VHF
COMMUNICATIONS

76GHz Transverter

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Contents

Some very interesting articles in this issue especially if you are constructing equipment for the higher frequency bands.

The new RSGB book that I have edited,called Microwave Projects, is now available from the RSGB. There is an advert for it on the inside back cover of this magazine. I have already had several enquiries about buying copies, this one is only available from The RSGB but if you have difficulties please contact me

73s - Andy

Bernd Neubig, DK1AG

VCXOs with wide pull-in range using alternatives to quartz

Frequency generation using a quartz crystal oscillator offers a high degree of stability over a relatively wide range of temperature. However, alternative materials to quartz offer a wider pull-in range with similar stability, making it possible to create VCXOs with ten times the pull-in range.

1. Introduction

Quartz crystal oscillators are distinguished by a high degree of stability, but the pull-in range for VCXOs (Voltage Controlled Crystal Oscillators) is relatively narrow, typically ±50 ppm to ±150 ppm.

The author described the various circuit techniques that can be used to increase the pull-in range [1]. To do this, high-inductance coils must be used in series and/or parallel with the crystal. The higher the series inductance, the wider the pull-in range. But the quality, and therefore the stability, falls drastically, since the characteristics of the coil become more and more important in specifying the frequency.

Research has been going on into alternatives since quartz started to be used to generate frequencies, but no material has been found to date that can really replace quartz. In recent years great progress has been made with research and development projects involving piezo-electric crystals. These are almost as stable as quartz, but allow for wider pull-in ranges (and/or wider bandwidths in filters). Also much progress has been made into the industrial production of these crystals. Here are some of these alternative materials to quartz:

- berlinite (aluminium phosphate $AlPO_4$)

- gallium phosphate(V) ($GaPO_4$), together with the lanthanum compounds

- langasite (LGS, lanthanum gallium silicate $La_3Ga_5SiO_{14}$)

- langanite (LGN, lanthanum gallium niobate $La_3Ga_5NbO_{14}$) and langatate (LGT, lanthanum gallium tantalate $La_3Ga_5TaO_{14}$).

Some of the materials listed are nowadays commercially available as synthetically pulled crystals (Fig 1) [2], [3].

Resonators made from these crystals have a considerably higher pulling capability than quartz with an identical or similar crystal structure. Thus voltage-controlled oscillators can be manufactured which have ten times the pull-in range of a quartz VCXO, also broadband filters.

Fig 1: Gallium Phosphate crystal. Photgraph from AVL List [2].

2.
Basic principles

The pulling capability of a piezo-electrical resonator is determined by the so-called coupling factor, k. Its value is pre-set by the physical characteristics of the material, and it varies depending on the angle of cut. Table 1 shows the maximum value of k for various materials.

The equivalent circuit diagram of a piezo-electric resonator is shown in Fig 2. C_1, L_1 and R_1 form the frequency-determining dynamic branch and C_0 the static inherent capacitance.

The coupling factor, k, is related to the equivalent circuit components by the relationship:

$$k^2 \approx \frac{C_1}{C_0 + C_1} \approx \frac{1}{r}$$

Where: r = C_0/C_1 described as the capacity ratio.

For high frequency stability combined with greater pulling capability, the resonator disc must be cut out of the crystal at an angle at which the temperature coefficient of the frequency is as low as possible, and simultaneously the coupling factor is relatively high. The most

favourable situation is when a temperature coefficient, TK_f, of zero is present in the middle of the working temperature range.

In contrast to the working temperature cutting of quartz, the most favourable temperature coefficient curve for thickness shear oscillators using $GaPO_4$ and the lanthanum compounds LGS, LGN and LGT is a second-order parabola with the form:

$$\frac{\Delta f}{f_0} = b \cdot (T - T_0)^2$$

Where: T_0 is the arrest point temperature ($TK_f = 0$) and b the parabolic coefficient. The latter is negative, i.e. the parabola is "opened downwards".

An arrest point at $T_0 \approx 25°C$ can be attained using $GaPO_4$ with a cutting angle of $\theta \approx -16°$ and using LGS with a Y cut ($\theta = 0°$).

The parabolic coefficient, b, for $GaPO_4$ is approximately 2×10^{-8} K^{-2}, and for LGS approximately 6×10^{-8} K^{-2}. In practice, this means that for $GaPO_4$ a frequency stability of $\pm$ 20ppm can be attained in the temperature range from 20°C to +70°C and $\pm$ 40ppm in the range from 40°C to +85°C.

The possibilities for resonators and VCXOs using alternatives to quartz can be shown through two examples at 5MHz and 10MHz.

Table 1: Maximum values of coupling factor k for various materials.

Material	k
Quartz	7%
Langasit (LGS)	15.8%
Gallium Phosphate (GaPO$_4$)	19%
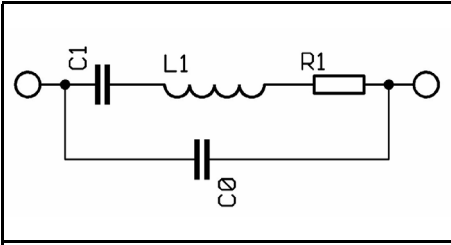
Fig 2: Equivalent circuit diagram of a piezo-electric resonator.

3.
Example 1: 10MHz resonators

Resonators were manufactured in the HC-52/U housing, using gallium phosphate (GaPO$_4$) with a Y-16° cut, and the following typical data were obtained:

- Dynamic capacitance $C_1 = 80fF$
- Static capacitance $C_0 = 4.3pF$
- Resonant impedance $Rr = 2.3\Omega$
- Quality $Q = 87,000$

In contrast, quartzes with the same frequency and size display the following typical values:

- Dynamic capacitance $C_1 = 10fF$
- Static capacitance $C_0 = 2.8pF$
- Resonant impedance $Rr = 14\Omega$
- Quality $Q = 114,000$

The capacity ratio is thus 5.3 times greater for the GaPO$_4$ resonator than for a quartz oscillator. In addition, the former has considerably lower impedance, though the quality is somewhat less.

The frequency-temperature characteristic of the GaPO$_4$ resonator is represented in Fig 3. The frequency stability is better than $\pm$ 20ppm / 20°C to +70°C, the typical stability for an oscillating quartz in the same temperature range being $\pm$ 10ppm.

The pulling characteristic measured with the GaPO$_4$ resonators, i.e. the frequency change plotted against the load capacitance, C_L , is shown in Fig 4. The curve demonstrates a possible pull-in range of 2,500ppm, without the use of coils in the pull-in circuit, which is 10 times the range of a normal quartz crystal VCXO! (Fig 4)

With an additional coil in series with the GaPO$_4$ resonator, a pull-in range of 5,000ppm or more can be obtained.

For comparison purposes, resonators made from langasite (LGS) were measured, which display the following properties:
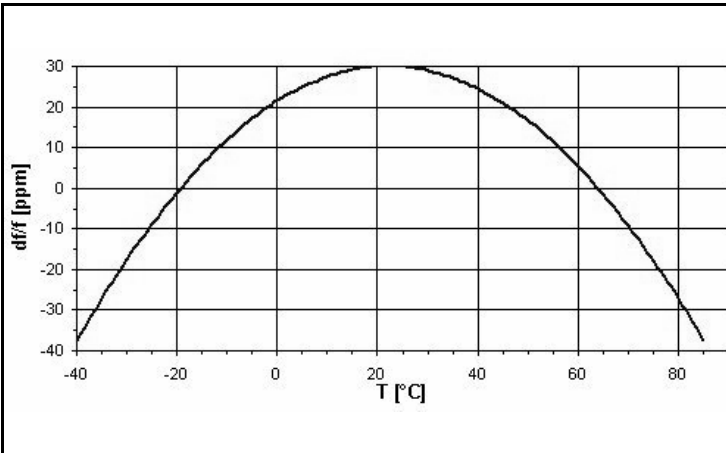
Fig 3: Temperature coefficient of a 10MHz GaPO$_4$ resonator.
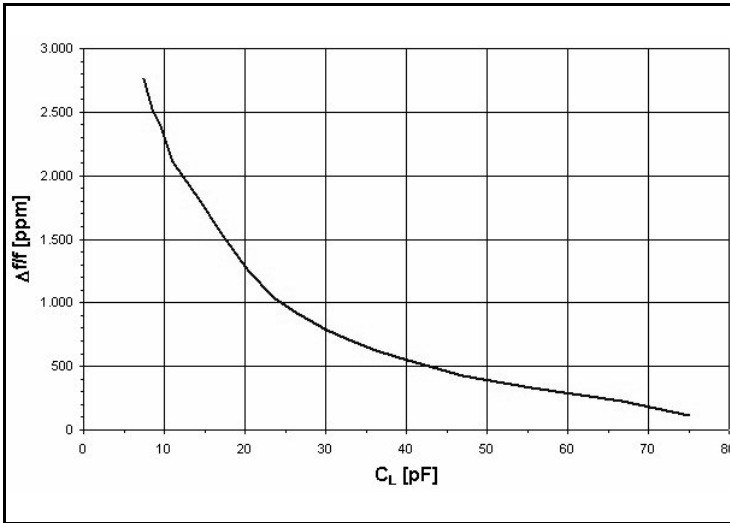
**Fig 4:
Frequency
change plotted
against load
capacitance for
a 10MHz
resonator.**

- Dynamic capacitance C_1 = 4 fF

- Static capacitance C_0 = 3.3pF

- Resonant impedance   Rr = 25Ω

- Quality   Q = 12,500

The capacity ratio of the LGS resonator is four times as high as that of the quartz crystal, but it is 32% lower than that of the $GaPO_4$. The resonators have a higher impedance and a markedly lower quality than the comparison materials.

4.
Example 2: 5MHz VCXO with 1300ppm pull-in range

Below 10MHz, quartz crystals in a small housing such as the HC-52/U are more difficult (and thus more expensive) to manufacture, since the quartz resonators must be formed as convex discs in order to obtain sufficient quality and stability. However, this is associated with a reduction in the dynamic capacitance, C_1, and thus with a reduced pulling capability.

Alternative materials to quartz such as

$GaPO_4$ have a thinner resonator disc at the same frequency, which can be manufactured with parallel faces and can thus deliver a more favourable capacity ratio.

The following equivalent circuit data was obtained using $GaPO_4$ resonators in Y-16° section at 5MHz:

- Dynamic capacitance C_1 = 58fF

- Static capacitance C_0 = 2.8pF

- Resonant impedance   Rr = 6Ω

- Quality   Q = 90,000

The capacity ratio, r , amounts to approximately 48; with fundamental crystals of the same frequency and size, r has values approximately 7 or 8 times greater.

These resonators were used to manufacture VCXOs in a miniature SMD housing with dimensions of 9mm x 14mm. The frequency / voltage characteristic obtained with commercially available capacitance diodes and without coils is shown in Fig 5. For a pull-in voltage range of 0.25V to 4.75V, a pull-in range of approximately 1,300ppm was obtained.
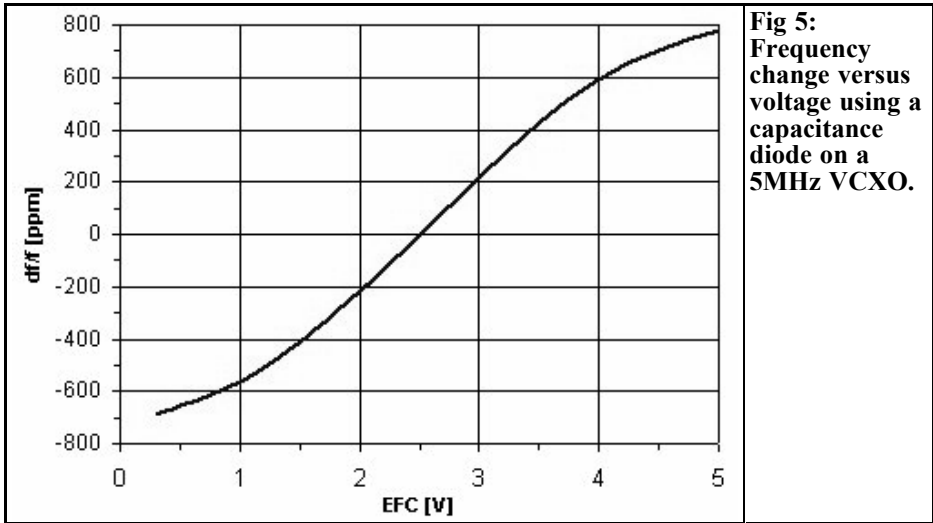
Fig 5: Frequency change versus voltage using a capacitance diode on a 5MHz VCXO.

5.
Summary

Piezo-electric crystals are now industrially produced that can be used to manufacture oscillators that exceed the limits of quartz crystal oscillators. Thus frequency stable VCXOs can be manufactured with pull-in ranges up to 10 times greater than those of quartz crystal oscillators.

Using such resonators and oscillators, made from alternative materials to quartz, opens up new possibilities both for professional applications and for amateurs wishing to experiment. Wide range VCXOs in SMD format are already commercially available today [4]. Resonators can also be obtained from the same source.

6.
Literature

[1] Neubig, B.: Design of crystal oscillator circuits, VHF Communications, issues 3 and 4/1979; downloadable from the authors website, www.qsl.net/dk1ag

[2] GaPO$_4$ website of AVL List web site www.gapo4.com

[3] Website of Roditi International Corp. www.roditi.de

[4] Website of AXTAL Advanced XTAL Products www.AXTAL.com: AXIS30 range.

Alexander Meier, DG6RBP

Digital mW meter

The revised version of a digital mW meter [1] is described below. Almost any diode detectors not needing a Townsend current can be used as sensors for this equipment. Depending on the detector, measurements up to 30GHz and beyond are possible. In addition to a measurement range between -45 and +20dBm, the equipment has additional interesting features such as an IEC bus interface.

1.
Functioning

The DC output voltage from a diode detector is dependent on the RF input power. Due to the non-linear diode characteristic, this DC voltage follows the square of the input voltage for low input power levels (< -20dBm), it is therefore proportional to the input power. From an input power level of approximately 10dBm, the output voltage, in contrast, is proportional to the input voltage. In between, there is a gradual transition from one range to another. This condition makes a great deal of correction necessary, but this is very simply achieved with the help of a micro-controller.

The digital mW meter first amplifies the initial DC voltage from the diode detec-tor. Four different amplification factors are available for this purpose. A switch-able low pass filter suppresses residual modulation (AM), together with 50Hz interference. The DC voltage signal is then digitised using a 12 bit A/D converter. In order to avoid earth loops (50Hz hum), the pre-amplifier is isolated from the rest of the equipment (micro-controller, interfaces, power supply).

A rapid, 8032 compatible type from DALLAS is used as the micro-controller. This compares the measured detector voltage with a look-up table in order to determine the measured RF power. In this table, values from the calibration of the detector with an output voltage going in 1dB steps from 45 to +20dBm are saved in the EEPROM. There is linear interpolation between the 1dB calibration points. The readings are displayed with a resolution of 0.1dB or 0.5dB at very low input levels (below approximately 40dBm). The measurements are started at adjustable intervals (1/2/5) of between 10ms and 20s. If this triggering is switched off, the equipment starts a new measurement as soon as the previous one ends. Approximately 1,000 measurements a second can be carried out with this equipment.

An analogue recorder output was integrated, together with an IEC bus interface (laboratory standard). [2] contains a simple converter between the IEC bus
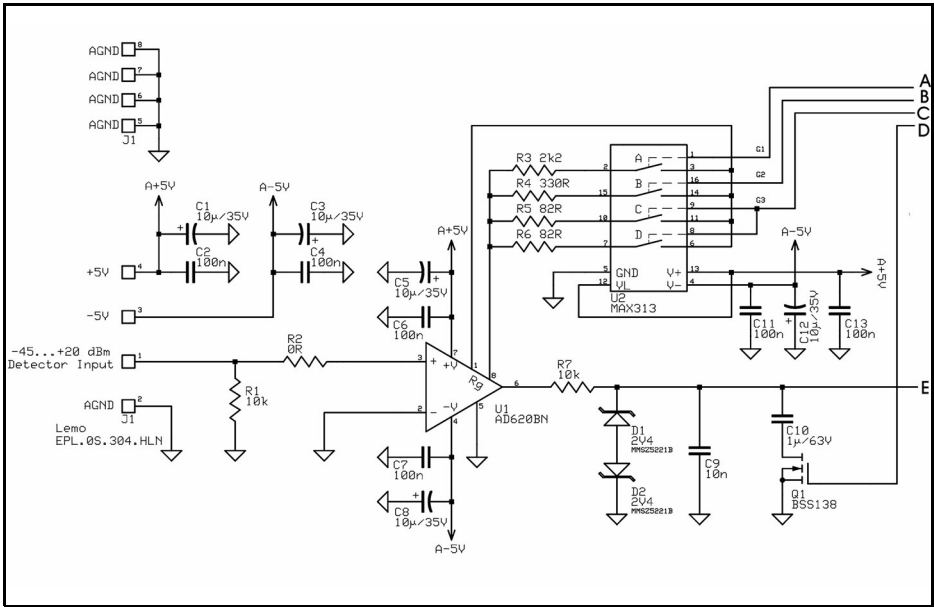
Fig 1: Input amplifier for the digital mW meter.

and the RS-232 serial interface.

However, the use of a micro-controller also brings some additional advantages to operating the equipment. It is possible to add an adjustable offset to the reading in order, for example, to take a directional coupler used in the measurement into account. Relative measurements using an adjustable reference level, or one which can be saved from current measurement, are possible. This also allows adjustment measurements to be carried out using a measuring bridge or a directional coupler. If the measured relative level corresponds to the return loss, the mW meter can use this to calculate and display the corresponding SWR.

An additional special feature is the switchable display from dBm and mW to dBμV and Volts (for 50Ω detectors).

The LCD display contrast can be digitally adjusted using keys on the front panel. The technical data from the equipment are summarised in Table 1.

2.
Circuit description

The complete circuit diagram of the digital RF mW meter is shown in the Fig 1, Fig 2 and Fig 3. The micro-controller unit is not shown as this is outside the scope of the circuit diagram. Those interested can email the publishers for a copy of the circuit diagram.

The DC output voltage from the detector is fed to the pre-amplifier through a 4 pin precision socket and given a load of 10kΩ (R1). A ±5V supply voltage is also connected to this socket so that extensions can be used (e.g. special sensors).

The pre-amplifier used is an AD620 instrument amplifier from Analog Devices. The amplification here is adjusted between PIN 1 and PIN 8 here using a resistor, with a high level of amplification requiring a small resistor.

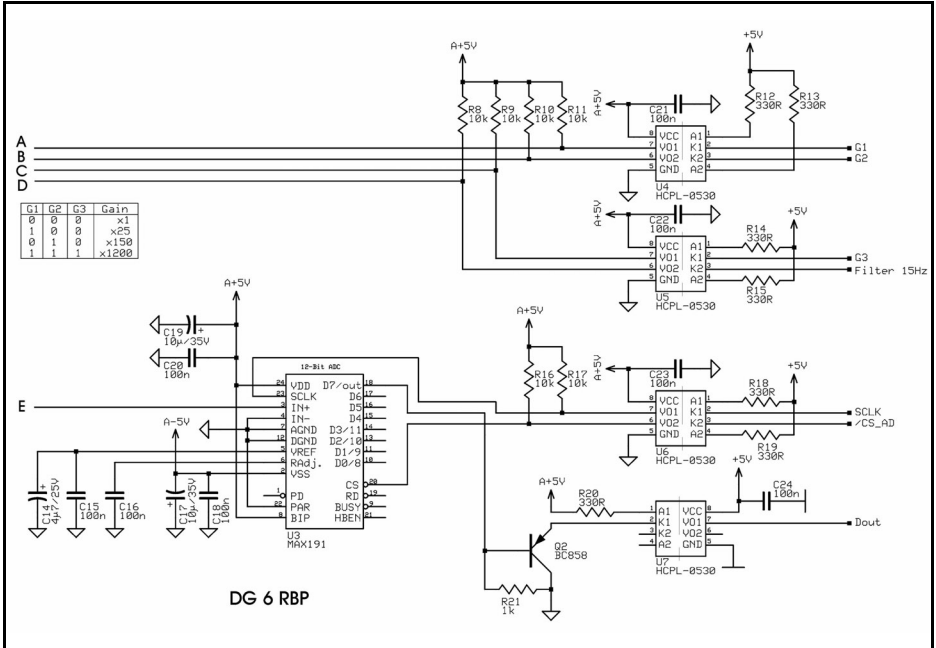Using the U2 analogue switch, the ampli-

Fig 2: A/D converter for the digital mW meter.

fication can be switched between approximately 1200/150/25 and 1.

