to-noise ratio. For digital transmission the Bit Error Rate (BER) is used to determine the quality. The Bit Error Rate is the relationship of the number of bits that were incorrectly demodulated to the total number of bits sent. A bit error rate of 6 * 10^{-6} means that on the average 6 bits can be wrong if 1 million bits were sent.

A detailed search of the receiver sensitivities in the literature showed that often they couldn't be compared because the Bit Error Rates are defined differently. To be able to compare these systems on the basis of receiver sensitivity, the receiver sensitivities were converted for a Bit Error Rate of 10^{-3}. For DECT no conversions were necessary since the standard Bit Error Rate is referred to 10^{-3}.

3.1. Receiver sensitivities

The BER for an AWGN channel has the

form $P_b = a \cdot Q\left(b \cdot \sqrt{c \cdot \gamma_b}\right)$ {2}

Where constants a, b and c constants depend on the modulation system and the signal to noise ratio

$$\gamma_b = \frac{E_b}{N_0} \quad \{3\}$$

with the signal energy of a bit Eb and the noise performance density N_0. Q () represents the Q-function.

3.1.1. Q-function

The Q-function {4} [15] is often used in communication systems instead of the complementary error function. The probability for an incorrect (BER) is defined with the help of the Q-function.

$$Q(x) = \frac{1}{2} erfc\left(\frac{x}{\sqrt{2}}\right) \quad \{4\}$$

The following are calculated values of the Q-function:

Q(x) = BER	x
10^{-5}	4.26489
10^{-3}	3.09023
2×10^{-3}	2.87816
10^{-2}	2.32635
1.2% = 0.012	2.25713
1.5% = 0.015	2.17009
2.4% = 0.024	1.97737
2 x 0.024	1.66456

3.1.2. Tetrapol

Modulation: GMSK BT = 0.25

Sensitivity is indicated for Tetrapol for a BER of 1.5% [2]. The error probability is calculated according to [16]:

$$P_b = Q\left(\sqrt{\frac{2 \cdot \alpha \cdot E_b}{N_0}}\right) \quad \{5\}$$

a ≈ 0.68 for BT = 0.25

BER = 10^{-3}:

$$3.09.23 = \sqrt{2 \cdot 0.68 \cdot \gamma_{b2}} \Rightarrow \gamma_{b1} = 7.02 \rightarrow 10\log\gamma_{bidB} = 8.46dB$$

BER = 1.5%:

$$2.17009 = \sqrt{2 \cdot 0.68 \cdot \gamma_{b2}} \Rightarrow \gamma_{b2} = 3.46 \rightarrow 10\log\gamma_{b2dB} = 5.39dB$$

Correction factor = $\gamma_{b1dB} - \gamma_{b2dB} = 3.07dB$

3.1.3. GSM/GSM-R/GSM-BOS

Modulation: GMSK BT = 0.3

With GSM sensitivity for a BER of 2.4%, based on a data sheet of the Sony Ericsson K610 mobile telephone. In accordance with [26] the error probability is computed:

$$P_b = 0.5 \cdot Q\left(\sqrt{d^2{}_{\min} \cdot \frac{E_b}{N_0}}\right) \quad \{6\}$$

d^2min = 1.7874 for BT = 0.3

$$P_b = 0.5 \cdot Q\left(\sqrt{d^2{}_{\min} \cdot \frac{E_b}{N_0}}\right) \rightarrow 2 \cdot P_b = Q\left(\sqrt{d^2{}_{\min} \cdot \frac{E_b}{N_0}}\right)$$

BER = 10^{-3}:

$$2.87816 = \sqrt{1.7874 \cdot \gamma_{b1}} \Rightarrow \gamma_{b1} = 4.63 \rightarrow 10\log\gamma_{b1dB} = 6.66dB$$

BER = 2.4%:

$$1.66456 = \sqrt{1.7874 \cdot \lambda_{b2}} \Rightarrow \gamma_{b2} = 1.55 \rightarrow 10\log\gamma_{b2dB} = 1.90dB$$

Correction factor = $\gamma_{b1dB} - \gamma_{b2dB} = 4.75dB$

3.1.4. TETRA

Modulation: p/4-DQPSK (coherent). Sensitivity is defined for TETRA as a BER of 1.2% [1]. The error probability amounts to [16]:

$$P_b = Q\left(\sqrt{2 \cdot \gamma_b}\right) \quad \{7\}$$

BER = 10^{-3}

$$3.09023 = \sqrt{2 \cdot \gamma_{b1}} \Rightarrow \lambda_{b1} = 4.77 \rightarrow 10\log\gamma_{b1dB} = 6.79dB$$

Table 2: System types and receiver sensitivity.

Reference standard	BER= 10^{-3}	20dB	---------------- BER = 10^{-2}			--------------------
	DECT	FM speech	TETRA	Terapol	DIIS	GSM, GSM-R GSM-BOS
Receiver sensitivity	-94dBm	-115dBm	-104dBm BER= 1.2%	-111dBm BER= 1.5%	-110dBm	-102dBm BER> 2.4%
Modulation	GMSK	FM	p/4 DQPSK	GMSK	CP-4GFSK	GMSK
Correction factor for BER=10^{-3}		+2.7dB		+3.1dB	approx. +2.5dB	+4.8dB
Corrected sensitivity BER=10^{-3}bBm	-94	-115	-101.3	-107.9	-107.5	-97.1

BER = 1.2%:

$$2.25713 = \sqrt{2 \cdot \gamma_{b2}} \Rightarrow \gamma_{b2} \rightarrow 10 \log \gamma_{b2dB} = 4.06dB$$

Correction factor = $\gamma_{b1dB} - \gamma_{b2dB} = 2.73dB$

3.1.5. Interpretation

In Table 2 all systems described in chapter 2 are compared based on receiver sensitivity as described in chapter 3.1. The systems are arranged according to their BER. It can be seen that none of the digital systems approaches the good receiver sensitivity of a similar FM voice transmission. This shows that FM voice transmission has receiver sensitivity better than the state of the art problem free systems. It can be seen that the narrow band systems Tetrapol, TETRA and DIIS exhibit the best receiver sensitivities of the digital systems. This is because the sensitivity of a receiver is directly related to the noise of the received spectrum.

