VHF
COMMUNICATIONS

A modern precision automatic SINAD meter, distortion meter and true RMS voltmeter

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Contents

There are some very interesting articles in this issue, I now know how a YIG oscillator works! The SINAD meter was published in three consecutive issues of UKW Berichte but I decided to put the whole article in this issue.

For the readers who often ask for some 2m projects there is a nice 2m PA, the D1030UK transistor is a bit expensive but full power without a dangerous HT supply is worth the price. Zeljko Bozic has come up with a neat design for a GPS controlled frequency standard, a quick check on eBay and I found the GPS units for about £25. Finally a welcome return of the spectrum analyser by Matjaz Vidmar, I found his web site while surfing for something else and spotted the LCD oscilloscope and thought it would be a good idea to print the article. As Matjaz mentions on his web site he can put many more pictures (and in colour) on his site, so please take a look, but not everyone has access to The Internet even if you do there is no guarantee you will find the site.

73s - Andy

Alexander Meier, DG6RBP

How YIG oscillators work and are driven plus an example YIG oscillator driver

YIG oscillators have been used for several years in expensive measuring instruments, so they are now available as used parts at reasonable prices. However the way that these oscillators work and why they are suitable for use in RF Wobblers or spectrum analysers, because of their excellent characteristics, is not well known by radio amateurs. The way that they are driven is quite different from other oscillators, the basics are described in the following article.

1.

Characteristics of YIG oscillators

Among the most important characteristics of a YIG oscillator, also known as a YTO (Yig Tuned Oscillator), are the large tuning range of more than one octave and the linear tuning characteristic. They are different from varactor tuned oscillators that require a specially characterised tuning voltage to give linear tuning. YIG oscillators are tuned with a variable current. To avoid temperature effects most YIG oscillators are heated and held at a constant temperature. YIG oscillators are available for frequencies

from approximately 1.5GHz to over 24GHz with power outputs typically +13dBm (20mW). The first harmonic is usually suppressed by approximately 10 to 20dB.

2.

Operation of YIG oscillators

The circuit diagram of a simple YIG oscillator is shown in Fig 1. The interesting part is the YIG sphere in the emitter circuit of transistor T1. But how does a YIG resonator functions?

A YIG resonator consists of a coil with a small black sphere sitting in the centre as shown in Fig 2. The abbreviation YIG stands for Yttrium Iron Garnet ($Y3Fe5O12$). The sphere is approximately 0.5 to 1mm diameter. If this is exposed to an external magnetic field (H_0) the spinning electrons in the YIG that form magnetic dipoles start to align with the field and precess around that field. This generates a small alternating field or RF signal that is coupled to the sampling coil. This resonant frequency (f_0) is independent of the YIG sphere diameter and can be computed by:

$$f_0 = \gamma\left(H_0 + H_a\right) \qquad \gamma = 352 \cdot 10^3 \left[\frac{Hz}{A/m}\right]$$
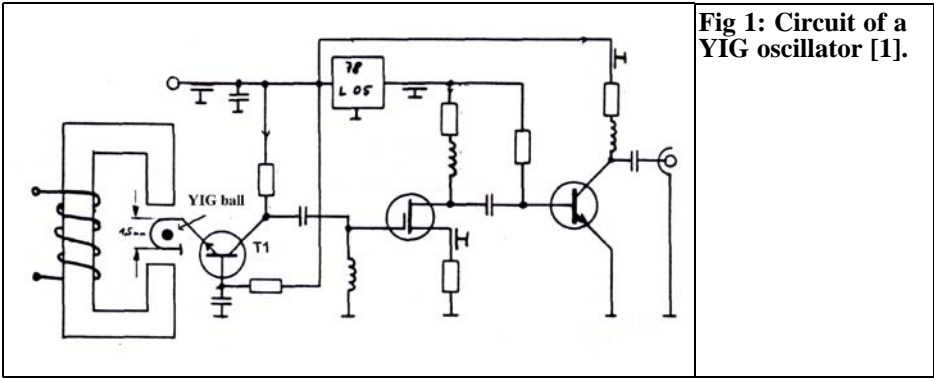
Fig 1: Circuit of a YIG oscillator [1].

Where:

γ = gyro magnetic relationship, a constant dependant on the charge mass relationship of electrons.

H_0 = Continuous external magnetic field.

H_a = Anisotropic field, this is the difference between the applied field and the very small (nearly negligible) field internal to the YIG. This is temperature dependent but can be minimised by an adjustment of the YIG sphere.

If this YIG arrangement, with its periodical characteristics, is used in the circuit of an oscillator (Fig 1), then the frequency can be controlled by an external magnetic field H_0. This magnetic field is generated by a direct current in field coils, this is the tuning current. If the very small anisotropy field is disre-

garded, then f_0 is proportional to H_0 giving a tuning range linear to the tuning current.

3.

YIG connections

YIG oscillators are usually built into round or square metal housings, see Fig 3. The SMA socket is the RF output, the other solder pins are for the supply voltage and the tuning current. The connections are not standardised therefore care is needed when buying a used YIG oscillator at a flea market. The following data refers to the connections of a typical YIG oscillator like the one shown in Fig 4.
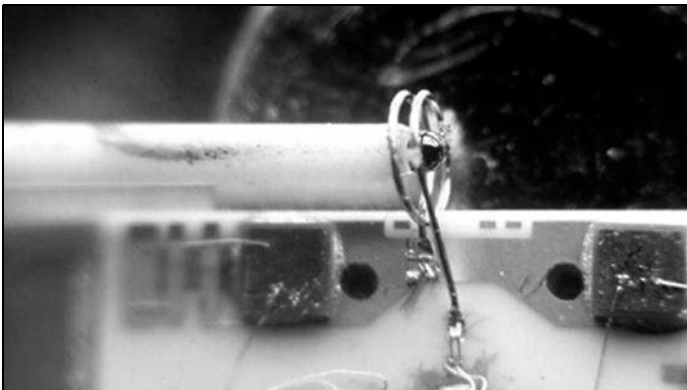
Fig 2: Picture showing the YIG sphere.

Fig 3: A 4 to 8GHz YIG oscillator.

- Heating (not always available)
- Operating voltage (positive, negative or both)
- Tuning coil (Tune), typically 20MHz/mA
- Tuning coil (FM), typically 300kHz/mA
- RF output, typically +13dBm

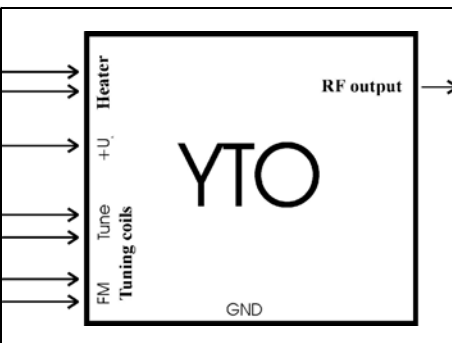Frequently there are two connections for the heater, one should be connected to ground and the other to the appropriate DC voltage (usually +24V). Initially a high inrush current will flow (often > 0.5A), this settles down within a few


Fig 4: Typical connections for a YIG oscillator.

seconds to approximately 50 to 100mA. This heats the YIG sphere to 80 - 125°C and is kept constant at this value. The heating is necessary to minimise temperature affects. Some YIG oscillators only require a positive supply voltage of 15V at 100 to 300mA, others require a negative supply and some require both a positive and a negative supply. Also unusual supply voltages are sometimes required e.g. -4.3V or -5.1V. The ground connection is usually missing, therefore this must be connected to the housing.

Two coils are used for frequency tuning:

- The main coil "Tune" (approximately 10Ω) is used for coarse frequency adjustment. The tuning rate is usually 20MHz per mA! Thus a current of (2000MHz/20MHz) = 100mA is necessary for an output frequency of 2GHz. For 18GHz this is then 900mA! This high value of current can cause problems with control, if this varies by ±2.5mA, then the output frequency varies by ± 50MHz! The control current should therefore be as stable and as low noise as possible. The tuning linearity is very good, typically ±0.1%. To achieve better stabilisation of the output frequency a capacitor of approximately 500µF can be connected in parallel with the tuning coil. However the YIG oscillator cannot be tuned so fast and is not suitable for a Wobbler or spectrum analyser and is only suitable as a CW oscillator.

- The second coil is the "FM coil" (approximately 1Ω) and a tuning rate of typically 300kHz/mA. Thus the YIG oscillator can be frequency modulated or this coil can be used for small adjustments of the output frequency. It has a smaller inductance than the main coil, so that faster tuning is possible. The small FM coil is wound with much thinner wire so the control circuit should have a safety device to limit the current to <250mA.
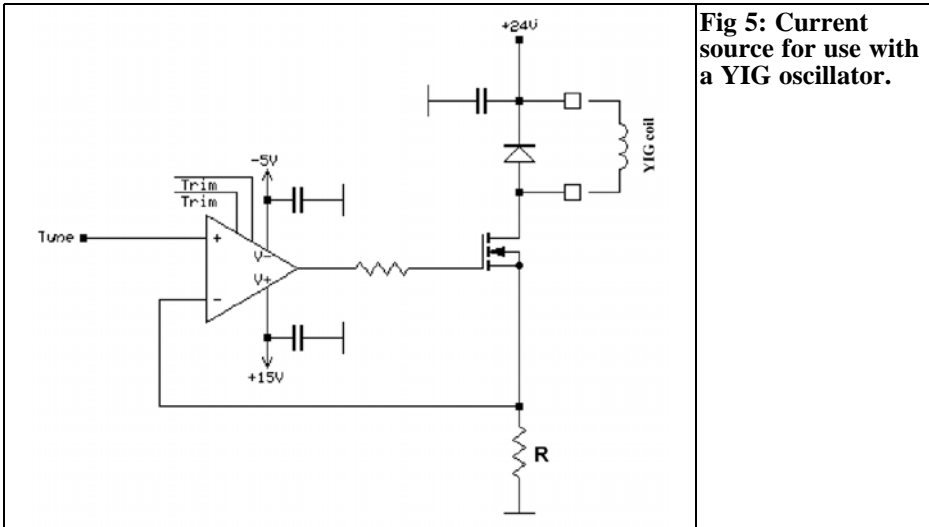
Fig 5: Current source for use with a YIG oscillator.

If only the tuning coil is used it does not matter which connection is positive. If both tuning coils are used the polarity is important so that both magnetic fields are additive. The tuning coils should never be used with a constant voltage because the coils warm up due to the current flow changing their resistance, hence the current changes which changes the output frequency.

4.

Control of a YIG oscillator

To control the frequency of the YIG oscillator with a voltage, a current source is used as shown in Fig 5. The current flowing in the YIG coil also flows through the resistance "R" to ground. The current produces a proportional voltage drop across the resistance "R" that is fed to the inverting input of the operation amplifier. This compares the actual current with the tuning voltage on the non-inverting input and regulates the current. The control current is given by:

$$I_{coil} = \frac{V_{tune}}{R}$$

It is advisable to use a field effect transistor instead of a bipolar transistor because the base current of a transistor would flow through resistance "R" causing a small error. The diode in parallel with the YIG coil is necessary, if the control voltage exhibits fast jumps e.g. with a ramp signal or when abrupt switching occurs, because an induction voltage would be developed that could destroy the transistor. The resistor "R" should only be wound from Constantine wire. This gives very good temperature stability and very good frequency stability. The characteristics of other wire wound resistors are not as good. The resistance should be fitted so that it is kept at a constant temperature if possible. An alternatively circuit can be used that gives an inversion from input voltage to output current. This type of circuit is frequently used and has some advantages but needs a negative control voltage. The circuit is not described here because the principle is the same.

A low pass filter should be used in front of the current source. If the YIG oscillator is used for a Wobbler, the filter should be designed for approximately 50kHz so that is does not to deform the sawtooth. For CW operation the filter

should be a much lower frequency, approximately 10Hz to suppress any low frequencies e.g. 50Hz mains. Additionally a capacitor with a small series resistor can be used in parallel with the tuning coil for CW operation. The capacitor should be approximately 500µF. If the YIG oscillator is used for a Wobbler, the FM coil can be used to control the frequency in the same way as the main tuning coil. For small span sweepers, the current in the main coil is set to a fixed current to give the starting frequency and then the FM coil used to vary the frequency. The FM coil can only change the frequency around that set by the tuning coil. If the FM coil is not used for a Wobbler, it is advisable to use a DC coupled push-pull operational amplifier in the circuit of Fig 5. This will give the best low frequency response and allow the frequency to be varied in both directions.

Finally it should be emphasised that absolutely low noise and stable supply voltages are necessary for the control circuits! The smallest fluctuations, e.g. by stray effects of transformers make the output frequency vary very greatly! High quality, low noise amplifiers should be used e.g. OP07 or OP27. The wiring between the control circuit and the YIG oscillator must be laid out with care avoiding stray effects from transformers and other cables. The wires to the tuning coils should be twisted. Unfortunately 50Hz modulation is nearly always present on the RF output caused by stray effects finding their way into the wiring of the control circuitry.

5.

A universal driver circuit for YIG oscillators

The circuit diagram of the driver circuit is shown in Fig 6 and Fig 7. The control voltage between 0 and 4V is applied at pin 2 of J1. The trimmer R2 sets the tuning rate for the YTO between 0.3 and 2.3V/GHz. For example this would be set to 1.5V/GHz for a YTO with a frequency range from 2 to 8GHz. So that the oscillator is set to its lowest frequency when the control voltage is 0V, an offset is applied to the non inverting input of the operational amplifier U1. This is produced by a low noise precision reference (U4) and inverted with the operation amplifier U3. The trimmer R8 is used to adjust the offset voltage, which produces an offset current in the main coil (Tune) of the YIG oscillator from 50 to 650mA. For our example oscillator from 2 to 8GHz the offset current is adjusted to 100mA.

The low pass filter (R12, C12) has a cut off frequency of approximately 50kHz. This does not cause distortion even with very fast sawtooth sweep speeds. For CW operation the cut off frequency can be reduced to 15Hz with the switch Q4 and the capacitor C11, this will reduce the 50Hz hum.

The filtered control voltage is fed to the current source U2. That control current for the YIG coil is proportional to the voltage drop across the resistor R18. The operation amplifier compares the actual value with the desired value (control voltage) and regulates the current accordingly. The resistor R18 should only be wound from Constantine resistance wire. Only this wire will give outstanding temperature and thus frequency stability. The current in the main coil (Tune) of the YIG oscillator can be measured with a voltmeter across the test connection TP1 at 1V per Ampere.

The diode D2 limits the affects of abrupt changes in the coil current due to jumps in the control voltage e.g. sawtooth used for a Wobbler and protects the control circuit.
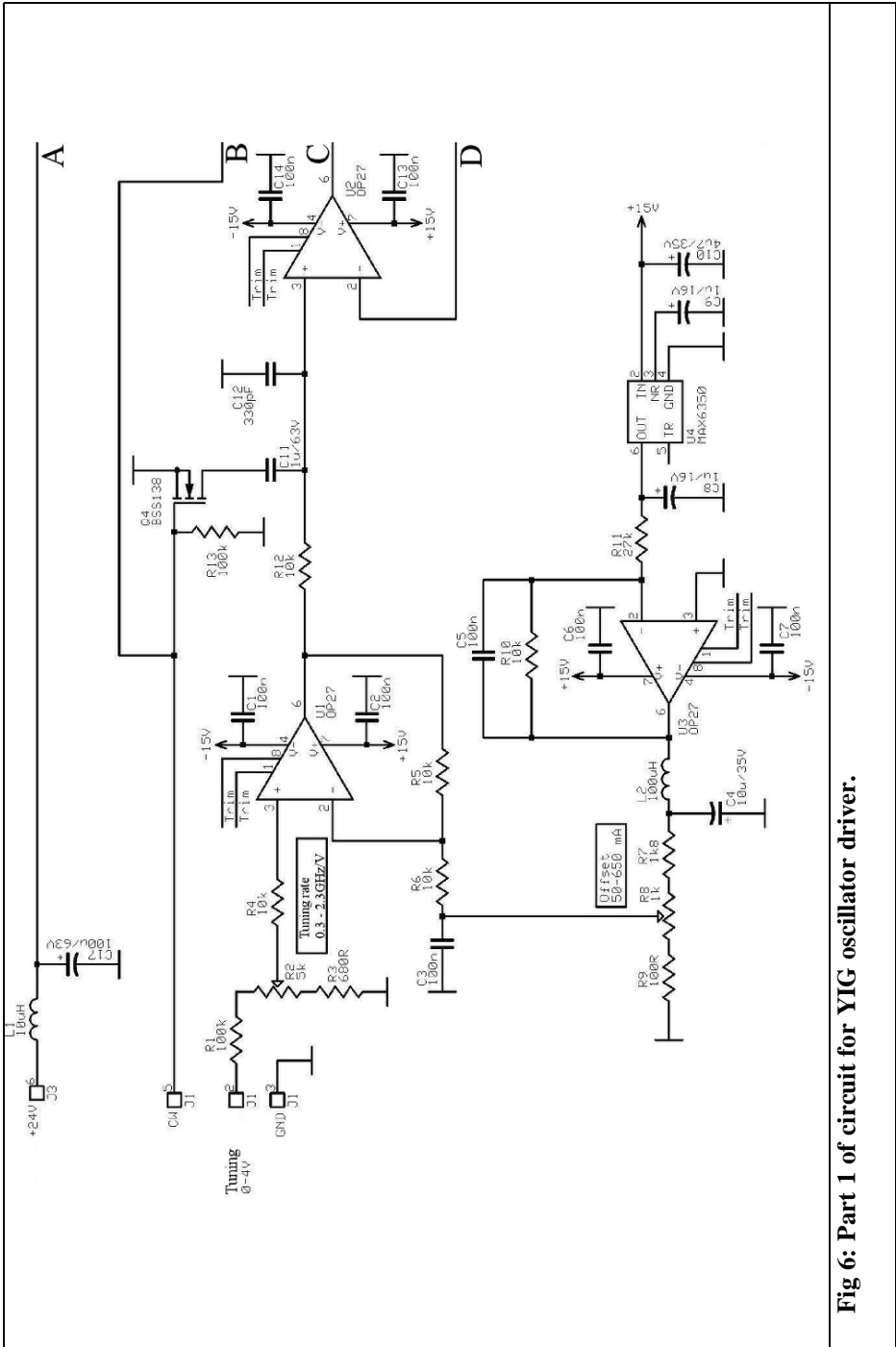
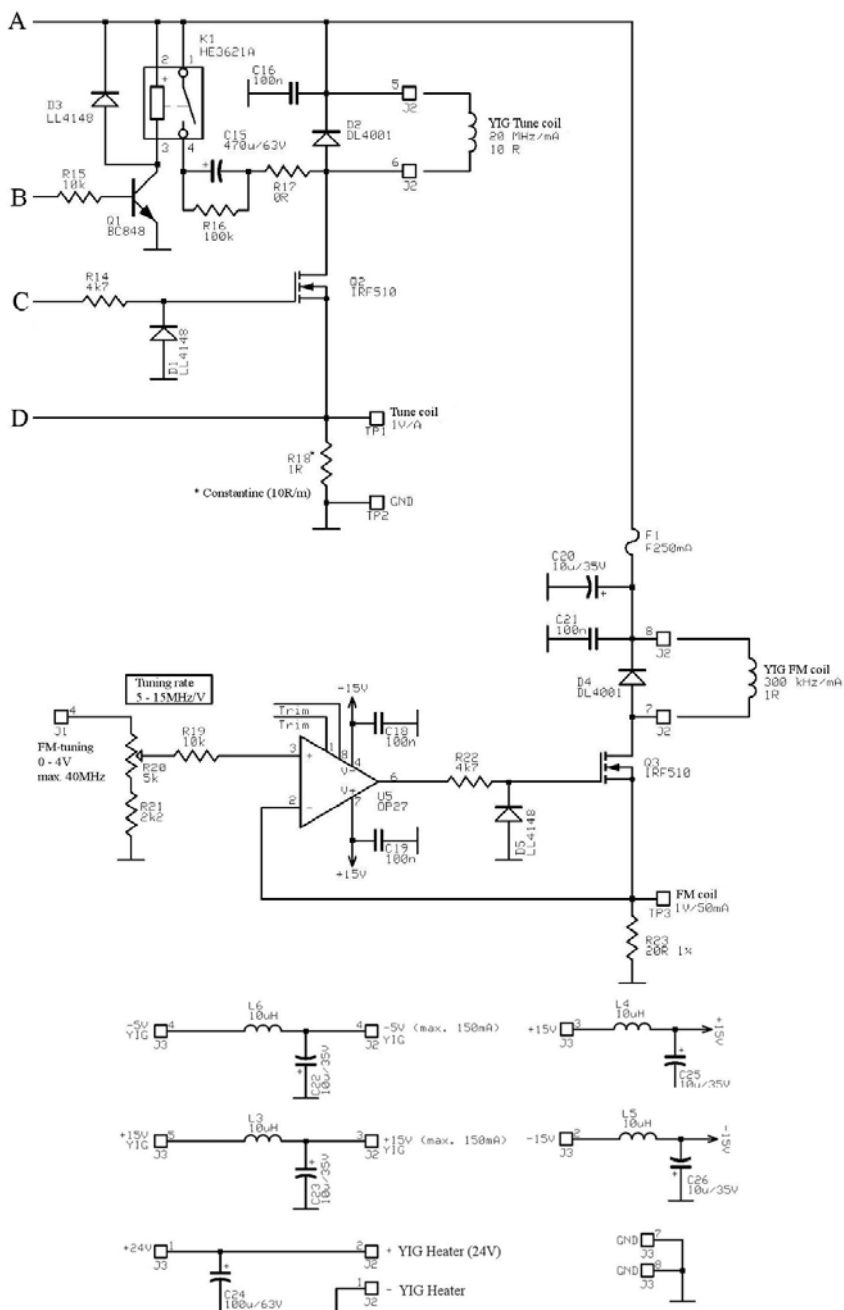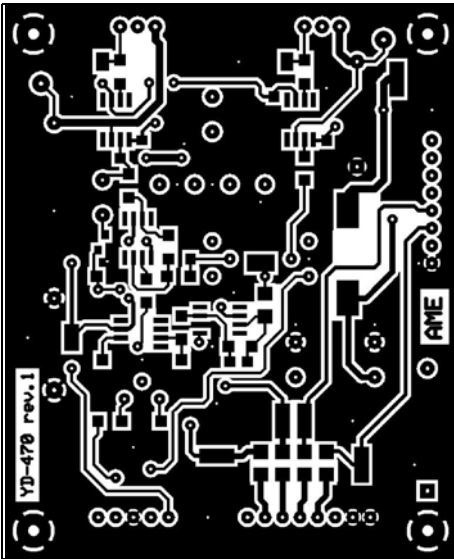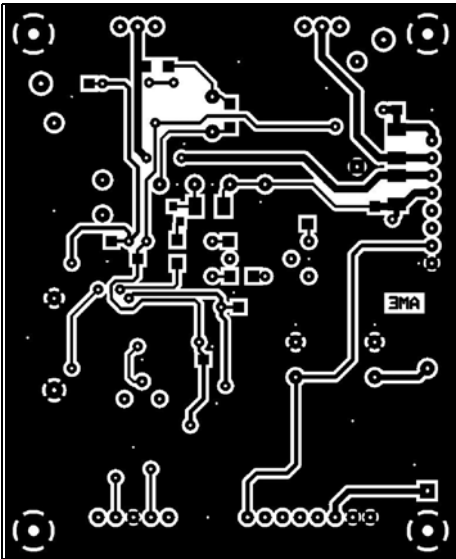Fig 6: Part 1 of circuit for YIG oscillator driver.