The AD620s offset voltage, even for an amplification of 1,200 is only a few mV at the output, so that it is sufficient to offset this from the measured voltage through the micro-controller.

The output voltage from the U1 instrument amplifier is compressed by means of R7, D1 and D2 and fed to the analogue/digital converter. In spite of the characteristics of the detector diode, a large input power range can be covered without switching the amplification using the A/D converter.

At two or less measurements per second, a 15Hz RC low pass (R7, C10) is automatically connected, with the help of the MOS-FET Q1. This suppresses any possible 50Hz interference.

The analogue/digital converter (U3) used is a MAX 191 from MAXIM. This is wired via PIN 8 for bipolar operation.

The input voltage level from 0 to ±2048 mV is digitised to 11 bit accuracy (resolution 1mV). The twelfth bit signifies the polarity. Thus diode detectors with both positive and negative output voltages can be used.

The opto-couplers, U4 to U7, provide isolation between the pre-amplifier and the A/D converter from the rest of the digital section. This is absolutely necessary, since otherwise undesirable earth loops may occur (e.g. via an earthed test object and an earthed piece of equipment at the recorder output). The isolated earths are connected to one another using a high value resistor R39.

The recorder output is realised using the 12-bit digital/analogue converter (U8). The output voltage of this converter covers a range between 0 and 4,096mV in steps of 1mV. The software can be used to scale the relationship between the output voltage and the measurement re-
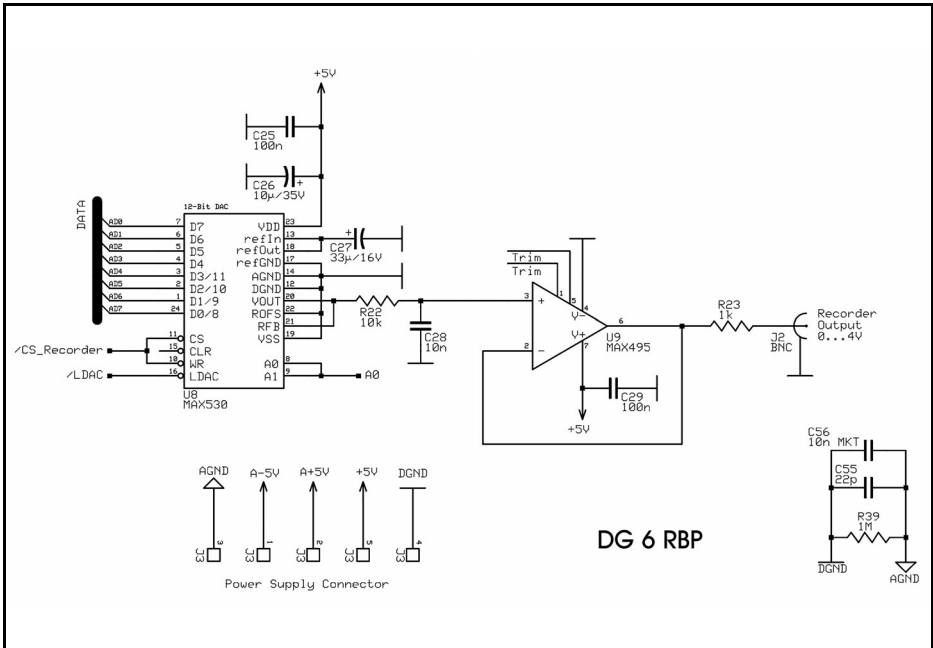
Fig 3: Recorder output for the digital mW meter.

sult. After a buffer and a 1kΩ series resistor, the voltage is available at the BNC socket on the rear panel of the equipment.

The control via the micro-controller is made up of the usual configuration: micro-controller, U12 address/data latch, U17 address decoder, an EPROM for the software and a RAM as memory. The calibration data and equipment settings are saved in the non-volatile EEPROM (U10), so that they are retained, even after the equipment is switched off. When the equipment is switched on again, the data is copied into the faster U14 RAM from the EEPROM.

The second U18 micro-controller is needed only for interrogating the keys. This makes it possible to run the software better, since not only must a keystroke be recognised but the pressure dwell too. A recognised and evaluated keystroke is serially transferred to the U11 controller

at 4,807 Baud. The U18 keyboard controller already has a flash memory for its software integrated in the chip.

A two-line LCD display with background lighting is used for the display. The U15 digital potentiometer (10kΩ) is used to set the contrast.

The U19 latch serves to control the light-emitting diodes on the front panel. The resistors, R34 to R38, serve to limit the current for the LEDs.

The IEEE-488 interface consists merely of an IC and a quartz oscillator. The i7210 (U21) used is compatible with the earlier industry standard μPD7210 from NEC, which is no longer obtainable. Even the SN75160 / SN75162 driver modules, which are still difficult to obtain and are usually in SMD format are already integrated in the new i7210 IC.

The equipment address is written into the corresponding register of the i7210 when the mW meter is initialised. An 8MHz

Table 1: Technical data for the digital mW meter.

Usable power sensors	All diode detectors which need no Townsend current
Polarity	Positive or negative (reversible)
Frequency range	Determined by detector
Input impedance	Approximately 10 kΩ
Input socket	4-pole. Detector voltage, earth, ± 5V for special sensors
Stored calibration data	For a maximum of 7 detectors
Measurement range	- 45.0 to + 19.9dBm (or 62 to 127dBμV)
Measurement resolution	0.1dB (> -40.0dBm); 0.5dB (< -40.0dBm); Depending on detector!
Measurement interval	10ms to 20s in 1/2/5 steps Continuously repeating measurement (1,000 per second) Measurement can be halted by pressing key
Precision	Approximately 0.5dB (dependent on detector & cal.)
Filter	15Hz low pass filter (20dB / decade) automatic when 2 or fewer measurements per second)
Display range	- 99.9 to + 99.9dBm (or 7.1 to 206.9dBμV)
Display format	dBm and mW dBμV and Volts (referred to 50Ω) Relative to reference value Relative to reference value + SWR
Display resolution	0.1dB
Display speed LCD	4 measurements per second
Null balance	Automatic when key is pressed
Offset	Adjustable 99.9 to +99.99dB, switched on/off by key
Reference value	Storable from measurement in progress or adjustable from 99.9 to 99.9dBm
DC output	0 - 4 Volts corresponding to 99.9 to 99.9dBm 0 - 4 Volts corresponding to 99.9 to 99.9dB relative to reference value 0 - 0.8 Volts corresponding to 50.0 to + 30.0dBm
Interface	IEEE-499 (IEC bus) (SH1, AH1, T8, L4, SRO, RL2, PP0, DC1, DT1)
Power supply	+ 8 to 20 Volts DC, max. 500mA

Fig 4: Completed digital mW meter by DG6RBP.

quartz oscillator (U20) is used as the clock generator. However, the i7201 in the 40-pin housing has recently been discontinued. It is, though, still obtainable in the smaller SMD housing and can then be pinned into the 40-pin base with an adapter board.

A plug-in power pack which should deliver at least 500mA at 12V is used for the power supply. The diode D1 acts as reverse battery protection.

3.
Literature

[1] Digital HF power measurement equipment;
Alexander Meier, DG6RBP, Paper read at 45th VHF Congress, Weinheim 2000

[2] Jirmann, Jochen: IEC-Bus-Interface, VHF Reports, issue 3/ 1999

Sigurd Werner, DL9MFV

A simple concept for an efficient 76GHz transverter

Following the successful testing of the recently described 76GHz beacon [1], an interesting transverter module was developed by adding a W band mixer (CHM2179a from United Monolithic Semiconductors). The original project was expanded to include a second oscillator (for the RX signal) and an optional phase modulator (for the TX signal). If we put all this together, we have a simple concept for an efficient 76GHz transverter.

1.
Introduction

I described a frequency octupler in this magazine that generates a stable 76GHz signal from a 9.5GHz input frequency [1]. The 38GHz signal generated by the quadrupler (CHX2092a) is fed to the CHU 2277 doubler (with integrated amplifier). This chip has two output ports for 76GHz. The power of the main port is coupled into a WR12 waveguide and acts as a beacon output and/or as a TX port (approximately 12 - 14dBm). The signal supplied at the secondary port, which is approximately 9dBm, was terminated with a 50Ω load in the beacon project. This RF output from the secondary port is now used as the LO signal for a mixer.

2.
Concept of a 76GHz transverter

The circuit proposed uses almost all the components from the beacon project, and is expanded by a mixer, a second oscillator (for the RX signal) and, if applicable, by a phase modulator (for the TX signal).

The schematic diagram is shown in Fig 1. Two oscillators are used to produce transmit and receive signals. The OCXO in the transmitter branch has a frequency of approximately 99MHz (e.g. 99.0001MHz gives a theoretical final frequency of 76,032.0768MHz). If a modulated output is required instead of a CW signal, a phase modulator can be introduced. (A suitable, simple circuit is in preparation).

The signal passes through the frequency multiplier stages, and is finally extracted at the main port of the CHU2277 chip (doubler) and coupled into a WR12 waveguide (for details, see Fig 5).

The OCXO in the receiver branch is tuned to a frequency of approximately 98.8125MHz. The signal is multiplied as above and a frequency of 75,888 MHz is taken from the secondary port of the CHU 2277 chip (doubler), see detailed diagram in Fig 2. This signal is now fed to a mixer as an LO signal. A CHM2179 W band mixer from UMS was chosen.
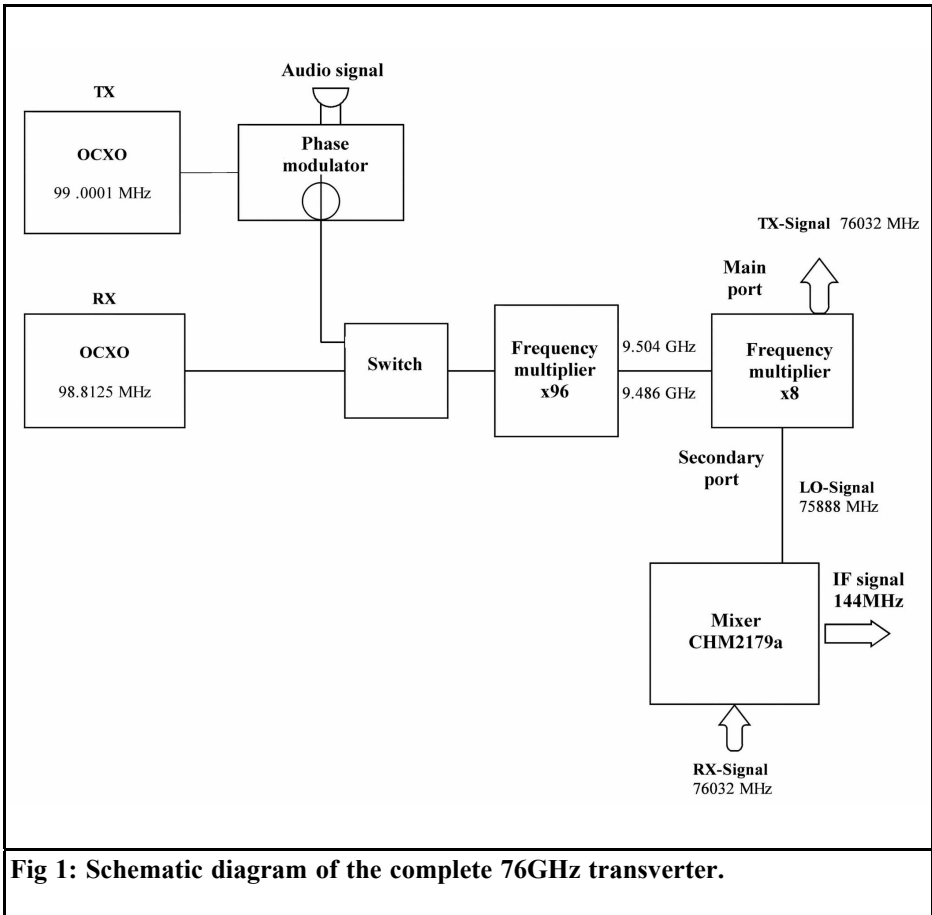
Fig 1: Schematic diagram of the complete 76GHz transverter.

This is manufactured in a rat race format on a GaAs substrate. The input signal is fed through a separate WR 12 waveguide to the corresponding input port of the mixer and the mixed product is extracted at the intermediate frequency port.

3.
Mechanical and electrical assembly

The housing (dimensions 38mm x 42mm x 13.5mm) was milled from brass and gold plated. The chips for the quadrupler

(CHX2092a) and the doubler were initially inserted into the cavity created (4.1mm. deep), with an integrated 4 stage amplifier (CHU2277), as described [1] (see Fig 3).

The signal supplied to the secondary port of the CHU2277 chip (f = 75,888MHz; approximately 6 - 9dBm) is fed in a wide arc to the mixer using a 50Ω microstrip (on an Al_2O_3 substrate) (see detailed photo in Fig. 4).

The mixer chip (CHM2179a) is mounted on a small milled out platform (0.15mm high), in order to compensate for the difference in heights with the aluminium oxide substrate. It was fixed with silver

Fig 2: Detailed schematic diagram of the mixer circuit.

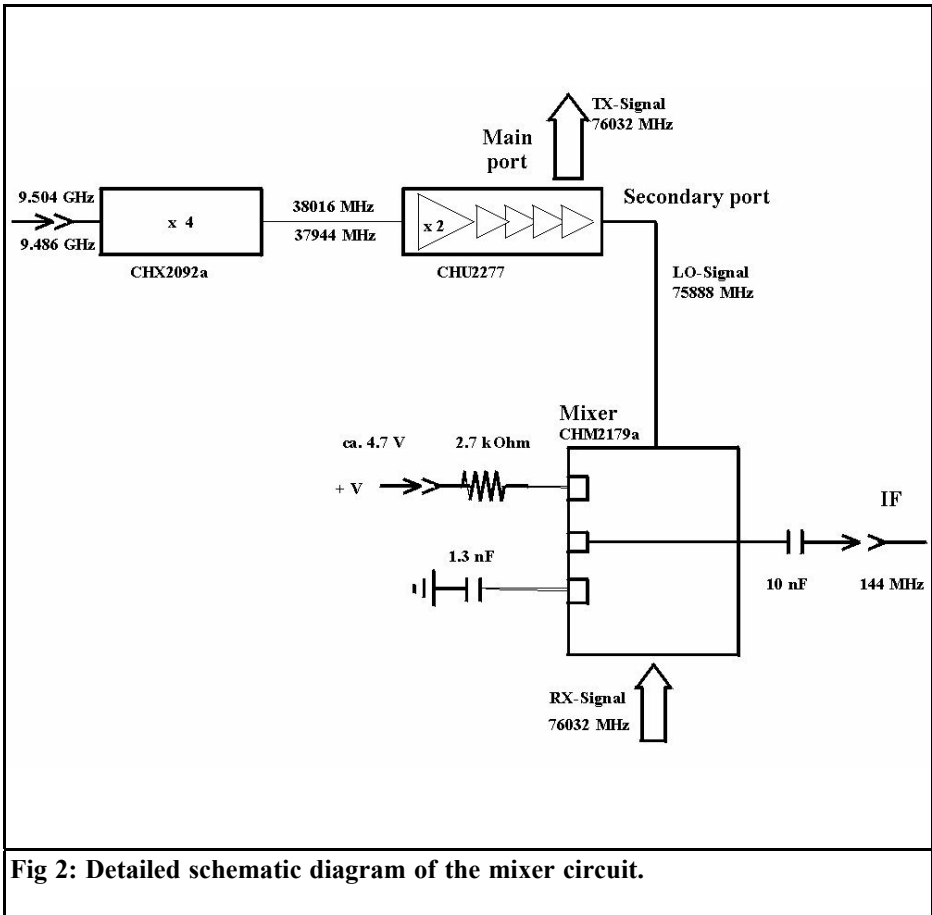conducting adhesive.

The received signal is de-coupled from the second WR12 wave guide through a short microstrip (like the transmit output, see Fig 5) and fed to the receive port of the mixer. All connections between chip and substrate were created using bonded 18μm gold thread [2].

The bias de-coupling capacitor required for the mixer was glued to the housing base as a 1.3nF single-layer type (on the left near the mixer chip in Fig 4) and was connected using bond wires.

The bias supply voltage is fed to the mixer through the housing base using a feedthrough capacitor (on the right at the top of Fig 4). A small SMD resistor (series 0402) of 2.7kΩ was also wired in series.

The intermediate frequency signal is connected to a 50Ω microstrip through two bond wires leading over the chip and decoupled from the DC voltage using a 10nF ceramic capacitor and output via an SMA jack on the housing (on the right in Fig 3).

The underside of the housing, with the feedthrough capacitors for the various supply voltages and the two WR12 waveguides for the transmit and/or receive signal are shown in Fig 6.
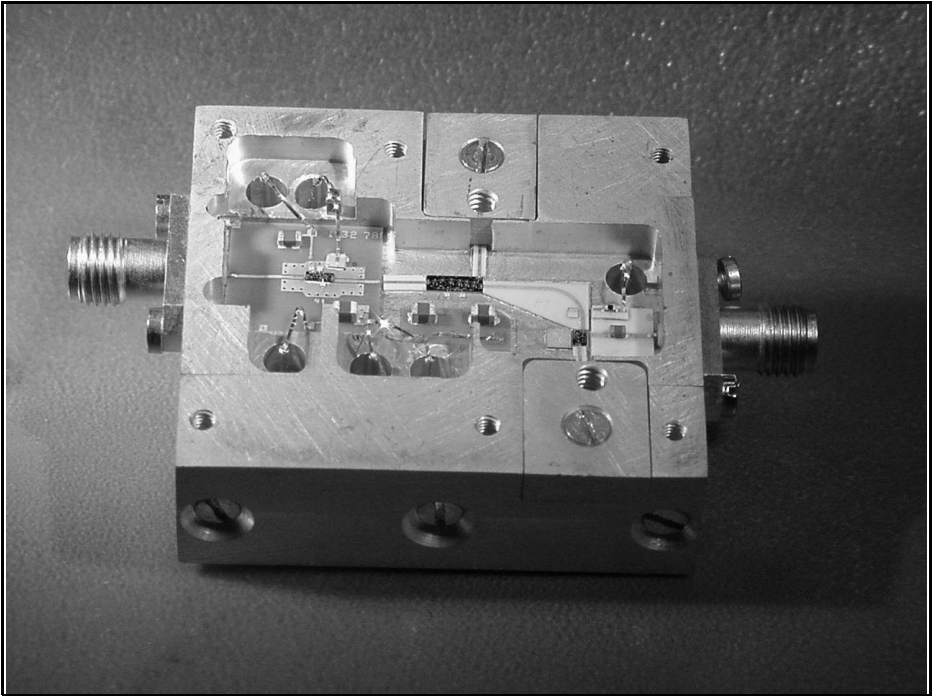
Fig 3: The completed transverter. On the left is the 9.5GHz input. On the right is the IF signal output. The three chips can be seen clearly.

4.
Results

4.1. Tuning

Experience with the 76GHz beacon has shown that some serious tuning errors are still being made, correcting these could provide considerably better results. This is true, in particular, for the transitions to the two output ports of the doubler chip (CHU2277) and the coupling of the microstrips into the waveguides.