Receiver sensitivity of similar transmission modes is referred to by a certain signal/signal-to-noise ratio of the demodulated signal and often defined as Signal, Noise and Distortion to Noise and Distortion (SINAD). SINAD describes the logarithmic relationship of signal to noise performance in dB, where distortions and all other defects are considered. A further comparison assumes that for a similar voice transmission with a BER of 10^{-3} corresponds to 20dB SINAD.

3.2. Maximum range

The maximum possible attenuation between transmitters and receivers is shown in Table 3. It shows the result of the difference in transmitter power with receiver sensitivity corrected for a BER of 10^{-3}.

$$a_{max} = P_{Smax} - P_{Emin} \quad \{8\}$$

where a_{max} is the maximum attenuation in dB, P_{Smax} is the maximum transmitter power in dBm and P_{Emin} is the receiver sensitivity in dBm. For the case for obstacle-free propagation between the transmitter and receiver

$$a_{max} = a_0 \quad \{9\}$$

with the free apace attenuation a_0dB and thus the maximum range r_{max}

$$r_{max} = \frac{\lambda}{4\pi} \cdot 10^{\frac{a_{max}}{20}} \quad \{10\}$$

in metres when using isotropic antennas.

3.2.1. Interpretation

Table 3 shows what is evident from

Table 3: System comparison on the basis maximum possible range.

	DECT	FM speech or BOS		TETRA	Tetra pol	DIIS	GSM GSM-R GSM-BOS
Corrected sensitivity BER=10^{-3}dBm	-94	-115		-101.3	-107.9	-107.5	-97.2
Permitted transmitter power							
Base station W	0.25	25		25	25	25	25
Base station dBm	24.0	43.98		44	44	44	44
Mobile W	0.25	5		1	1	5	2
Mobile dBm	24	336.99		30	30	37	33
Maximum attenuation between transmitter and receiver							
Base station dB	118	158.98		145.3	151.9	151.5	141.2
Mobile dB	118	151.99		131.3	137.9	144.5	130.3
Maximum range due to free space attenuation							
Used freq. MHz	1890	150	75	390	390	390	900
Wavelength m	0.16	2	4	0.77	0.77	0.77	0.33
Range km	10	6328.6	12657.1	224.1	482.3	1026.4	86.4

Table 2 that none of the digital systems reaches the range of the similar system. The theoretical range for FM voice transmission is high because on the one hand the transmitting power is higher in the comparison to for instance DECT and on the other hand due to the different free space attenuation of the frequency bands used.

For an objective comparison of the systems a more exact analysis is required.

3.3. Comparison for the same frequency and same transmitting power

For this comparison the maximum range was compared using the equation {10} that uses the free space attenuation under ideal conditions. The frequency range used for this comparison was not the range used by the systems but the 4m and 2m bands that are used by BOS. The results are shown in Table 4.

Table 5 shows the systems after standardising them all to a transmit power of 5W.

3.3.1. Interpretation

The translation of the frequency ranges used by BOS shows the DECT standard cannot be compared because its bandwidth requirements are too large. GSM would be possible, however the number of channels is very small and one GSM channel would use the worst frequency.

In systems like TETRA, Tetrapol, DIIS and GSM it can be seen in Table 3 that compared with the values in Table 2 that a substantially higher range is to be expected with these systems on the bad frequency ranges. The systems are easier to compare if the transmitting power is standardised. For this standardisation an output of 5W was selected because hand held radios are specified with a maximum transmitting power of 5W. This comparison of the systems on the same frequency and same transmitting power are shown in Table 5. In this comparison the FM voice transmission is clearly the best for range.

The conclusion from this analysis is that systems that have many decades of tech-

Table 4: System comparison through frequency bands used by BOS.

	DECT	FM speech	TETRA	Tetrapol	DIIS	GSM GSM-R GSM-BOS
2m band (150MHz) max. bandwidths 0.48MHz & 1.82MHz	Not possible bandwith too large	25 & 92 channel	Possible 19 & 72 channel	Possible 38 & 145 channel	Possible 38 &145 channel	Possible 2 & 9 channel
Range in free space km	126.12	6328.56	582.62	1254.02	2668.73	518.23
4m band (75MHz) bandwidth 3.26MHz	May be possible no more than 1 channel	164 channel	Possible 130 channel	Possible 260 channel	Possible 260 channel	Possible max 16 channel
Range in free space km	252.24	12657.13	1165.24	2508.04	533.47	1036.45

nological development are to be preferred. In addition similar radio connections do not fail abruptly like digital transmissions at a given Bit Error Rate threshold.

Table 5: System comparison for the same frequency and same transmitting power.

	DECT	FM speech	TETRA	Tetrapol	DIIS	GSM GSM-R GSM-BOS
2m band (150MHz) max. bandwidths 0.48MHz & 1.82MHz	Not possible bandwith too large	25 & 92 channel	Possible 19 & 72 channel	Possible 38 & 145 channel	Possible 38 &145 channel	Possible 2 & 9 channel
Range in free space km	564.03	6328.56	1302.78	2804.08	2668.73	819.39
4m band (75MHz) bandwidth 3.26MHz	May be possible no more than 1 channel	164 channel	Possible 130 channel	Possible 260 channel	Possible 260 channel	Possible max 16 channel
Range in free space km	1128.07	12657.13	2605.56	5608.16	533.47	1638.78

4.

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Franco Rota, I2FHW

10MHz - 10GHz Noise source diodes

1. Introduction

In VHF Communications 1/2007 [1] I described a simple 10MHz to 3.5GHz noise source, the purpose of that article was to explain how to build a very simple noise generator using the NS-301 noise diode, either for applications like noise figure measurement or for a broadband noise generator for scalar applications with a spectrum analyser.

Now I will describe a 10MHz - 10GHz noise source generator with an improved bias network that uses the NS-303 noise diode.

This project was born some months ago for the 13th E.M.E. (moon bounce) conference in Florence during August 2008, the organisation asked me to cooperate to build some noise source generators to give to participants during the conference.

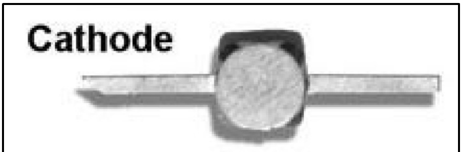


Fig 1: NS-303 noise diode.
Specification:
Case : Metal-ceramic gold plated
Frequency range: 10Hz - 8GHz (max 10GHz)
Output level: about 30dBENR
Bias: 8 - 10mA (8 - 12V)

Tests and measurements are supported by 20 pieces of noise source generators built for this conference, so I think that results are very reliable and repeatable.