**Fig 7: Part 2 of circuit for YIG oscillator driver.**

Fig 8: PCB layout for component side of YIG oscillator driver.

Fig 9: PCB layout for SMD side, actual size is 65mm x 80mm.

The reed relay K1 switches the capacitor in parallel with the main coil (Tune) for CW operation. This greatly improves the frequency stability. The value of the capacitor and the series resistor R17 (optional) can be selected individually. For a first attempt, select C15 of 470µF
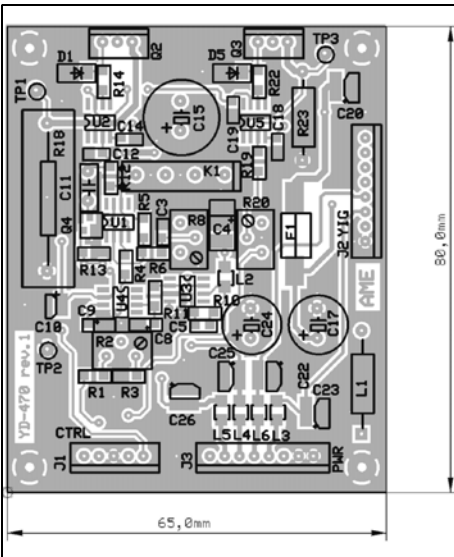
Fig 10: Component layout for component side of PCB.
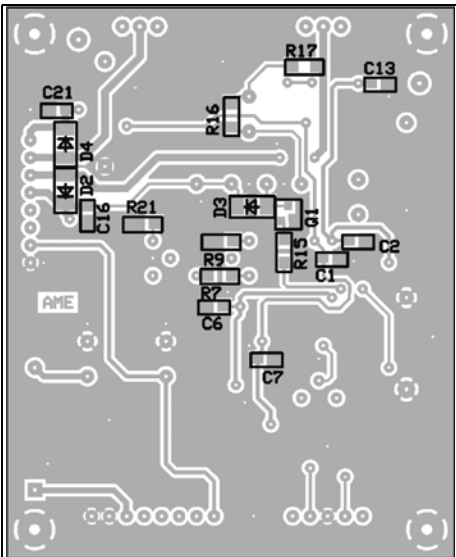
Fig 11: Component layout for SMD side of PCB.
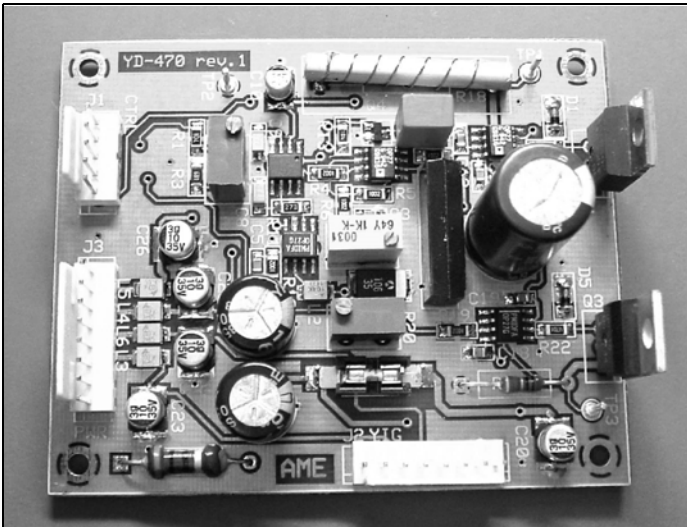
Fig 12: Finished PCB for the YIG oscillator driver.

and R17 of 0Ω.

The control of the FM coil was developed for Wobber applications with small frequency changes. It is only possible to change the frequency in one direction but this is sufficient for a Wobbler. The starting frequency is set by the main coil. The circuit can be used for FM modulation by overlaying the modulation voltage onto a DC voltage of 2V at pin 4 of J1.

The FM coil has a smaller tuning rate of 300kHz/mA therefore a precision metal film resistor is sufficient for R23. The current in the coil can be measured with a voltmeter across test point TP3 at 1V per 50mA. The fuse F1 provides protection for the thinner wire of the FM coil in the event of a fault.

The supply voltage for the oscillator used (+15V/-5V) is fed to the circuitry via a low pass filter (L6 and C22, L3 and C23). The supply voltage for the control circuit (±15V) is filtered separately. The only other supply required is +24V for the heater of the YIG oscillator and the current supply for the YIG coils. The power supply for the control circuit must

be able to supply sufficient current for the YIG oscillator. All output voltages of the power supply must be protected in order to save the expensive oscillator in the event of a fault. It should only be a high quality, linear regulated power supply (no switch mode power supplies).

The PCB is a 65mm x 80mm double sided board with plated through holes, the layout is shown in Figs 8 and 9.

6.

Construction

The component layout is shown in Figs 10 and 11. First all of the SMD parts are fitted followed by the wired components. The Constantine wire wound resistor R18 is wound on a thin aluminium rod covered with insulating tape. The completed resistor is fastened onto the PCB using two-part adhesive. The transistors Q2 and Q3 are fitted to suitably sized heatsinks using thermal paste. The circuit can be tested with resistors in place of the YIG of oscillator coils, 10Ω for the Tune coil
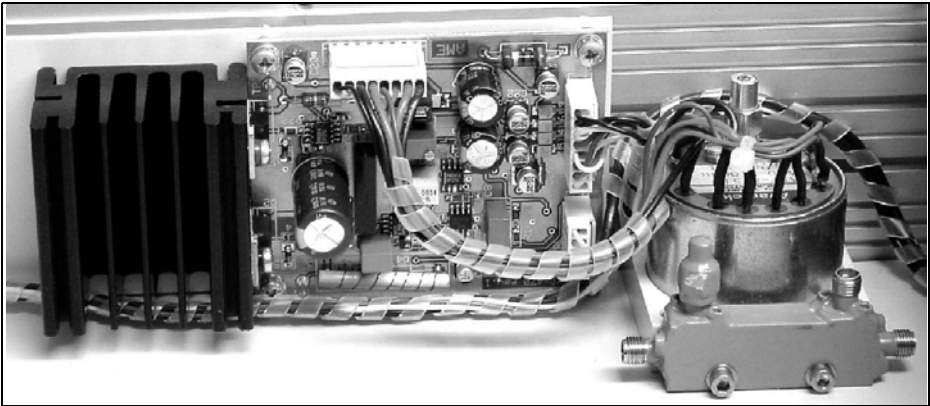
Fig 13: YIG oscillator driver built into a microwave Wobbler.

and 1Ω for the FM coil. With the supply connected all of the supply voltage for the YIG should be tested on connector J2. Next the control voltages are connected and the current in the 10Ω and 1Ω resistors measured. If everything functions corectly, the YIG oscillator can be attached.

For the wiring of the driver circuit to the other building blocks (YIG oscillator, power supply, etc.) a suitable arrangement to minimise stray effects (e.g. 50Hz hum) must be adopted. The control voltage should be connected to the control circuit using shielded coaxial cable.

A finished printed circuit board is shown in Fig 12 and a printed circuit board with YIG oscillator and heatsinks, built into a microwave Wobbler, is shown in Fig 13.

7.

Parts list

R17	0Ω, 1206
R9	100Ω, 1206
R3	680Ω, 1206
R7	1.8kΩ, 1206
R21	2.2kΩ, 1206
R14, R22	4.7kΩ, 1206
R4, R5, R6, R10, R12, R15, R19	10kΩ, 1206
R11	27kΩ, 1206
R1, R13	100kΩ, 1206
R8	1kΩ, trimmer 64Y
R2, R20	5kΩ, trimmer 64Y
R18	1Ω, Constantine
R23	20Ω, 1%, metallic film 1/4W
C12	330pF, 0805
C1-C3, C5-C7, C13, C14, C16, C18, C19, C21	100nF, 0805
C8, C9	1μF/16V, SMD tantalum
C11	1μF/63V, RM5, MKS
C10	4.7μF/35V, SMD Electrolytic
C4, C20, C22, C23, C25, C26	10μF/35V, SMD Electrolytic
C17, C24	100μF/63V, Radial
C15	470μF/63V, Radial
L1	10μH, wired, 1 A
L3, L4-L6	10μH, SIMID 1210
L2	100μH, SIMID 1210

U1, U2, U3, U5	OP27 SMD
U4	MAX 6350, SMD
D1, D3, D5	LL 4148
D2, D4	DL 4001
Q1	UC848
Q4	BSS138
Q2, Q3	IRF510
K1	HE 3621 A, Reed Relay 24V
F1	Fuse F250mA + holder
J1	5 pole plug
J2, J3	8 pole plug
1 x	receptacle for 5 pole plug
2 x	receptacle for 8 pole plug
21 x	crimp pins for plugs
2 x	Test pins
1 x	Heatsink
1 x	PCB DG6RBP YIG control

8.

References

[1] Spectrum analyzer from LF to 48GHz, Vieland Carsten, Lecture to the Weinheim VHF conference 1987

9.

Literature

For further reading about YIG oscillators, the following literature is recommended:

[a] YIG oscillators and their control, Alexander Meier, Lecture on the VHF, Weinheim conference (2003), The current version can be found on the AME Engineering web site, www.ame-engineering.de

[b] A simple approach to YIG oscillators, Bernd Kaa, VHF Communications Magazine 4/2004, pp 217 - 224

[c] YIG resonators and filters, Joseph Helszajn, John Wiley & Sons Ltd. (1985), ISBN 047190516

[d] Microwave filters, Impedance Matching network, and Coupling Structures, Matthaei Young Jones, Artech House Inc., 1980, ISBN 0890060991

[e] Microwave engineering, Günter Käs and Peter Pauli, Franzis Publishing House, Munich 1991, ISBN 3-7723-5594-3

[f] Spectrum analyser from LF to 48GHz, Vieland Carsten, Lecture to the Weinheim VHF conference 1987

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A modern precision automatic SINAD meter, distortion factor meter and true RMS voltmeter

In all receiver engineering with similar types of modulation the data for the receiver characteristics refer to signal/signal-to-noise ratio and noise figure either directly or indirectly. The quality of the audio signal is evaluated by the distortion factor or signal breakdown. To simplify the measurement procedures a SINAD meter is used. The equipment presented in this article combines a distortion factor meter with a true RMS voltmeter with over 102dB range, in fact all the capabilities of a laboratory instrument (for the AF side) to make measurement precisely and efficiently. A CCITT weighting filter permits, if required, correct evaluation according to the standard.

1.

Introduction

The characteristics of a radio receiver are usually measured by two general measurement methods. The receiver (RX) is considered as a two port device. This practice permits the examination of data during maintenance or in service without the need to interfere with the test specimen. The procedure for determining many fundamental characteristics e.g. the receiver sensitivity, is shown in the schematic diagram, Fig 1.

To determine the behaviour in the scenario when several transmitters are received at the antenna socket, the two or multi-transmitter measuring procedure is used as shown in Fig 2. Typically this is used to examine the large signal characteristics for adjacent channel suppression, blocking or intermodulation.

Further information on the fundamental measurements is available in [2]. [3] deals with the special case of frequency modulation F3E.

From the block diagrams it can be seen that the measuring instrument described in this article needs to be able to evaluate the AF output signal.

Measuring the frequency of the demodulated output is only required in some special cased. Because even inexpensive digital meters can measure frequencies up to 10MHz with a resolution of four places it was a conscious decision not to include an AF counter.

To make the instrument easy to use it was decided to use a moving pointer meter. This is particularly useful for alignment work, it is much easier to follow a moving pointer rather than a flickering number display. This is also true for changes in measurements
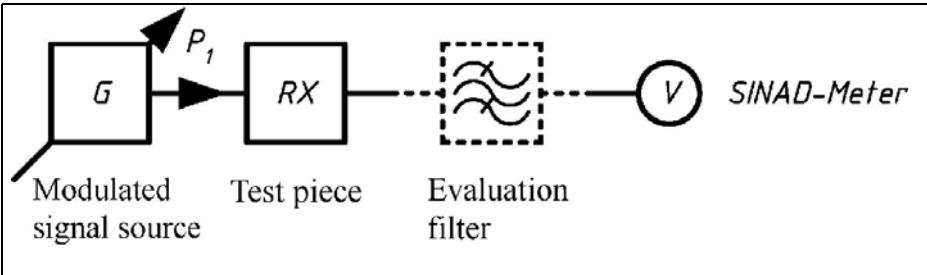
Fig 1: Main method for testing the characteristics of a radio receiver.

2.

Design of the measured variables

A signal V_{GW} splits down into the dominant mode plus harmonics (they are a measure of the signal distortion). Additionally to the signal distortion there is also more or less noise. The dominant mode represents the actual information signal. This must be larger by a given margin (expressed in dB) than the unwanted signals otherwise the information content is impaired too much.

2.1. True RMS

If DC voltage is applied to a resistor, then a constant current flows through the resistor and it warms up. The electrical power causing this heating is determined by the well-known equation:
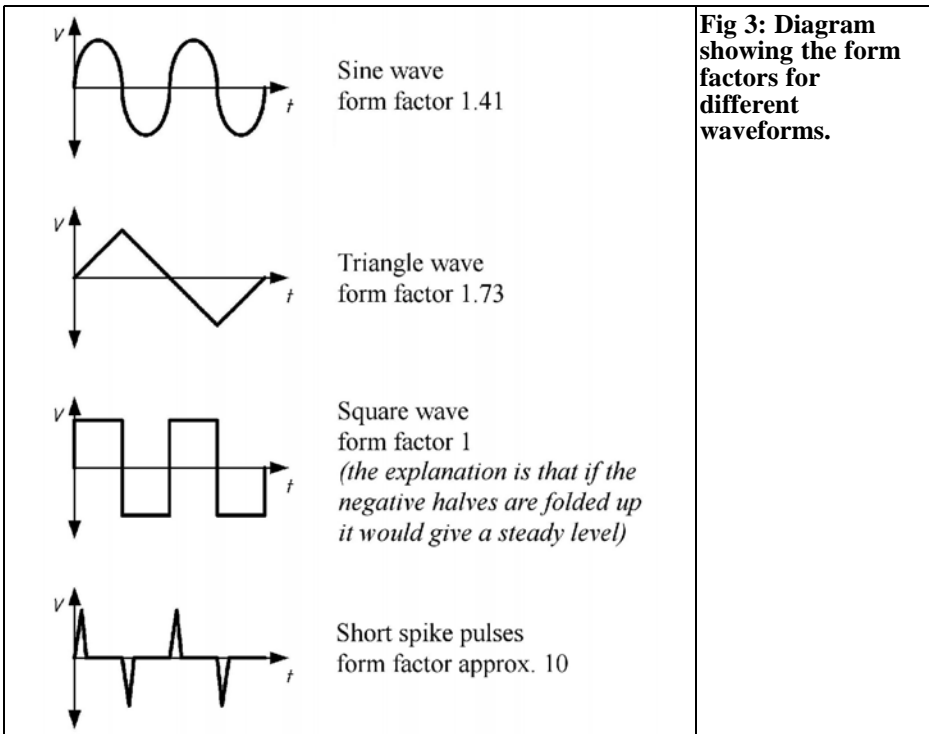
$$P = \frac{V^2}{R} \qquad (1)$$

Where:

P is in Watts (W); V is in volts and R is in Ω

If the current flowing through the resistor is alternating rather than direct then the value of current flowing is dependant on time. The resistor heats up and cools down with time but the temperature will
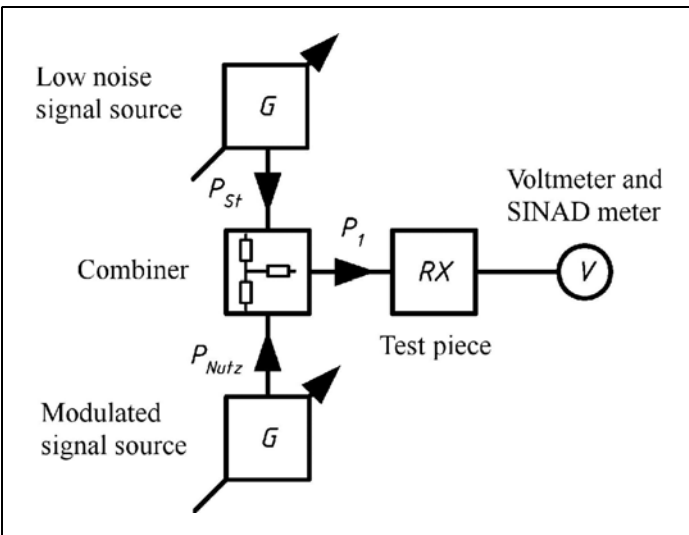
Fig 2: Method of measuring receiver characteristics using two signal sources.

Fig 3: Diagram showing the form factors for different waveforms.

settle down to an average value. In electronics the power dissipation is of interest in connection with the thermal load (see typical data sheet: max. energy dissipation $P_{tot} = \ldots$ W), it is thus usually only this average value that is of importance. The power dissipation is derived from the square of the voltage waveform. It is called RMS (Root Mean Square) and is defined as follows. The RMS is that portion of the alternating current that would cause the same thermal effect as an equivalent direct current! This can also be applied to the voltage. Thus the RMS of an alternating voltage must be smaller than the peak value. Dependent on the shape of the waveform the RMS value is related to the peak value, see Fig 3. This relationship defines the Form Factor:

The Form Factor is periodic (thus always recurring) and is value of the relationship between the peak value and the RMS value. For sinusoidal alternating voltages this is 1.41. The display on moving pointer measuring instruments using a rectifier is calibrated in RMS only for sinusoidal signals. The same applies to many simple digital instruments. However if the signal is not sinusoidal then a genuine RMS measuring instrument is needed (also called true RMS) [6].

Because noise is indefinite and constantly changing, measuring its RMS value gives a measure of the noise performance rather than trying to measure the peak value. From [8] a measuring instrument must be able to measure white noise having a form factor of 5 with a 1% error.

2.2. S/N, (S+N) /N

For a small information signal, noise is the dominant degradation factor. The relationship of the noise to a signal of the same level can be described as signal-to-noise ratio (S/N) of 0dB - both components have the same signal level. The
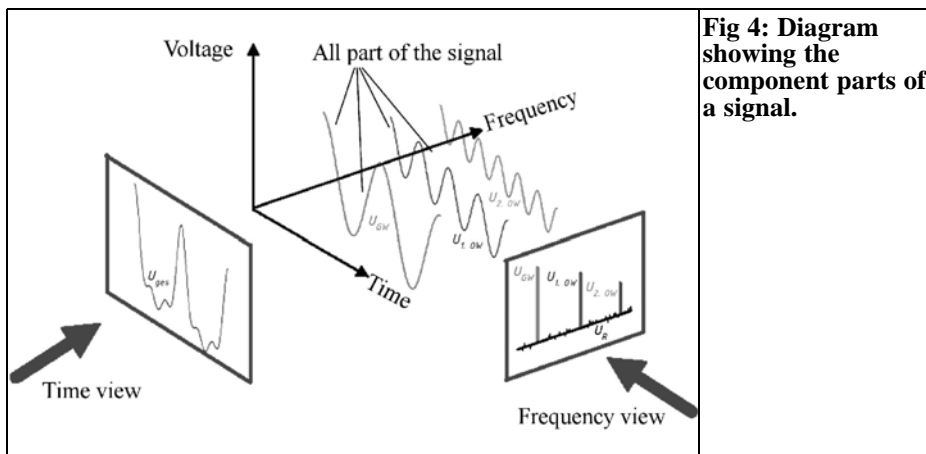
Fig 4: Diagram showing the component parts of a signal.

term Signal-to-Noise-Ratio (SNR) is common.

In practice it is not possible to determine the S/N by measuring the signal then the noise because the noise cannot be switched off. Thus the signal must be measured with the noise to determine the Signal plus Noise to Noise (S+N)/N which is the auxiliary form that surrounds us in the natural environment [9]. The larger that the signal-to-noise ratio is, the smaller the actual difference. For a moderate signal with (S+N)/N of 7dB the S/N will be 1dB lower.

2.3. SINAD

For larger signal-to-noise ratios, other parts of the signal become more dominant than the noise. These can cause harmonics during the demodulation of the AF signal that can impair the comprehensibility or "quality of understanding". Measuring the noise and all other signal impurities is the familiar SINAD value. This stands for "Signal, Noise and Distortion to Noise and Distortion", (S+N+D)/N+D. This mathematical equation (2) is:

$$SINAD = 20 \cdot \log \left(\frac{V_{ges}}{\sqrt{V_R{}^2 + V_{1OW}{}^2 + V_{2OW}{}^2 + V_{3OW}{}^2 .... + V_{nOW}{}^2}} \right)$$

Fig 4 shows how a signal can be split into it's component parts by frequency in accordance with the equation above.

Measurements of receiver sensitivity using a SINAD meter are quite efficient. For example the capture of the noise and AF signal can be automated by measuring the noise when the demodulated information fades out. The SINAD in dBs can be directly displayed. Importantly a notch filter is used in the SINAD meter to measure only the signal information at the centre frequency of the filter to avoid measurement errors. This is because most of the information in an AF signal is contained at 1kHz. In the case of poor signal processing in a radio receiver the same values are obtained for (S+N) /N and SINAD.

For telegraph (A1A modulation – Morse) a full amplitude signal is available when the code is transmitted, this is not the case for other transmission modes e.g. AM (A3E) or Single Sideband (J3E) where the amplitude of the modulation envelope varies. Since many transmission modes are complex with much more than a single tone present (e.g. the whole speech band) a certain minimum signal to noise ratio is needed for acceptable transmission. Therefore the operating sensitivity for a radio receiver is specified for a certain (S+N)/(N) or SINAD at the AF output (for a reference power output). Normally 10dBs, 12dBs or 20dBs is acceptable and are often specified in tests. A signal-to-noise ratio of 33dBs corresponds to broadcast quality.
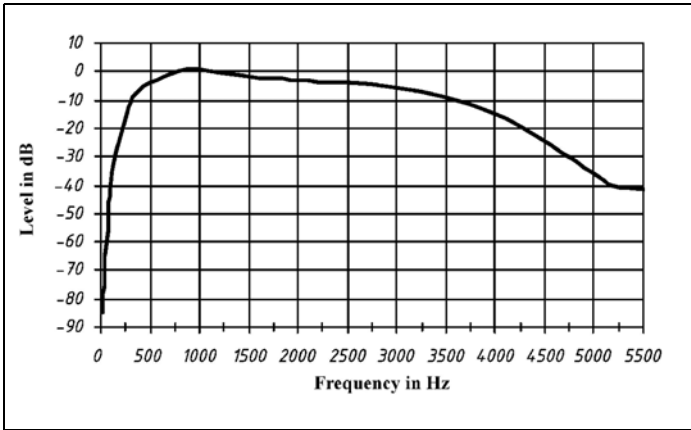
Fig 5: The response curve for a CCITT filter.