The bond wires between the chips and the substrates were kept extremely short (approximately 150 - 200µm). As a further step towards reducing the inductance, two wires were parallel bonded whenever this was possible. The addi-

tional tuning required was achieved by fitting extremely small "gold flags" (0.4mm x 0.25mm; thickness 15µm) on the tracks. In a routine operation, the "flags" were pushed into position under a stereo microscope with a magnification of approximately 50. The result was monitored using a wattmeter and the metal was fixed on the track with a tiny drop of silver conducting adhesive.

An important factor here is that about half the adhesive is positioned on the gold layer on the track and the other half on the "flag". On no account must the adhesive flow under the metal, since when the components harden this will lead to additional resistive contact which will falsify the results.

In order to avoid the latter difficulties, I briefly tried out an alternative procedure. The "flags" were fixed with a small drop of a photo-activated adhesive based on
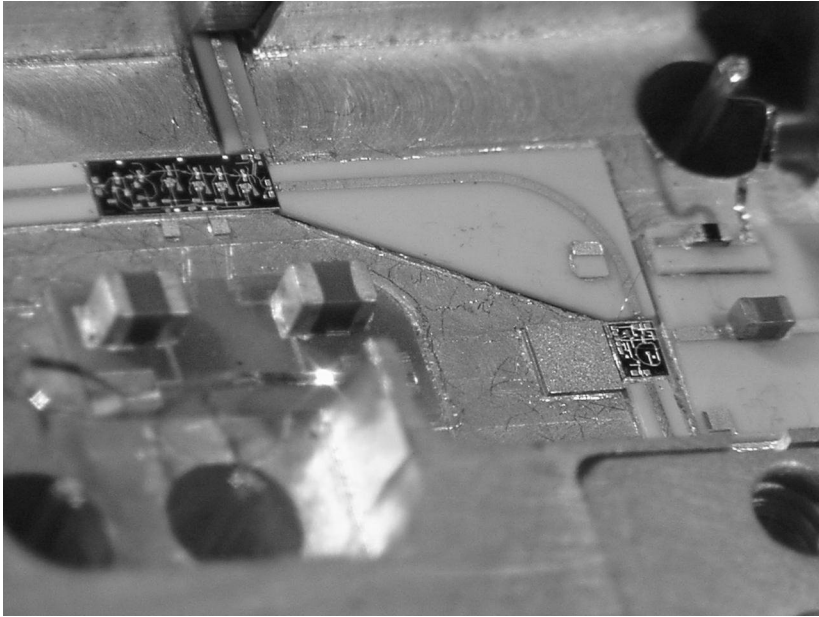
Fig 4: 2: Detail view of the receive circuit. The curved microstrip connects the LO signal to the mixer. On the left of the mixer is the bias de-coupling capacitor and on the right is the bias supply. The IF signal is fed via a ceramic capacitor across to the connector in the housing on the right.

methacrylate (e.g. Vitralit 5603 from Panacol-Elosol GmbH). As soon as the final position of the "gold flag" is reached (no time limit), the adhesive is cured by means of UV radiation. The most elegant way of doing this is to use a UV dot source with a fibreglass line, as used by dentists.

4.2. The transverters behaviour

The power supply voltages for the frequency multiplier are identical to those required for the beacon project [1]. Using various samples, CW outputs of up to 14.5dBm (28.4mW!) could be attained at the TX waveguide output.

For a bias current of 1mA to flow, a voltage of 4.7V had to be applied.

To measure the conversion attenuation at the mixer, it was initially operated as a transmitter mixer (reversed), i.e. a defined intermediate frequency signal was fed in and simultaneously the output at the receive wave guide was measured. It was initially determined that the mixer operated in a wide DC frequency range of up to 200MHz (specification is DC - 100MHz data sheet [3]) without any restrictions. It could also be used up to approximately 450MHz with some slight limitations. For a feed frequency of 144MHz (range 0 to +2dBm) a mixer attenuation of at least 7dB was obtained. The output power here amounted to approximately 350 - 400μW.

Since the precise measurement of the transverters noise factor was not possible
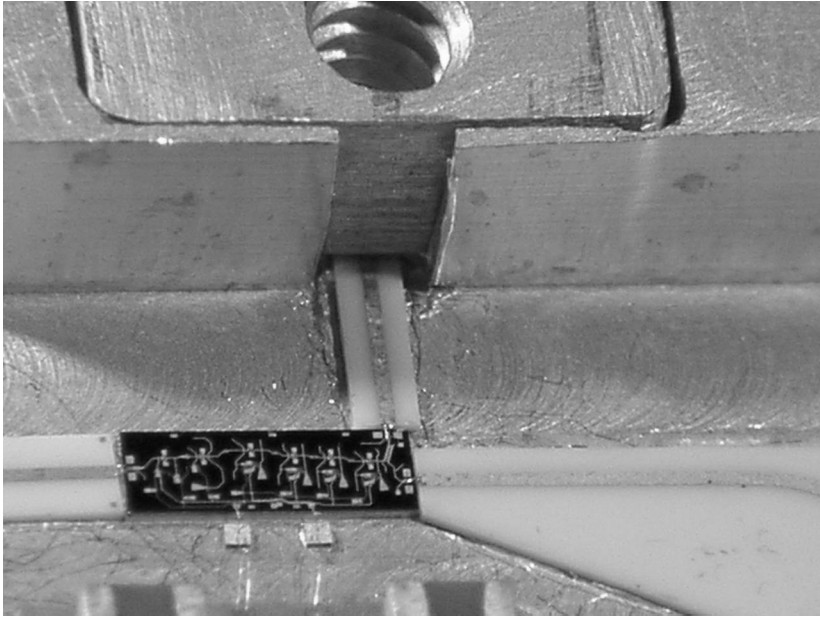
Fig 5: Detailed view of the microstrip coupling to the CHU2277 chip showing the main and secondary ports.

in the short term, a rig was assembled to estimate the reception sensitivity for a 76GHz beacon signal. Using a chain of attenuators (a combination of various variable and fixed waveguide attenuators), the signal strength was defined as being reduced to values < -100dBm.

A tip: to avoid cross-talk on wave guide/ flange connections, make sure the connections are neat and precise for any attenuation exceeding 80dB!

The strongly attenuated signal at 76GHz was fed to the receiver input of the transverter, and the intermediate frequency product generated was extracted, fed through a 2m pre-amplifier and finally displayed through a spectrum analyser (HP 8565 A). (noise factors of pre-amplifier and analyser added together approximately 3 dB).

The incoming signal was measured using

a bandwidth of 100kHz; the noise floor for the measurement rig was -120dBm. The 76GHz signal could be reduced all the way to 10dB above the noise. That corresponds to an equipment sensitivity of approximately -110dBm. The values determined using this assembly demonstrate that the receive section of the transverter has a high degree of sensitivity, and make it possible to estimate a noise factor of approximately 10dB (or better).

4.3. Outlook and acknowledgements

The 9.5GHz transverter module has just three MMICs, that are in a matchbox sized housing. A special feature of the circuit is that separate waveguide ports are used for transmit and receive. Since
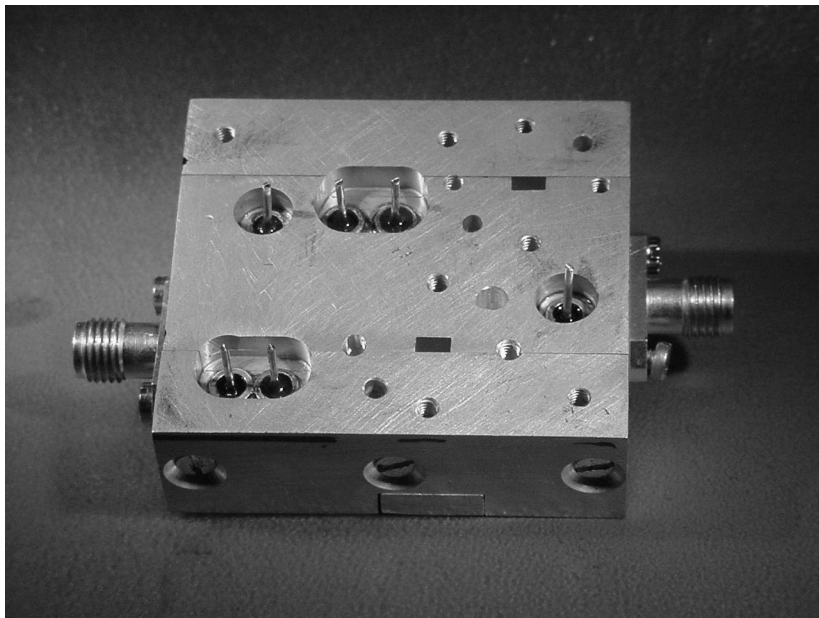
Fig 6: A back view of the completed unit showing the WR12 ports and supply feeds

the mixer is used only in receive mode, it could be "blocked" during transmission by switching off the bias voltage.

Naturally, the mixer can be used as a receive or a transmit converter. But in transmit mode we can expect an output of approximately 400µW, i.e. approximately 200µW per sideband. In this case a transmit amplifier must be used. Work is being done on just such a project.

Once again, I would like to thank the employees of Rohde & Schwarz (Münich) for their continuing support, particularly Dr. Jünemann (DK7AH) and most of all W. Hohenester. I am also much indebted to Jochen Ehrlich (DF3CK) for his help with the mechanical side of things, and with the determination of receiver sensitivity and, something particularly important for me, for always being willing to talk things over. My thanks to Rudi Wakolbinger

(OE5VRL) for producing the oscillators so promptly.

5.
Literature references

[1] Sigurd Werner, DL9MFV, Frequency multiplier for 76GHz with an integrated amplifier, VHF Communications 1/2003, pp. 35 - 41.

[2] Sigurd Werner, DL9MFV, Amplifier for 47GHz using chip technology, VHF Communications 3/2002, pp. 160 - 164.

[3] Data sheet from United Monolithic Semiconductors, S.A.S.; Ref DSCH21790192, 22nd June, 2000.

Dr Volker Grassmann, DF5AI

Transatlantic transmission on 144MHz

Ionospheric long distance contacts over distances exceeding 3,000 kilometres are not unknown in amateur VHF radio and thus give grounds for the conjecture that transatlantic VHF contacts can not be ruled out. To provide experimental proof of terrestrial transatlantic contacts in the VHF range, radio amateurs on both sides of the ocean have already erected transatlantic beacons. The transmitter beacons F5XAR (France) and VE1SMU (Canada) are used to determine the potential reception areas on each side of the Atlantic and the geographical position of the reflection areas in the E layer in the event of a "double hop" propagation (double reflection of radio waves in the ionosphere).

1. Introduction

We are still waiting for the sensational news but there has been no indisputable report of a 144MHz terrestrial transatlantic contact as yet. The report of a meteor scatter connection between Newfoundland and Ireland, said to have been achieved on 20th March 2002 [1], has so far been received with scepticism in the pages of amateur radio literature [2].

However, many VHF radio amateurs consider a radio bridge across the Atlantic as being overdue. In relation to this idea, people are pinning their main hopes on unusual tropospheric overshoots, or on multiple reflections between the Earths surface and the ionosphere.

2. Distance between continents

The most favourable geographical preconditions for terrestrial radio links across the Atlantic can be found in Brazil and/or on the West coast of Africa.

The radio path to be traversed between the Brazilian province of Rio Grande do Norte on the Eastern tip of South America and the West African states of Gambia, Senegal, Guinea-Bissau, Guinea, Sierra Leone and Liberia is less than 3,000 kilometres long (Fig 1). So VHF radio amateurs living in these places could secure victory in the "radio races" to cross the Atlantic.

In the North Atlantic, the shortest distance is that between Newfoundland and Ireland, the minimum gap being approximately 3,050 (Fig 2), which corresponds to the distance between Frankfurt and the Canary Islands (EA8). After all, the distance between Hawaii and California (approximately 3,800 kilometres) has already been bridged several times at
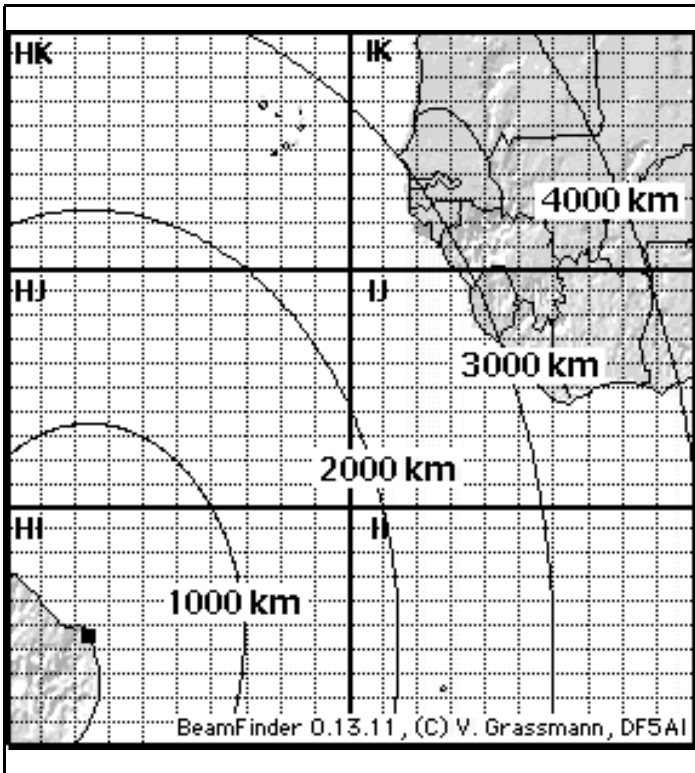
Fig 1: The shortest distance between South America and Africa is less than 3000 kilometres.

144MHz. However, comparably favourable tropospheric propagation conditions can scarcely be expected under North Atlantic climatic conditions.

3.
Results for transatlantic beacon F5XAR

The possibility of double reflection between the ground and the ionospheric E layer (referred to in what follows by the customary radio amateur term of "double hop") has recently been investigated, using the example of transatlantic beacon F5XAR in Brittany (Quimper, IN87KW, 144.405MHz). The beacon transmits a signal with a transmission power of 400 Watts ERP in the direction 290° azimuth

[3]. With an assumed reflection height of 105 kilometres, potential reception areas can be identified in Labrador, Newfoundland and Nova Scotia (see representation of regions in Fig 3). The grey areas in Fig 3 indicate the geographical position of the associated reflection areas in the ionosphere. It should be noted that the contours of these areas resemble the reception areas referred to above see, for example, the contours of Greenland which, due to the assumed double hop propagation, makes its appearance once over Ireland and again over Iceland.

The reception from beacon F5XAR in Newfoundland can be described through the following propagation route:

The signal from the beacon is reflected in the E layer in the South Western part of locator field "IO" and comes back down in the lower centre of the field "HO"
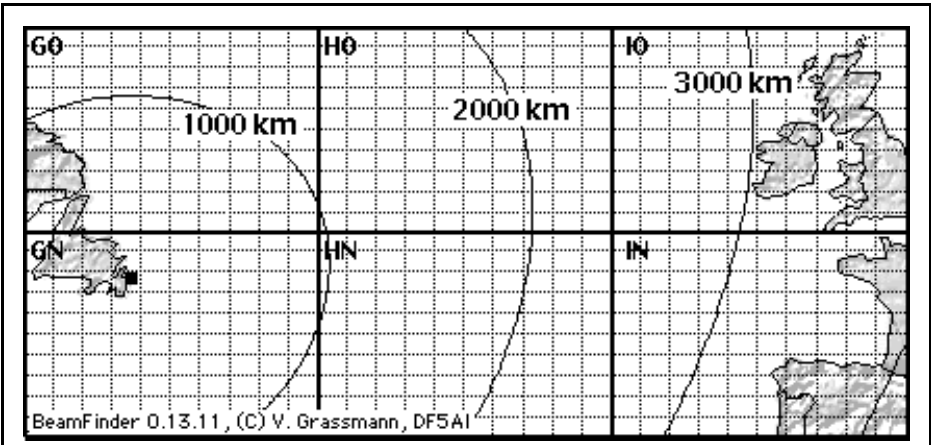
85

Fig 2: In the North Atlantic, Newfoundland and Ireland no more than 3050 kilometres apart.

(first "skip"). From the surface of the sea, the radio waves go back into the ionosphere and are reflected for the second time in the South Eastern part of the field "GO". This second "skip" comes back to earth in Newfoundland.

4.
Results for transatlantic beacon VE1SMU

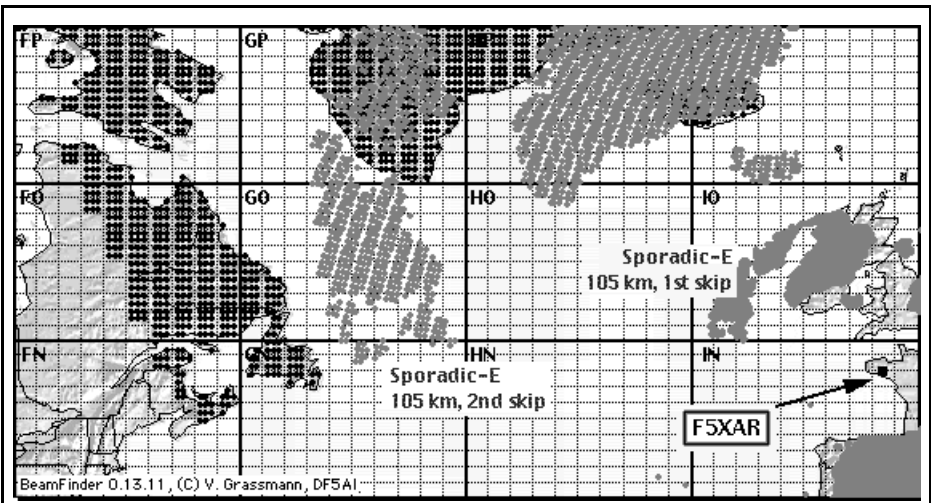The reception area calculated for transatlantic beacon VE1SMU (Halifax/Nova

Fig 3: Possible reception of the beacon F5XAR is Labrador, Newfoundland and parts of Nova Scotia (black). The grey areas show the reflection areas in the ionosphere (105km, "double hop" distance).
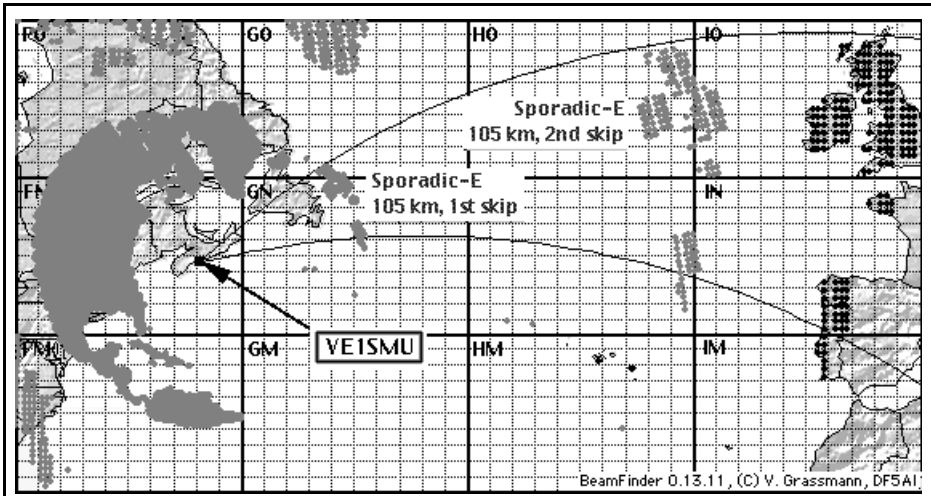
Fig 4: Possible reception of beacon VE1SMU is Portugal, Spain, France, Ireland and Great Britain (black). The reflections from the ionosphere are to the east and south of Newfoundland (first skip) and on the edges of locator squares HO and IO plus in the east the HN square (second skip).