2. Schematic diagram and components

The noise generator uses the NS-303 diode (Fig 1) that is guaranteed up to 8GHz but following the descriptions below it will be very easy to reach 10.5GHz making it possible to use it up to the 3cm band (10.368 GHz), using a diode of moderate price.

The aim of this article is to explain how to build a noise generator using easy to find components.

The circuit diagram, Fig 2, is very simple, the power supply is 28V pulsed AC applied to connector J1 which is normalised in all the noise figure meter instruments. U1 is a low dropout precision regulator to stabilise the voltage for the noise diode to 8 - 12V, the current through the diode can be around 8 to 10mA set by trimmer RV1.

2.1 R3, R4 and R5 resistors

These resistors can be a total of 100 - 220Ω, the total value is not critical, the 0603 case is very important in order to keep the stray capacity as low as possible, it would be better to solder the

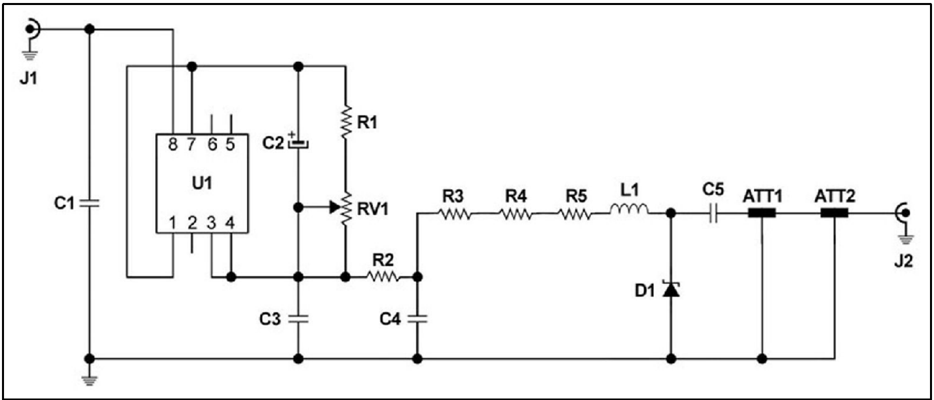


Fig 2: Circuit diagram for a noise source 10MHz - 10GHz using a NS-303 noise diode.

resistors without using copper track on the PCB see Fig 5.

2.2 ATT1, ATT2 Attenuators

These attenuators are very important to obtain an output level of about 15dBENR but more important to obtain an output return loss as low as possible.

In my previous article in issue 1/2007 I described this concept very well, the mismatch uncertainty is the main cause of errors in noise figure measurement [2].

The total value of attenuators ATT1 + ATT2 can be around 14dB, the pictures in Fig 5 – 6 show a 6dB chip attenuator mounted on the PCB and a 7 or 8dB external good quality attenuator, in fact the output return loss depends mainly on the last attenuator (ATT2). The first attenuator (ATT1) can be less expensive and built directly on the PCB because it is less important for the output return loss.

I used a 7 or 8dB external attenuator in order to obtain the best output ENR value because every diode has it’s own output noise.

Everyone can change the output attenuator depending on the ENR that is needed; in this project I chose an output level of 15dBENR so the attenuators have a value of 14dB.

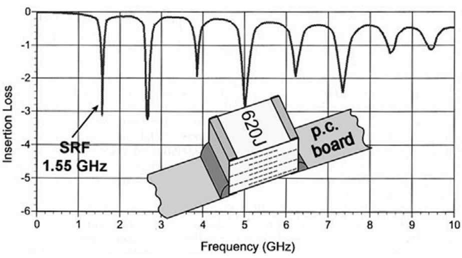
2.3 C5 dc block output capacitor

The selection of this capacitor is very important to flatten the output level; in the previous article I only quickly mentioned this fact because we were only talking about 3.5GHz, now in order to reach 10.5GHz I will do a better description.

The DC blocking capacitors are used to

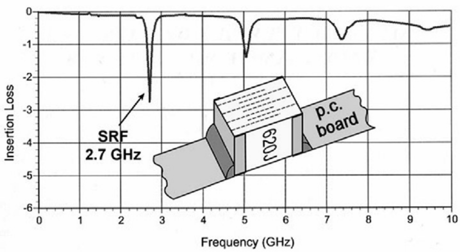
Table 1: Parts list.

D1	NS-303 noise diode	NS-303	J1	BNC female connector	
U1	LP2951CMX SMD SO8 case	LP2951CMX	J2	SMA male panel mount connector	SMA-24A
C1	10nF 0805			Suhner 13SMA50-0-172	
C2	1µF 25V tantalum		R1	100Ω 1206	
C3	100nF 0805		R2	18Ω 0805	
C4	1nF 0805 COG		R3, R4, R5	33Ω to 68 0603	
C5	2 x 1nF 0805 COG in parallel	see text	L1	6.8 or 8.2nH 0603	BCG-6n8-A
ATT1	6dB chip attenuator DC-12GHz		RV1	100Ω trimmer multi turn SMD	POT-SM-101-M
ATT2	7 or 8dB external attenuator				
	CD - 12GHz or better DC - 18GHz		PCB	25N or RO4003 or RO4350	see text
				30 mils, εr 3.40, 11 x 51mm	



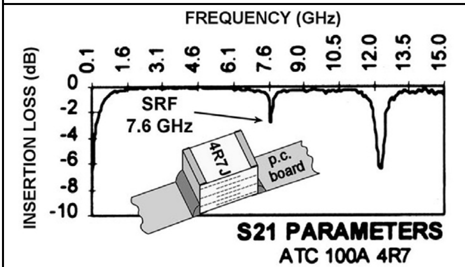
ATC100B 62pF 110mils = 3mm

The manufacturer guarantees an SRF > 900MHz, in fact the network analyser shows an SRF of 1.55GHz with parallel orientation.