The maximum signal-to-noise ratio marks the highest attainable information signal quality. This is usually due to the noise but SINAD also measures AF harmonics that can be very meaningful. Besides the AF section of a radio receiver developing distortion, noise voltage can come from the power supply. Single sideband can generate close in noise from the different conversion oscillators [5].

2.4. Distortion factor

Instead of measuring the maximum signal-to-noise ratio the distortion factor can be measured. The distortion can come from many sources but mainly the audio stages. For a sine wave test it is assumed that all harmonics except the of dominant signal cause a decrease of the signal quality. The mathematical description of this is (3):

$$k_{\%} = \frac{\sqrt{V_{1OW}{}^2 + V_{2OW}{}^2 + V_{3OW}{}^2 ... + V_{nOW}{}^2}}{V_{ges}} \cdot 100\%$$

The distortion factor is the ratio of the RMS value of all the harmonics, except the dominant signal, to the RMS value of the whole signal, V_{ges}, expressed as a percentage (Fig 4). This is normally called the Total Distortion Factor or THD (Total Harmonic Distortion) [4]. A value of 10% for k% means that the harmonics of a signal amount to at the most 10% of the total signal.

For a radio receiver the demodulation distortion factor is an important parameter. This is measured for a reference power output for several defined RF signals at the antenna socket. The RF signals are modulated with nominal modulation. Good designs exhibit demodulation distortion factors of under 1% with large input signals.

2.5. Weighting filter

The frequency range of the measuring instrument is restricted with a weighting filter. In the strictest case this is a bandpass filter shaped to the response of the human hearing as used for telecommunication transmission. This is usually the standard curve described by the CCITT standard P53A (general known as a CCITT filter) as shown in Fig 5. As a result of using a weighting filter the signal-to-noise ratio measured will be higher and the distortion factor will be lower. This is due to a decrease in the noise and the missing harmonics that are outside the filter pass band, thus on paper receiver sensitivity is better. For radio receivers with different AF output signals, i.e. special decoders, the results will not be meaningful. For meaningful comparisons it is important to specify the conditions of the test.

Fig 6: Data sheet for the Analog Devices [14] ADF536A.

3.

The concept

3.1. Handling different RMS voltages

The Analog Devices [14] AD536A shown in Fig 6 is an integrated circuit containing a true RMS to DC converter. The output voltage is proportional to the RMS value of the input voltage. For a Form Factor of 7 the error is less than 1%, even with a Form factor of 10 it is still only 2.5%. This transfer function of the IC can be described in principle as:

$$V_{OUT} \propto \sqrt{\frac{1}{M} \cdot \sum_{m+1}^{M} V_{in}^{\,2}(t_m)} \qquad (4)$$

Where:

V_{out} = output voltage proportional to the rms value of the input signal

M = number of samples over the averaging period

V_{in} = input signal at any point in time (t_m)

This is the correct mathematical definition of the internal operation. The length of time that is taken to measure the average can be determined simply by an external capacitor. An internal buffer can be used to drive a moving pointer micro-ammeter with a current proportional to the RMS value. The scale of this meter will be linear in RMS volts.

A wide band RMS converter is needed to measure all of the harmonics in the signal. From the response of the human ear, 16kHz is the obvious AF range. The AD536 has an error of less than 1% up to 45kHz for an input of 100mV. With smaller input voltages it's range drops rapidly so that at 10mV input it is 5kHz for less than 1% error. This means that an amplifier will be required for measuring small signals. Using a change of amplification at the correct point the resolution can be maintained by stretching the input range. A proven way to stagger the changes in amplification is to use switched gains of 1/3.169/10/31.69. This gives a downward step of 10dB each time the gain is switched.

$$20 \cdot \log\left(\frac{31.69}{10}\right) = 20 \cdot \log\left(\frac{10}{3.169}\right) = 10dB$$
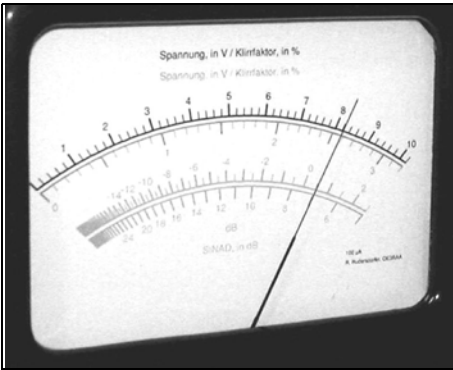
Fig 7: The output meter has a special scale to show all of the ranges that can be measured.

This is shown in the block diagram, Fig 8, it can be bypassed when the switch is set to V_{RMS}.

3.2. Development to the automatic SINAD meter

As described in the introduction, the signal to be measured consists of the dominant mode plus signal distortion and the noise. The signal to be measured (V_{ges}) can contain three signal components so using equation (2) we can develop equation (5).

$$SINAD = 20 \cdot \log \frac{\sqrt{V_{GW}^2 + V_R^2 + V_{1OW}^2 + V_{2OW}^2 + V_{3OW}^2 ... + V_{nOW}^2}}{\sqrt{V_R^2 + V_{1OW}^2 + V_{2OW}^2 + V_{3OW}^2 ... + V_{nOW2}}}$$

Because all values are RMS they can

only be summed by squaring and then taking the square root. It can be seen that the term on the top of the equation differs from the bottom term by the addition of V_{GW}^2. If the measured signal can always have a constant RMS voltage then the term on the top of the equation will have a constant value.
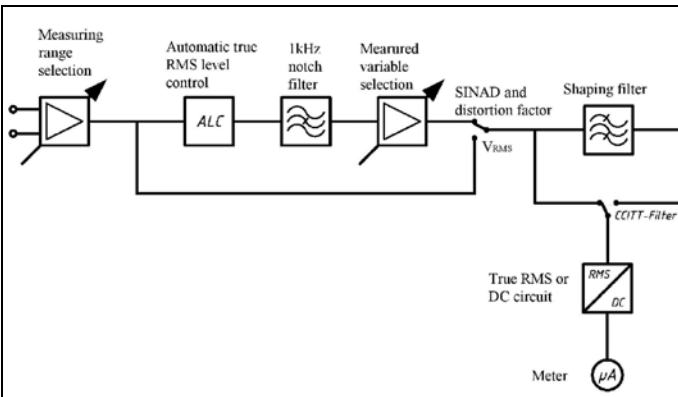
This can be realised by using an ALC (automatic level control) circuit, this must handle high Form factors. If the primary signal V_{GW} is held constant or "locked out" then the RMS value of the remaining signal is the direct relationship between the two RMS signals and can be displayed on an output scale. The SINAD is higher for smaller values of noise and harmonics so a high SINAD gives a small pointer movement. The sensitivity for measuring SINAD values is good and is predominantly determined by the self-noise of the meter circuit.

To represent SINAD correctly in dB the logarithm of the result must be taken and then multiplied by 20. This is a mathematical process that is applied to all values. This can be achieve using a logarithmic scale on the meter as shown in Fig 7.
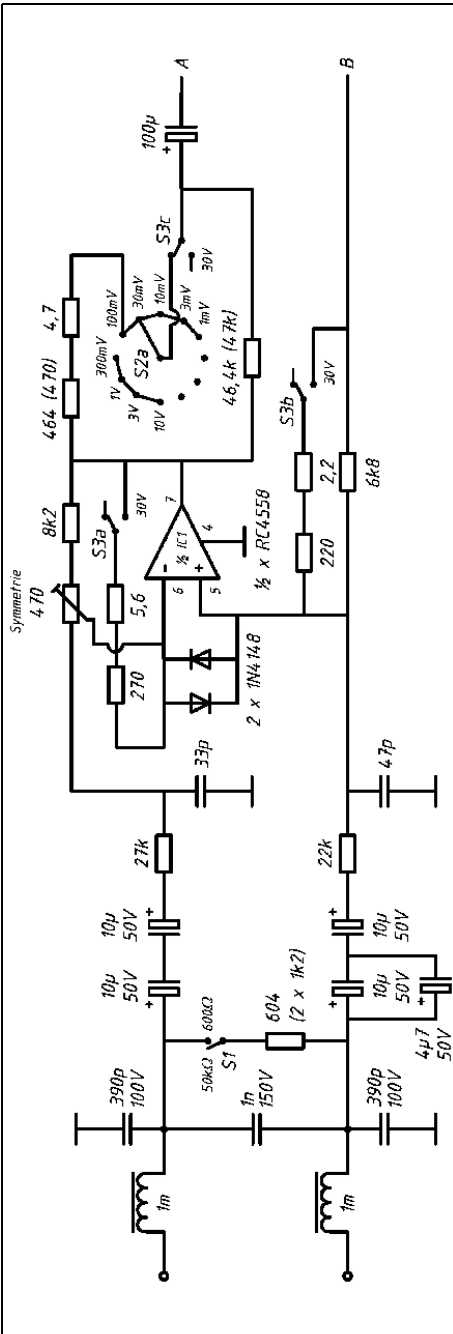
3.3. Expansion for distortion factor measurement

Looking at the equation for distortion, (3) can be expanded to give equation (6).

$$k_{\%} = \frac{\sqrt{V_{1OW}^2 + V_{2OW}^2 + V_{3OW}^2 ... + V_{nOW}^2}}{\sqrt{V_{GW}^2 + V_R^2 + V_{1OW}^2 + V_{2OW}^2 + V_{3OW}^2 ... + V_{nOW}}} \cdot 100\%$$

Fing 8: Block diagram of the SINAD meter with facilities to measure distortion factor and true RMS voltage.

Fig 9: Input circuit of the SINAD meter.

Comparing equations (5) and (6) there are some clear similarities, however:

- The denominator in (6) is the same as the numerator in (5)
- The numerator in (6) is the same as the denominator in (5) but the term V_R^2 is missing

Therefore the considerations already made in the preceding section apply. The situation is now reversed so that the smaller the measured RMS value of the harmonics with locked out primary signal, the smaller the distortion factor will be. This means that good values of distortion factor will result in a small pointer movement on the meter. The scale for this measurement is linear and shown as the two upper scales in Fig 7. These are also used for the RMS voltage measurement.

As mentioned in section 2.2, noise cannot be switched off and it cannot be isolated from the signal to be measured. This is a problem because it is not in the definition of distortion factor. Fortunately the harmonics are usually much larger than the noise so it has an infinitesimal influence on the measured result. Even if this is not the case, the measured value is meaningful, since it gives information about the signal quality.

3.4. Concept summary

The block diagram of the instrument is shown in Fig 8. The signal is amplified with a switchable gain amplifier, this makes it possible to make good measurements of SINAD and distortion factor by stretching the measurment range. An additional weighting filter can be used if required.

4.

Circuit description

A balanced input is used to prevent hum loops and the effects of other input disturbances [10].
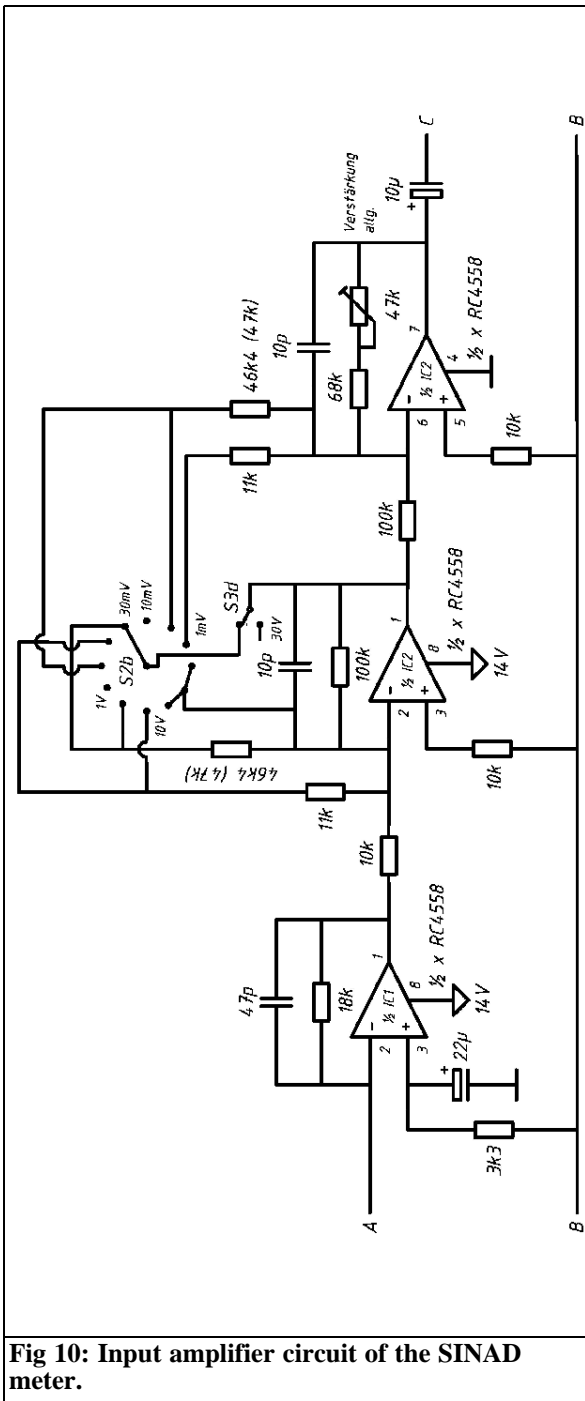
Fig 10: Input amplifier circuit of the SINAD meter.

4.1. Circuit diagram: Fig 9

The 1mH chokes and capacitors at the input prevent any RF that may be on the input signal from getting into the measuring circuit. The switch S1 can be used to select between a high impedance input impedance of approximately 50kΩ and a load of 600Ω usual used for audio systems. The two 10µF electrolytic capacitors form a unipolar capacitor giving galvanic isolation from the following circuit. The high capacitor values are critical for a low frequency response. The 470Ω trimmer potentiometer can be adjusted for accurate symmetry. If necessary the 4.7µF electrolytic capacitor can be changed for a somewhat smaller value.

The input of the operational amplifier is protected from large voltage differences by two back-to-back diodes. S2 in Fig 9 and 10 is a standard two wafer rotary switch with twelve positions but only nine positions are used. It is used to gradually change the amplification and thus the measuring range. Signals under $250\mu V_{eff}$ can actual be measured perfectly. The largest measuring range is 32V full scale giving a signal processing range of more than 102dB.

The 0Ω connections for the measuring of ranges 1mV to 100mV, and 300mV to 10V on the rotary switch S2a are not on the printed circuit board, they are soldered directly at the switch.

An option is provided to change the measuring range to the largest value at any time while leaving all the other
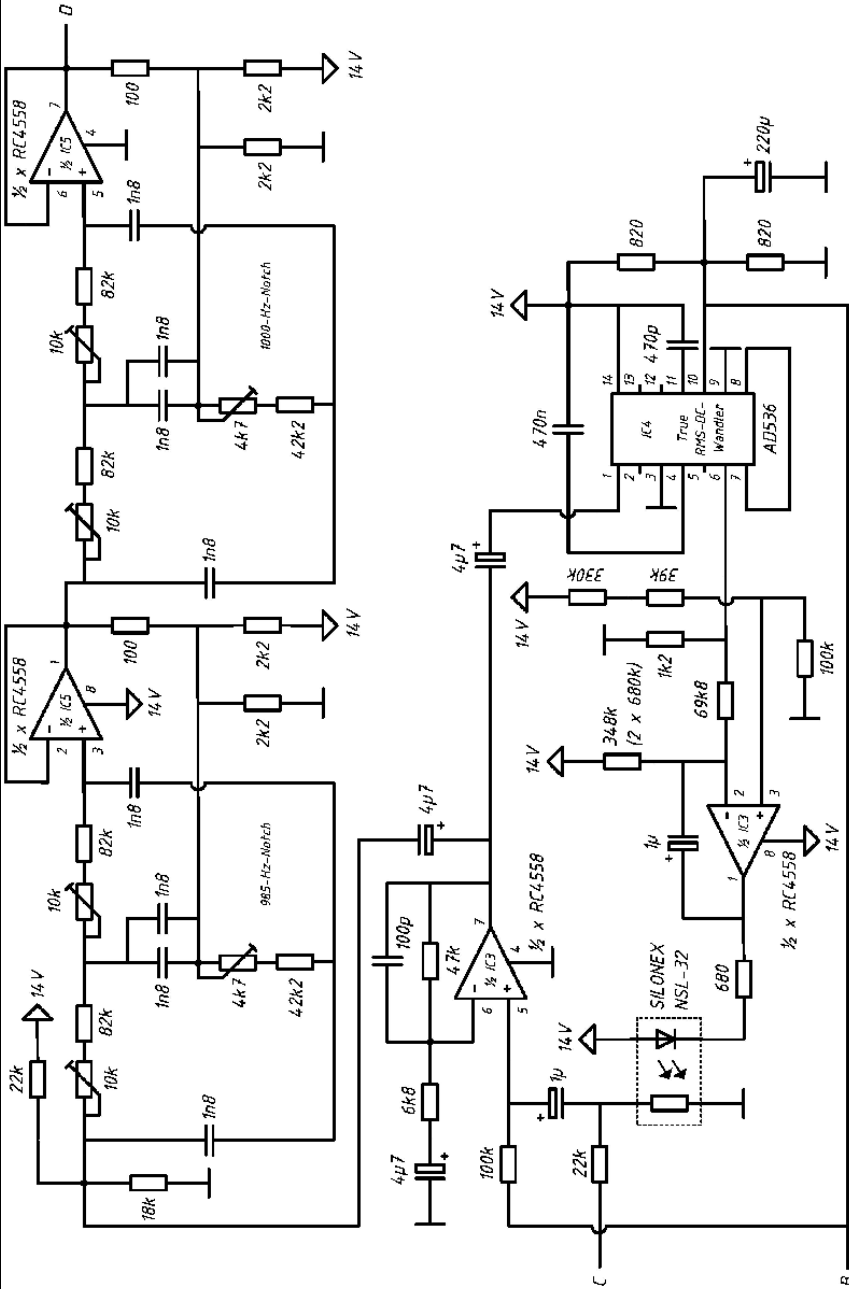
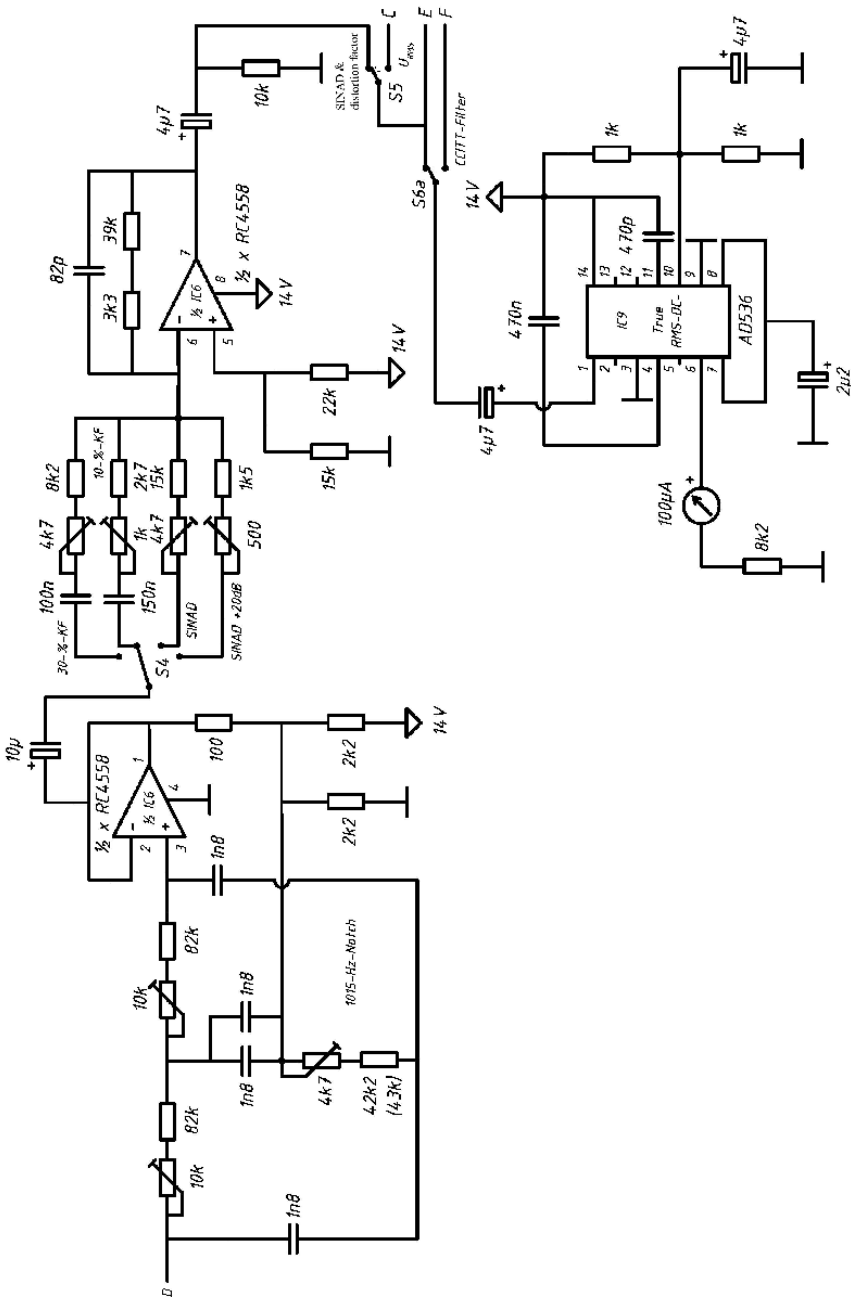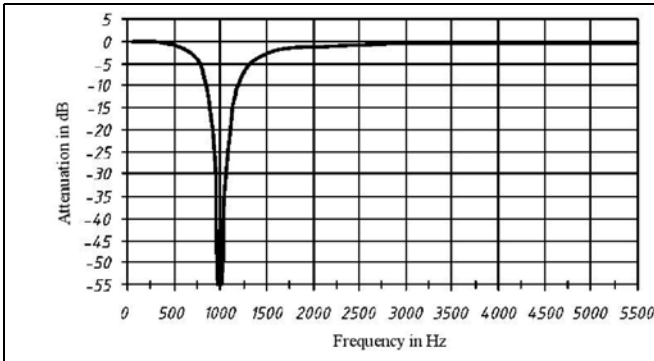Fig 11: Notch filter and ALC circuits of the SINAD meter.

Fig 12: Range selection and output amplifier circuit of the SINAD meter.

Fig 13: Frequency response of the 1kHz notch filter.

setting the same. This uses the four-pole rocker switch S3. This helps to protect the moving coil meter when measuring unknown signals. S3 can be switched to take the pointer off of the end stop while the least sensitive range is selected.