Scotia, FN84CM, 144.300MHz) extends from the West Coast of Portugal and Spain as far as Ireland and Great Britain (Fig 4).

With a transmission power of 250 Watts, a 4 x 10 element antenna array and a direction of beam of 61° (see the apex angle drawn in Fig 4), the beacon represents an extremely interesting observation object for radio amateurs in the areas referred to. Since VE1SMU is transmitting on the international call frequency 144.300MHz, it would have been entirely possible for a chance observation to have occurred already. So are transatlantic 2m contacts in fact a vain hope? Or have any opportunities that have happened remained un-observed so far because the potential observers did not have their antennas aligned towards the Atlantic often enough?

Incidentally, the investigations also involved transatlantic beacon W1RJA (FN42CJ; 144.282MHz), but double hop propagation seems to exclude any reception of this beacon in Europe.

5.
Discussion

The assessments referred to above are from an investigation of the model of an ionospheric radio propagation (multiple reflection between the surface and the ionosphere) across the Atlantic. To this end, the potential reception areas for transatlantic beacon F5XAR in North America and the reception areas for the Canadian beacon VE1SMU in Europe were determined. The results make it clear that the observation of a transatlantic double hop propagation appears possible, and not merely from the locations in Newfoundland and Ireland.

It is hoped that more VHF radio amateurs will be encouraged to observe the transatlantic beacons on the basis of this assessment. By laying out the concepts of the reception areas identified in Figs 3 and 4, we may already have encouraged a considerable number of VHF radio amateurs who could contribute to regular observation of transatlantic radio paths.
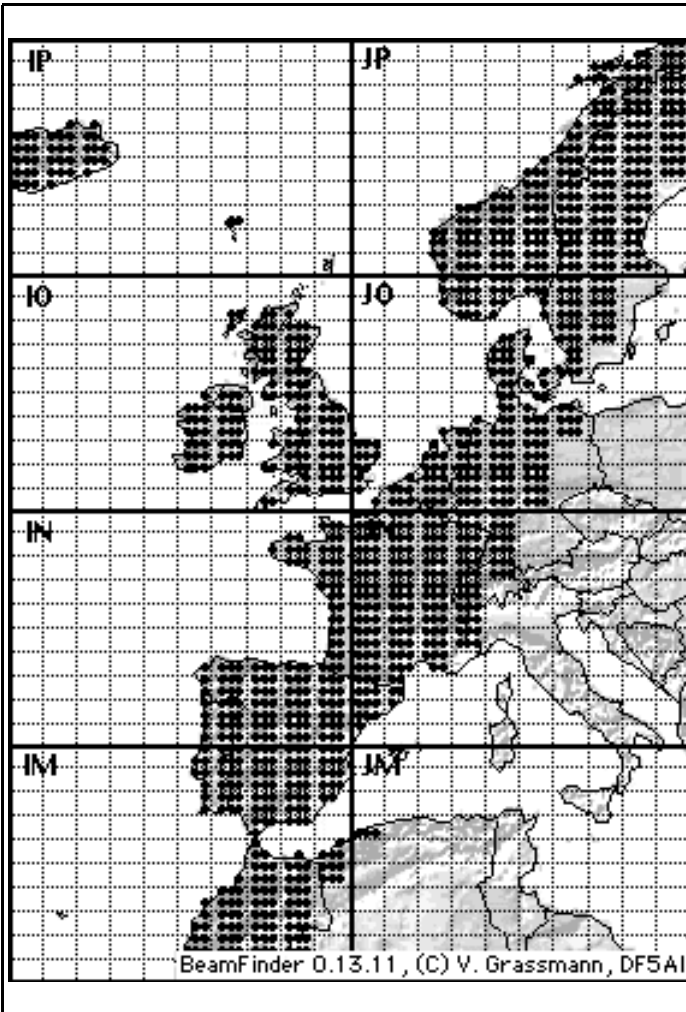
Fig 5: Possible reception areas for a transatlantic beacon at the most eastern point of Newfoundland.

If even one transatlantic beacon was available at the extreme Eastern tip of Newfoundland, even radio amateurs from Africa to Scandinavia would be called up to take an active part in the observations see Fig. 5

Sporadic E double hop contacts are not unknown on the 2m band, but they happen very rarely. But this may also be due to the geographical distribution of VHF radio amateurs. It is presumably comparatively improbable that there would be a chance discovery of a path

over 4,000 km by European radio amateurs because at the end of the transmission path, there are usually no receiving stations available! So the reports in the "144MHz Toplist" [4] appeared all the more remarkable, as twenty of the 214 European reporters had realised ionospheric long distance contacts from more than 3,000km. away (see column "Es" in [4]), which corresponds to an astonishing proportion of nine per cent.

Thus at the moment we can scarcely even judge whether bridging the Atlantic at

144MHz has failed to date for physical or statistical reasons. The discovery of a transatlantic 2m path is difficult for two reasons:

- For a transatlantic double hop connection, we must have suitable ionospheric conditions at the specified locations (see the reflection areas indicated in Figs 3 and 4). These two conditions considerably limit the observation probability.

- There are not enough observation stations available on either side of the Atlantic so far to guarantee any regular radio monitoring of the North Atlantic.

We are therefore seldom on the lookout but we are expecting the discovery of a rare phenomenon. Our situation perhaps resembles that of the observer of a storm, who opens his or her closed eyes only occasionally, for split seconds, and thus never catches sight of a flash of lightning.

The assessments shown in the diagrams were prepared using the BeamFinder analysis program. The software is available for downloading from the Internet [5].

6.
Literature references

[1] Atlantic overcome
http://www.tesnetwork.cz/ok2kkw/atlantikp.htm

[2] First 2m QSO across the Atlantic!?
Funktelegramm, 4/2002, S. 8

[3] Transatlantic beacon F5XAR
http://f6kpq.ifrance.com/f6kpq/f5xar.htm

[4] 144MHz Toplist
DUBUS 3/2002, Pp. 92 - 93

[5] Amateur Radio Propagation Studies
http://www.df5ai.net

Gunthard Kraus, DG8GB

An interesting program APLAC

1.

Brief description

Name: APLAC

Area of application:

APLAC is a universal program for ana-
lysing and synthesising every conceiv-
able circuit or system for telecommunica-
tions, electronics and physics in a spe-
cific time or frequency range. This in-
cludes EM simulators, using the FDTD
method.

Most important positive characteristics:

There is practically nothing that cannot
be investigated using APLAC.

Just to show the application options, the
parts list contains not only all the elec-
tronics components previously known,
but also models for mechanical oscilla-
tion systems and shock absorbers. And
the manual even provides an example of
the simulation of RF energy absorption
in the human skull when a telephone call
is made on a mobile phone.

What I liked:

All manuals are available for download-
ing at no cost and constitute complete
textbooks. All known microwave compo-
nents and models of irregularities (from
the microstrip bend through the interdig-

ital capacitor and length coupler through
to the circulator) are also released in the
free student version. An unbelievably
comprehensive library relating to func-
tions and options, not just in electronics,
but also in mathematics and physics.

A good, easy to operate editor for "draw-
ing" wiring diagrams.

For experienced users and if all the
effects playing a role are included, the
simulation results tally precisely with
reality.

Problem areas:

The program is a universal tool and is
thus not equipped with easy to operate
button control (like ANSOFT SER-
ENADE). It operates in its own "APLAC
language", which initially requires the
operator to set up several "control ob-
jects". These are then combined to form
a "simulation file" and executed. In plain
language this requires a lot of writing
and separate programming for the solu-
tion description, the sweep co-ordinates
and for displaying the results in dia-
grams, text files or tables. In principle,
this all represents a mixture of PSPICE
and ARRL radio designer command
lines, together with elements of the pro-
gramming language "C".

File type and file size:

A download of approximately 45MB is
required for all program sections and

manuals. Following installation, that gives a Windows file of approximately 65MB.

No problems could be detected during operation using WIN 95/98/ME/ 2000/ NT or fast computers.

Bugs or serious miscalculations detected:

Only insignificant little details or idiosyncrasies (e.g. in the graphics structure or the print out), which could have been clarified through the "APLAC Internet hotline" using email. One pleasant thing is that any query to this hotline (whether it is from a student or from an industrial firm) is treated in an impartial and friendly manner, and is handled competently and free of charge.

Desirable improvements:

Intensive further development of what has already been started, to make inputting using wizards (small additional programs with graphics user guides) considerably easier.

Expansion of permissible simulation content in free student version (see also under "Limitations").

User friendliness:

As already pointed out, a very intensive familiarisation phase is required before the number of error messages following the start of a simulation is reduced. Perfect control is not achieved until some time has elapsed and then only with continuous usage. Nevertheless, the operator has more to do than for similar products.

Help:

Very comprehensive online PDF manuals in 7 volumes, each with approximately 1,800 pages. Many examples of circuit and EM simulation as files on CD, together with discussion in the manual. Very friendly support, free of charge, obtainable via Internet or email.

Limitations:

The free student version restricts the number of components to a maximum of 12, although resistors, capacitors and coils are not counted. In addition, the main memory in use during simulation must not exceed a specific size, or else the processing is interrupted.

One way out for experienced users in such a case is to request the transfer to a 45 day full test version free of charge.

Acquisition options:

Download student version from http://www.aplac.com or through "Internet Treasure Trove CDs" from VHF Communications.

2.
A brief overall view

Professor Marti Valtonen developed APLAC (Analysis Program for Linear Active Circuits) back in 1972 in its basic form. Since then, it has grown and flourished in the fertile soil of Helsinki University (and no doubt within the countrys NOKIA culture as well). As already stated, the basic concept clearly goes beyond the limits of the rival products, which mainly specialise in certain specific simulation areas and then look for business from customers with sophisticated operating systems.

APLAC, on the other hand, is designed to handle every question, and brings together everything conceivable or known in the world regarding simulation tools. This covers the PSPICE module for linear circuit simulation through the harmonic balanced simulator to non linear large signal analyses, from the communications system simulator through the regulation of noise sidebands in oscillators right through to the EM simulation of microstrip antennas or RF absorption in the human skull. It is only
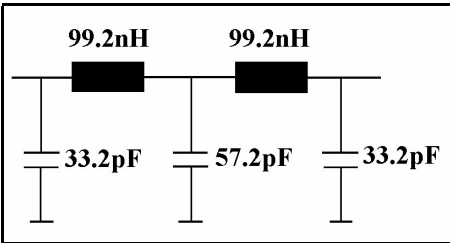
Fig 1: The results of a filter program for a 110MHz Chebyshev low pass filter.

after an intensive study of the comprehensive manuals that you discover everything you can do with this system.

Universality like this naturally has to be paid for, in this case with low user friendliness. This means that you have to assemble everything yourself and you have to tailor everything to the application in question. For this, you have to use the systems own "APLAC language" that you may well take some time to get into your head and which you'll probably never understand completely.

Setting up the necessary simulation files is similar to C programming. On the first run through there will be any number of

error messages, but with increasing experience they can be eradicated ever more rapidly. Once this hurdle has been overcome, not only can we enjoy the precise results, but also the numerous investigation and display options very quickly console us for the relatively laborious operation.

Now some simulations to explain the right way to use the software and simultaneously demonstrate its efficiency.

2.1. First simulation example: S Parameter of a 110MHz Chebyshev low pass filter

2.1.1.Setting up the circuit diagram

Before we can start up APLAC, the component values of the desired low pass filter are determined, using one of the free DOS filter programs from the Internet (e.g. "fds" or "faisyn20"). The following attributes should apply here:
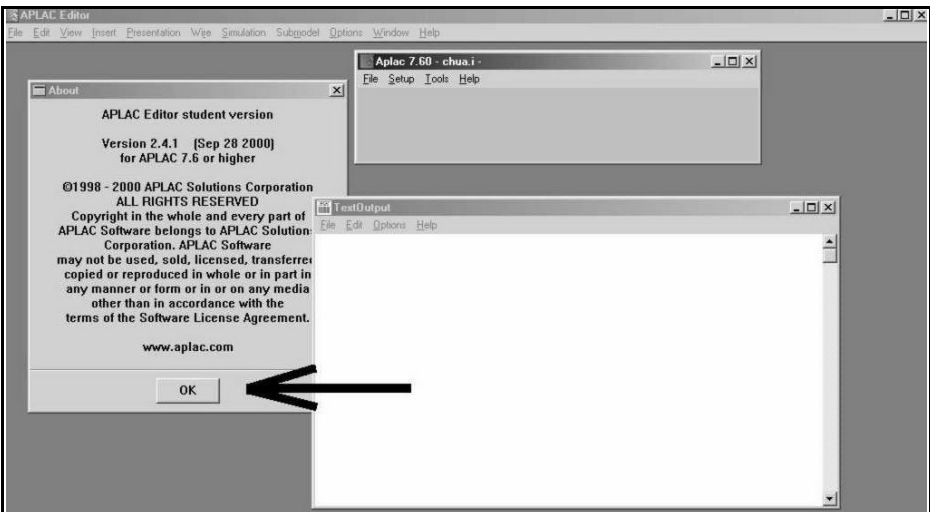

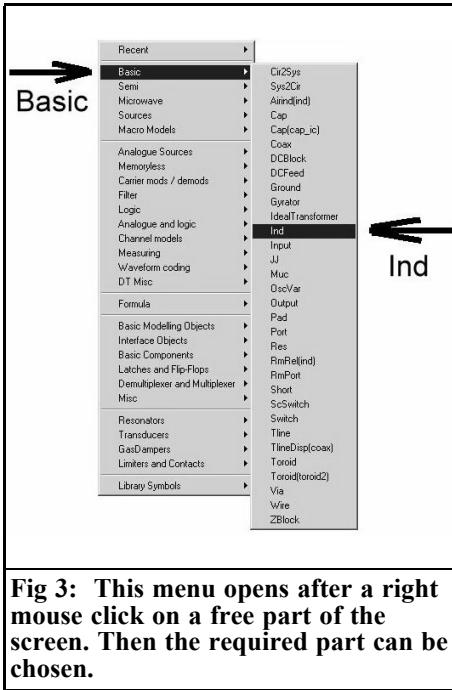**Fig 2: After starting APLAC press OK to start the editor.**

Fig 3: This menu opens after a right mouse click on a free part of the screen. Then the required part can be chosen.

- Ripple = 0.1dB.

- Ripple limiting frequency, fg = 110MHz

- Low inductance Implementation

- Degree of filtration, n = 5

- System resistance, Z = 50Ω

The result can be seen in Fig 1.

Now we start up APLAC, the screen is shown in Fig 2. This shows how the circuit diagram editor is started up. Use the white drawing surface that appears and make it full screen. We then right click the mouse somewhere in the free area of the screen. A new menu appears and we select "Basic". This calls up the list of basic components, and we can select an inductance, using the term "Ind" (Fig 3). Following confirmation, using the <ENTER> key, the coil is suspended from the cursor and is deposited in the wiring diagram. Repeat the sequence and you will have the two colis

required. Any component used is also listed under "BASIC / RECENT" and can be called up from there.

Now we call up the three capacitors under the term "Cap". However, each time they are placed they must be rotated through 90° using the key combination <CONTROL> + <R>. We also need three earth symbols, in the basic parts list the earth symbol is shown as "Ground".

Finally we double click on the graphic for each component and enter its value in the "Attributes" field. Do not forget to make the numbering of the component and its attributes visible using "Show Name" and "Show all Attributes". When a value is entered always use decimal points (American notation) and not commas!

And now for the wiring, just double left click anywhere in the free area of the wiring diagram, and a "wire coil" will suddenly be suspended from the cursor. Click on the connection of a component and the wire then uncoils to the next connection desired.

The connections here may, and should, run in curving, oblique lines over the circuit. Just left click again, and the editor itself will automatically turn the wiring into very neat right angled shapes!

For input and output, right click to call up a port via "Basic" and deposit the symbol on the input side. Then repeat on the other side for the output port. If you connect the two ports to the circuit and to earth, then you should have a wiring diagram corresponding to Fig 4. (Please always make all attributes visible after double clicking on a symbol). Looking carefully you will notice that the system impedance is pre-set to 50Ω for both ports.

To see the details better you should zoom in on the circuit to make it full screen. To do this, go into the "View" menu and select "Zoom Whole Diagram". You have now finished drawing the circuit.
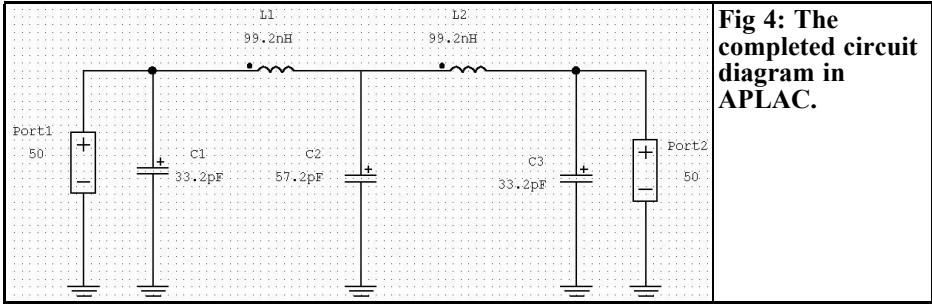
Fig 4: The completed circuit diagram in APLAC.

In addition to this, the simulated data for the low pass filter can also be written into an S parameter file in Touchstone format (with the ending "*.s2p"). To do this, click on the output port symbol and enter the corresponding requirement

STORE "LPF110MHz.s2p" GHZ MA

for the attributes (in a new line following the 50Ω system impedance).

2.1.2. Circuit simulation

First, open the "Object List Box" (Fig 5) and double click on it. In the menu which opens, first select the setting "Isweep" and then remove it with "Delete", because this is an "interactive Windows sweep", that should not be used here. Replace this by a new sweep type object and type the following settings:

```
"110MHz_low_pass"
LOOP 1001 FREQ LIN 100KHZ 0.3GHZ
WINDOW=0    Y   ""   "dB"    -50    0

...
SHOW W=0    DB    S(1,1)
+           DB    S(2,1)
```

Explanation:

1st line:

The term "110MHz_low_pass" is automatically used as the name of the project for each diagram and results print out and is included on the print out

2nd line:

Carry out a simulation in the frequency range and divide the range from 100kHz to 3GHz linearly into 1001 simulation steps.

3rd line:

Enter "Window=0" for the results print out set up for a plot diagram with the vertical axis numbered from -50 to 0dB.

Note:

The blank line separates the simulation section from the print out section and should therefore never be forgotten!

5th line:

Show the curves for S11 and S22 in dB in the window number zero.

Now re-open the "Presentation" menu. The sweep file has to be made visible on the screen. To do this, click on "Show Control Object", select "Sweep" from the list and close with "OK".

Two little things should be mentioned before starting the simulation. First, you must save the current project under a suitable name in a suitable directory. Secondly, make a final visual check of the object box list, as the sweep file must always be right at the bottom of the list (otherwise you will receive an error message which is difficult to understand). Fig 6 shows what everything should look like.

If all settings are correct, start the simulator, using the key combination <Control + S>. The results can be zoomed to full
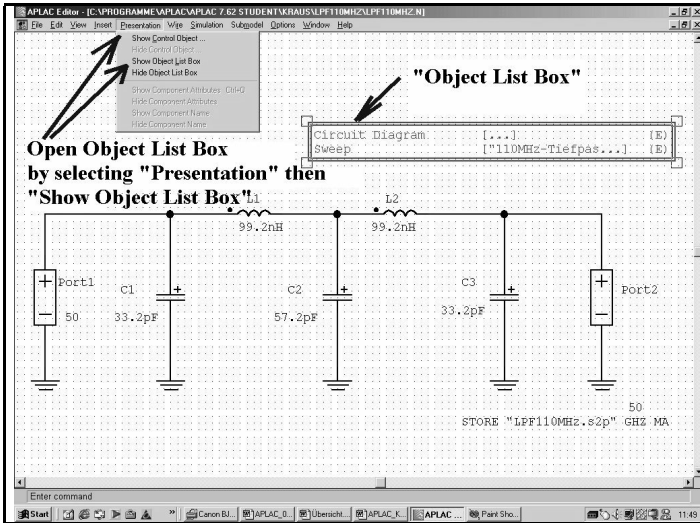
Fig 5: Opening the Object List Box to set up the required sweep object.

screen in the usual Windows maner as shown in Fig 7.