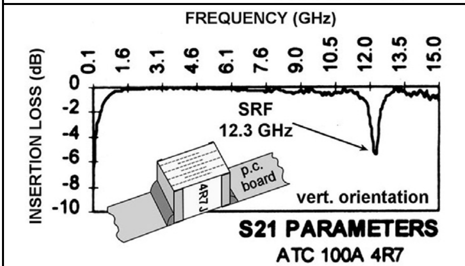
ATC100B 62pF 110mils = 3mm

With the same capacitor the network analyser shows an SRF 2.7GHz with vertical orientation



ATC100A 4.7pF 55mils = 1.5mm

The manufacturer guarantees an SRF > 4GHz, in fact the network analyser shows an SRF of 7.6GHz with parallel orientation.



ATC100A 4.7pF 55mils = 1.5mm

With the same capacitor the network analyser shows an SRF 12.3GHz with vertical orientation

Fig 3: Examples of SRF frequency and its improvement.

block the DC voltage and to pass the RF signal with the minimum possible attenuation. If you use the ATC100A or 100B capacitors they have a very low insertion loss but have the problem of self resonance in ultra wide band applications, the graphs in Fig 3 show 2 examples how you can improve the SRF with vertical orientation.

Fig 3 shows 4 graphs of the SRF frequencies for ceramic capacitors and how to improve the SRF of ATC100A or 100B capacitors for ultra wide band applications.

My decision was to avoid ATC capacitors and to find some capacitors without any SRF and lower Q, after many at-

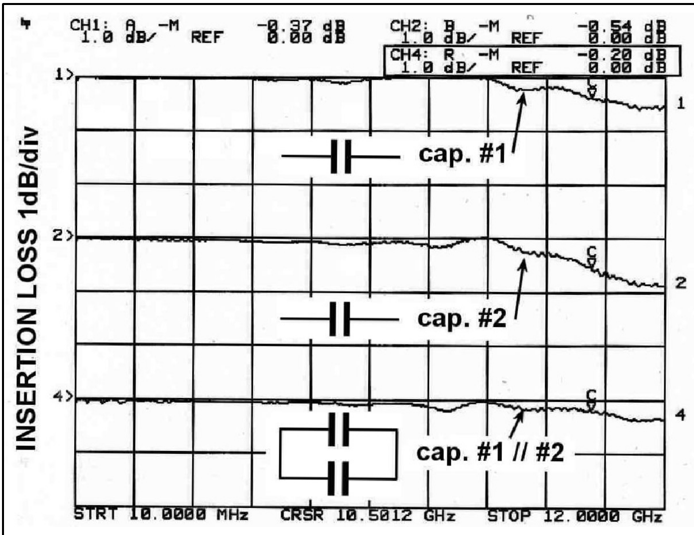


Fig 4: C5 capacitor CCB 1nF. Insertion loss of 1nF NP0 class 1 capacitor with a span from 10MHz - 12GHz, 1dB/div.

It is demonstrated that there are no SRFs in the entire band.

C = 10,5GHz marker

tempts and researches I found that NP0 class 1 multi-layer capacitors with an 0805 case have the best performance referred to low level applications (not to be used in RF power applications or in low noise amplifiers).

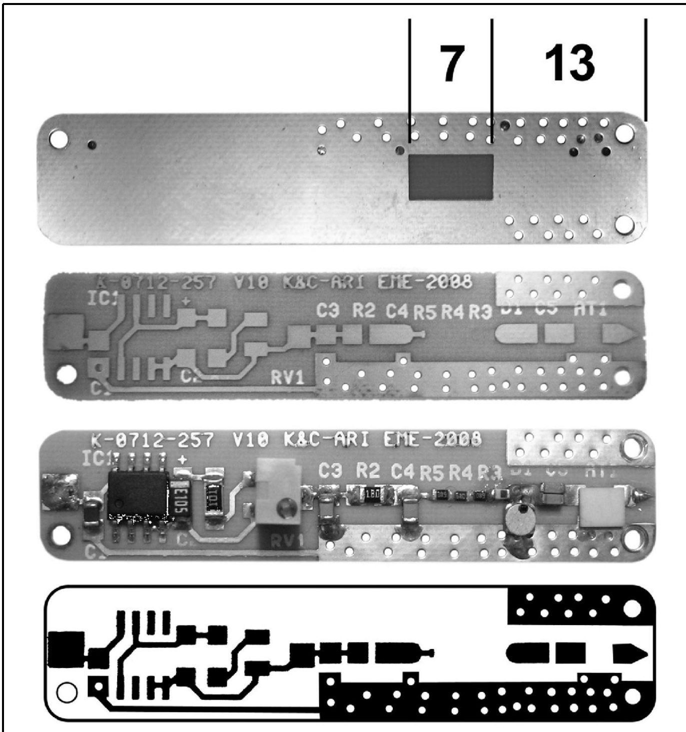


Fig 5: PCB and component layout.

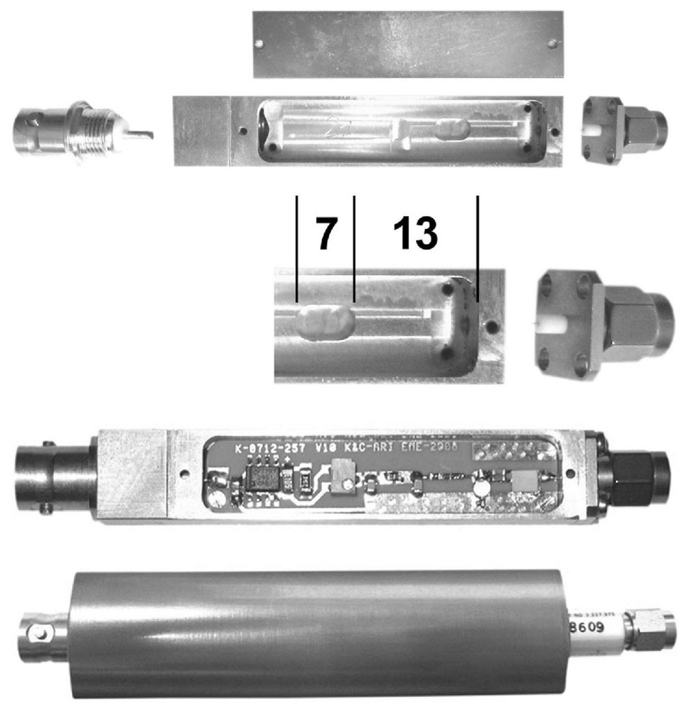


Fig 6: Box and final release.