4.2. Circuit diagram Fig 10

The non inverting inputs of the following operational amplifier chain are connected to half of the supply voltage by a resistor divider, therefore a single positive supply is all that is required. The amplification range can be set using the 47kΩ trimmer potentiometer in the feedback path of IC2. The parallel 10pF capacitor minimises the high frequency response. The same applies to the other 10pF and 47pF capacitors. The AC portion of the signal is coupled to the true RMS level control using a 10µF electrolytic capacitor.

4.3. Circuit diagram Fig 11

The 22kΩ resistor forms a variable voltage divider with the photoelectric cell (like a light-dependent resistor) in the opto-isolator NSL-32 from Silonex [16]. This opto-isolator is ideally suited as described in the product description [17] it says: "The resistive output of this type of device has very low distortion characteristics". The true RMS to DC conveter, IC4, gives an output proportional to the true rms value of the signal from IC3, pin 7. It is available at pin 6 across the 1.2kΩ resistor.

The following circuit formed by the second half of the operational amplifier

IC3 works as integrator. The output at pin 1 feeds current to the LED in the opto-isolator as long as the voltage on the negative input is not equal to the 3V on the positive input. The illumination of the LED changes the resistance of the photo-electric cell between some hundred ohms and over 200kΩ. The changes of resistance take place slowly with the duration to the final value in the seconds range. This give the ALC it's control system, it is impossible for it to react to short voltage peaks and the response is to the measured signal.

A signal with a constant level of approximately $710mV_{eff}$ is fed via a 4.7µF electrolytic capacitor to the input of the notch filter. It corresponds to an accurate full-scale image of the measured signal. Measurements confirm that the level is kept extremely constant at this point, even if the measuring range were wrongly selected (see Fig. 16).

The 1kHz dominant mode of the measuring signal is eliminated by three cascaded band pass filters (Fig 11 and Fig 12). They can be easily adjusted to different notch frequencies so that faulty measurements are excluded by slightly deviating from the 1kHz frequency. The literature [11], shows that these filters require very little calculation. The 100Ω resistors at the output of the operational amplifiers in the feedback loop, easily expand the maximum stop band attenuation. This gives an accurate band pass filter with steep sides, Fig 13 shows the actual filter
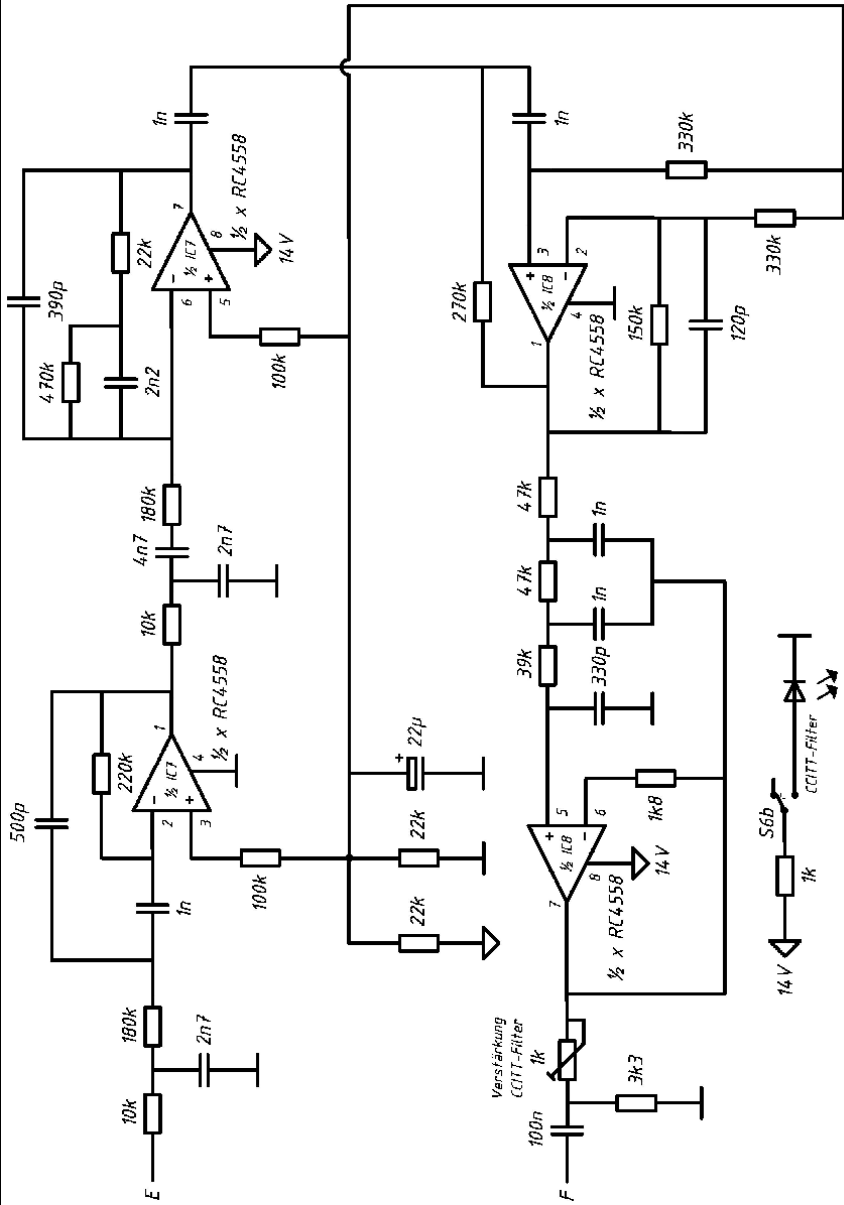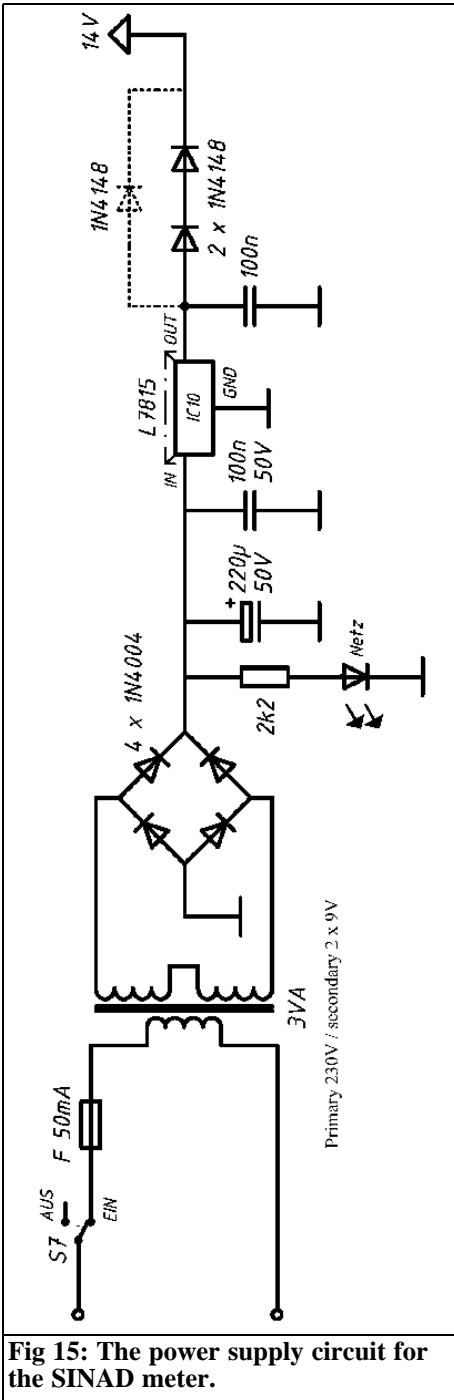
Fig 14: The CCITT filter circuit of the SINAD meter.

Fig 15: The power supply circuit for the SINAD meter.

characteristic. It is characterised by high stop band attenuation and very high quality. Also the first harmonic of the measured signal is covered. The small pass band attenuation at 2kHz meets the requirements for SINAD measurements.

4.4. Circuit diagram Fig 12

Switch S4 is used to select the different distortion factor measuring range and SINAD measuring range by switching the amplification of the following inverting amplifier. Each of the four trimmer potentiometers ("30% -DF", "10% -DF", "SINAD", "SINAD +20dB") are for fine adjustment of each individual measuring range.

The capacitors in series with the two branches for distortion factor measurement limit the frequency response to low frequencies. Signals below the first harmonic of the measured signal are without meaning. Therefore the dynamics of the distortion factor measurement range is increased slightly. The 82pF capacitor in parallel with the resistors in the feedback branch of the operational amplifier limits the high frequency response.

Switch S5 selects the measured variable, for RMS (V_{RMS}) the signal before the ALC is measured. The signal decoupled from the DC voltage is the true RMS to DC transducer, IC9, input signal. It directly feeds the moving coil instrument with 100μA full-scale.

4.5. Circuit diagram Fig 14

If required the CCITT filter can be connected using switch S6. The circuit with operational amplifiers IC7 and IC8 gives a frequency response in accordance with CCITT P53A. The trimmer potentiometer at the output stages can be used to align the filter.

4.6. Circuit diagram of the power supply (Fig 15)

The circuit of the power supply is shown in Fig 15 with the PCB layout in Fig 17 and the component layout in Fig 18. The actual stabilised DC voltage required is 14V. It is obtained using an economical
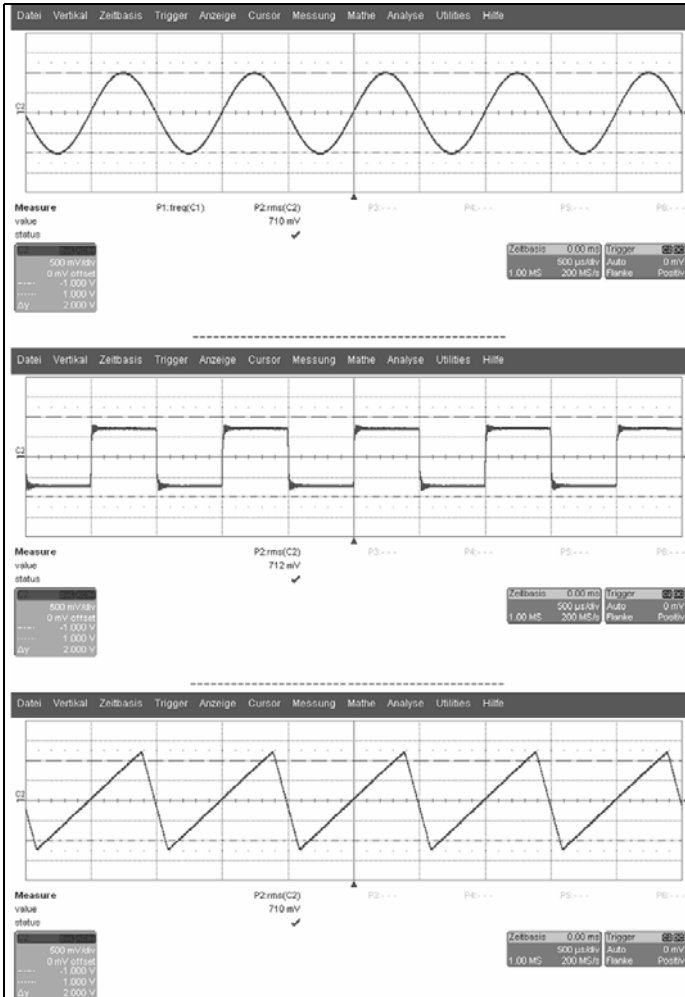
Fig 16: The signal at the output of the ALC on pin 7 of IC3 with different inputs. For example with signals between 140mV and 3V fed into the 1V measuring range the output is independent of the signal form and always a constant RMS value. The output is always 710mV.

15V regulator and depending on the output voltage either one or two silicon diodes.

5.

Construction

5.1. Selection of suitable components

Since this is a measuring instrument, resistors with 1% tolerance should be used. Some of the resistors shown in the circuit diagrams with a value in brackets are quite rare so the value shown in brackets can be use with a small loss in accuracy. Similarly capacitors with close tolerances save problems with alignment. This particularly applies to components in filters where foil capacitors or small value ceramic capacitors with NP0 dielectric should be 5% tolerance. If not otherwise stated components should be rated for at least 16V. The 500pF capacitor on pin 1 of IC7 is made up from a
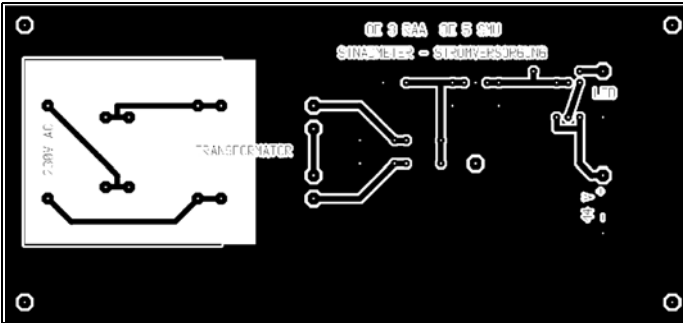
Fig 17: PCB layout of the power supply for the SINAD meter.

470pF capacitor with a 27pF capacitor in parallel, these could be SMD 0804 or 0603 size.

The PCB for the power supply has been designed to accept various transformers with E-I cores from 3 to 3.8VA. Using different transformers may result in some unused solder connections on the PCB.

To ensure that the reading accuracy and measurement tolerance are maintained a large 100µA moving coil meter with 1% to 2% accuracy should be used.

The AD536A is available in various packages, the PCB shown in Figs 19 and 22 are designed for the DIP14 package. The more economical TO100 package can be used for the AD536AJH IC4 and IC9, an adaptor PCB, shown in Fig 25, was designed to adapt this package to a DIL footprint. To use the adaptor fit solder pins for the DIL positions and form the leads of the TO100 package into an oval shape before fitting to the adaptor. The rest of the instrument uses the well-proven RC4558 low noise op amplifier from Texas instruments [15].

5.2. Scale layout

The scale for the meter was produced using software available from James Tonne, WB6BLD, [19]. It was written to design meter scales and has the facilities to accommodate various meter scale sizes, there is an application note (How to do it appnote) on the web site [19] that describes exactly how to re scale a meter. Three indicator scales with different division are required, a fourth for general use to read off is dB is useful. Two scales need linear divisions with an initial value of 0 and a full scale of 10.2 and 3.2 (for voltage and distortion factor). Because the switching of voltage measurement is in 10dB steps these two ranges have their full scale deflection at the same pointer angle. In the menu "Scale Design" of the design software the "Linearity Exponent" should be set to 1 for these two scales.
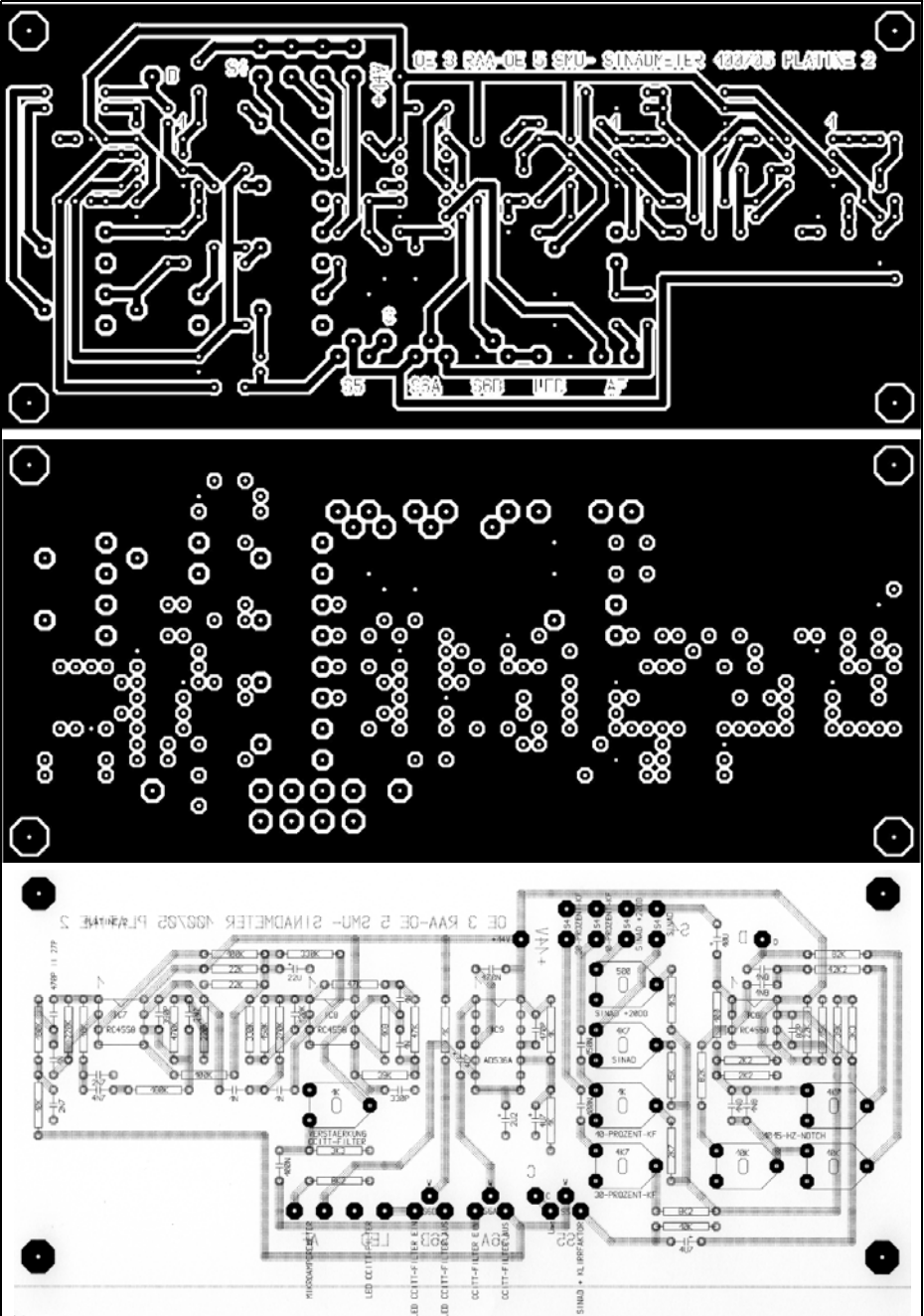
The two further scales require a logarithmic layout with increasing compression towards the left end division and thus a ever greater compression to the left end. The beginning of the scale represents
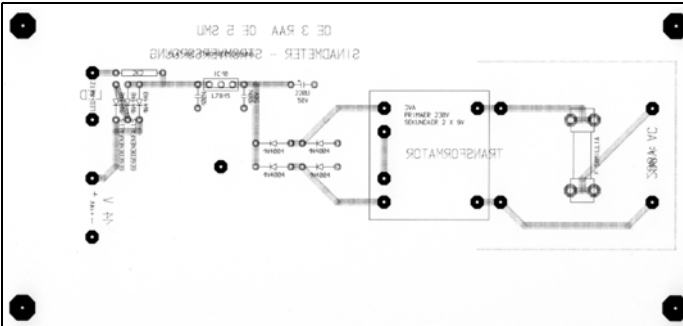
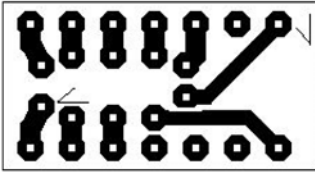Fig 18: Component layout of the power supply for the SINAD meter.
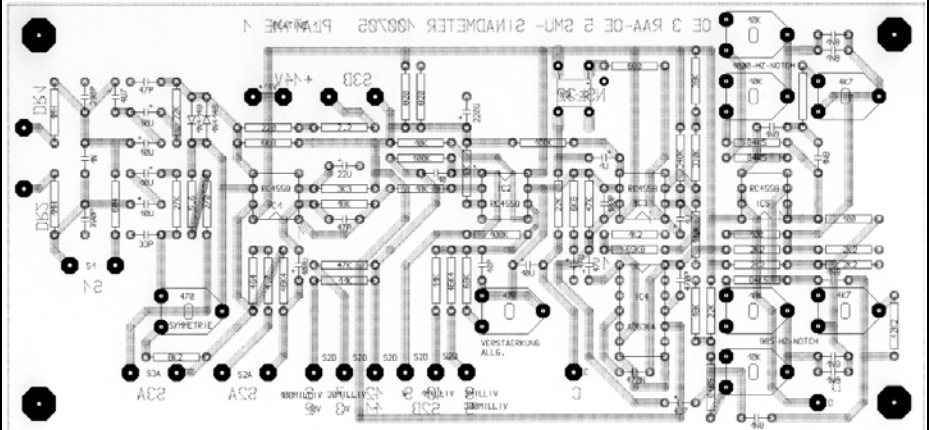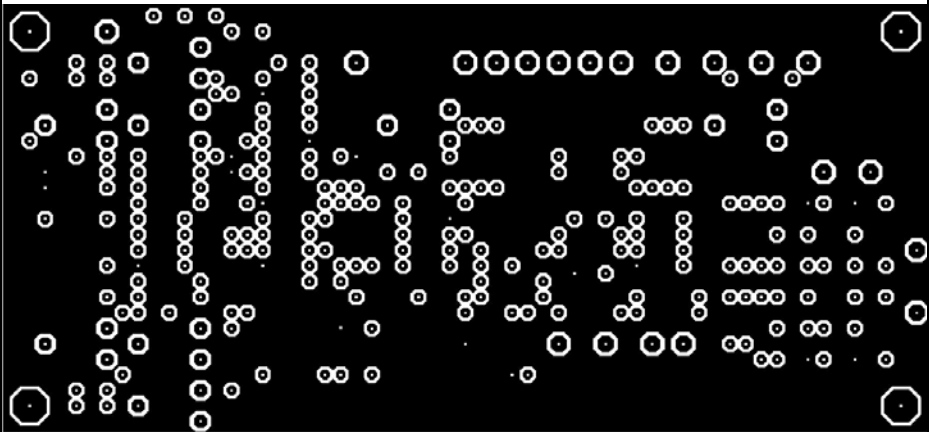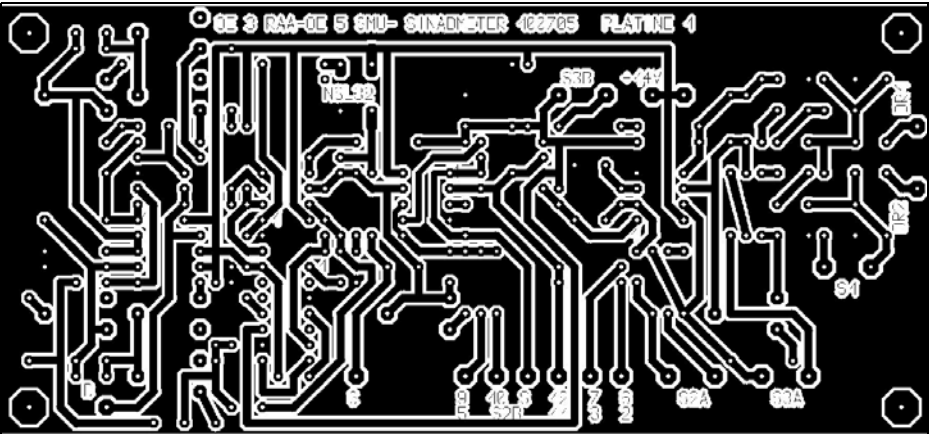
Figs 19 - 21: Layout of PCB 1 for the SINAD meter. The full size is 71.1 x 152.5mm.