To see the Chebyshev ripples in S21 should move the cursor to the lowest value in the vertical scale, the cursor then has a small additional directional arrow. As an example, click on the value "-50 dB", a short menu opens and type "-0.4 dB". Enter "OK" and you will see Fig 8, thus demonstrating that the filter programs calculations are correct.

Anyone wanting more details should investigate the two menus "Scales" and "Options", there are several options for altering the results diagram.

2.2. Second simulation example: low noise MMIC amplifier

The full circuit diagram for a GPS
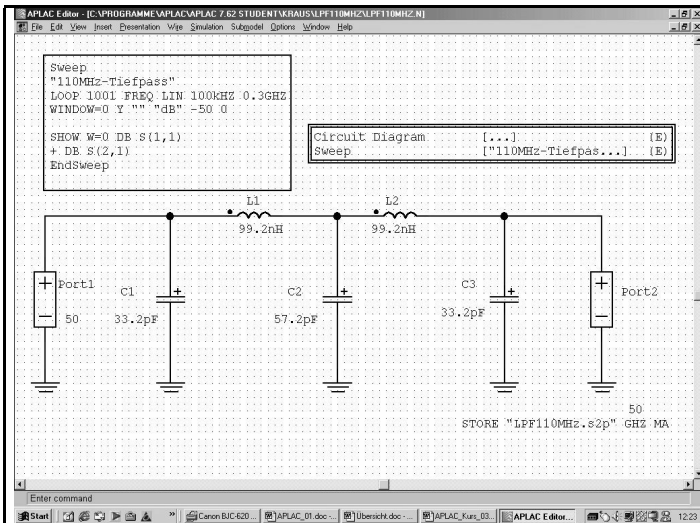
Fig 6: The completed settings for simulation of the 110MHz low pass filter.
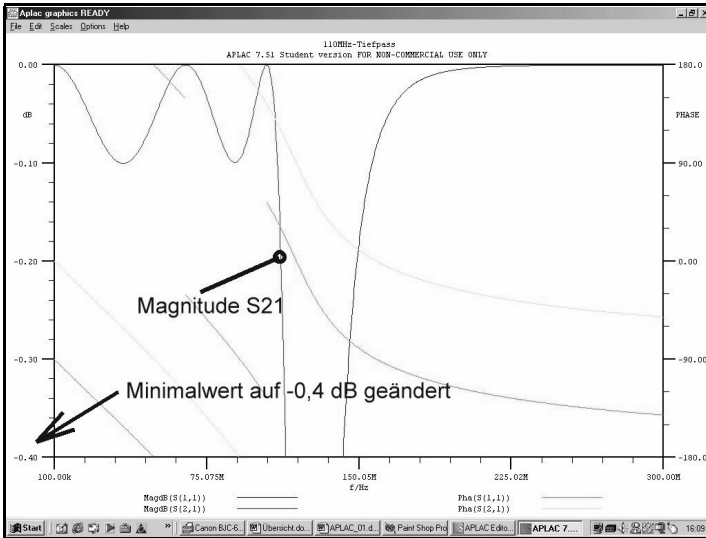
Fig 7: APLAC simulation results for the 110MHz low pass filter showing S11 and S22.

pre-amplifier (mean frequency: 1,575.42MHz) can be seen in Fig 9. The noise matching required has already been carried out by two microstrips at the input. The printed circuit board material is to be double sided Rogers 0.813mm thick (25 MIL). The underside forms a continuous earth surface, with the earth connections on the topside being connected by a sufficient number of feed-throughs.

Thus a brief APLAC task list for this simulation might be:

The behaviour of the circuit in the frequency range from 0.1 to 10GHz should be described through the S parameters and the curve for the noise factor, F (in dB). You should also check whether any stability problems can be expected with the circuit.
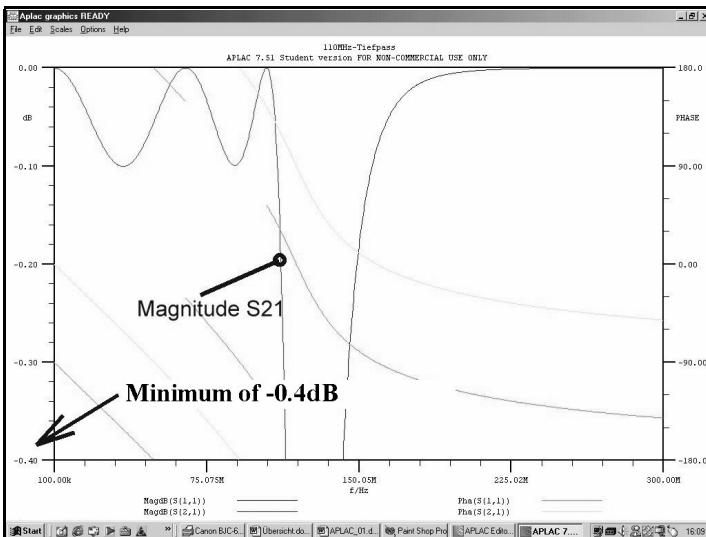
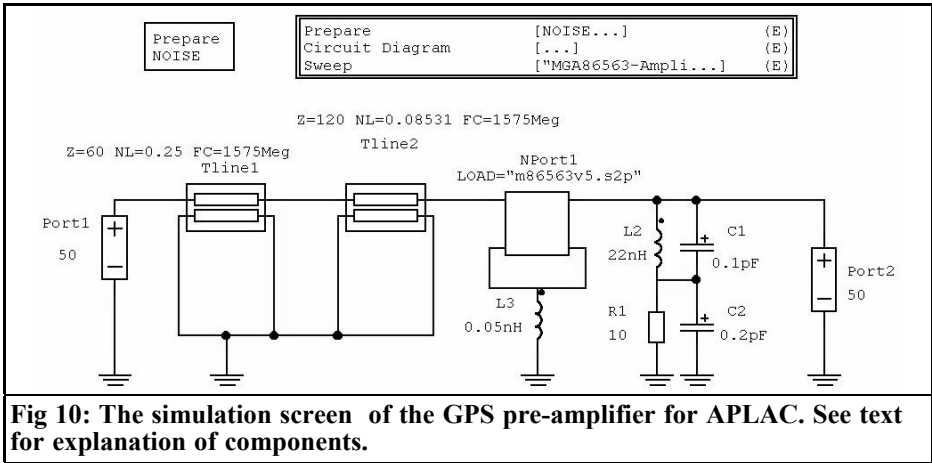**Fig 8: Repositioning the display clearly shows the Chebyshev ripples in the S21 curve.**
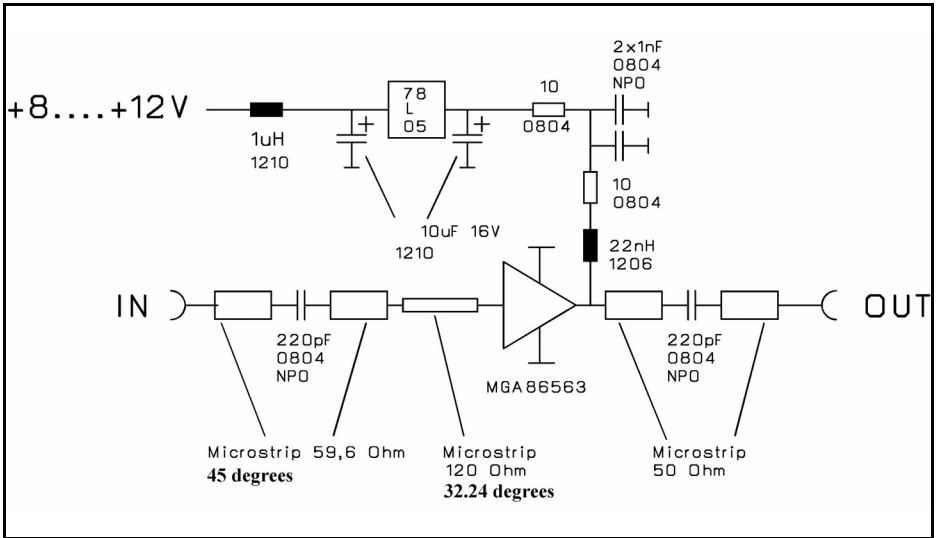
Fig 9: Circuit diagram of GPS pre-amplifier for simulation by APLAC.

2.2.1. Simulation screen

The screen required for the simulation can be seen in Fig 10, and some explanations are probably needed.

- The amplifier IC is represented by an "Nport" (Two_Port) from the "macro library" that can be accessed by right clicking on the screen area.

- The S parameter file taken from the Agilent companys homepage, which also includes the noise data, is connected, through its attributes using "LOAD=m86563v5.s2p". This file must be copied into the valid project file for our circuit.

- A "Prepare" object should be positioned in the object list box ("Prepare Noise"), which will prepare the noise simulation.

- The two microstrips used for the noise matching are first simulated as ideal lines which are "Tline" in the basic menu (right click

mouse). The impedance has to be entered in the attributes. The electrical length is in wavelengths of the frequency, FC=1,575MHz (for the first line, 90 degrees long, you need to enter "0.25").

- On the output side, the supply voltage is fed to the MMIC through a small SMD 22nH choke (size: 1,206) and a 10Ω SMD resistor. The self capacitance of the choke coil is also accounted for as 0.1pF (it can be calculated from the self resonant frequency specified in the data sheet) also 0.2pF for the capacitance of the solder pad linking the resistor and coil.

- The two earth connections for the MMIC must be connected to the lower earth surface using as many feed-throughs as possible. The inductance of these feed-throughs is usually the reason for any tendency to oscillate and any instability in such amplifier circuits, it must therefore be kept

Fig 10: The simulation screen of the GPS pre-amplifier for APLAC. See text for explanation of components.

as low as possible! If we set a value of approximately 1nH per feed-through per 1mm of printed circuit board, it then follows that with 10 feed-throughs on each side, we can expect a total inductance of about 0.05nH. This inductance is shown as component L3 in the circuit, so that it can be used to monitor the stability.

• A 50Ω port is connected to the input and another to the output as terminations.

2.2.2. Sweep file to simulate S parameter and noise factor

This file (as always, as the last position in the Object Box List) must look like:

```
"MGA86563-Amplifier"
LOOP 1001 FREQ LIN 0.1GHZ 10GHZ
WINDOW=0   Y ""   "dB"   -40   0 GRID
WINDOW=1   Y ""   "dB"    0   40 GRID
WINDOW=0   Y ""   "dB"    0    4 GRID

SHOW W=0   Y   MagdB(S(1,1))
+          Y   MagdB(S(2,2))
+     W=1  Y   MagdB(S(2,1))
+     W=2  Y   NoiseFigure
+          Y   MinNoiseFigure Marker=1
```

Explanation:

In the first half (before the blank line) we have first the circuit name "MGA86563-Amplifier" (it will be shown in every result following the simulation). This is followed by the requirement to simulate a maximum of 1001 points in the frequency range from 100MHz to 10GHz (maximum value for free student version).

In the following three diagrams (Window = 0 to Window = 2) a grid is programmed for the vertical axis instead of dB divisions, in order to make the values easier to read off. The display range in the first diagram (for S11 and S22!) goes from 40 to 0dB, whereas a range from 0 to + 40dB makes more sense in Window 1 for S21. A scale from 0 to 4dB is the best choice for the noise factor. Now (never forget this!) a blank line is required.

The subsequent display notes for the diagrams are probably self explanatory. Window 2 is programmed to display both the noise factor actually obtained and the theoretical minimum possible (NFopt). Finally, press <Control + S> on the keyboard to start the simulation.
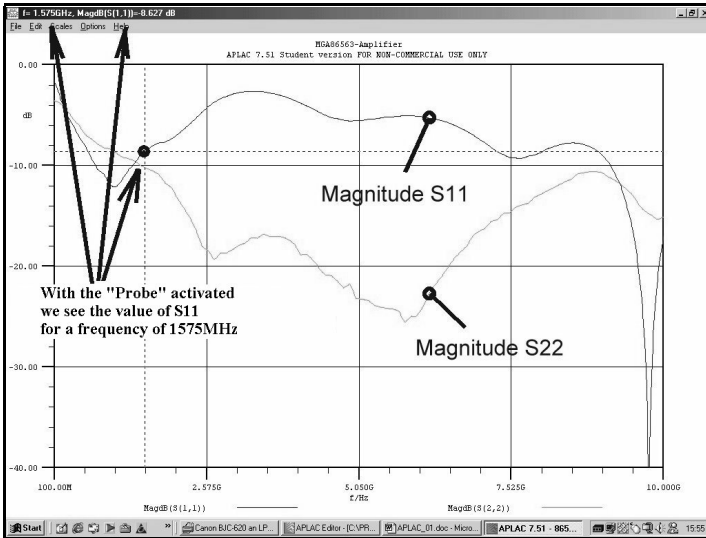
**Fig 11:
Simulation results
for the GPS
pre-amplifier
showing S11 and
S22 up to 10GHz.**

2.2.3. First simulation results

Following some calculation time, the
results appear on the screen. Click to
select the desired window into the fore-
ground and zoom. The values for the
input reflection, S11, and the output
reflection, S22, are shown in Fig 11. The
curve for S22, in particular, gives no
cause for complaint, and only gets better
as the frequency becomes higher. S11, on
the other hand, is determined by the
noise matching circuit and must be ac-
cepted as it is. Under "Options", you will
find the option "Probe", using this you
can go to each point of a curve and the
associated value will be shown in the top
left hand corner of the screen.

In the S21 curve, shown in Fig 12, the
only cause for complaint might be that
the maximum amplification is at a higher
frequency than ideal. We get a very nice
result from Fig 13, where the noise
factor, between 1GHz and 2.5GHz, is
very close to the minimum value

2.2.4. Stability control

2.2.4.1. Basic requirements for stability

The best circuit will be of no use if it has
a tendency to uncontrollable self excita-
tion, i.e. if it oscillates after assembly. In
the specialist literature and in practice,
various checks are standard, which all
make use of the S parameters of the
circuit using specific guidelines.

The oldest and best known method will
now be examined in greater detail. It
operates using the stability factors "K"
and "B1"

These two variables are dependent only
on the S parameters of the two port
network used and not on the termina-
tions. They can be used to express the
condition for "absolute freedom from
oscillation under all conditions and load
and/or internal resistances" (uncondi-
tional stability) in the following manner:

K > 1 and B1 > zero
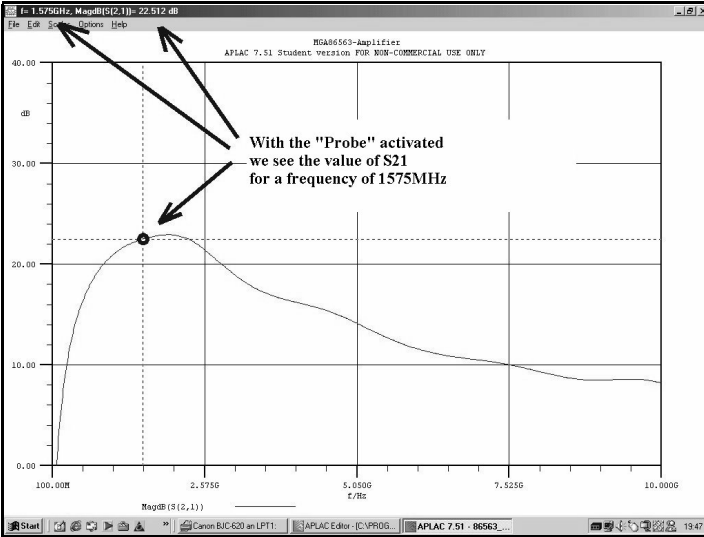
The two variables are calculated from the

99

Fig 12:
Simulation results for the GPS pre-amplifier showing S21 up to 10GHz.

S parameters in the following manner (but we naturally leave this to a program....):

$$K = \frac{1 - |S11|^2 - |S22|^2 + |\Delta|^2}{2|S12 \cdot S21|}$$

and

$$B1 = 1 + |S11|^2 - |S22|^2 - |\Delta|^2$$

where:

$$|\Delta| = |S11 \cdot S22 - S12 \cdot S21|$$

Note:

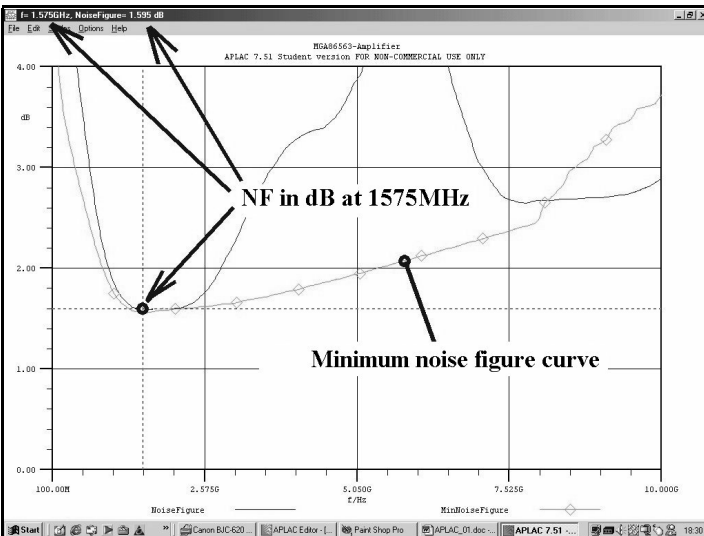APLAC no longer uses this technique, but instead makes a total of three more modern but very different control functions available:

Fig 13:
Simulation results for the GPS pre-amplifier showing noise figure.

First option:

Following the method just discussed (and perhaps also because many people are still so used to it), the following variants can be used by calling up the "S_K" and "S_D" functions:

The stability factor K must exceed 1 AND the S matrix determinant D = Δ must be less than 1, if the circuit is to be stable overall (unconditionally stable).

Second option:

We can work with only a single stability factor "μ" and exercise control using the function "S_u". Then we have:

The stability factor μ must exceed 1 if the circuit is to be "unconditionally stable" overall.

Third option:

This most modern solution would be the one to fulfil the "Nyquist criterion". It is described as an "Advanced stability analysis" (see APLAC Users Guide, Volume 2, pp. 3-26) and represents the "standardised determinants function" in the complex plane, using the option "NDF". It maintains that:

As soon as the origin ("zero point") of this complex plane is not embraced by the NDF curve, the circuit remains stable overall.

But for this an analysis must first be carried out in the time domain e.g. using an AC source because results cannot be obtained by just using the S parameters. So this option has been omitted.

2.2.4.2. Stability control using K and Δ and also using μ

Methods 1 and 2 are combined for the amplifier project. Naturally, a few changes are needed in the sweep file for this purpose:

- In the simulation section, we need a third window for the three stability control factors. The graduation of the scale should go from 0 to 4.

- In the display section, we arrange that "D", "K" and "μ" are not only displayed but are also marked differently.

This sweep file can be seen In Table 1 (the additions are marked in bold):

(Note: In this section of the sweep file, the change from the instruction "SHOW" to "DISPLAY" is deliberate. "DISPLAY" is the complete function for all types of representations that require more calculations to be carried out but also offer more options. "SHOW" is the rapid, slimmed down version with restricted capabilities, it is not adequate for the stability factors.

The result of this simulation is shown in Fig 14. In spite of the total feed-through inductance of 0.05nH transferred into the circuit as a precaution, the circuit remains stable over the entire frequency range (K and μ exceed 1overall, whereas D is less than 1 or 0.5).