I choose to put 2 1000pF capacitors in parallel in order to reach a minimum frequency of 10MHz.

For ultra broadband applications the ATC manufacturer has a capacitor of 100nF with 16kHz to 40GHz frequency operation in a 0402 case [3] but I prefer to avoid this special component and use more easy to find one.

In Fig 4 the 1nF capacitors show a low insertion loss, with 2 capacitors in parallel, the marker C shows an insertion loss of about 0.2dB at 10.5GHz that is appropriate for this project at low price.

2.4 PCB

The noise generator is considered a passive circuit so it is not necessary to use very expensive Teflon laminates, moreover the track length is so short that the attenuation introduced makes it unnecessary to use Teflon laminates. I selected ceramic laminate, that is very popular in RF applications, with εr 3.40. It is available in several brands and they all have the same performance, Rogers RO4003 or RO4350, Arlon 25N etc…, with a thickness of 30mils (0.76mm).

In order to easily reach the 10GHz band it is necessary to remove the ground plane around R3, R4, R5 and L1, the size is 7 x 4mm (Fig 5)

2.5 Metallic box

As shown in Fig 6 the components of the noise source generator are enclosed in a very small milled box. Every box behaves like a cavity excited by several secondary propagation modes. For higher frequencies or in medium size boxes the RF circuit will also have many secondary propagation modes at various frequencies. Since every box is different in size, shape and operating frequency the calculation of secondary propagation modes is very difficult. To avoid this problem

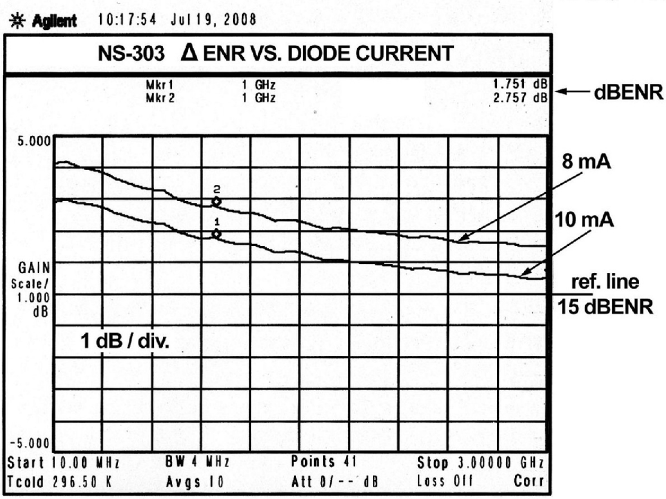


Fig 7: Shows the variation of output noise level vs. current. Span 10MHz - 3GHz, 1dB/div.

microwave absorbers are very often used placed into the cover of the box to dampen the resonance [4].

I selected a very small box in order to avoid both the secondary propagation modes and the microwave absorber; the size that I used gives no trouble up to 10GHz.

If someone wants to increase the size of the box (internal size) it will be necessary to use a microwave absorber.

It is also necessary to remove part of the ground plane in the metallic box by milling a 7 x 4 x 3mm deep slot corresponding to R3, R4, R5 and L1.

3. Bias current

The nominal current should be 8mA, during my tests I found that the output

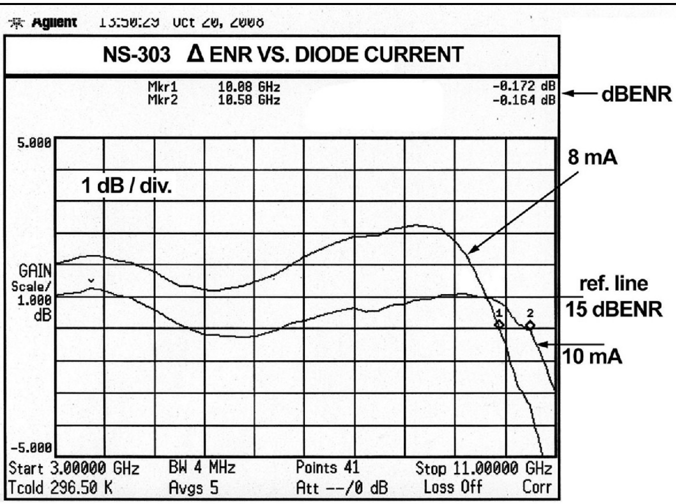


Fig 8: Shows the variation of output noise level vs. current. Span 3GHz - 11GHz, 1dB/div.

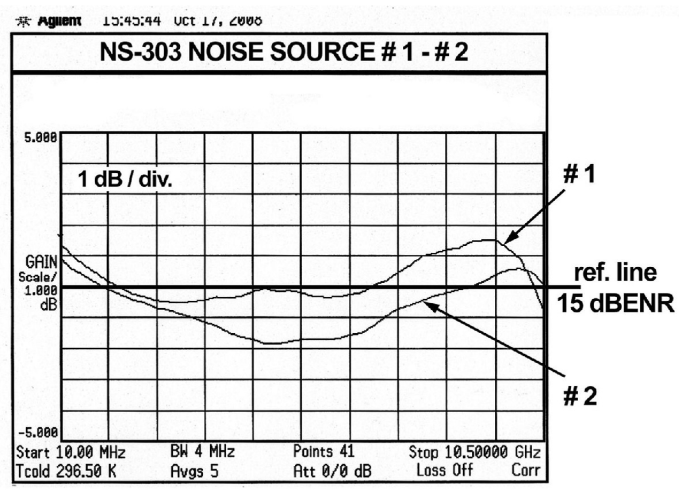


Fig 9: Typical output noise from 2 different noise sources. Span 10MHz - 10.5GHz. Reference line 15dBENR, 1dB/div.

noise level has a quite strange but interesting variation: increasing the diode current the output noise level decreases by about 0.5dB/mA up to about 9GHz, beyond this frequency the effect is exactly the opposite.

Fig 7 shows the difference in output ENR of about 1dB with 8 and 10mA bias current and Fig 8 shows a little improvement of frequency range by about 500MHz with 8 and 10mA bias current.

Fig 8 shows the decrease of about 1dB of ENR level with 10mA instead of 8mA maintaining the same shape in the diagram.

During the calibration it is possible to play with the current to "tune" the ENR level, if you can loose 1dB of ENR level, you will have a more extended frequency range which is exactly what is needed to reach the 3cm amateur radio frequency band (10.4GHz).