Figs 22 - 24: Layout of PCB 2 for the SINAD meter. The full size is 71.1 x 152.5mm.

Fig 25: Adaptor PCB for TO100 case. PCB size 10.3 x 19mm.

minus infinity.

In common with the AF standard, the db scales use 775mV as the 0dB reference. Measuring –34dB on the 30dB range reads –4dB which is smaller than 0dB. A mathematical relationship is required to represent this difference and calculate the linearity exponent to draw the scale. In addition the reference point for full scale must be maintained.

where:

$$x_b = \frac{V_{Bez}}{V_{max}} \qquad (7)$$

x_b = reference factor
V_{Bez} = reference voltage on that linear voltage scale, in V
V_{max} = full scale on the linear voltage scale, in V

To draw the logarithmic scales with the same full scale deflection angle of the pointer with the correct initial value in dB the following formula is needed:

where:

$$y_s = \frac{\log(x_b)}{\log\left(\dfrac{y_1}{20 \cdot \log(x_b) + y_1}\right)} \qquad (8)$$

y_s = linearity exponent
x_b = reference factor from (7)
y_1 = initial scale value, in dB

In order to avoid the unmanageable minus infinity, for y_1, a manageable value is selected. A good compromise is -68dB. Any errors are only small [6]. An example is:

$$x_b = \frac{0.775}{1.02} = 0.7598$$

therefore the linearity exponent becomes:

$$y_s = \frac{\log(0.7598)}{\log\left(\dfrac{-68dB}{20 \cdot \log(0.7598) + (-68dB)}\right)} = 7.97$$

A linearity exponent of 7.6 is the best to use with the "Meter" software [19] With divisions every 0.5dB gives 141 division lines (-68dB to 2.5dB).

The fourth scale, is SINAD in dB showing full dBs bearing in mind that a SINAD of less than 5dB hardly makes sense. Using the same linearity exponent of 7.6 the scale in dB has 71 divisions (76dB to 5dB SINAD).

The software only permits evenly distributed legends to be printed over the entire scale, for example at each third scale line. For the logarithmic scales this causes the legends to be squeezed up at the left hand end. To overcome this the finished scale is printed to a pdf or other picture format file and the unwanted legends deleted. The scale shown in Fig 26 was produced by this method.

5.3. Construction and mechanical details

Great importance was paid to the housing of the prototype particularly for serviceability and ergonomic layout of the controls. The complete interior is accessible after the cover plate and bottom plate have been removed, see Fig 27. This gives the best situation for changes and experiments. The front and back plate form part of the chassis, closer details are shown in Fig 28 and 29. The front panel was manufactured from copper clad FR4, which combines the advantages of good mechanical workability and stability, as well as the possibility for the front panel to be designed with a panel layout program. After polishing the copper side is covered with protective varnish. The controls are arranged on the basis of use and technical criteria, see Fig 30. The impor-
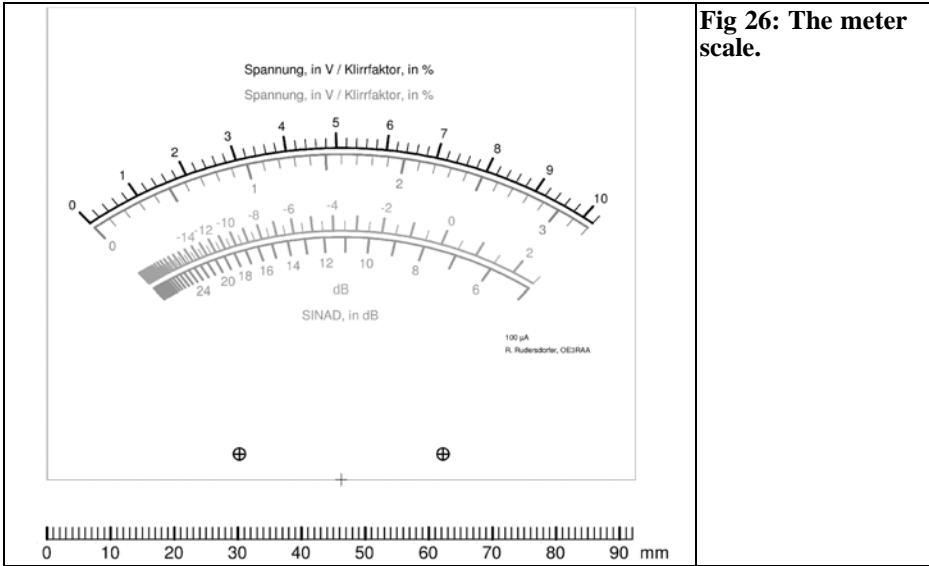
Fig 26: The meter scale.

tance on good accessibility was attached to individual buttons and switches. A further criterion was that each control function should have its own physical control.

The complete metal housing is grounded, which results in good shielding as well as good safety practice. The individual printed circuit boards should be isolated and have their ground surfaces connected to the common ground with a short piece of wire from one point. Thus the inputs are floating and their symmetry is maintained. The input socket on the front panel should be connected to the input on PCB 1 with wires of the same length to preserve the symmetry. The points "C" and "D" on PCB 1 and PCB 2 should be connected with a short piece of coaxial cable with the screen connected to ground at one end. In some cases the wiring from S3 can cause undefined signals on the 3mV measuring range. The solution to this is to solder a 4.7pF (10pF maximum) across the connections to S3a on PCB 1. For the remaining signal wiring in the instrument, screened cable such as RG174 should be used.

The ideal connectors for the symmetrical input are BNO connectors from Huber+Suhner [20]. Outwardly BNO plugs resemble BNC plugs with their bayonet fixing but they have two internal connections, see Fig 31. Suitable test leads can be made, as shown in Fig 32, that have screened leads almost up to the tip. A bundled test lead was planned for the prototype including two leads with 4mm banana plugs, this can be used in the laboratory.

6.

Switch on, alignment and calibration

All preset potentiometers should be set to their central position. Before the two main PCBs are connected to the power supply, check that with two diodes connected in series with the power supply output as shown in Fig 15, the output voltage is 14V with a 80mA load. With PCB 1 and PCB 2 connected and 14V supply voltage the supply current should be between 75mA and 95mA.
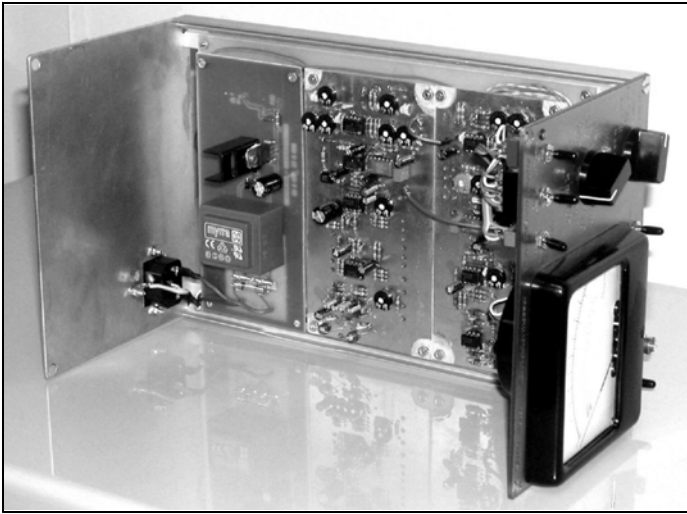
**Fig 27: Picture of the completed prototype with the cover removed.**

6.1. Alignment of the true RMS voltmeter

*Front panel Switch positions: Voltmeter, CCITT filter out, 1V or 0dB, 50kΩ symm.*

Equipment needed: Signal generator (sine wave generator), 1kHz and adjustable output level

Feed 775mV into the symmetrical input and adjust the preset potentiometer marked "Verstärkung allg" (general gain), see Fig 21, to give a reading of exactly 0dB. If the signal generator used does not have an output meter, use a suitable measuring instrument; it does not need to be a true RMS instrument. To align the circuit for the maximum possible symmetry (common mode rejection)

the same test signal should have one side connected to the common ground at the input socket The preset potentiometer marked "symmetrie" (symmetry), see Fig 21, should be varied until the point of the smallest pointer deflection is found on one of the sensitive voltage measuring ranges. The same procedure is recommended at 50Hz and 20kHz, then the best compromise between the three points can be found. With this adjustment the common mode rejection should over 45dB. In critical cases the 4.7µF electrolytic capacitor in the lower input arm, see Fig 9, can be changed for 2.2µF or 1µF. A fine adjustment of the general gain potentiometer may be required after this.
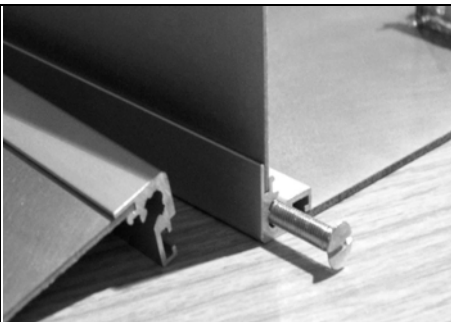
Fig 28: Detail of panel assembly.
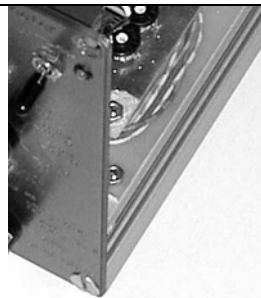
Fig 29: Detail of panel mounting.
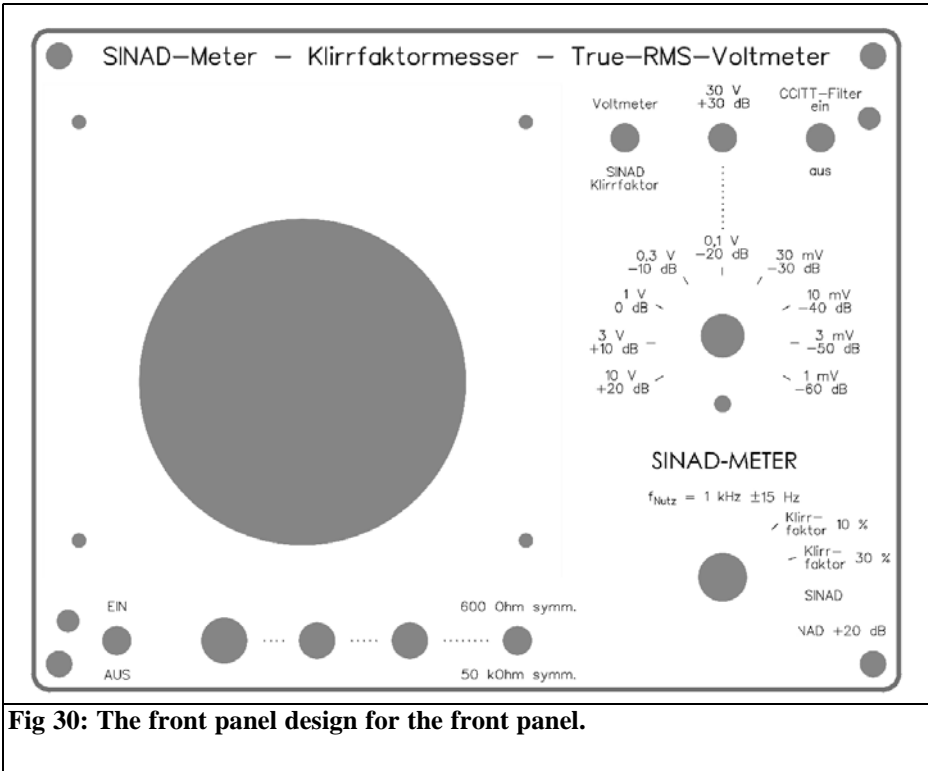
Fig 30: The front panel design for the front panel.

If the recommended components have been used, no further adjustments should be needed on the other measurement ranges. If an examination of individual ranges is required, this is best made using the following data with a 1kHz input signal:

Measuring range	Full scale input
30V or +30dB	24.5V
10V or +20dB	7.75V
3V or +10dB	2.45V
0.3V or –10dB	245mV
0.1V or –20dB	77.5mV
30mV or –30dB	24.5mV
10mV or –40dB	7.75mV
3mV or –50dB	2.45mV
1mV or –60dB	775μV

For each range the meter reading should be within ±4dB of that for 0dB input. The frequency response can be examined in the same way with an input of 2.45V on the 3V or 10dB range at frequencies of 30, 300, 10,000 and 20,000Hz. Again the meter reading should be within ±4dB

of that for 0dB input.

6.2. Alignment of the CCITT filter

Front panel switch positions: Voltmeter, CCITT filter, 1V or 0dB, 50kΩ symm.

Equipment needed: Signal generator (sine-wave generator) with 800Hz adjustable output level

Feed 775mV into the symmetrical input and adjust the preset potentiometer marked "Verstärkung CCITT-Filter" (CCITT filter gain), see Fig 24, to give a reading of exactly 0dB.

If the recommended components have been used, no further adjustments should be needed on the other measurement ranges. If an examination of the frequency response of the CCITT filter is required, this is best made using the following data with a 775mV input signal:
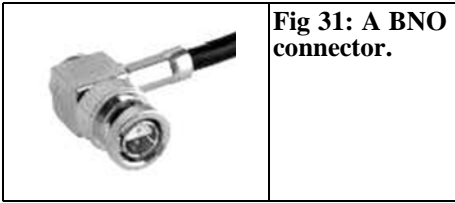
Fig 31: A BNO connector.

Measurement range	Measurement frequency	Reading
0.1V or -20dB	200Hz	-1dB (±2dB)
1V or 0dB	400Hz	-6.3dB (±1dB)
1V or 0dB	1kHz	1dB (±1dB)
1V or 0dB	3kHz	-5.6dB (±1dB)
0.3V or –10dB	4kHz	-5 dB (±dB)

The filter is within the permissible tolerance if these measurements are obtained. They are defined in CCITT P53A (now also ITU-T recommendation O.41).

6.3. Alignment of the 1kHz notch filter

Front panel switch positions: SINAD & distortion factor, CCITT filter out, 1V or 0dBs, SINAD +20dB, 50kΩ symm.

Equipment needed: Signal generator (sine wave generator) with 985/1000/1015Hz and adjustable output level, (oscilloscope with low ranges and probe)

Feed 1V at 985Hz into the symmetric input. Adjust all three preset potentiometers in the 985Hz notch (Fig 21) alternately in small steps until the voltage measured on the pole of S4 is the smallest possible value. Repeat this procedure with signals of 1000Hz and 1015Hz and adjust the relevant three preset potentiometers in the 1000Hz and the 1015Hz notch filters until the optimum result is obtained. With the filters correctly aligned the measured signal should be less than 10mV, see Fig 33. Use a 1:1 probe on an oscilloscope to measure the signal; this will give the best resolution. If a digital storage oscilloscope is used, the average of several traces can be used to get extra resolution. If an oscilloscope is not available the filters must be adjusted for minimum meter reading.
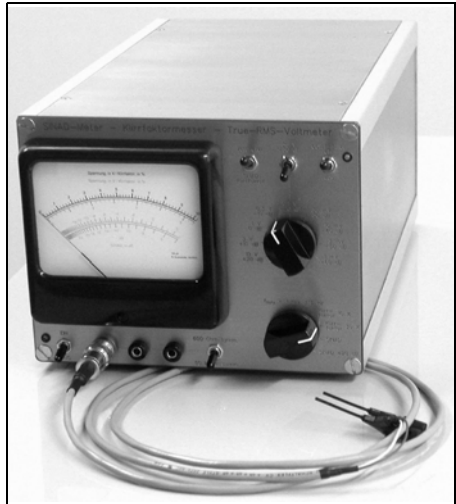
Fig 32: Picture of the completed SINAD meter with a set of test leads.

6.4. Alignment of distortion factor measurement and SINAD meter

Front panel switch positions: SINAD & distortion factor, CCITT filter out, 3V or +10dBs, measured variable selector switch see below, 50 kΩ symm.

Equipment needed: two signal generators (sine wave generators) with adjustable output level - one with 1kHz and a second with a single frequency between 2kHz and 10 kHz, power combiner

Feed a 1V signal at 1000Hz and a 250mV signal at the second frequency into the symmetric input using a power combiner. With selector switch set to "SINAD" and use the "SINAD" preset potentiometer, see Fig 24, to set the measured value to 12dB SINAD. Set the selector switch to "distortion factor 30%" and use the preset potentiometer "30-Prozent-KF" (30% distortion factor), see Fig 24, to set the measured value to 25% distortion factor. Now reduce the 250mV signal to 70mV, set the selector switch to "distortion factor 10%" and use the preset potentiometer "10-Prozent-KF" (10% distortion factor), see Fig 24, to set the measured value to 7% distortion factor. Reduce the 70mV to 39mV and set the

Fig 33: The ikHz signal after the notch filter. (1mV/Div., 500μs/Div.)
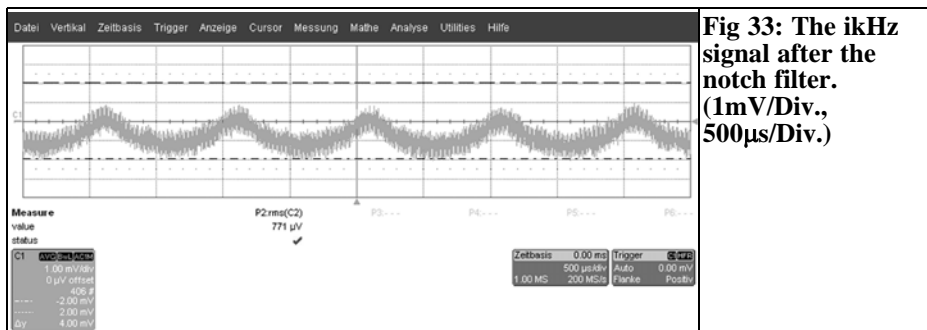
selector switch to "SINAD +20dB" and use the preset potentiometer "SINAD +20dB", see Fig 24, to set the measured value to 8dB SINAD (actually 28dB SINAD).

If the output of the signal generators and the power combiner used are designed for 50Ω, then use two 100Ω resistors in parallel to provide the correct load. The attenuation of the combiner and the low impedance must be considered; the voltage values given always apply directly at the symmetrical input.

7.

Conclusion

When measuring distortion factor or SINAD the measured signal should be within the limits of the measuring range being used, this should be at least one fifth of the total scale. The complete dynamic range can only be used on the 1mV measuring range with the CCITT filter activated.

This is a valuable instrument for use in the workshop to measure radio receivers. The circuits used have been explained in great detail. The components required can be sourced easily and there are no complications when making the PCBs because all wired components are used, this means the meter is easily reproduced. The achievable results are summarised in Table 1 and they speak for themselves!

The book shown in [13] is a rich source of suggestions and circuits. Thanks are due to Markus Pichler and Erwin Schimbäck of "The Linz centre of Competence in Mechatronics" for long discussions about constructional and technical detail.

Table 1: Performance data.

Ranges:	
Voltage	1mV/3mV/../10V/30V (0dB, 775mV)
Distortion Factor	10% / 30%
SINAD	5dB...>43dB
Filter for Distortion factor and SINAD:	
Notch filter 985Hz...1015Hz	
Measurement error:	
Voltage	<0.4dB + tolerance of meter
Distortion factor	<0.4dB + tolerance of meter
SINAD	<1dB in the upper part of scale otherwise <2dB + tolerance of meter
Error with high form factor:	
	<0.1dB for form factor up to 7 and ,0.3dB up to 11
Own distortion factor:	<0.2%
Frequency response:	<0.4dB between 20kHz and 45Hz for 30V to 0.1V. 20Hz to 22kHz with CCITT filter
-3dB frequency:	6Hz
Input:	Symetrical 600Ω or 50kΩ
Common mode rejection:	>55dB at 50Hz
Tempearture range:	5°C to 40°C
Power supply:	230V AC < 1.8VA

8.

References and sources of supply

8.1. Literature

[1] Complete measuring methods for collecting important AF parameters in receiver engineering, Ralf Rudersdorfer, Hans Otto Modler, Conference volume for amateur radio conference Munich 2006, pp 135 - 155

[2] The radio measuring regulation for HF receivers, transmitters, transceivers, receiving and sending devices, Franc Sichla, Ralf Rudersdorfer, Funk 2/2002, P. 68-71; ISSN 0342-1651

[3] The radio measuring regulation for FM receivers, transmitters, transceiver, receiving and transmitting devices, Franc Sichla, Ralf Rudersdorfer, Radio 3/2002, pp 68-70, ISSN 0342-1651

[4] Measuring technique in the communications technology – Edition 1, Ulrich Freyer, Carl Hanser publishing house 1983, ISBN 3-446-13703-3

[5] Paperback of high-frequency engineering – Edition 5, Hans Heinrich Meinke, Friedrich Wilhelm Gundlach, Springer publishing house 1992, ISBN 3-540-54717-7

[6] Circuit analysers use and understanding, Ralf Rudersdorfer, QSP - The Austrian transmitter federation 5/2004 - 29.Jhg. /5, pp 72-79

[7] The handling of modern multirange instruments, Ralf Rudersdorfer, Leisure Electronics 1/2003, P. 16-19, __ISSN 0342-1651

[8] Noise, Hans Hellmuth Cuno, Script of the lecture practical electronics, FH-Regensburg 9/2003, P. 6.1-6.11

[9] Im Größenrausch der Rauschgrößen, Franc Sichla: Funk 12/2001, P. 59-63; ISSN 0342-1651

[10] Electrical measuring techniques, Reinhard Lerch, 1st Edition, Springer publishing house 1996, ISBN 3-540-59373-X

[11] The large electronics workbook, Dieter Nührmann, 6th Edition, Franzis Publishing House 1994, ISBN 3-7723-6546-9

[12] All you need to know about SINAD (and its use with 2023 signal generator measurements), Dave Owen, IFR Application note 4/1999

[13] Service manual for Stabilock 4020 Series radio, Hrsg. Acterna/Schlumberger ENERTEC, Edition 1980, part 3 display section and operating sector

8.2. Sources of supply

[14] Analog Devices , No 3 Horizon Business Village, 1 Brooklands Road, Weybridge, Surrey, KT13 0TJ, web: http://www.analog.com

[15] Texas Instruments, 800 Pavilion Drive, Northampton Business Drive, Northampton, NN4 7YL, web: http://www.ti.com

[16] Silonex Inc., 5200 pc. Patrick Street, Montreal, Quebec, H4E 4N9, Canada, Tel.: ++1 514/7688000, fax.: ++1 514/7688889, Internet: http://www.silonex.com

[17] Farnell InOne, Canal Road, Leeds, LS12 2TU, web: http://uk.farnell.com

[18] Reichelt Electronics e. Kfr, D-26452 Sande, Tel.: +49 (0) 4422/955-333, fax.: +49 (0) 4422/955 111, web: http://www.reichelt.de

[19] Tonne Software, James Tonne, WB6BLD, 118 Garnet Lane, Oak Ridge TN 37830, USA, web: http://www.tonnesoftware.com

[20] Huber+Suhner GmbH, flour berry road 6, D-82019 baptizing churches, Tel.: +49 (0) 89/61201-0, fax.: +49 (0) 89/61201-162, Internet: http://www.hubersuhner.de/co-de-tk/de-tk-index

[21] proMa Technology GmbH, Im Leibolzgraben 16, D-36132 Eiterfeld, Tel.: +49 (0) 6672/898 600, fax.: +49 (0) 6672/898 - 688, Internet: http://www.proma-technologie.com

Konrad Hupfer, DJ1EE

400W power amplifier for 2m

Because the 2m band is enjoying an increase in popularity a power amplifier with the maximum output of 400 - 450W with a supply voltage of 28 – 32V is described in this article. The active device was chosen as a proven "VHF workhorse" from the manufacturer SEMELAB. It is very robust; the load VSWR may vary up to 20:1 thus giving more scope for output network optimisation without destroying the semiconductor. Full data for the transistor can be found in the datasheet [1]. The circuit is not complicated yet has good characteristics.