2.2.4.3. Additional control using the stability circuits

The minute you get involved with circuits that have a tendency to oscillate only at specific internal resistances or loads (potentially unstable devices), the method described in the previous section, using the three factors K, μ and D, is no longer adequate. We know only that a tendency to oscillate can arise as soon as K or μ falls below 1 or D exceeds 1, but we do not know the critical conditions yet. More comprehensive analyses are needed here, which give a result in the form of two circles showing precisely these unstable areas of the associated reflection factors in the Smith diagram.

One circle shows how the stability, K,

Table 1: Modified sweep file.

```
"MGA86563-Amplifier"
LOOP 1001 FREQ LIN 0.1GHZ 10GHZ
WINDOW=0   Y ""    "dB"   -40    0 GRID
WINDOW=1   Y ""    "dB"    0    40 GRID
WINDOW=0   Y ""    "dB"    0     4 GRID
WINDOW=3   Y ""    ""      0     4 GRID

SHOW W=0   Y   MagdB(S(1,1))
+          Y   MagdB(S(2,2))
+      W=1 Y   MagdB(S(2,1))
+      W=2 Y   NoiseFigure
+          Y   MinNoiseFigure Marker=1
DISPLAY WINDOW 3 Y      "D"  S_D
+              Y "K"  S_K   Marker=1
+              Y "µ"  S_u   Marker=2
```

depends on the load (more correctly, on the associated reflection factor, ⌈LOAD), the associated technical expression is:

Stability K (Load Plane)

The other circle shows how the stability, K, depends on the internal resistance of the source (better: once again, on the associated reflection factor, ⌈SOURCE).

Here the technical expression is:

Stability K (Source Plane)

Yet both these "stability circles" are only valid for *a single frequency* and the simulation must be repeated if the frequency is changed.

Thus the circuit can often still be stable for a specific frequency if the internal resistance or load is known and constant, although the stability factors, K, µ and D are already warning of a potentially unstable situation.

A very simple rule applies for the evaluation of the circles:

• If both circles (for ⌈SOURCE or ⌈LOAD) run completely outside the Smith diagram, the circuit is guaranteed to be unconditionally stable.

• It is likewise unconditionally stable if the Smith diagram (for ⌈SOURCE or ⌈LOAD) is completely surrounded (i.e. enclosed) by a stability circle. That sounds a little crazy, but the background is as follows. The centre of the stability circle then lies in the infinite and its circular
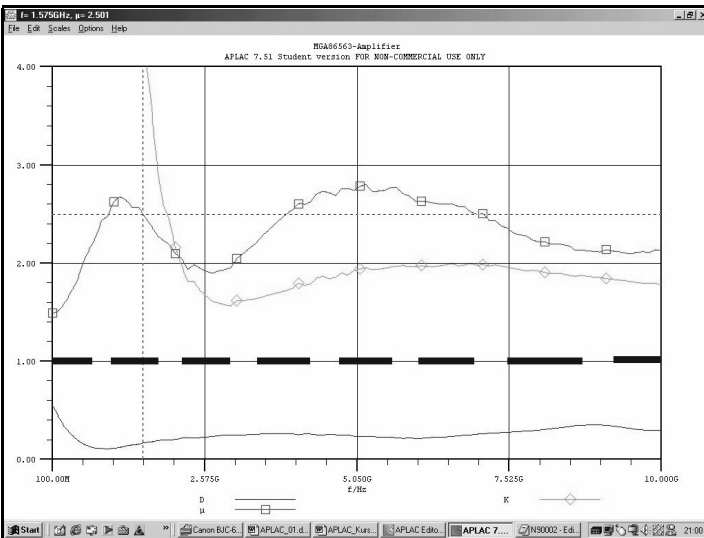
Fig 14: Simulation results for the GPS pre-amplifier showing stability.
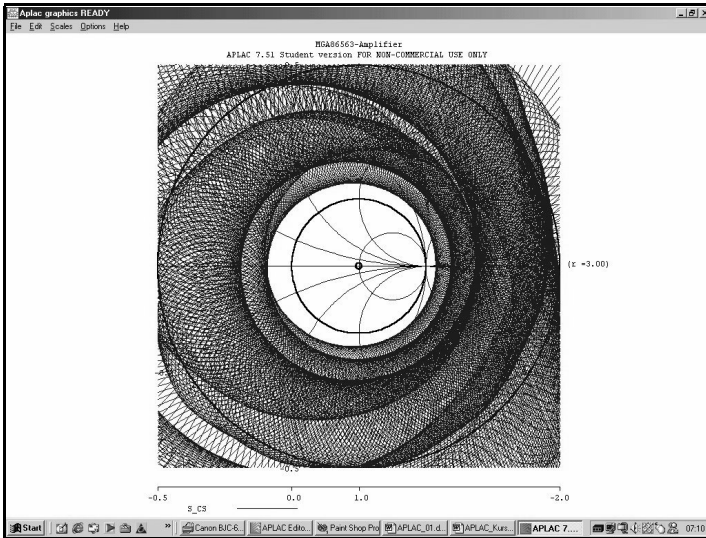
**Fig 15:
Simulation results
for the GPS
pre-amplifier
showing the
smith diagram
for the source
side.**

area covers the entire universe, but still leaves the Smith diagram free.

To help you visualise this, you can imagine a globe. Push a pin into Germany, which represents the centre of the Smith chart. Stick another pin into the other side of the globe e.g. in California, which represents the centre of the stability circle.

If we now continually increase the radius of the "Californian stability circle", then it will finally approach the pin in Germany from all sides around the globe. Eventually it is pinned into a tiny little free area and that is the stable region expected, in which there is still no oscillation.

In both cases, "conventional analysis" simultaneously gives a value of K > 1, μ >1 and D<1.

But if any parts of the circles are inside the Smith diagram, then the circuit oscillates if the internal resistance or load lies in the intersection zones. Then K is also < 1 or μ < 1 or D > 1!

It is thus the developers task to check whether, in practical operation, these

critical resistances can occur. If not, then the circuit can be implemented by careful design and can function reliably.

APLAC is able to do the calculation of functions corresponding to these circles which we must call up. They are:

S_CS for the centre and S_RS for the radius of the source stability circle, together with S_CL for the centre and S_RL for the radius of the load stability circle.

And now here's an interesting thing, which is probably not something any of APLACs competitors can offer:

Using "manual programming" of the APLAC sweep file, you can easily (e.g. by entering the following):

LOOP 1 FREQ LIN 1.57542GHz 1.57542GHz

have the ratios and all selected parameters displayed for a single individual frequency (here 1.57542GHz). But only APLAC will also calculate and show all the associated stability circles for a wider frequency range during sweeping!

All other products (e.g. APPCAD, Ansoft, etc.) are restricted here to a single selectable design frequency.

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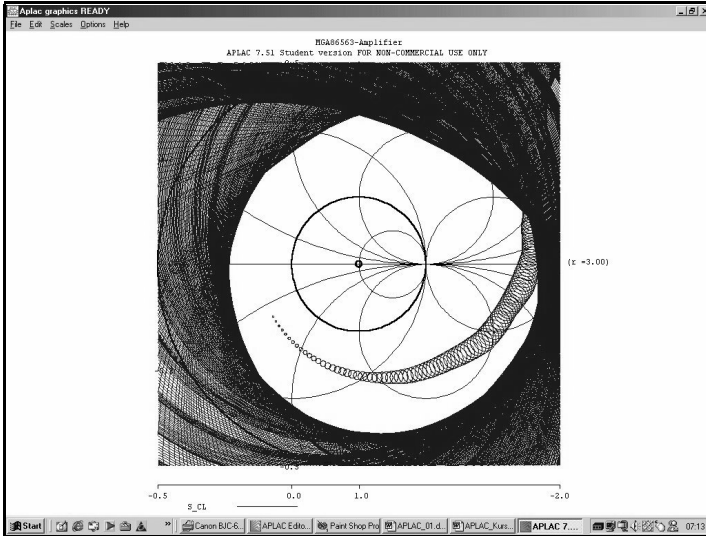
**Fig 16:
Simulation results
for the GPS
pre-amplifier
showing the
smith diagram
for the load side.**

As a final step in the amplifier project, this can be viewed. To do this, expand the sweep file again adding two Smith diagrams with the radius R=3:

The result of the work can be seen in Fig 15 for the source side, and in Fig 16 for the load side. It has already been established that the circuit is stable, but now we can see the safety margins for the different frequencies and can, if necessary, pick individual frequencies out, using "Probe", for more precise analyses.

Nor would it be a problem to look at unstable regions and work out which source impedances or load impedances to avoid and then permit "restricted operation", e.g. with precise 50Ω loads.

That can become important for most PHEMTs that have been designed for use at 10GHz and almost always have a tendency to oscillate at low frequencies. We can use just such tricks for such projects as a super low noise Meteosat or 2 meter pre-amplifier with a noise figure of less than 0.5dB to permit clean operation.

2.3. Simulation using real microstrips

In conclusion, Fig. 17 shows the amplifier circuit, where the ideal transmission lines have been replaced at the input by two actual microstrips with different widths and (to correct the effects when lines with different widths meet together) a "symmetrical Mstep". Microstrips can be called up using: "right click mouse / Microwave / Microstrip / Mlin(stripline).

The circuit dimensions required were calculated using several programs for comparison. First using PUFF, then using the stripline calculator TRL85, then using the microstrip calculator from the new AppCAD, version 3.0, and finally using a self written APLAC microstrip calculator (APLAC merely prepares the necessary functions for this users have to compile the calculation instructions themselves). There were no significant differences. Likewise, the new amplifier simulations fully coincide with the values from the previous section.
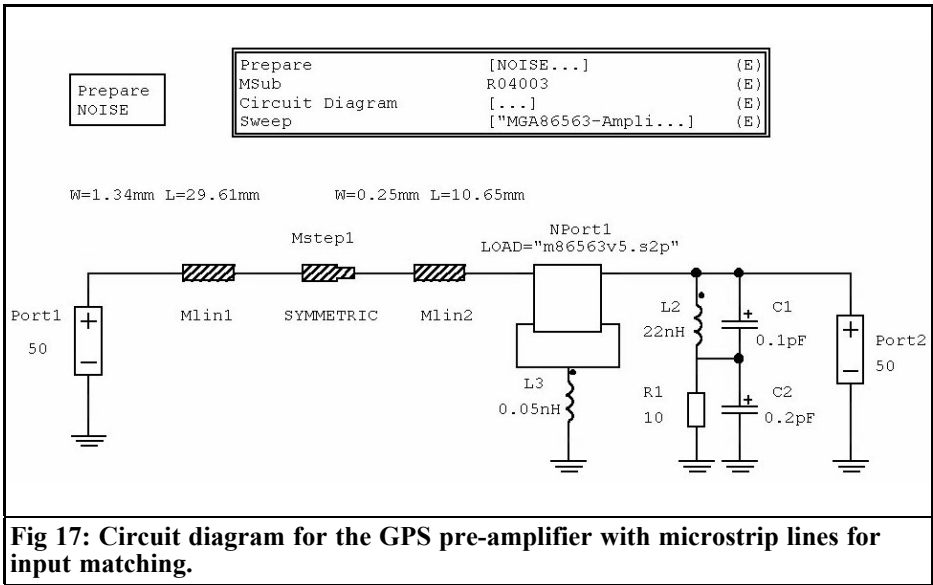
Fig 17: Circuit diagram for the GPS pre-amplifier with microstrip lines for input matching.

3.
Summary

APLAC offers an almost incredible number of options, but presents itself with a very plain and simple user interface. That means that, in comparison with other products, considerably more training is needed and routine use is decidedly harder work. Yet if we consider that all the simulations and investigations presented were undertaken exclusively using the free student version, things look much better. The author therefore uses this student version for the simulation of all those things which are not available in the familiar PUFF system, and often exchanges simulation results or partial solutions between the two programs in the form of S parameter files.

Even using the free version, there is no barrier to such things as impedance jumps or open ended extensions of microstrips, interdigital capacitors, bends, etc. Yet the message "Memory restricted in this version" often puts a

damper on ones enthusiasm in such applications.

However, the author considers himself fortunate finally to have convinced his department to procure a full version under the APLAC university programme. As a consequence of this, a 120 page APLAC tutorial has been created in four sections:

• Part 1: Training + simulations in time domain and frequency rang

• Part 2: Investigation of digital circuits

• Part 3: Passive microwave and microstrip circuits

• Part 4: Active microwave circuits

Part 1 has also been translated into English for an international introductory course in our school.

A copy of this "APLAC-CD" can be obtained from the author by email at cost price: krausg@elektronikschule.de

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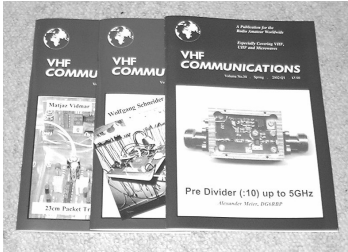
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PUFF Version 2.1 Microwave CAD Software
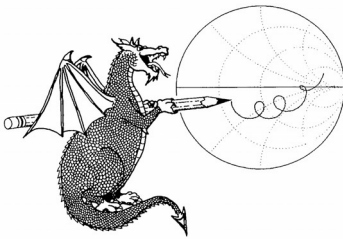
This software is used by many authors of articles in VHF Communications. It is supplied on 3.5 inch floppy disc with a full English handbook. PUFF is £20.00 + postage. (UK £1.00, Surface mail £1.30, Air mail to Europe £1.50, Air mail outside Europe £2.50)

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Prof. Gisbert Glasmachers

Synthesisers

1.
Introduction

Signal sources with adjustable frequencies are used in many places. Sources of this kind are needed in every mobile phone, in radios, in televisions etc. to select the required channel. In electronics, we normally use adjustable capacitors or coils to alter the frequency of resonance of a tuned circuit and therefore the frequency of an oscillator that is controlled by such a tuned circuit. This principle has been known for many years and it is still in use, even today, in the most modern integrated circuits. Older radios still have a knob that you turn to select the station. The disadvantage of such oscillators lies in their lack of frequency stability. To tune such an oscillator to receive transmission as clearly as possible you have to have an "ear" for these things and then you need to re-adjust it after a while. Modern receivers no longer have this problem, for they benefit from the excellent frequency stability of a crystal oscillator. This has a frequency error of a few ppm (parts per million). Even relatively high variations in temperature do not affect this outstanding stability. The only disadvantage is that you cannot vary a crystals frequency. A radio with such an oscillator would only be suitable for receiving a single transmitter, and would therefore

be practically unusable. The requirement is for a tunable oscillator with the accuracy of a crystal oscillator. This requirement can be fulfilled, with some limitations, by the synthesiser.

2.
Operating principle and applications

Even if the frequency of a crystal cannot be altered, it is still possible to generate another frequency from the first one by multiplication and division. The D flip-flop is probably the only means of halving a frequency. By combining flip-flops and gates a circuit can be assembled with which, within certain limits, any whole numbers can be generated. Two processes for multiplication are:

- Frequency multiplication by whole numbers using non-linear characteristic curves, see also [1]

- PLL , the Phased Locked Loop.

Due to the high cost of filters, frequency multiplication is excluded for most applications. PLLs, on the other hand, are the core element of any synthesiser. The limitation that multiplying and dividing the crystal frequency can produce only specific frequencies cannot be overcome

Fig 1: Block diagram of the Phased Locked Loop.
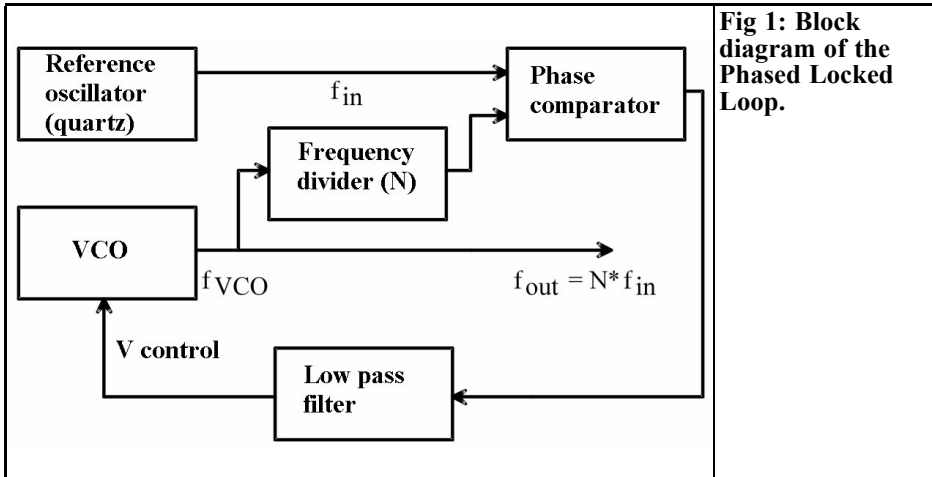
using a synthesiser. But for many applications this is an advantage rather than a disadvantage. Thus in a VHF radio, the synthesiser can be planned for the channels of the VHF transmitter. Thus whoever is using the radio is unable to tune it incorrectly. In modern telecommunications equipment, in something like a mobile phone, the synthesisers frequency is set by the system itself. The synthesiser is thus also remote controllable. This can be used to change the frequency, even while a link is in progress.

2.1. The phased locked loop (PLL)

The PLL consists of a reference oscillator, usually with a crystal, a voltage-controlled oscillator (VCO), a frequency divider (if applicable), a phase comparator (phase detector) and a low pass filter (Fig 1). Its function is to convert an input frequency from a crystal oscillator into a higher output frequency.

We are dealing with a control circuit, adjusted in such a way that the two voltages have an identical frequency at the phase comparator input. The reference is from a very stable frequency, f_{in}, from a crystal oscillator. The adjusted frequency is from a variable frequency

oscillator, the VCO (voltage controlled oscillator). A frequency divider, with a whole number division factor N, is used to make the frequency of the VCO higher than the reference oscillator by the division factor. When the equipment is switched on, a control process begins which alters the frequency of the VCO until it is precisely N times as high as the reference frequency. This is described as "latching". All the components in modern synthesiser chips are integrated, right down to the low pass filter and the crystal oscillator and often even the VCO.

2.2. The VCO

The voltage controlled oscillator (VCO) corresponds to the "manually adjustable" oscillator in the old radio because there is a capacitor adjusted as part of a tuned amplifier circuit. In the PLL a capacitance diode is used, the capacitance of this can be altered by means of a control voltage i.e. a DC voltage in the reverse bias direction.

A phase comparator and a RC low pass filter generate the control voltage. The control characteristic (usually non linear) can often be simplified in such a way that
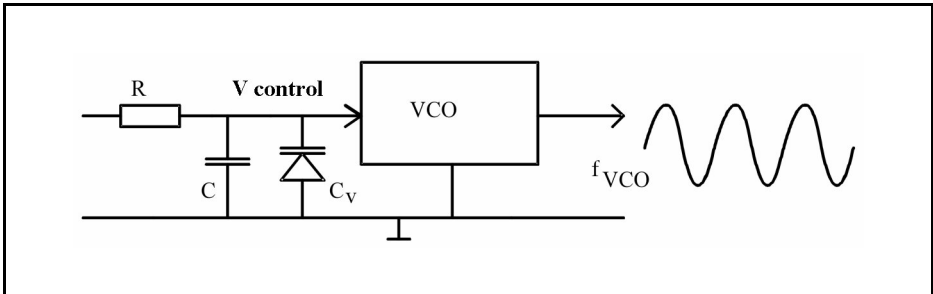
Fig 2: Block diagram of the VCO.

a linear equation precisely describes the relationship between the frequency and the control voltage (Fig 3). The gradient, k_{VCO}, of this straight line is described as the VCO gain and is expressed in MHz/V. It is an important variable for calculating the low pass filter.

$$f_{VCO} = f_0 + k_{VCO} \cdot V_{CONTROL}$$

where f_0 is the idling frequency without any control voltage.