The bias current can be measured easily directly on the BNC input connector with +28V DC from a normal power supply; the input current is more or less the same current through the noise diode.

4.

Test results

I tested 20 pieces of the noise source generator and they all gave nearly the same results, the measurement in Fig 9 refer to the use of a 6dB internal attenuator plus a 8dB external attenuator (MaCom or Narda DC - 18 GHz).

A typical output noise level can be 15dBENR +/-1.5dB or 15dBENR +/-2dB or 15dBENR +1/-2dB, a ripple of +/-1.5dB or +/-2dB is a normal values.

The output return loss depends mainly on the external attenuator; I measured a 30dB return loss up to 5GHz, 28dB up to 8GHz and 25 to 28dB at 10GHz.

We have to consider that each 1dB more of external attenuation will improve the output return loss by 2dB, so if you can use, for instance, an attenuator of 17/18dB you will reach a very good return loss (>30/35dB) with an output noise around 5dBENR.

5.

Calibration

Unfortunately the calibration of a noise source is not an easy thing to do.

We know very well that RF signal generators have an output level precision of typically +/-1/1.5dB and this doesn't worry us, we also know that our power meter can reach +/-0.5dB precision or even better. We need a very high precision for a noise generator used with a noise figure meter. For the classic noise source 346A, B and C, Agilent gives ENR uncertainty of +/-0.2dB max. (< 0.01dB/°C). The new N4000 series are used for the new noise figure analyser N8975A with ENR uncertainty of +/-0.15dB max.

In my lab I used the new noise figure analyser N8975A with the precision noise source N4001A so I can guarantee a typical precision of +/-0.1dB up to 3GHz and 0.15dB up to 10GHz.

It means that the calibration must be done with a good reference noise source, it can be a calibrated noise source compared with the one you have built with a low noise preamplifier and a typical noise figure meter.

Example: you have a low noise amplifier with 0.6dBNF and your calibrated noise source indicates a 15.35dB of ENR, now you can change the noise source to the one you have built and for instance you measure 0.75dBNF, it means that your noise source has 15.35 + (0.75dB - 0.6dB) = 15.50dBENR.

6.

Other application

As I described in the previous article [1] that the noise source can be used as a broadband noise generator combined with a spectrum analyser like a "tracking generator" for scalar applications.

This is not a true tracking generator because it works in a different way (read my previous article [1]). The problem here is to reach 3 decades of frequency range, 10MHz to 10GHz, with a flat amplifier of at least 50dB.

Today some MMICs are available that can do this work like ERA1, ERA2, MGA86576 etc…, the problems can be to reach a flat amplification and to avoid self oscillations with such high amplification.

This device can be very interesting because it can be a useful tool to use with any kind of obsolete spectrum analyser to tune filters, to measure the return loss etc… up to 10GHz.

For more information regarding noise source diodes see:
www.rfmicrowave.it/pdf/diodi.pdf (from page A 14)

7.

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Richard Giles, G4LBH

Review of Mini-Kits 6cm 1 Watt PA (EME141-5800)

1.0 Review

Mini-Kits [1] in Australia, produce several SMD kits of interest to the keen microwaver.

In addition to a range of preamplifiers and ATV equipment there is a range of three similar kits producing 2W on 13cm, 2W on 9cm and 1W on 6cm respectively.

Reviewed here is their 1W PA requiring only 3.2mW (+5dBm) for the rated output at 1dB gain compression. This particular amplifier uses a Transcom TC3531 PHEMT MMIC IC whereas the 9cm amplifier uses the TC3341 device. These MMIC ICs are 50ohm devices and therefore require minimal input and output matching. As a result, the circuit is very compact, on a single 48mm x 37mm board and requires very little setting up (Fig 1).

The kit is supplied complete with a fibreglass reinforced ceramic PC board, all parts, screws and optional heatsink. As an additional option, the TC3531 can be supplied ready mounted on the PC board and this is recommended unless you are confident that excess solder will not flow through the plating holes to the underside of the board (needs to be flat and in direct contact to the heatsink) and that you will not blow the device!

Building is very straightforward but you will need to drill and tap the heatsink for the SMA connectors and, unless you are to use bolts and nuts through the heatsink, you will also need to drill and tap the heatsink to affix the completed board to it. It is here that I deviated from the instructions by cutting down the heatsink, by stripping the cooling fins, as the bottom of the supplied heatsink was to be fitted to an aluminium box with an external heatsink mounted on the outside. The board, the modified supplied heatsink, case and external heatsink were then bolted together.

A 140uH inductor has to be wound using the enamelled copper wire and core supplied. This feeds all power to the PHEMT MIMC from a 7V output switched mode regulator run off any suitable 5 to 14V supply. The total current taken is less than 1A.

Setting up requires adjustment of the bias to the MIMIC and making sure the maximum drive of 10mW is not exceeded. In my case, several attenuators were needed to bring the 200mW of my DB6NT transverter output down to below this level.

The total cost of this kit at the time of writing this review is $121.10AU plus shipping.