1.

The circuit

The critical part of a transistor power amplifier circuit (Fig 1) is the output network. A mismatch at the output represents a danger to the semiconductor. Therefore the matching circuits shown in Fig. 2 will be described in more detail in this article.

A narrow band solution would be sufficient for the 144MHz to 146MHz frequency range, it is nevertheless advisable to use a wider band solution in order to allow for adjustment tolerances and alignment sensitivity. The input and output matching circuits NW1 and NW4 consists of $\lambda/4$ matching transformers (25Ω coax cable), they also provide a 50Ω asymmetric to 12.5Ω symmetric transformation. The networks NW2 and NW3 are very simple L/C circuits with inductances made from individual striplines.

1.1. Design of the circuit for network NW3

The $\lambda/4$ matching transformer NW4 is made from 25Ω coaxial cable (flexible or Semirigid cable should be used with an outside diameter not less than 3mm because of the 400W output power that it must handle) that transforms the asymmetric 50Ω load R_L to 2 x R'_L = 2 x 6.25Ω = 12.5Ω.

To make the computation a bit clearer, Fig 3 shows that the real load is divided from 12.5Ω to 2 x 6.25Ω. The network NW3 must be designed to match the optimal load resistance R'_{opt} at half of the total output of 200W with the internal values of the transistor shown in Fig 4. The value R'_{opt} is shown in the SEMELAB data sheet. The internal load resistance R'_{DS} can be determined from the effective RF Drain voltage V'_{RF} at 200W. The effective RF voltage is close to the supply less "the bottoming voltage" V_K (assumed to be 3V):

Fig 1: Circuit of the 400W 2m power amplifier.

L1, L2 = 3 turns of 1mm diameter copper wire, 6mm diameter

L3 = 5 turns of 0.5mm copper wire wound on a 5.6Ω 2W resistor
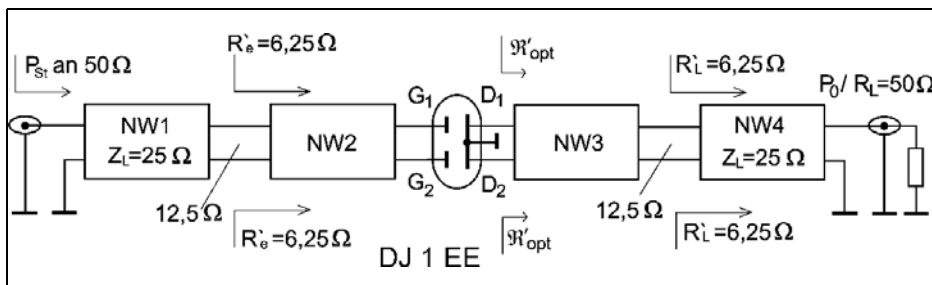
Fig 2: Block diagram of the 400W 2m power amplifier.

$$R'_{DS} = \frac{(U - U_K)^2}{2 \cdot P_o} = \frac{(28-3)^2}{400} = 1.56\Omega$$

In parallel with this is an output capacitance $C_{ob} = 190pF$ (see data sheet). The effective capacity is:

$$C'_{ob} = 1.3 \cdot C_{ob} = 1.3 \cdot 190 = 250\,pF$$

The inductance of the drain connection is taken from the SEMELAB data and is 0.63nH.

1.1.1. Calculating the value of L1 (Fig 4)

L1 is made from a small low impedance stripline. The transformation procedure can be seen using a Smith chart. The Smith chart program by Fritz Dellsperger [2] was used for this process and found to be extraordinarily helpful for this task. The Smith chart in Fig 5 is standardised for the terminal resistance $R'_2 = 6.25\Omega$, the internal load resistance $R'_{DS} = 1.56\Omega$ and the goal value $R'_L = 6.25\Omega$ can be seen. The chart is simplified to make it

Fig 3: Network NW3.

easier to understand. Point 1 (Z_1) on the real axis represents $R'_{DS} = 1.56\Omega$. $C'_{ob} = 250pF$ is point 2 on the conductance circle. The bond inductance is a series inductor $L = 0.63\mu H$ and is represented as $Z_2 = 1.4 + j0.5\Omega$ giving point 3. This is the impedance at the drain connector lug; $Z_3 = 1.4 + j0.1\Omega$. This "transistor connection resistance" is now transformed with an inductance, formed by striplines, S3 and S4, on the circle around the centre of the diagram to point 4 giving $Z_4 = 1.4 + j2.6\Omega$. This is the intersection with the conductance circle corresponding to the target resistance $Z_5 = 6.25 + j0\Omega$.

The stripline width W was selected as 8mm and the substrate thickness, H, used was 0.83mm. The characteristic impedance Z_0 is given by:

$$\frac{W}{H} = \frac{8mm}{0.83mm} \quad \text{gives } 28.4\Omega$$

This is for a dielectric constant $\varepsilon_R = 1$. In reality the characteristic impedance of the striplines S3 and S4 are 16.5Ω because the substrate has a dielectric constant $\varepsilon_R = 3.3$. The inductance of this stripline is $2.8\mu H$ and is "quasi-stable" the length of the line is $l = 30mm$. The inductance of this very short line is independent of ε_R!

If a 320pF capacitor is connected from point 4 to ground then a real resistance of 6.25Ω is achieved (point 5 on the Smith chart). This matches one half of the
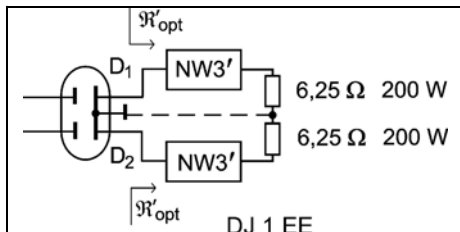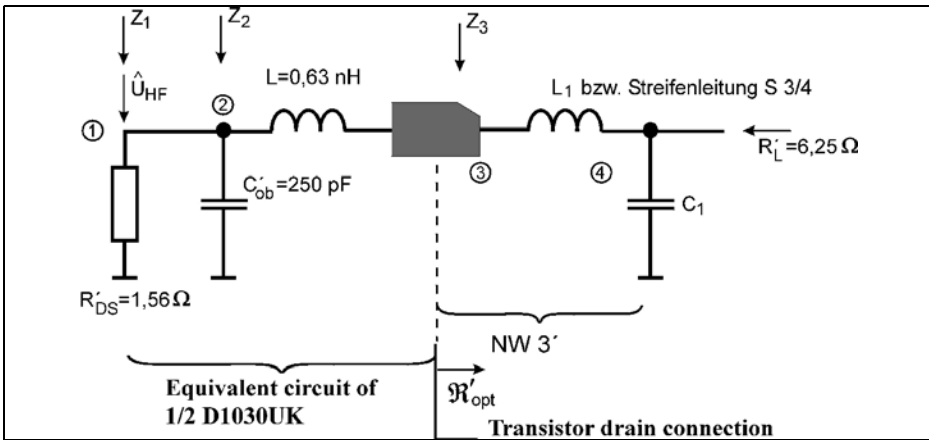
Fig 4: The equivalent circuit of 1/2 D1030UK.

power transistor to 6.25Ω. For the push-pull circuit twice this value is used, see Fig 6. Because the "virtual ground" does not exist, the two single capacitors in Fig 6 are combined into a single capacitor (2 x 320pF in series = 160pF).

If the output networks are used with the calculated values the output power will be approximately 250 to 300W. The transistor equivalent circuit used do not exactly correspond to the actual values partly because of differences in the mounting of the transistor. In order to

achieve the maximum output power of 400W with good efficiency (~70%), changes in the striplines S3 and S4 as well as the capacitor C5 are required. This is shown in Fig 1 by the variable arrow on these components. This tuning was carried out while watching the efficiency and the values shown in Fig 7 were the result.

The characteristics of the output stage with a supply voltage V_{DS} of 28V and 32V are shown in Fig 8.
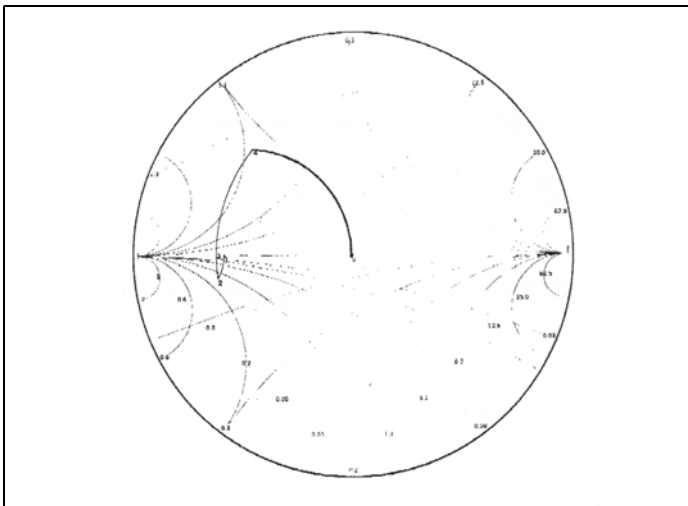
Fig 5: Smith chart showing the transformation to 6.25Ω.
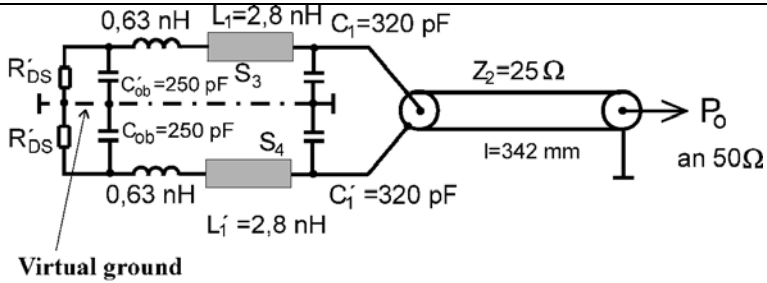
Fig 6: Circuit showing the two the two halves of the push-pull amplifier.
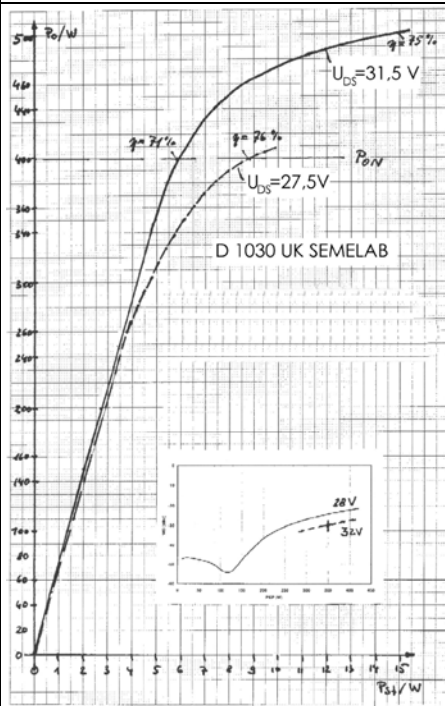
Fig 8: Output characteristics of the D1030UK.

1.2.1. Measured performance

The performance of the amplifier was measured with the equipment shown in Fig 13. The results for a supply voltage of 28V are shown in Table 1 and for a 32V supply voltage in Table 2.

1.3. Input matching

The input impedance of ½ D1030UK has the typical values shown in Fig 9.

R'_{GS} is found, like R'_{DS}, in the data sheet [1]. R'_{GS} can be transformed to 6.25 +j0Ω using a simple L/C circuit. The inductor is again made from a stripline. The transformation to 50Ω is achieved with λ/4 matching transformers made from 25Ω coax cable; thin cable is sufficient here because the power is low. Using the values calculated using a Smith chart, the input matching from 5:1 to 2:1 is achieved. The striplines and capacitors must be optimised as shown in Fig 10. Unlike the output the input can be trimmed to give a return of zero.
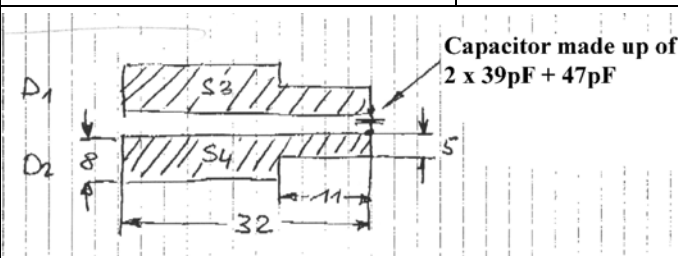
Fig 7: Dimensions of striplines S3 and S4 after optimisation.

Table 1: Measured parameters at 28V.

U_{DS} = 28V, I_D = 1.7A, f = 145MHz

CW:	P_o	400W
	$P_{ST.}$	8.9W
	Efficiency	75%
	U'_{DS}	58V
Harmonics:	2 x f_0	-49dB
	3 x f_0	-44dB
SSB power:	PEP_o	300W
	PEP_{ST}	4.7W
	FMD_3	-29 to 31dB
	FMD_5	-35B
	Efficiency	49%

Table 2: Measured parameters at 32V.

U_{DS} = 32V, $I_D \approx$ 1.8A, f = 145MHz

Po/W	P_{ST}/W	I_D/A	G/dB	Effic%
50	0.7	6.8	18.7	24
100	1.35	9.2	18.9	
200	2.7	12.7	18.7	
300	4.3	15.9	18.5	
350	5.0	17.0	18.5	
400	5.9	18.1	18.3	71
450	8.2	19.6	17.4	74

| SSB power | PEP_o | 350W |
| | IM_3 | -30dB |

Maximum drain voltage:

(DC + RF) $\approx$ 65V for P_o = 480W
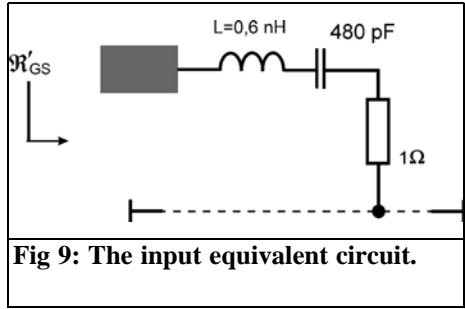
Fig 9: The input equivalent circuit.

2.

Mechanical construction

Fig 12 shows the construction of the amplifier. The 50 x 150mm baseboard is fitted to the large 150 x 120 x 80mm heatsink as shown in Fig 11. The baseboard is 1.6mm thick copper clad FR4 material; the upper surface is the RF and DC ground. So that fitting the power transistor does not interrupt this ground, a very thin copper foil (~ 0.1mm) is fitted into the cutout for the transistor. The striplines are individually cut from 0.83mm thickness material and soldered to the baseboard ground surface. The DC wiring can take place as desired. Good heat transfer between the foil, transistor and heatsink is extremely important. Spread thermal compound on the indi-
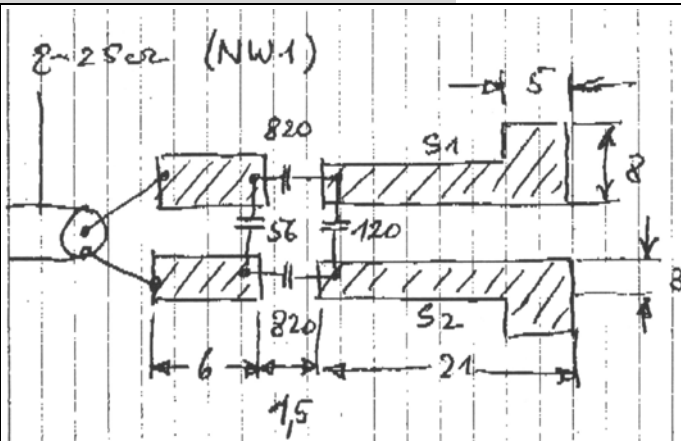
Fig 10: Dimensions of striplines S1 and S2 after optimisation.
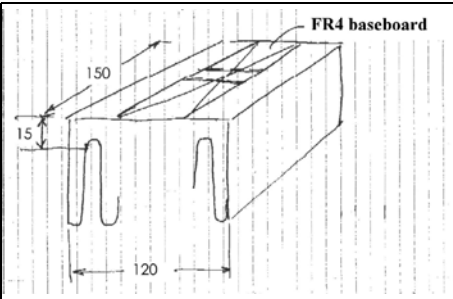
Fig 11: The heatsink and baseboard.

vidual surfaces very thinly because only the pores in the metal are to be filled! The heatsink should be well cooled with the aim of a maximum flange tempera-ture of 60°C with a PEP output of 300W.

3.

References

[1] SEMELAB, web: www.semelab.co.uk. The data sheet for the D1030UK can be downloaded from http://www.semelab.co.uk/pdf/rf/D1030 UK.pdf

[2] Fritz Dellsperger, Smith chart program and Smith chart tutorials, email: fritz.dellsperger@isb.ch, web: http://www.fritz.dellsperger.net/
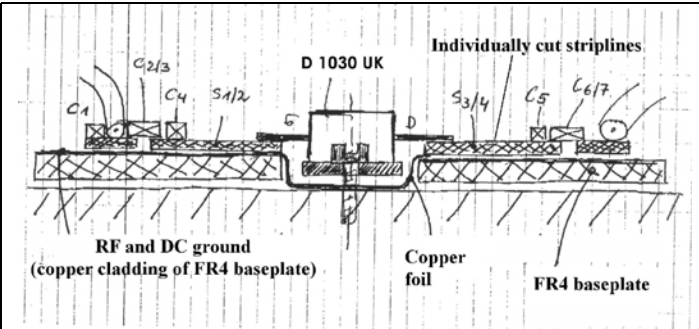
Fig 12: The mechanical construction of the 400W 2m power amplifier.
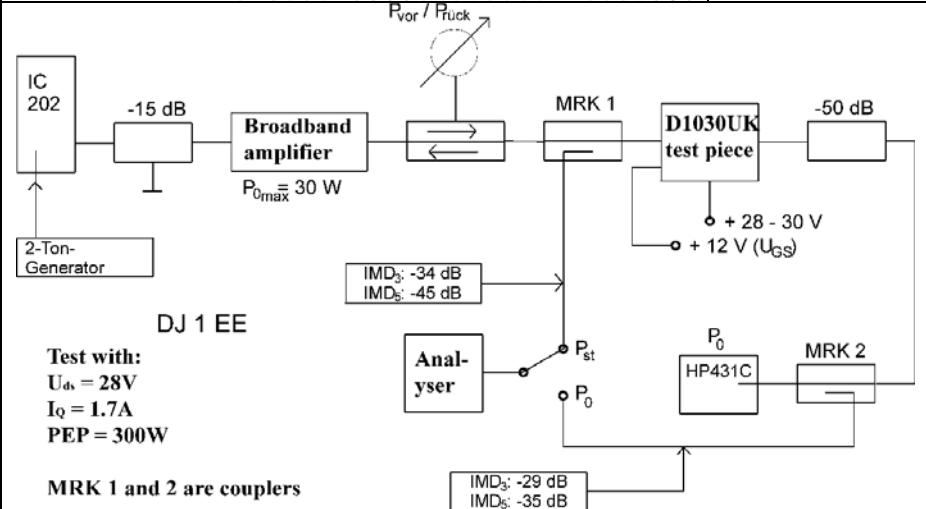
Fig 13: Block diagram of the performance measuring equipment.

Zeljko Bozic, S52ZB

A GPS controlled frequency standard

1.

Introduction

Most of radio amateurs have frequency counters with reference oscillators that are not temperature compensated and they are made with bad quality crystals. A special problem is the ageing of crystals that can take up to 10 years. Because of this frequency counters can deviate by more than a hundred kHz at a frequency of 10GHz and are really only suitable for 100MHz or less.

A simple and inexpensive solution is a GPS controlled frequency standard, often referred to as a GPSDO (GPS Disciplined Oscillator). It is a frequency stand-ard that calibrates itself using GPS.

2.

Block diagram

The block diagram of a GPS controlled frequency standard is shown in Fig 1.

3.