An oscillator of this kind can be built using a reverse biased transistor and a tuned amplifier circuit. The capacitance of the tuned amplifier circuit is formed, in whole or in part, by the capacitance diode. A DC voltage, $V_{CONTROL}$ at the diode, changes its capacitance, typically in the ratio of 1 to 3 to 1 to 5. Since the RF output voltage is also applied to the

Fig 3: Response curve of VCO.

diode, make sure that it is not short circuited by the DC voltage $V_{CONTROL}$. Integrated VCOs take all these requirements into account. Only the capacitance diode must be connected to the pin provided. In the more recent chips even this is integrated. Even the tuned amplifier circuit inductance is "on chip". This considerably simplifies the amount of circuitry required. VCO chips are available from several manufacturers e.g. Motorola or Maxim. Capacitance diodes can be obtained, among others, from Siemens (Infineon).

2.3. The phase comparator

The reference vs. actual comparison in the PLL in fact takes place in a phase comparator. There are analogue and digital phase comparators. The former include multipliers and the latter ExOr gates. But most standard synthesisers contain a considerably more complicated digital circuit. Multipliers and ExOr gates actually both have the disadvantage that they only supply an output when signals with the same frequency are of different phases. However, this condition is created only when the PLL is in the latched condition but when the equipment is switched on the reference source and the VCO always have different frequencies. So a PLL can therefore latch only if there is also a frequency sensitivity in addition to the phase detection. A

Fig 4: Low Pass or Loop Filter for the PLL.

phase comparator of this type is a digital circuit with flip-flops and gates. It does not have digital logic at its output, instead it has a charge pump. This is a bi-polar pulsed current source used charged and discharged a capacitor. The current from this source is proportional to the phase difference of the two input signals.

$$I_{PC} = k_{PC} \cdot \left(\phi_{REF} - \phi_{OUT} \right) \qquad (2)$$

where the phase reference comes from the reference oscillator. (2) is essentially a periodic function since the phase is periodic every 2π. The amplification, k_{PC}, of the phase comparator, with the unit mA/rad, is again an important variable for calculating the low pass filter. At the output of the phase comparator, an earthed capacitor is used (C1 in Fig 4), which converts the pulsed current from the charge pump into a control voltage for the VCO. The higher frequency elements are attenuated using the low pass filter.

2.4. The low pass filter

As already stated, the low pass filter, also known as a Loop Filter, has to convert the rectangular current pulses from the charge pump into a DC voltage. The simple RC circuit shown in Fig. 3 is able to do this, but not in an optimum way. A

circuit corresponding to Fig 4 is used instead to give the control circuit better stability.

With a PLL it is not easy to maintain stable operation. As in all control circuits, the system becomes unstable if a phase shift of 360° is reached. The shift of the phase comparator is 180° and the VCO a further 90°. Thus a balance remaining for the low pass filter of less than 90°. At the same time it has to suppress the AC elements with a frequency corresponding to the reference oscillator. It is thus operated far above its cut off frequency. The simple RC low pass filter has a phase shift of 90° so a circuit with a lower phase displacement must be used.

In Fig 5, we see the magnitude and phase of the low pass filter when the current is fed into the input and the voltage is measured at the output. The phase shift starts at 90° at low frequencies but then reduces to a minimum (bulge). The reference frequency of the PLL should be near this frequency. In the example, there is a phase shift of 45° at approximately 20kHz. The difference from 90° is described as the phase margin. It plays an important part in the calculation programmes for PLLs.

2.5. Phase noise

A significant criterion for the assessment of oscillators is their frequency stability. Here we distinguish between the long term stability, the short term stability and the phase noise. Only the phase noise should be considered here.

According to Fourier, a sine function with constant frequency can be represented as an infinitely narrow line representing the output on the frequency, see the left hand side of Fig 6.

In practice, an absolutely constant frequency cannot be created. The spectrum
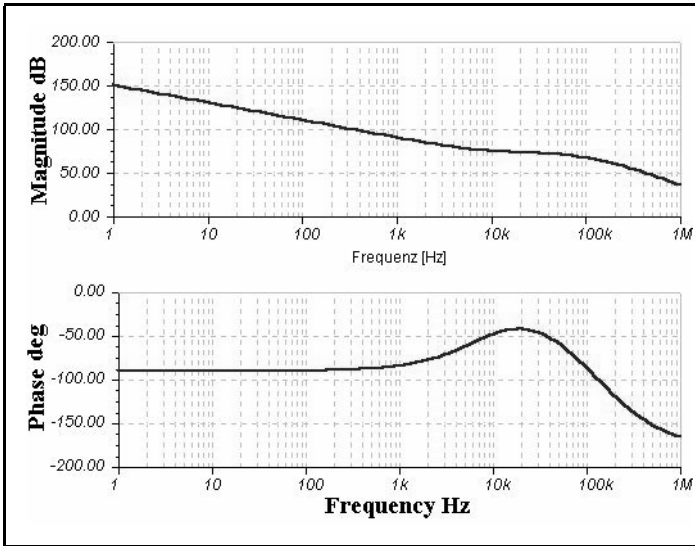
Fig 5: Frequency response showing magnitude and phase for the loop filter shown in Fig 4.

line "dances" randomly right and left of the frequency (Right side of Fig 6) around its mean value. Since we are dealing with a stochastic (random) process, we speak of phase noise. If the power density is plotted as a function of the frequency, a distribution is obtained which is similar to a bell curve.

The phase noise is defined mathematically as the quotient of two power levels.

$$PN(\Delta) = \frac{\int_{\Delta}^{\Delta+1Hz} \frac{dP}{df} df}{\int_{-\infty}^{+\infty} \frac{dP}{df} df} \quad (3)$$

The numerator is the Δ, with a defined frequency separation, measured relative to the centre, with a frequency margin of 1Hz. This corresponds to the small vertical bar in Fig 7. The denominator is the total output of the oscillator. It corresponds to the shaded area below the curve. This relationship is expressed in dBc (Decibels to the "Carrier"), usually with the suffix "Hz", which indicates the bandwidth of one Hz in the numerator.

As an example: For a good VHF oscillator, the phase noise is specified as:

Phase Noise = -95 dBc/Hz at Δ = 100kHz Offset.

3.
From PLL to synthesiser

A PLL generates only a single frequency. Further development to the synthesiser involves programmable frequency dividers. There are numerous different synthesisers on the market, but their architectures do not differ very greatly. A typical example would be a synthesiser like that from ATMEL (previously Temic) and National Semiconductors shown here in various circuits.

In digital technology, counters and frequency dividers are basically identical circuits. The counters used in the synthesiser count backwards to zero from a starting value. The starting value itself is written into the synthesisers memory by an external processor via a serial data bus.

Four such counters can be defined (See Fig. 8):
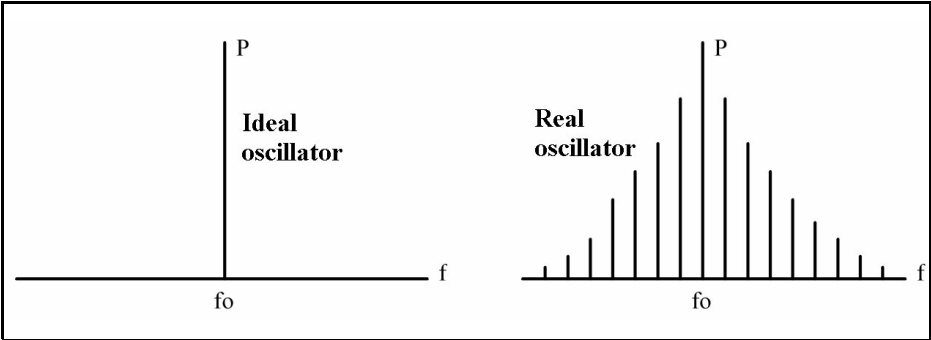
Fig 6: Spectrum of an ideal oscillator and a real oscillator.

- The Reference Counter R

- The Main Counter M

- The Swallow Counter N

- The Prescaler P

The letter R for the reference counter means that the input frequency is divided by the whole number R. The situation is similar for the counters M, N and P. The following number ranges apply for the LMX2325 from National:

```
R (14 Bit)    from 1 to 16383
M (10 Bit)    from 1 to 1023
N (7 Bit)     from 0 to 127
P = 32/33 or 64/64 or 128/129
```

The function of the Reference Counter, R, is the easiest to understand. For the PLL in Fig 1, a crystal frequency from

Fig 7: Phase noise spectrum.

the reference oscillator is applied directly to the input of the phase detector. In the synthesiser, this frequency can be very much reduced by whole number division by the counter, R,

$$f_{REF} = f_{OSC} / R.$$

The following figures are a suitable example for a VHF radio:

- f_{OSC} = 10MHz

- R = 200

- thus f_{REF} = 50kHz

The last value corresponds to the frequency spacing used for VHF radios.

The frequency f1 of the VCO is divided by the prescaler. The division factor is determined by the bit width. For the LMX2325, this can be selected as 5 to 7 bits. This gives a division by the numbers 32, 64 or 128. The prescaler has a digital control input, C, with the following function:

C = Lo means: Division by 32, 64 or 128
C = Hi means: Division by 33, 65 or 129

The circuit with the series connected AND gates before the Swallow Counter prevents an underflow of this counter. When the counter reaches zero, it simply stops. This also alters the level of the control output from Hi (Swallow Counter has not yet reached zero) to Lo.

Fig 8: Block diagram of a programable frequency synthesiser.

The entire division factor between the VCO and the phase comparator can be obtained from the product of

P · M   or (P+1) · M.

The two parallel counters, Main Counter M and Swallow Counter N, begin counting backwards simultaneously, since they both receive their clock signal from the prescaler. However, the starting values, M_{set} or N_{set}, are different, with the condition $M_{set} > N_{set}$. Consequently, the Swallow Counter N reaches zero earlier than the Main Counter. The starting values are not registered until the main counter has reached zero. This is the initial point in Table 1 where all combinations are listed.

The total number of impulses from the VCO is thus:

$$Z_{VCO} = (P+1) \cdot N_{SET} + P \cdot (M_{SET} - N_{SET}) = P \cdot M_{SET} + N_{SET}$$

(4)

The two signals at the phase comparator input have the same frequency when latched, and thus the same period, which gives us:

$$Z_{VCO} = P \cdot M_{SET} + N_{SET} = R \cdot Z_{OSC} \quad (5)$$

The frequency as the reciprocal value of the period is then obtained as:

$$f_{VCO} = f_{OSC} \cdot \frac{P \cdot M_{SET} + N_{SET}}{R} \quad (6)$$

Equation (6) can be interpreted in the following manner. The possible VCO frequencies are whole number multiples

Table 1: List of VCO pulses and counter settings.

Pulse from VCO	Main counter	Swallow counter	Comment
	M=0	N=0	Start with both counters zero
	M=M$_{SET}$	N=N$_{SET}$	When the main counter is zero, use the preset. The prescaler is P+1
(P+1)*N	M=M$_{SET}$-N$_{SET}$	N=0	When M$_{SET}$ > N$_{SET}$, N=0 while M > 0. When N=0 the prescaler is switched to P
additionally P*(M$_{SET}$-N$_{SET}$)	M=0	N=0	Complete cycle

of a basic frequency, f$_{OSC}$/R. Thus frequency channels can be constructed as is used for much telecommunications equipment.

3.1. Generation of frequency channels

In practice (television sets, radios, mobile telephones, video recorders, etc.), we frequently need channel selection using specific spacing, e.g. in the VHF range (Fig 9).

The lower series of numbers indicates the transmitter frequencies. The upper series of numbers shows the associated frequencies of the local oscillator. These frequencies must be generated by the synthesiser, in accordance with equation

Fig 9: VHF radio channels showing transmitter frequency on the bottom and the associated local oscillator frequency on the top line.

(7) as whole number multiples of 50kHz.

An example:

$$\frac{102750 kHz}{50 kHz} = 2055 = 32 \cdot 64 + 7 \quad (7)$$

If a crystal oscillator is used with $f_{OSC} = 10MHz$. The reference counter is set to the value R = 200. This gives us a reference frequency, $f_{REF} = 50kHz$. This corresponds to the required spacing. For the Main Counter Mset is then set to 32 and for the Swallow Counter Nset to 7, with a prescaler division factor of 64.

Important: the condition M$_{set}$ > N$_{set}$ is fulfilled, otherwise the synthesiser would not function satisfactorily. The calculation for all values of the VHF range can be carried out using software. There are relevant programs for this, which we will come back to later. As regards higher frequency ranges, e.g. for 433 or 868MHz, there are higher values for the counters. This gives us the higher operating frequency is in relation to the channel spacing. The synthesiser must be able to cope with the higher frequencies. Nowadays this can be overcome without any problems using modern digital technology. The problem is rather in the phase noise and in the dynamic behaviour during frequency jumping.

Fig 10: Sample screen of the National Semiconductors synthesiser programmer.

3.2. Programming the synthesiser

No synthesiser needs to be programmed until it has been switched on. In modern communications equipment, there is usually a small processor concealed behind the operator interface that can be activated by pressing a button. When a radio is switched on, the previous frequency is initially selected. The scroll key can be used to search and to let the processor program the next higher or lower frequency. The synthesiser from National Semiconductors (already mentioned several times) has a four wire bus for programming the counters, earth, data, clock, enable. It is a laborious task to go into the murky depths of hardware programming yourself. For this reason, NATIONAL, ATMEL and other companies offer software available as freeware, which runs on Windows on a PC and uses the parallel interface as bus circuits. For details, see the manufacturers Internet pages. The freeware from NATIONAL can serve as an example here.

The programming is simple. The crystal frequency is entered once, followed by the output frequency. The software calculates the starting values for the counters, displays them as decimal and binary numbers and simultaneously transmits them to the synthesiser. The screen photo (Fig 10) shows the interface of such a program.

The counters N and M are referred to here as A and B. You can also scroll along the frequency axis. The only disadvantage is that a comparatively enormous PC is controlling a tiny chip, the synthesiser. This works well in the trial phase, but is not suitable for incorporation into equipment. A second programming procedure is the building up of a processor circuit of your own. For this we have the "Programmable Interface Controller (PIC)" from MIRCOCHIP. Even the smallest of these modules has a program memory, a RAM memory, some digital peripheral connections and a timer. This is enough to assemble a small circuit with two operating keys for scrolling, two LED displays for displaying the channel number and a bus link to the synthesiser.

3.3. Calculating the loop filter

The designer must construct the low pass filter, it is not contained in the chip referred to above. Unsatisfactory results in the behaviour of a PLL can very often be traced back to wrongly calculated low pass filters. Because a mathematical calculation is difficult, it is preferable to use
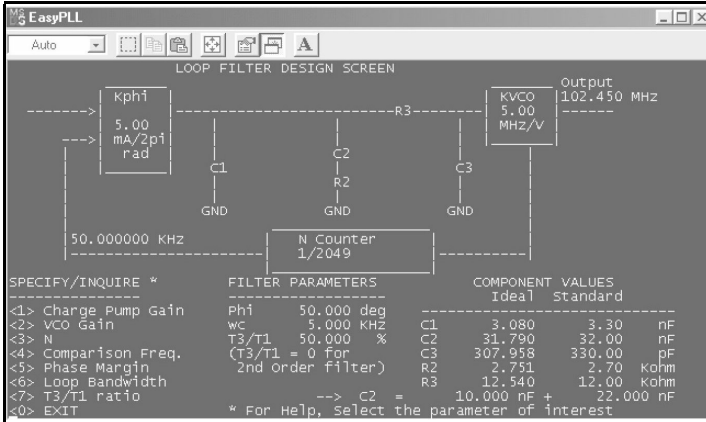
Fig 11: Sample screen display from EasyPLL.

freely available simulation software [2]. Two examples for the interfaces of such software are:

• EasyPLL is a relatively old DOS program (Fig 11) with an old fashioned interface, but supplies very usable results for filter design. However, the simulation results are very poor.

• LoopfilterDesign is a Windows program, which has a modern operator interface (Fig 12), but does not perform as well. It offers less freedom than the older

software. The author has had no experience, so far, of the loop filter calculations.

Considerably better features are offered by more modern simulation tools such as ADISimPLL from Analog Devices [3].

3.4. Analyses of output signal

The essential quality characteristics of a synthesiser are the phase noise, the behaviour during frequency jumping and the FM residues at multiples of the
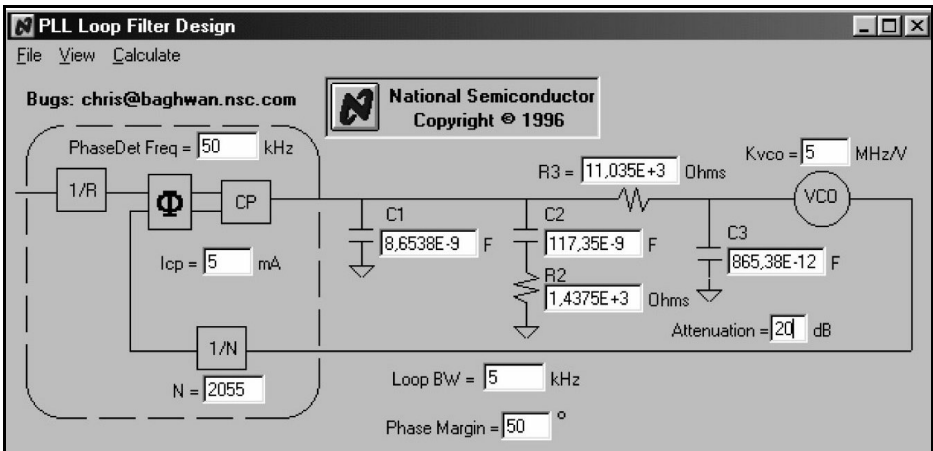
Fig 12 : Sample scree from PLL Loop Filter Design.
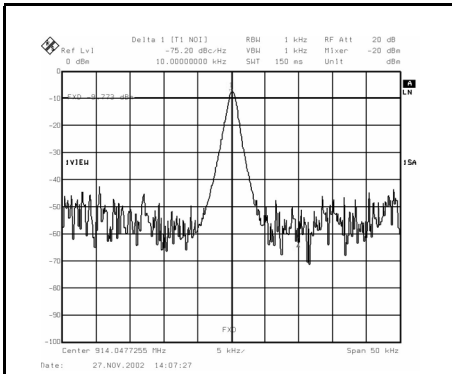
Fig 13: Spectrum analyser measuring phase noise.

Fig 14: Spectrum analyser showing frequency jumping performance.

reference frequency. None of these performance features can be measured properly without a spectrum analyser. But for the radio amateur with no spectrum analyser there is still an option to test the function of a synthesiser, a simple frequency counter with a crystal time base can do the job.

A look at the screen of such an analyser is interesting (Fig 13). We see the narrow curve of the oscillator in the centre on a wide noise platform. The measuring apparatus shows the reading:

(-75dBc/Hz at 10 kHz offset) at the top of the display, defined by one cursor at the peak and a second at a distance of 10kHz to the right of the peak. This is different to the noise in Fig 7, with a VCO in the control circuit the behaviour is very different.

Fig 14 shows frequency jumping as a time function. The flank should preferably be as steep as possible, without any overshooting. In this example, which is typical, there is a jump of 100kHz, with a strong transient effect, which lasts for something over 200 microseconds. Having a loop filter with a larger bandwidth can reduce this time further. In addition, an optimal solution can be found for both the phase noise and the transient effect using a dynamic circuit change in the

loop filter, which is supported by many synthesisers.

The FM residues represent a frequency modulation that originates from the imperfectly smoothed control voltage at the input of the VCO. This modulation generates spurs at the reference frequencys interval. These spurs can be clearly seen in Fig 15.

4.
Outlook

The further development of the synthesiser is determined by the requirements, above all those coming from the telecommunications field. The trend there is towards ever higher frequencies, where the channel spacing is often set rather narrowly.