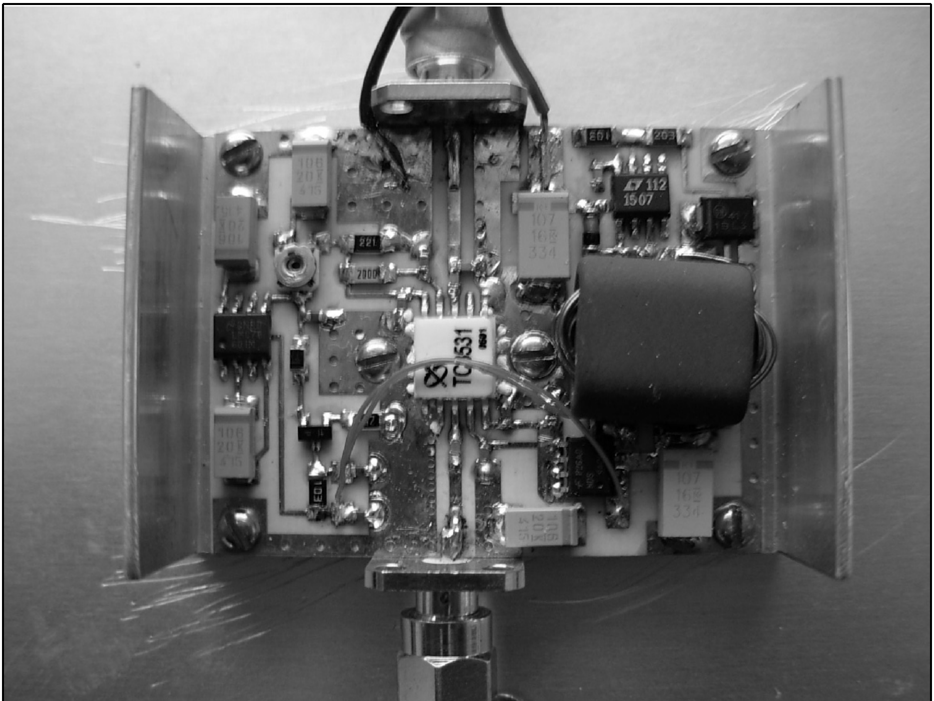


Fig 1: The completed 1W 6cm Power amplifier.

I particularly liked this kit because its diminutive size and weight opens up the possibility of mounting the amplifier close to the antenna feed. The onboard voltage regulator gives a stable board supply over a wide range of supply voltages, particularly useful when at the end of a long power lead from the shack. The output of 1W provides a useful output for a feed to a dish and, with more than a fourfold increase in power of my transverter, it has useful gain for very little money.

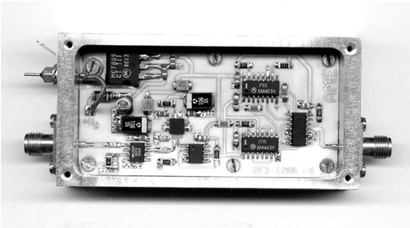
The Mini-Kits 6cm 1Watt PA is therefore a very cost effective solution to the problem of finding that medium power output amplifier at 6cms.

2.0

References

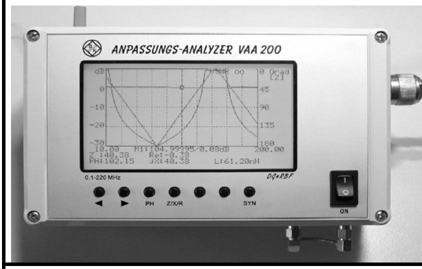
[1] Mini-Kits, South Australia. www.minikits.com.au

Ready built units, Kits and PCBs available for recent projects



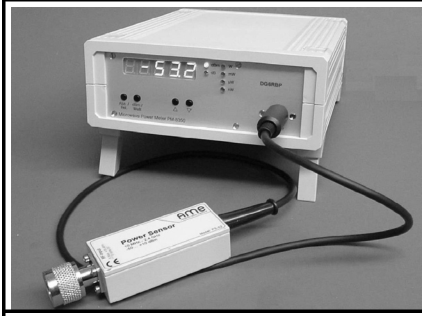
12GHz Divide by 1000 prescaler by Alexander Meier, DG6RBP from issue 4/2003

Ready built in diecast housing with SMA connectors	**£215.00**
Also availabe an 18GHz version of the divide by 1000 prescaler	**£330.00**



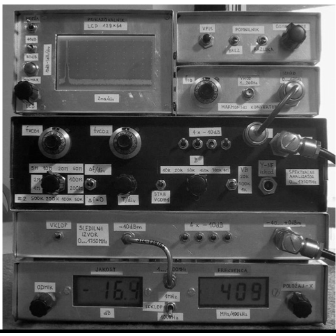
Vector analyser VAA 200 for the 0.1 to 220MHz range with graphical display by Bern Kaa, DG4RBF from issue 3/2007

PCBs	**£31.50**
Micro-controller with software	**£95.00**



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

PCB	**£25.00**
Micro-controller with software	**£65.00**



Spectrum Analyser by Matjaz Vidmar, S53MV

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

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

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

John Fielding, ZS5JF

John's Mechanical Gems Number 4

Setting out control shafts

Constructing an item for the shack requires a certain amount of mechanical skill. There is nothing more unsightly or depressing to the constructor than control shafts that are not aligned correctly and this spoils an otherwise good project. Recently the writer was constructing an HF linear amplifier and needed to ensure the tuning capacitor shafts were at the same height on the front panel.

Method 1

Using a flat surface as a reference we can simply measure the height of each control shaft. In the writers case the two tuning capacitors had different styles and so some additional metalwork was required to correct the situation.

Fig 1 shows the anode tuning capacitor bolted to a piece of aluminium angle. The loading capacitor needed a piece of 3mm thick plate to attach it to the same aluminium angle. This was set up on the drilling machine table and the back of the

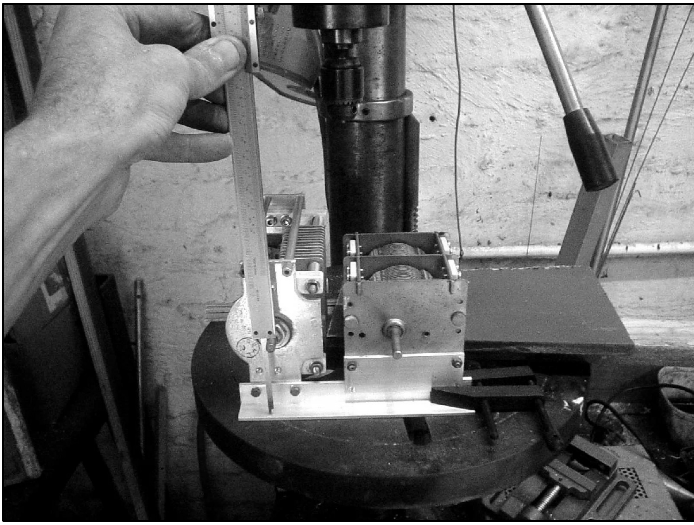


Fig 1: Measuring the height of the first capacitor tuning shaft.