Circuit diagram

The circuit diagram of a GPS controlled frequency standard is shown in Fig 2. The circuit diagram is simple as it can be.
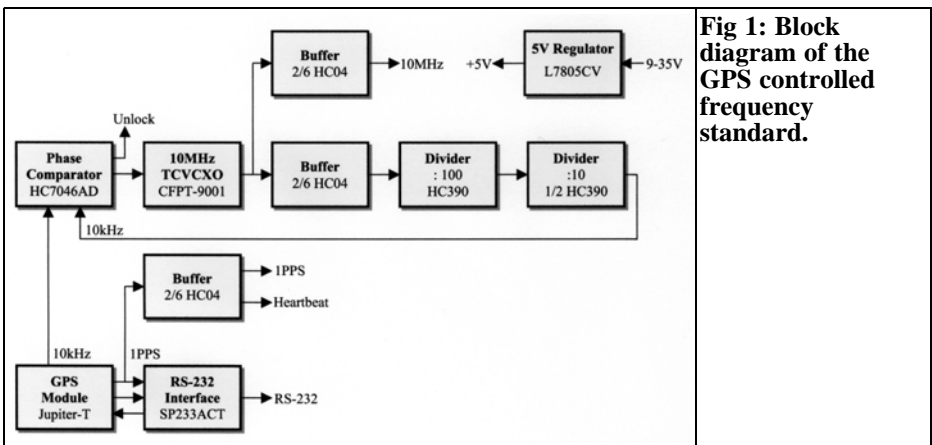
Fig 1: Block diagram of the GPS controlled frequency standard.
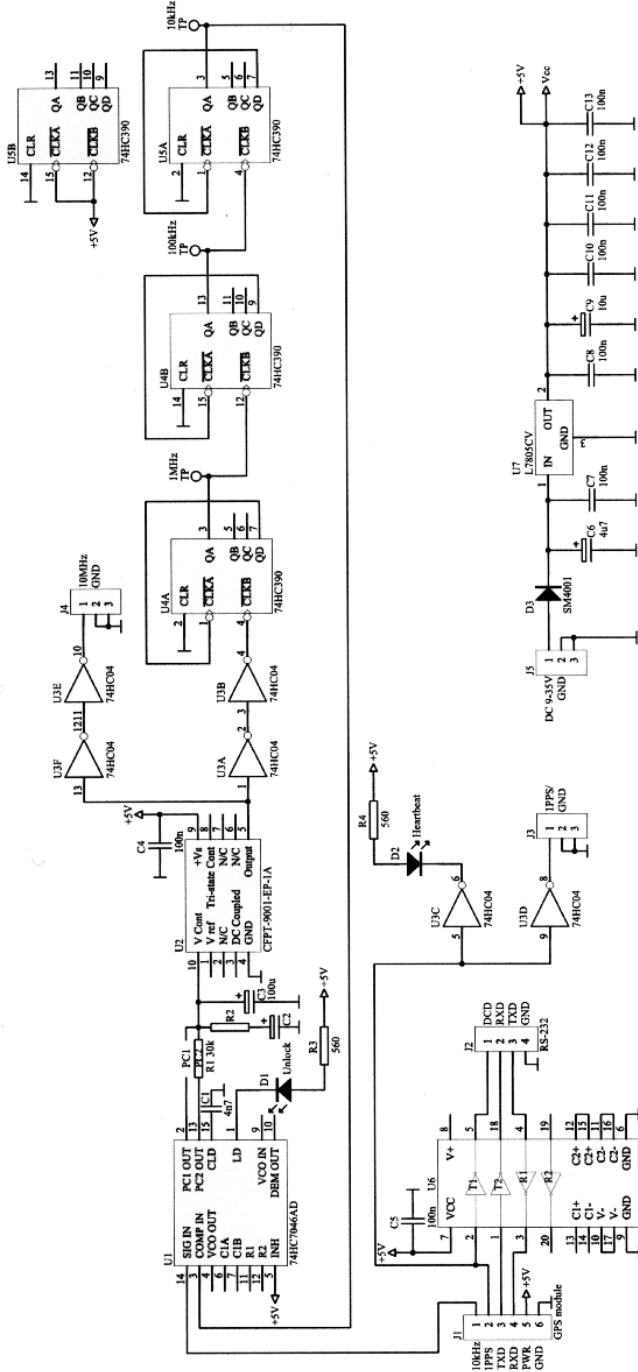
Fig 2: Circuit diagram of the GPS controlled frequency standard.
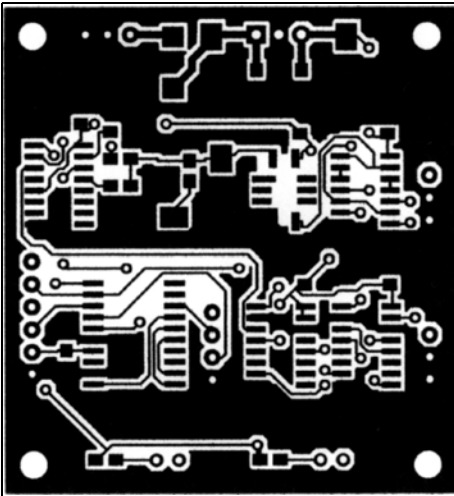
Fig 3: Top side PCB layout for the GPS controlled frequency standard.
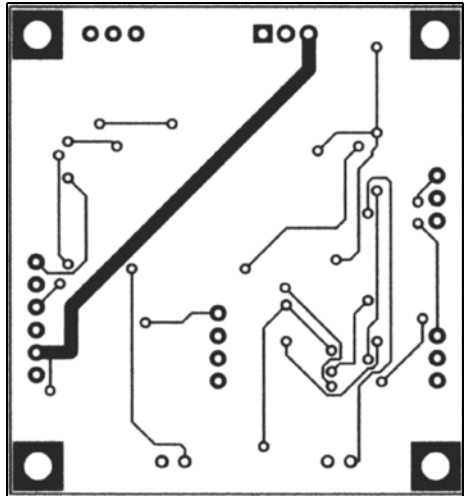
Fig 4: Bottom side PCB layout for the GPS controlled frequency standard.

The Navman TU60-D120-041 Jupiter-T GPS receiver [1] has a high accuracy 10kHz output synchronised to UTC. A 10MHz C-MAC TCVCXO CFPT-9001 [2] is divided down to 10kHz and phase locked to this GPS signal. Both outputs (10MHz and 1PPS) are buffered with inverters.

Fig 5: Component layout for the GPS controlled frequency standard.

The PLL filter components are designed for $\omega n=0.239$ rad/s, $\varsigma=0.707$ and Ko=67Hz/volt. R2 and C2 are not used. The GPS receiver can be controlled via the RS232 port and a 1PPS signal is available on DCD that can be used by a computer to keep precise UTC time to a few tens of ns.

4.

Construction

The frequency standard is constructed on a double sided PCB made from a 50.8mm x 55mm piece of 1.6mm thick FR4 laminate. It has plated through holes, solder resist and lead free tinning. The PCB layouts are shown in Figs 3 and 4, the component layout is shown in Fig 5 and the parts list in Table 1.

The GPS receiver used is a Navman TU60-D120-041 Jupiter-T which has both 1PPS and 10kHz outputs. It was designed especially for precision timing applications. In 2005 Navman discontinued manufacturing the Jupiter-T. Instead of the Jupiter-T you can use the standard
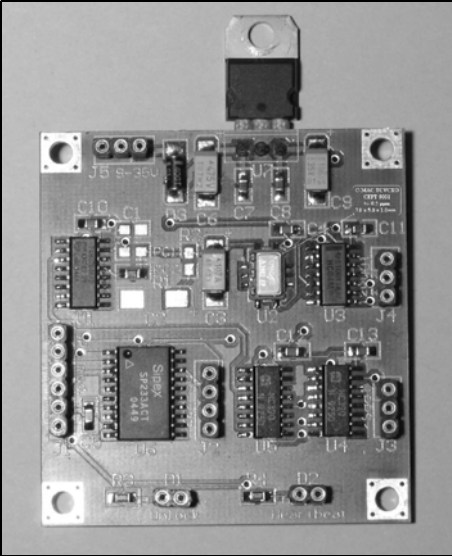
Fig 6: Picture of the GPS controlled frequency standard PCB.

Jupiter receivers TU30-D140-xxx and TU30-D410-xxx, which also have a 10kHz output. They are similar in size but the connector and pin out are different. These receivers are often available on eBay. Before purchase it is important to ensure that the firmware is v1.18 onwards. The firmware version is printed on a label and is also emitted at start up in an ASCII plain text NMEA message. The default protocol of the GPS receiver is Motorola UT+ "@@". Navman binary protocol and NMEA protocol are also available.

The GPS antenna is an external active antenna with 5V supplied through the coaxial cable. It is important that the GPS antenna has the fullest possible view of the sky. Getting the maximum number of satellites in view will get the best timing performance.

A better quality oscillator such as an OCXO can be used instead of the C-MAC TCVCXO but the PLL filter components must be recomputed. The oven-controlled oscillator has considerably better performance than a miniature SMD TCVCXO, but at the cost of greater power consumption and increased size. Good but big OCXOs can be found in old NMT base stations and also on eBay.

The 10MHz output is square wave at HCMOS level, an output level of
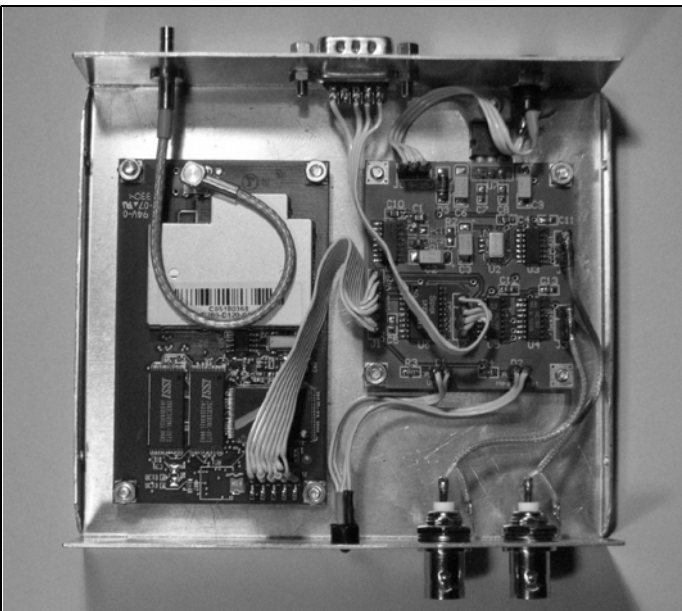
Fig 7: Top view of the completed GPS controlled frequency standard.
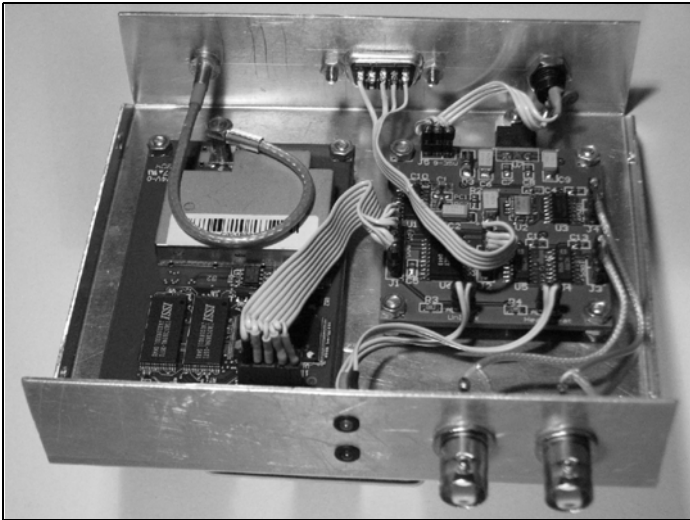
Fig 8: Front view of the completed GPS controlled frequency standard.

+15dBm. The PLL IC is a 74HC7046A [3] [4], it is an improved version of the familiar 74HC4046A. The PC3 function is replaced with an improved lock detector. The serial port driver IC is an SP233A [5], a low cost version of the familiar MAX233A.

The frequency standard status is shown by two LEDs. The green LED "Heartbeat" flashes at 1PPS and the red LED "Unlock" lights up when the PLL is unlocked. The value of the lock detector capacitor C1 is dependant on the PLL filter components. If the PLL filter components are changed, the value of C1 must also be changed.

The operating voltage of the frequency standard is 9 to 35V and the current consumption should be around 250mA including GPS receiver and GPS active antenna.

The frequency standard is built into a custom made housing 130mm wide, 110mm deep and 30mm high made from

Fig 9: Rear view of the completed GPS controlled frequency standard.

Table 1: parts list

R1	30kΩ	0805
R2	-	0805, not used
R3, R4	560Ω	0805
C1	4n7	0805 C0G (NP0)
C2	-	C, SMD tantalum 35V, not used
C3	100µF	C, SMD tantalum 35V
C4, C5, C7, C8, C10, C11, C12, C13	100n	0805
C6	4.7µF	C, SMD tantalum 35V
C9	10µF	C, SMD tantalum 35V
D1	LED	Red 3mm
D2	LED	Green 3mm
D3	SM4001	MELF
U1	74HC7046AD	SO-16
U2	CFPT-9001	SMD 7x5x2mm, 10-pin
U3	74HC04	SO-14
U4, U5	74HC390	SO-16
U6	SP233ACT	SO-20 wide
U7	L7805CV	TO-220
J1	SIL socket	6-pin
J2	SIL socket	4-pin
J3, J4, J5	SIL socket	3-pin
PCB	-	50.8 x 55mm, FR4 1.6mm
Housing	-	Aluminium plate 1mm

1mm thick aluminium plate. Figs 6, 7, 8 and 9 show the prototype GPS controlled frequency standard.

5.

Results

First measurements of the frequency standard show that it has accuracy of 1 x 10^{-10} after 1 hour and about 1 x 10^{-11} after 24 hours. The test equipment was an Agilent PSA E4445A (1Hz resolution) spectrum analyser locked on a 10MHz frequency standard OSA 5230 from Oscilloquartz. Measurements with the spectrum analyser were on the 100th (1GHz), 200th (2GHz) and 500th (5GHz) harmonic. More accurate measurements were with an oscilloscope (Lissajous figure).

GPS satellites can be monitored using one of the NMEA 0183 data programs such as Tac32 [6] or VisualGPS [7]. These programs give a graphically display of specific NMEA 0183 sentences.

I wish to thank James Miller (G3RUH) [8] for help and support, Robert Vilhar (S53WW) for measurements and Stojan Kuret (S51WI) for help in all phases of project.

6.

References

[1] Navman datasheet: Jupiter-T,web: http://www.jrmiller.demon.co.uk/projects/docs/10039C.PDF

[2] C-MAC datasheet: CFPT-9000 Series, web: http://fcp.rakon.com/pdf/cfpt-9000.pdf

[3] Philips datasheet: 74HC7046A, web: http://www.datasheetcatalog.com/datasheets_pdf/7/4/H/C/74HC7046A.shtml

[4] TI Application Note AN8823, web: http://focus.ti.com/lit/an/scha003b/scha003b.pdf

[5] Sipex datasheet: SP233A, web: http://www.sipex.com/Files/DataSheets/sp232_312a.pdf

[6] Tac32 software from CNS Systems Inc., web:

http://www.cnssys.com/cnsclock/Tac32Software.html

[7] VisualGPS software from Visual GPS LLC. web:
http://www.visualgps.net/VisualGPS/default.htm

[8] Simple GPS Stabilised 10 MHz Oscillator, James Miller, G3RUH, web:http://www.jrmiller.demon.co.uk/projects/ministd/frqstd.htm

[9] A GPS-Based Frequency Standard QST July 1998, Brooks Shera, W5OJM, web:
http://www.rt66.com/~shera/QST_GPS.pdf

[10] Time and Frequency Measurements Using the Global Positioning System, Michael A. Lombardi, Lisa M. Nelson, Andrew N. Novick, Victor S. Zhang:, Cal. Lab. Int. J. Metrology, pp. 26 -33, (July-September 2001), web: http://tf.nist.gov/general/pdf/1424.pdf
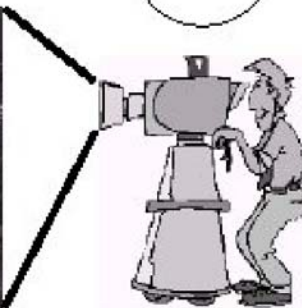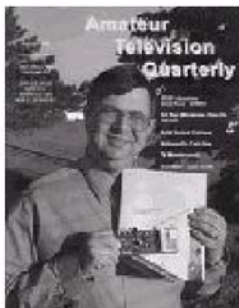
Matjaz Vidmar, S53MV

LCD oscilloscope for spectrum analysers. Plus update on spectrum analyser project

1.

Spectrum analyser project 2007 update

Since the development of the wide band VCO almost 10 years ago, the whole spectrum analyser project with all related accessories: tracking generator, harmonic converter, storage normaliser, marker counter and accessories developed by other experimenters (Darko S57UUD) have been published in many different places: magazines "VHF Communications Magazine issues 4/1998 – 4/1999", "AMSAT-DL Journal", "CQ ZRS" and the book "Beacon 99". It is reasonably believed that the project has been successfully reproduced in hundreds of units (33 sets of PCBs have been supplied to VHF Communications Magazine readers – Andy).

Since the spectrum analyser project is still very interesting for many radio amateurs and other radio and electronic experimenters, I decided to republish my original articles in English on my personal web page [1]. Unlike printed magazines, the web allows to publish many colour pictures of all interesting details of the project in addition to PDF, PCB and software files. Last but not least, updates to the project are really simple on the web. (This article has been taken from the web site, if you are interested in the project of want PCB or software files you should visit the site – Andy)

In this article a simple and inexpensive LCD oscilloscope to be used as a display for the spectrum analyser will be presented. Although a small LCD screen is unable replace a good analogue oscilloscope, an LCD may be very useful in field measurements under strong daylight conditions, for battery operation or simply when the available oscilloscope is required for a different measurement at the same time.

Before describing the LCD oscilloscope, some necessary updates to the spectrum analyser project will be presented. As expected and also according to the feedback, the most difficult part to reproduce is the wide band VCO. This difficulty is in partly due to component tolerances and in partly due to assembly (soldering) tolerances. In particular, the BB833 varactor diodes were found to have rather wide manufacturing tolerances both in the capacitance range and in the series (loss) resistance affecting the Q of the varactor.

As already explained in the VCO article, two different printed circuit boards are proposed for the wide band VCO. Starting with unknown components, one should always assemble the "narrow stripe" wide band VCO first (PCB#1). The "wide stripe" PCB#2 should only be
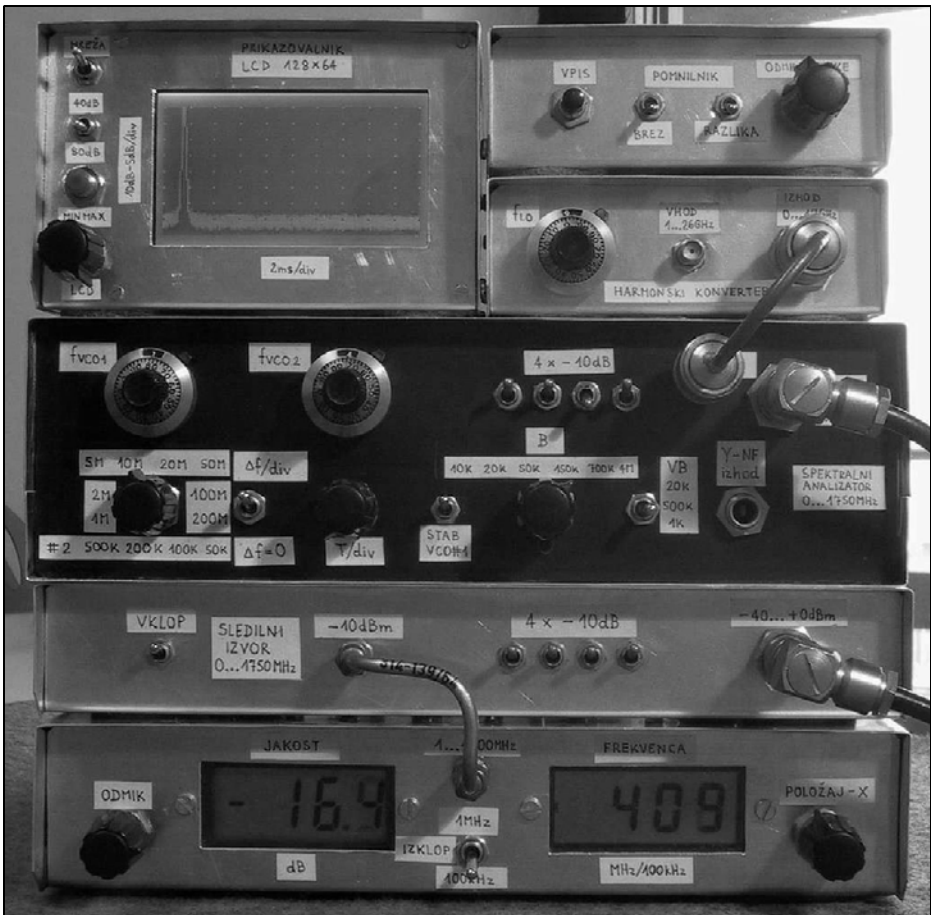
Fig 1: The spectrum analyser and all of the additional modules.

used if the frequency coverage obtained with PCB#1 is insufficient.

If output power of the wide band VCO drops in the middle of its frequency range between 2.5GHz and 3GHz or the VCO simply stops oscillating at all in this frequency range, excessive losses in the BB833 varactors are suspected. Besides poor varactor Q (try getting better BB833 varactors from a different source), the interdigital feedback filter may be misaligned due to soldering tolerances. In the latter case, a rather simple solution is to try to trim the length of the open end of the central finger of the interdigital filter.

Although the wide band VCO design was also published in a well respected and widely known professional magazine "Microwave Journal" back in 1999 [2], it took several years for the professionals to fully understand the operation and capabilities of this VCO design. Multiple resonator multiple varactor microwave VCO modules only became available recently on the professional market. Finally, there is a valid and inexpensive alternative to the YIG oscillator! Of course, these inexpensive commercial VCO modules can be used in place of the
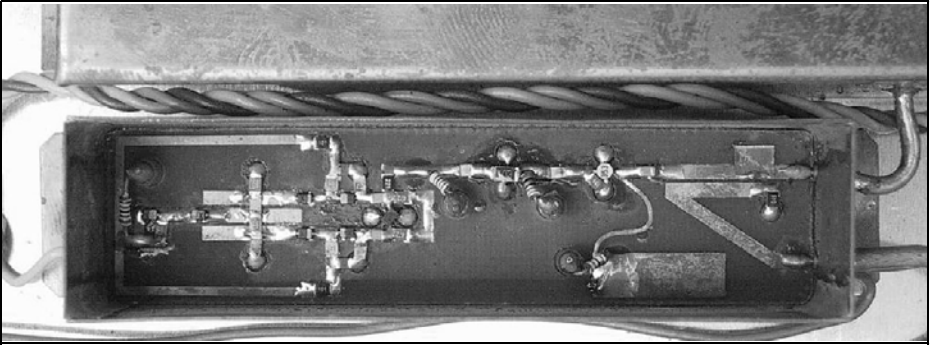
Fig 2: Picture of a wideband VCO after 10 year showing the corrosion on the microstrip circuit.

described wide band VCO.

After using the spectrum analyser and all related accessories shown in Fig 1 for many years, some long-term effects were noted as well. The worst seems to be the corrosion caused by outgassing chemicals from the antistatic foam used as a microwave absorber in high frequency shielded modules. Higher conductivity foam makes a better microwave absorber, but unfortunately more chemicals cause more corrosion as well! Fortunately, the microstrip circuits of the spectrum analyser continue to work correctly with little if any degradation, although they really look ugly, see Fig 2.

A less frequent problem is a long term failure of ATF35176 or similar HEMT devices. If the voltage between drain and source is kept too high, the drain current slowly decays. This decay may be very slow, just a few percent per week, but it is cumulative, irreversible and continues down to zero leading to a total failure of the circuit!

In this project, HEMTs are used in the buffer stages of wide band VCOs and in the buffer amplifiers inside the tracking generator. The described failure can be detected as a rise of the drain voltage. If the drain voltage rises above +3V...+3.5V, the decay will speed up and the HEMT will have to be replaced soon, possibly with better devices.

Some electronic parts became obsolete or hard to get. The most difficult seems to be the INA10386 MMIC amplifier that has no direct replacement. Worst of all, defective factory rejects are shipped for some obsolete parts, like the uA723 voltage regulator. Some uA723 regulators have a very noisy voltage reference with very large 1/f "popcorn" noise. The latter may be large enough to disturb even large signal circuits like the video amplifier inside the spectrum analyzer.

Besides getting better uA723 regulators, there is a simple solution to this problem. The internal 7V voltage reference, available on pin 6, should be filtered before use. If the latter is fed through a resistor to pin 5 (non-inverting input), then a single electrolytic capacitor from pin 5 to ground (pin 7) solves the problem.

2.