A further modern development is frequency hopping, as already in use, with Bluetooth. This calls for rapid frequency switching within a few microseconds. Here the conventional synthesiser comes up against its limits. For the requirements for narrow channel spacing at high frequencies calls for a high degree of spectrum purity, which again can be attained only through a very narrow band
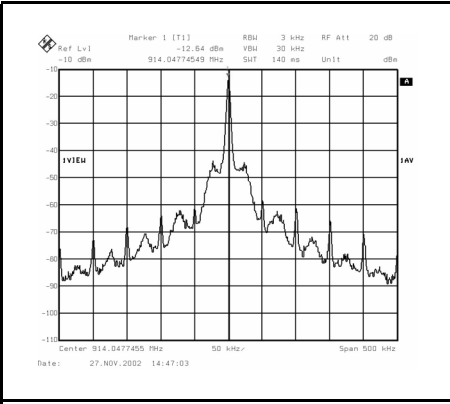
Fig 15: Spectrum analyser showing residual frequency modulation.

loop filter. But these in turn prevent any rapid frequency change. A solution can be sought through an increased use of digital technology. The VCO is replaced by a digital oscillator, the time base of which is controlled by a crystal.

"Direct Digital Synthesis" (DDS) is the keyword which is used more and more widely in technology. The very high frequencies in the GHz range have still not been included in this technique, but technological development is making progress in this direction.

The "fractional synthesiser" may be listed as a further development, which have been worked on for many years. This operates using frequency dividers, which also make possible division by fractions (numbers with decimal points). This can also be used to work with narrow channel intervals at comparatively high reference frequencies. This has an advantageous effect on the phase noise.

5.
Literature

[1] Sigurd Werner, Frequency multiplier for 76GHz with an integrated amplifier, VHF Communications 1/2003

[2] www.national.com/appinfo/wireless, click on pll-models

[3] www.analogdevices.com

André Jamet F9HX Gil Féraud F5CAU

Multiplication, division and addition of a 10MHz source to get a synthesised VHF signal

Here is an *original idea* to solve the frequency stability and repeatability problems encountered with crystal oscillators used as local oscillators for microwaves while getting a high quality spectrum close in and distant from the carrier.

1.

Introduction

Contacts at quite long distances are currently made on SHF using SSB and CW. That implies a rather difficult frequency stability goal both for reception and transmission. The magnitude of the relative stability df/f is more difficult to achieve than on the 2m or 23cm bands. Moreover digital modes like JT44 require narrower bandwidths than the classical modes so they need an increased stability. Likewise transmitter spectrum width must be limited to mode requirements.
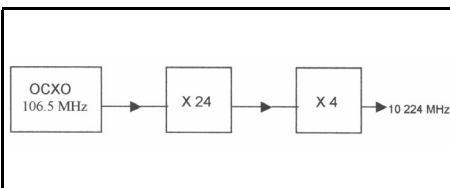
Fig 1: Block diagram of a local oscillator with a VHF OCXO source.

Local oscillator phase noise must be kept as low as possible in order to avoid any interference to week signals by strong adjacent signals.

2.

Frequency stability and requirements

An SSB or CW link needs 50Hz stability for a ten minute contact in spite of a few tenths of a degree temperature variation. For a long QSO a 100Hz variation can be accepted while temperature variation can reach several degrees. Through a day, one kilohertz is acceptable with significant temperature changes for outside usage.

Another point is the repeatability problem, this means the ability to get the same frequency after a short or long break. One kilohertz drift is acceptable owing to the receiver bandwidth.

To summarise for 10 GHz we need:

- Short-term 10 minutes : 50Hz 5.10^{-9} (5 ppb)
- For an hour: 100Hz 10^{-8} (10 ppb)
- Long period : 1000Hz 10^{-7} (0.1 ppm)

For the following ambient conditions:

119
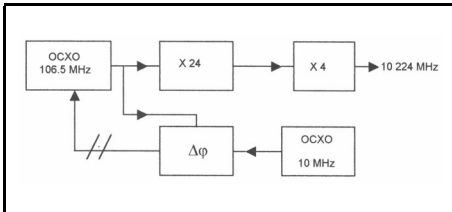
Fig 2: Block diagram of a local oscillator with VHF OCXO plus PLL with 10MHz reference.

Fig 3: Block diagram of a local oscillator with a 10MHz reference.

- Temperature : - 10 to + 45 °C
- Humidity : 0 to 100 %
- Height above sea level: 0 to 2000 meters

These stability values must be bettered for modes other than SSB and CW although we can take advantage of more restrictive values for CW.

For the higher SHF bands we have a goal five times harder for the 47GHz one and a lot more for 241GHz!

3.

Spectral purity

First we have to consider the close in noise, up to a few kilohertz due to amplitude and mainly to phase noises. Up to a megahertz we can have spurs, harmonics and mixing products. It is difficult to quantify the acceptable limits to the best of our ability and to measure them with amateur equipment. Measured values and results in the field will be covered later.

3.1. A well known solution: VHF OCXO

An example of that solution is given by in Fig 1 [1]. A VHF crystal oscillator drives several multiplier stages up to the wanted output frequency, for example 10,224MHz for a 10,368MHz transverter with a 144MHz intermediary frequency. The close in noise is good for a well designed oscillator. The phase noise is decreased when the crystal power is high but the ageing is worse, a compromise is required. That solution also needs a lack of spurious frequencies from the multiplier stages.

It is difficult using that solution to get the frequency stability needed. To put the XO into a constant temperature oven and using a crystal cut for 60°C allows a good stability in spite of the ambient temperature variations. But a repeatability effect gives a frequency shift after a break, worse when it is a long break. The only way to avoid that is to keep the power on the OCXO and get the ageing but not the repeatability problem.

3.2. Another well known solution: VHF OCXO + PLL with 10MHz reference

Several publications about this solution have been published in the amateur magazines [2]. The idea is to lock the VHF OCXO using a PLL to a very stable reference source (Fig 2). Stability and repeatability are the same as the reference that can be a professional or amateur 10MHz OCXO [3,4]. These devices do not suffer from the repeatability problems as much as VHF ones. After a break the frequency is within limits in a quarter
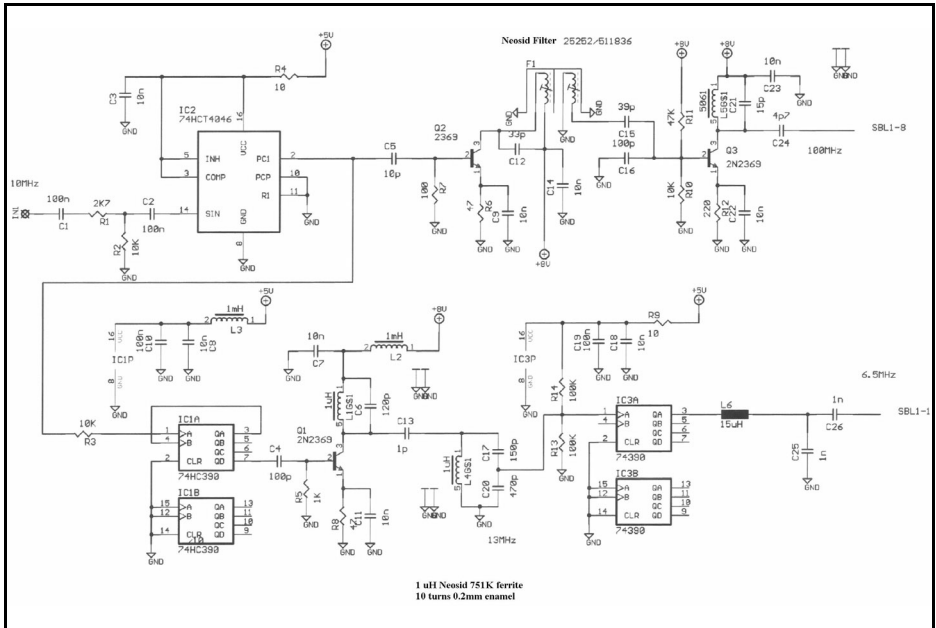
Fig 4a: Circuit diagram of local oscillator.

of an hour. We may also use a GPS signal, a standard radio frequency, a rubidium or a caesium source.

Since the frequency stability is achieved by the reference source we can increase

the crystal power in order to improve the oscillator signal/noise ratio.

Unfortunately any noise or spurs on the control signal from the PLL to a varicap diode can spoil the spectrum of the
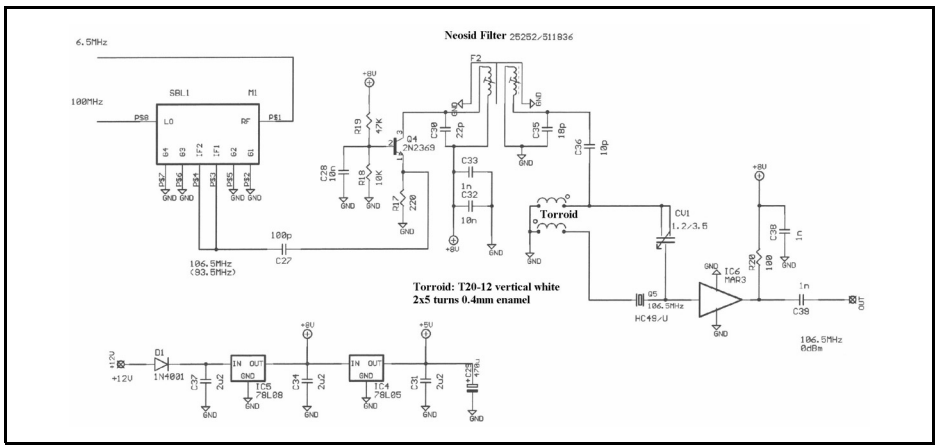
Fig 4b: Circuit diagram of local oscillator.

OCXO. A compromise is always needed between the loop stability of the PLL and any filter on the control signal.

4.

The original idea

It is based on multiplication, division and addition of a 10MHz source to get the VHF signal. If the frequency required is an integer of 10MHz, we only need multiplier stages. But to get 106.5MHz is not simple. Fig 3 shows the block diagram of the proposed device. It can seem complicated but you can't get away from it without a DDS.

The major difficulty is to get a pure VHF signal free from close in noise and spurs. Filters are included at each stage, single or coupled circuits and a *crystal filter* at 106.5MHz in the final stage. In fact that is achieved using a standard overtone 5 or 7 crystal as used in VHF OCXOs. That filter allows a very clean signal without too much loss of amplitude.

4.1. Description

Figs 4a and 4b give the complete circuit diagram. First of all, the source sine wave 10MHz signal is squared to a CMOS level. Followed by transistor harmonic multiplier stages with a strict selection of the harmonic needed. Divider stages work with digital signals. Adding the 100 and 6.5MHz signals is carried out by a mixer.

The crystal filter comprises an adjustable capacitor to compensate the inter electrodes and the case capacitance. The setting is quite tricky and we can see the spectacular effect on the screen of a spectrum analyser.

5.

Construction

The double side PCB is fitted in a 148 x 74 millimetres tined box. Components are SMD where available. Using the drawing, the construction is not difficult for a reasonably skilled OM.

6.

Set up

Cores and capacitor adjustment require some measuring equipment to avoid misleading tuning. It is valuable to follow the signal along the stages similar to tuning a multistage transmitter. A spectrum analyser and a 30MHz oscilloscope are quite useful or maybe a calibrated grid dip meter

The recommended sequence is as follows:

- Output of the 10MHz source: sine wave - 10dBm or square wave 5V p-p
- Output IC1: 10MHz square wave 5V p-p
- Output IC2A: 1MHz square wave 5V p-p
- Setting of F1 and L3: sine wave on SBL1 at 100MHz (not 90 nor 110MHz !)
- Setting of L1 and L2: sine wave on pin 1 of IC3 at 13MHz (not at 12 or 14MHz)
- Output IC3A: 13MHz square wave 5V p-p
- Output IC3B: 6.5MHz square wave 5V p-p
- Input SBL1 mixer: 6.5MHz sine wave 300mV p-p
- Output mixer: 93.5, 100 and 106.5MHz
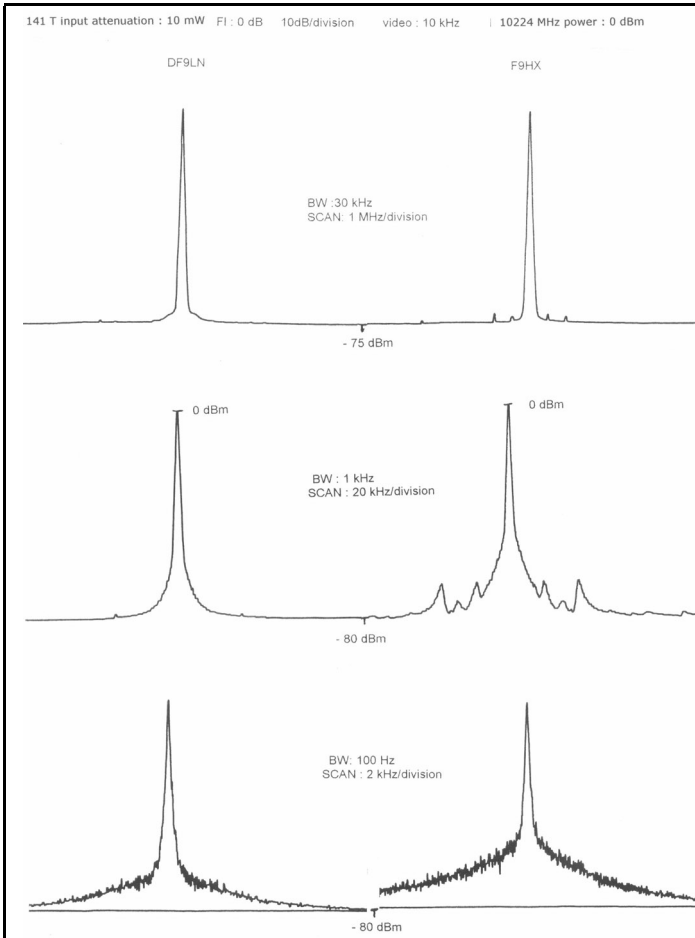
Fig 5:
The 10,224MHz
Spectrum.

- Setting of F2: 106.5MHz sine wave (not 93.5 or 100MHz)
- Setting of the adjustable capacitor to cancel all signals except 106.5MHz
- Power measurement on 50ohm load: + 0dBm 640mV p-p

7.

Results

Several 10MHz sources as TCXO, OCXO, synthesiser, have shown similar results in spite of their different spectrum grades owing to the crystal filter action.

The spectrum at 10,224MHz is shown in Fig 5. In the field tests do not show spurs on receive or transmit on the 3cm band.

No frequency stability test have been done as the 10MHz source is only responsible for that. In practice with a standard 10GHz OCXO we can reach a 108 stability i.e. 100Hz at 10GHz. Temperature tests from 0 up to 46°C showed a few dB of output amplitude variations which is acceptable for the following multiplier stages.
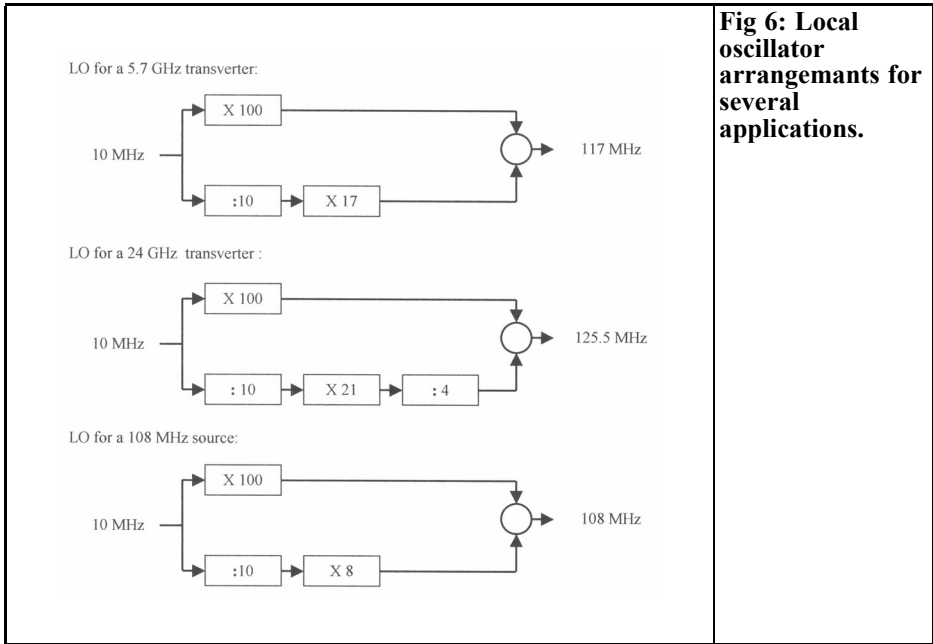
Fig 6: Local oscillator arrangemants for several applications.

8.

For other frequencies

Others SHF bands need different output frequencies. So we need to calculate which operations are required to get the right frequency always starting from 10MHz. Examples are given for some SHF bands in Fig 7.

Each time we have to use a custom cut crystal at the output frequency [6,7].

9.

Conclusion

The tests showed that this method is able to solve our frequency stability requirement at SHF with the availability of a high grade 10MHz source. This can be used for several outputs one for each SHF band.

10.

References

[1] Oven-Stabilised XO for VHF, DF9LN, DUBUS 3/1997

[2] Universal Microwave PLL System, WA6CGR http://ham-radio.com/wa6cgr/mwpll.html

The G8ACE Transverter for 76 GHz wa6cgr@ham-radio.com

PLL for Microwave Local Oscillators, VK5KK vk5kk@qsl.net

[3] DC9UP@germany.com

[4] A High Stability 10MHz Reference Oscillator, F9HX, Microwave Newsletter, May 1999

Gunthard Kraus, DG8GB

Internet Treasure Trove

Ansoft

The free download site of the well known supplier of simulation tools has expanded yet again. In addition to the student versions of Serenade SV (High frequency circuit design) and Ensemble (Wireless circuits and planar antennas), we now also have "Simplorer" (Multi-Technolgy Simulation for automotive, aerospace and power design). And the very latest thing is "Maxwell SV", an electromagnetic field simulator. These are stripped down versions of the professional tools written for Windows NT and cannot be guaranteed to operate satisfactorily with Windows ME or XP.

Address:
http://www.ansoft.com

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Address:
http://www.klmicrowave.com

Microwave Office

Another microwave simulation program, and another one that makes a free test version available for 30 days. It is prob-

ably not so interesting for the private user on grounds of cost alone, but the homepage does contain a lovely collection of application notes with many practical examples. It is worth more than a casual look and to see where the associated literature sources can be found (e.g. for hairpin filter design).

Address:
http://support.mwoffice.com/appnotes/listnote.aspx

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Address:
http://www.rfmd.com/prodTechnical.asp

Simtel

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maging around a bit here

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UK Microwave Group

You should drop in on these guys regularly, there is always something new. This time it's the free student version of the filter program "elsie", but there's also the "Hardware" section, and don't forget to look at the "scatterpoint" magazine.

Address:
http://www.microwavers.org/index.htm

Anadigics

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Address:
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LC

No, this is nothing to do with a tuned amplifier circuit or with filter circuits. We're talking about a free EM simulator (e.g. for calculating microwave structures or patch antennas). Free up to now for users of Cray or SGI super-computers, but suddenly there is also a LINUX / Intel x86 version available for downloading following registration. There must be lots of people who'd be interested in that.

Address:
http://lc.cray.com/

FIR filter software list

Anyone who has ventured into digital signal processing must have wanted to program digital filters. This site provides a comprehensive list of FIR design software that can be downloaded or used on line. And naturally there's an important link to other DSP pages.

Address:
http://www.inforamp.net/~poynton/Poynton-dsp.html

Micromanipulator

A small site admittedly, but a splendid one for all that. It has an outstanding list of the characteristics of standard contemporary microwave connectors, together with some good comments on correct microwave circuit terminology.

Address:
http://www.micromanipulator.com/RF-Notes.html

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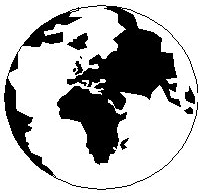
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