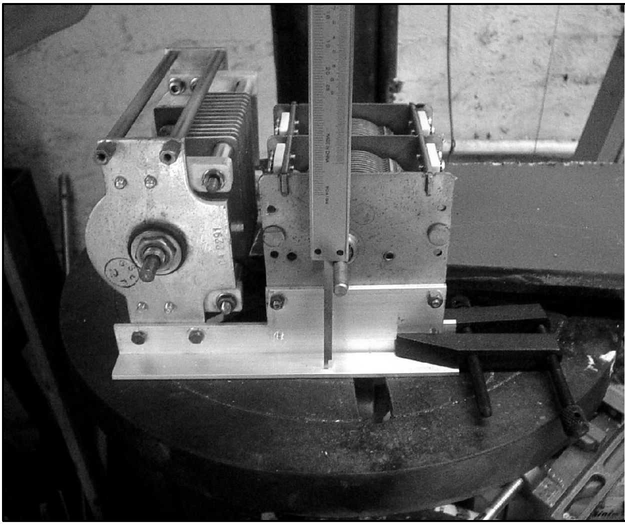


Fig 2: The second shaft set at the correct height.

anode tuning capacitor chocked up so the angle lay flat on the drill table. The loading capacitor with its piece of 3mm plate was positioned next to the anode tuning capacitor. A toolmakers clamp was attached to keep the loading capacitor in approximately the right place.

The vernier was transferred to the second capacitor shaft and the difference in height was noted. Since the toolmakers clamp is not tightened too much a gentle tap with a small hammer on the capacitor endplate can cause it to be moved down by small increments. Fig 2 shows the correct height.

Once the correct height has been attained the mounting holes in the 3mm plate can be transferred to the angle by a sharp scriber.

Method 2

An alternative method is to use a spirit level (Fig 3). Ensure the surface used is level by first checking with the spirit level. Place the spirit level across the two shafts and adjust the height until the level is correct.

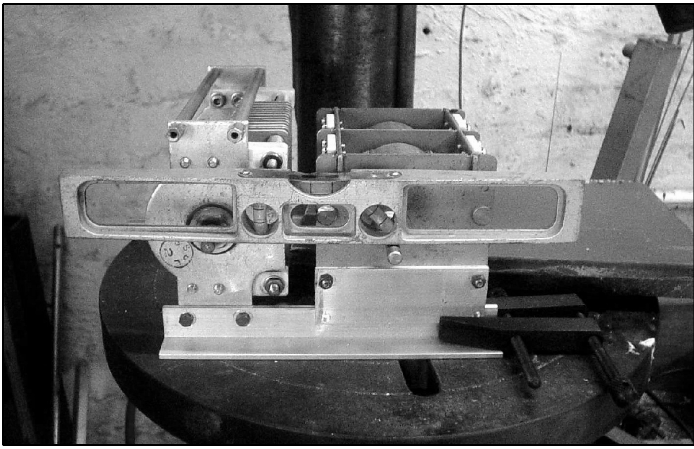


**Fig 3:
Using a spirit level to adjust the height.**

Gunthard Kraus, DG8GB

Internet Treasure Trove

Links to SPICE simulation resources

Anyone who has changed to simulate a circuit before it is constructed should look at this site. It gives links to sites with free SPICE models, training courses, software, magazine articles and application notes etc.

Address:
http://www.penzar.com/links.htm

VOACAP Quick Guide

"HF Ionospheric Communications Propagation Analysis and Prediction " this is a wave forecasting and propagation prediction program. It gives information on the current conditions and a free software download.

Address: http://voacap.com/

Watkins Johnson Historical information

For those who are interested in the products manufactured by the well known companies Watkins Johnson or Communications Engineering, this site gives a good overview of their history and products. There are manuals and information on items that you may still find in your shack.

Address:
http://watkins-johnson.terryo.org/

Various Scans of HP manual, etc.

If you are like the author always looking for manuals on measuring instruments, this site will be invaluable to find that special manual.

Address:
http://www.kennethkuhn.com/hpmuseum/scans/

Keiths Vintage RACAL Enthusiasts Site

This is a site that will bring back memories for the older generation and inform the younger readers. In particular the RA17 was a milestone in short wave receivers design with its special mixing technique. Now that the analogue receiver is fast disappearing in favour of the SDR receiver the achievements of the pioneers should not be forgotten.

Address:
http://www.recelectronics.demon.co.uk/

NEC LNAs for 2,4 GHz

There are lots of integrated LNA ICs available but this application note is interesting. There is an optimised circuit for a special application (wireless video camera) with information on using the IC and information on prototype. The results of measurement, the circuit and a photographs of the PCBs are published without restriction and therefore are most interesting for the circuit developer.

Address:
http://www.valontechnology.com/images/NEC_NE34018_113012LNA_report.PDF

An 18 to 40GHz double balanced mixer MMIC

How do you build a passive mixer for 20 to 40GHz? This question is not only answered in this article with measurement results, but it also gives the reasons for certain technologies used and circuit details. Very informatively.

Address:
http://www.plextek.co.uk/papers/BB%20Pass%20Mixer%20Full%20Paper2.pdf

But this article is not the only interesting item, at the address:
http://www.plextek.com/technicalpapers.htm
there is a complete collection of technical papers from this company e.g. Oscillator Design, RF mixer Design, Synthesiser Design.

Mini Circuits Application Notes

The information from this company is always very high quality and the application notes are exceptional. Thye have all the information required including a special mixture from basics, characteristics and applications. A very good site with a nice clean look.

Address:
http://www.minicircuits.com/pages/app_notes.html

Hittite

The products from this company cannot be ignored. They are an RF and microwave specialist and supply an almost infinite number of interesting products. They have a lot of information on the Internet This site has many interesting items in the literature section such as a PLL calculator and a mixer track calculator, available under the heading “engineering tools”.

Address:
http://www.hittite.com/literature/featured-articles.html

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

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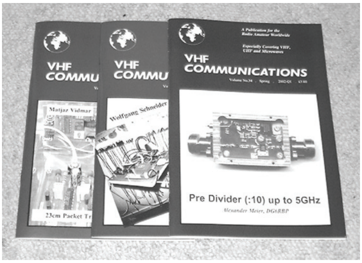

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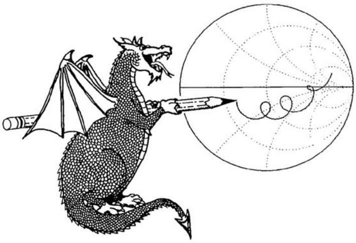

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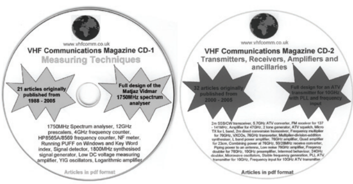

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