LCD oscilloscope for spectrum analyser

LCD modules are probably the most popular displays today, ranging from simple numerical displays with few single colour digits to large, graphical, high
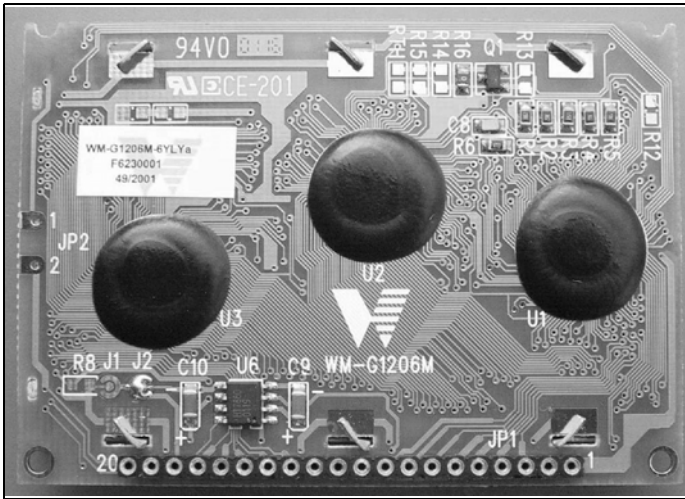
Fig 3: The rear view of an LCD display showing the bonded ICs covered with black resin.

resolution, full colour computer monitors. A medium resolution, single colour graphical LCD module is required in an oscilloscope display for a spectrum analyser. Of course, a simple solution was sought for the spectrum analyser project.

Medium resolution graphical LCD modules may have different interfaces. The simplest LCD modules have no built in controller. These modules only have shift registers associated with the columns and rows of the display. The user has to provide a continuous data flow to refresh and multiplex the display content. This requires a powerful microprocessor with lots of memory or in other words a complicated circuit with many chips.

LCD modules with a built in controller providing the refresh and multiplex of the display are much simpler to use. The most popular alphanumeric controller is certainly the Hitachi HD44780 that even has some very limited graphical capabilities. Older graphical LCD modules use the Toshiba T6963 controller with an external 8kbyte RAM (usually 6264). These modules also require a negative voltage supply for the LCD and a high-voltage AC source for the electro-luminescent backlight.

Recent graphical LCD modules use the

Samsung KS0107 and KS0108 chips. The KS0107 is the clock generator, scans 64 rows and may drive multiple KS0108 chips. The KS0108 drives the columns and includes storage for up to 64x64 dots. The microprocessor simply writes into the RAM inside the KS0108 chips. These modules usually include one or two 7660 chips to generate the required negative voltage LCD supply of -5V or -10V on-board the LCD module. Finally LEDs are used for the backlight so that the module can be operated from a single +5V supply.

A 128x64 graphical LCD module is used in the described LCD oscilloscope. This module includes one KS0107 and two KS0108 chips. All three controller chips are usually bonded directly to the printed-circuit board and covered with drops of black resin, see Fig 3.

The LCD oscilloscope is built around a PIC 16F876A microcontroller. The latter includes an A/D converter and steers the graphical LCD module directly. The circuit diagram of the LCD oscilloscope is shown in Fig.4.

The LCD oscilloscope requires two signals from the spectrum analyser; the analogue video (including blanking) and the trigger. Due to the limited resolution

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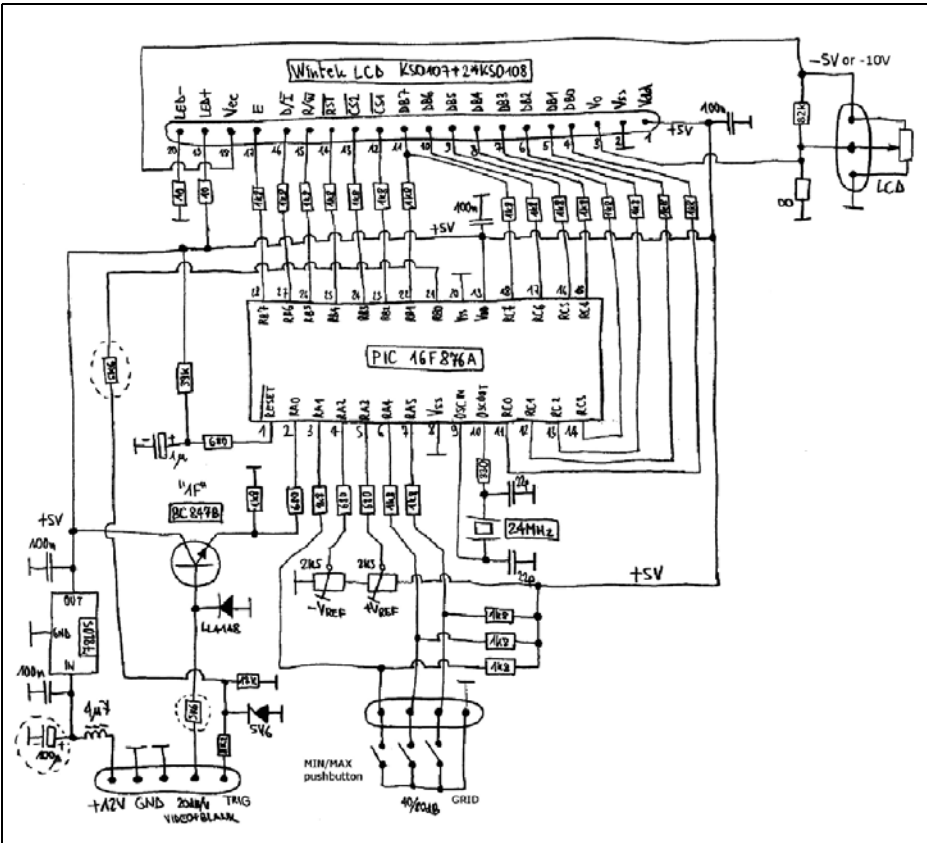
Fig 4: Circuit diagram of the LCD oscilloscope.

of the LCD, the analogue signal is over-sampled. Each column on the LCD is computed from eight consecutive samples and the last sample from the previous column. The A/D sampling rate is set so that the whole sweep (128 columns or 1024 samples) corresponds to about 20ms or in other words the fastest rate of the described spectrum analyzer.

Additional inputs to the microcontroller are provided for two switches and one push button. One switch is used to turn on the grid while the other selects a 80dB (4V) or 40dB (2V) full scale range. The scale is adjusted by two trimmers that define the reference voltages of the A/D converter. The pushbutton activates a MIN/MAX memory that is very useful when observing wide deviation slow FM signals or infrequent pulsed signals on the spectrum analyser.

All outputs of the PIC 16F876A have series 1.8kohm damping resistors to reduce the amount of radio interference radiated by the microcontroller. The same resistors are also used to operate the data bus to the LCD module bi-directionally, simplifying the programming of the PIC 16F876A. In particular, the RC0-7 pins are always programmed as outputs. In order to read the status of the KS0108 controllers, the DB7 line is
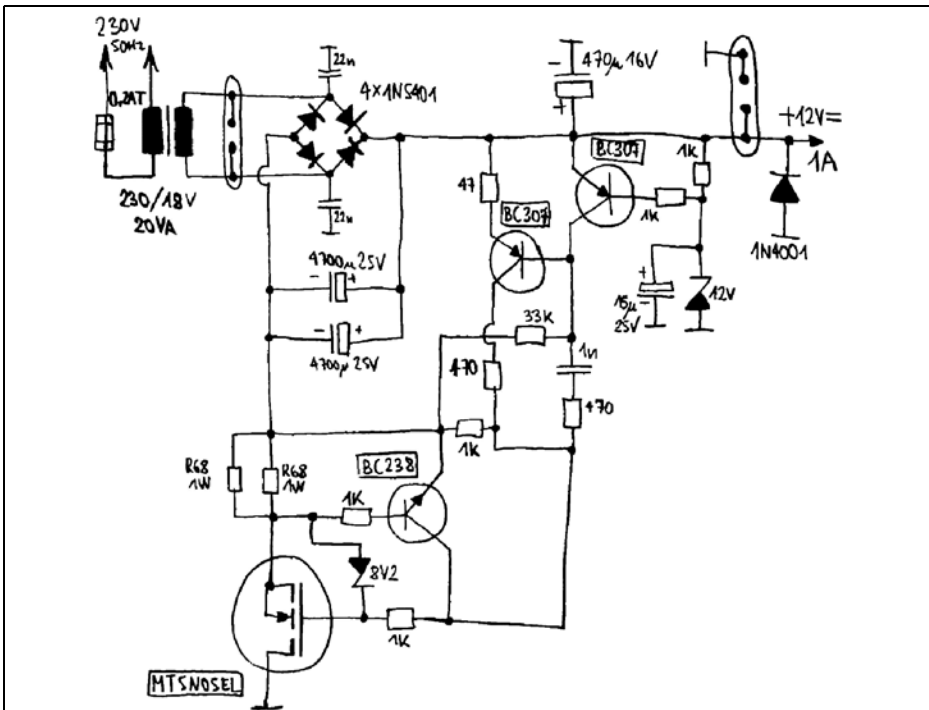
Fig 5: Circuit diagram of the power supply for the LCD oscilloscope and spectrum analyser modules.

also fed back to the RB1 input.

The display refresh rate is limited by three factors; the A/D conversion speed, the PIC computing power and the wait cycles required by the KS0108 controllers. In the LCD oscilloscope, all three factors are of the same order of magnitude limiting the refresh rate to about 35Hz - 40Hz and thus providing a live picture of the input signal. A minimum clock frequency of 20MHz is required for the PIC 16F876A for this purpose.

A 24MHz clock for the 16F876A was therefore selected to allow the spectrum analyser sweep time to remain at 20ms at all times with some safety margin. This is consistent with the resolution of the display (128 columns) and available filter bandwidths and spans of the spectrum analyser. Slower signals can be observed with the MIN/MAX function, using 256

bytes of the internal RAM inside the 16F876A. The minima and maxima are simply accumulated as long as the MIN-MAX pushbutton is kept depressed.

The same 256 bytes of internal RAM are also used for 128 minima and 128 maxima storage during normal operation without the MIN/MAX function. The A/D routine runs under interrupts triggered by an internal timer inside the PIC 16F876A. The main program is an endless loop computing and refreshing LCD content from the intermediate storage. Yet another interrupt is used for the trigger function.

The LCD oscilloscope is fully compatible with all described accessories; tracking generator, storage normaliser and marker counter. The 40dB scale is in fact intended for reflection measurements with the tracking generator and an addi-
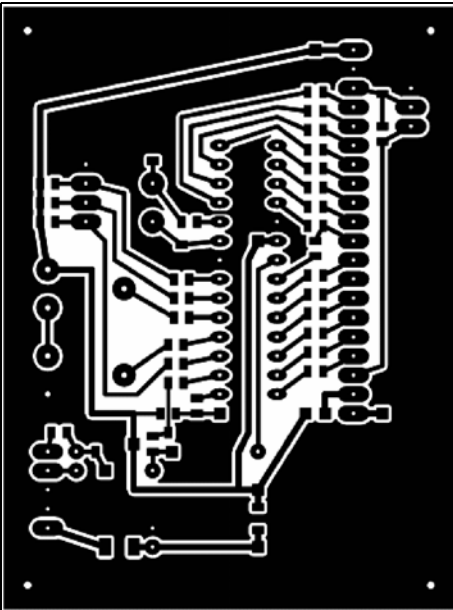
Fig 6: PCB layout fro the LCD oscilloscope.
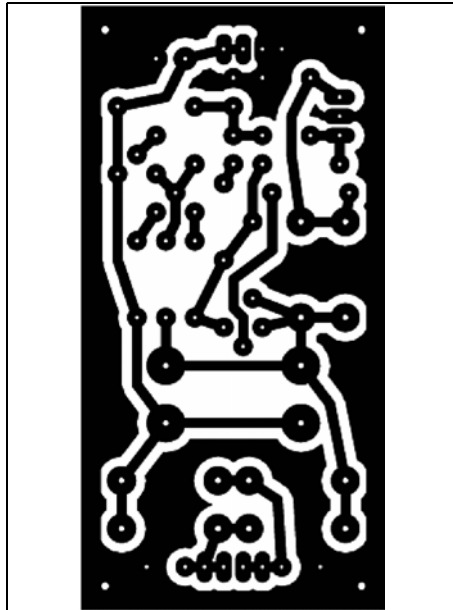
Fig 7: PCB layout for the power supply.

tional, external directional coupler. The over-sampling of the video signal allows a proper display even of the 455kHz marker.

The LCD oscilloscope includes its own +5V regulator (78L05). The latter also provides power to the backlight LEDs through two 10ohm current limiting resistors. Current limiting resistors and other protection components are also provided on all inputs to the module.

3.

230V mains power supply

Since the described LCD oscilloscope is small and simple, the remaining space behind the LCD module is used for a 230Vac mains power supply, providing 12V for the LCD oscilloscope, spectrum analyser and all accessories. The circuit diagram of the power supply is shown on

Fig.5.

The discrete component regulator has many advantages over integrated circuits; very low drop out voltage, grounded heatsink of the power device and last but not least, no additional protection diodes are required in the case of (parallel) 12V battery operation! Since the spectrum analyzer is sensitive to low frequency magnetic fields, it is imperative to use a toroidal core transformer in the power supply.

4.

Assembly of the LCD oscilloscope

The first thing to do is to check the availability and size of suitable LCD modules with the KS0107/KS0108 controller chips. Do not forget to check the
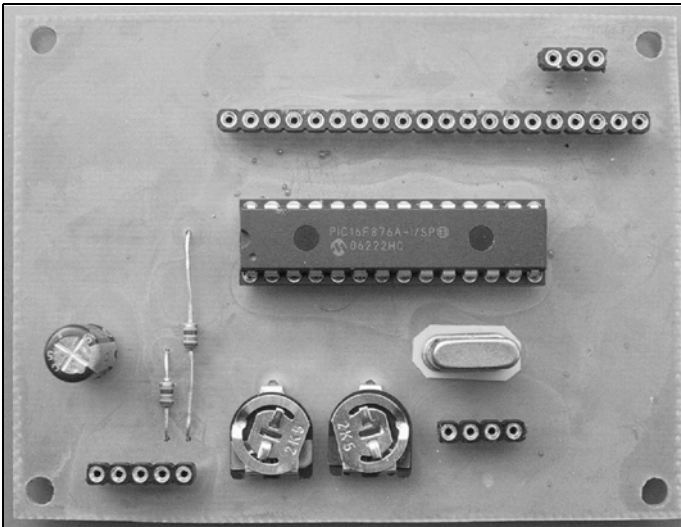
Fig 8: Picture of the LCD oscilloscope PCB fitted with components.

presence of the 7660 voltage generator as well as the type of backlight available. Last but not least, check the connections of the available LCD module!

The printed circuit board (Fig 6) was developed for Wintek LCD modules. Other manufacturers may use a different pin out requiring the swapping of some wires. In particular, the Vdd (+5V) and Vss (GND) connections may be swapped! CS1 and CS2 may be active low or active high, since each KS0108 actually has three such inputs (two in-

verted) and just one is brought out to the connector.

The LCD oscilloscope is built on two printed circuit boards as shown in Fig.6 and Fig 7. The PIC 16F876A and associated components are installed on an 80mm x 60mm single sided printed circuit board. The power supply regulator is built on an 80mm x 40mm single sided printed circuit board. Both circuit boards are etched on 1.6mm thick FR4 glass fibre epoxy laminate. Pictures of the
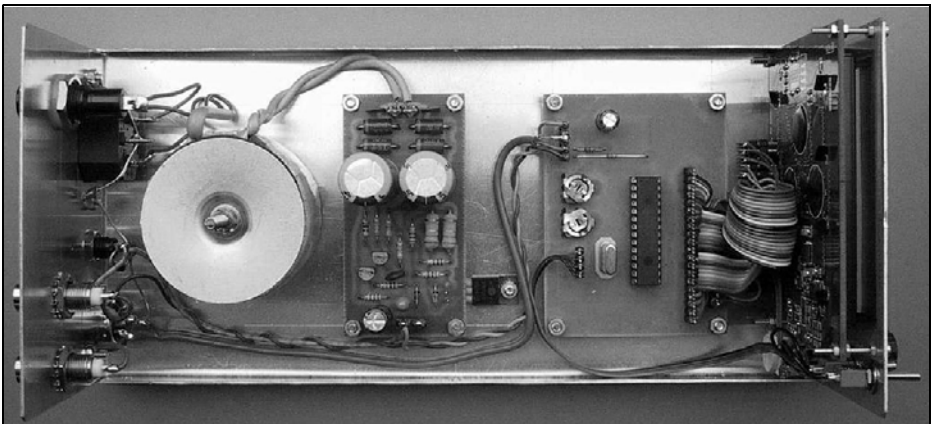
Fig 9: Picture of the chassis fitted with the LCD oscilloscope and power supply.
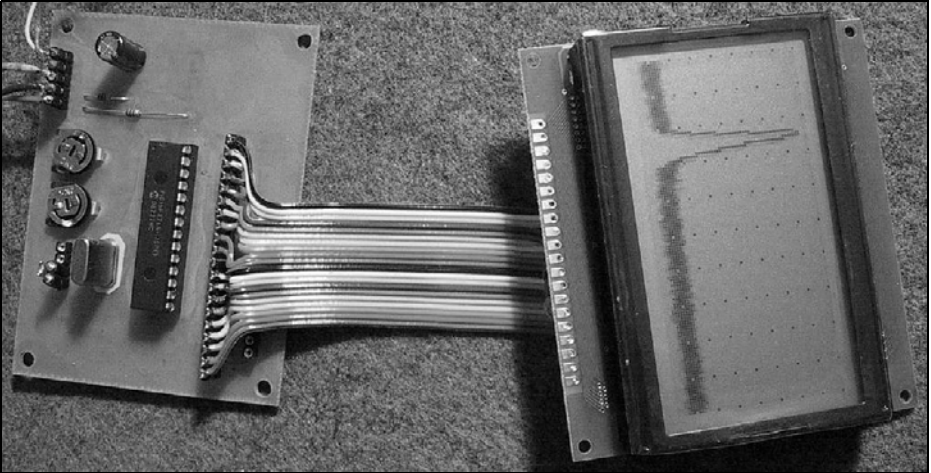
Fig 10: The LCD oscilloscope showing the PCB and display.

completed boards are shown in Figs 8 and 9. Figs 10 and 11 show the LCD screen in operation.

The LCD oscilloscope has the same depth (240mm) as the spectrum analyser [3] or [4]. The width is 110mm while the height is 70mm considering the size of the available LCD module. All modules are mounted in a single panel . The bottom of the box is simply a piece of 1mm thick aluminium sheet, bent in the

form of a "U". The cover is a similar "U" made from 0.6mm thick aluminium sheet.

As well as the DIN connector, two additional BNC connectors for the video + blanking and trigger signals are provided on the back panel to connect an additional, high quality, analogue oscilloscope. Although the LCD oscilloscope is protected from over voltages on all inputs even in the power down state, the LCD
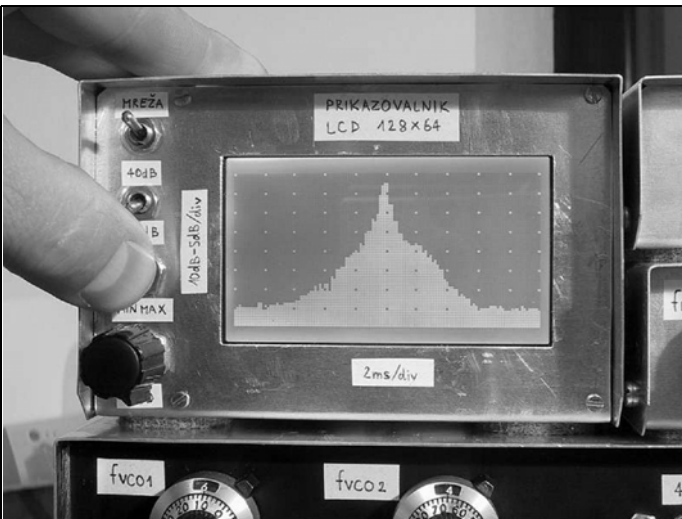
Fig 11: Picture of the LCD oscilloscope in use.

oscilloscope should be powered at all times to avoid loading the outputs of the spectrum analyser.

As well as the LCD module, two switches (grid on/off and scale 40dB/80dB), the MINMAX pushbutton and the LCD contrast potentiometer are located on the front panel. The latter could be omitted, since LCD modules from many different manufacturers were found to operate best at the maximum voltage. This means that the Vo input is simply tied to the Vee output through the 82kohm resistor on the printed circuit board and no additional potentiometer is required.

The LCD oscilloscope will display correctly input voltages between about +0.7V and +5.5V due to the emitter follower in front of the A/D converter. Therefore the trimmers inside the video amplifier of the spectrum analyser may need to be readjusted first, using a reliable analogue oscilloscope. The trimmers inside the storage normaliser and marker counter may need some minor readjustments as well.

Finally, the 80dB scale of the LCD oscilloscope can be adjusted with the two trimmers providing the reference voltages to the A/D converter. Please note that the operation of the LCD oscilloscope is slightly different between the 80dB mode and the 40dB mode. In the 80dB mode, the trace is always visible and saturates on the bottom or top of screen. In the 40dB mode, the trace runs off of the screen and only the central part of the original 80dB scale is displayed.

5.

References

[1] LCD oscilloscope for spectrum analyser, Matjaz Vidmar web pages, web: http://lea.hamradio.si/~s53mv/spectana/lcd.html

[2] A Wideband, Varactor-tuned Microstrip VCO, Matjaz Vidmar, Microwave Journal, 6/1999 pp 80-86. Also – Wideband and low noise microwave VCO, Matjaz Vidmar, VHF Communications Magazine, 4/1998, pp194 – 209.
[3] Spectrum analyser 0...1750MHz, Matjaz Vidmar, VHF Communications Magazine, 1/1999, pp 2 - 30.
[4] Matjaz Vidmar: "Spektrum-Analyzer von 0 bis 1750MHz, Teil 1: Aufbau der Baugruppen", pages 18-30/4-99, "Spektrum-Analyzer von 0 bis 1750MHz, Teil 2: Anzeigebaugruppen und Abgleich", pages 18-29/1-00, AMSAT-DL Journal.

Gunthard Kraus, DG8GB

Internet Treasure Trove

MicroStripes

This is not individual strip line simulator but a matured 3D simulation software package for draft layout and simulation of microwave modules and antennas

Address: http://www.microstripes.com

Microwaves101.com

An interesting and humorous collection information about "Microwaves". The operators of the site stress that not everything is fully developed, but it can be used! The site is well designed with many alphabetically sorted articles, topics, Application Notes and software. It is a wealth of information.

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Azimuth

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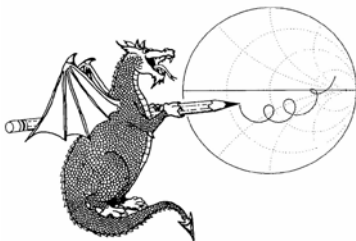
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