

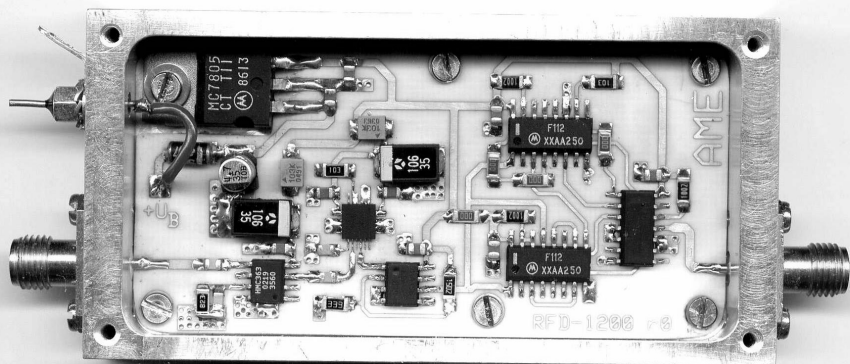


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12GHz Divide By 1000 Prescaler

Alexander Meier, DG6RBP

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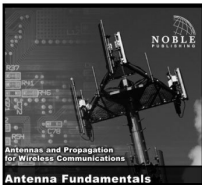
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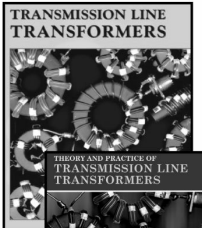
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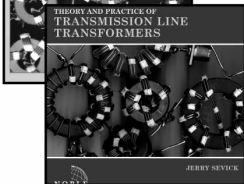
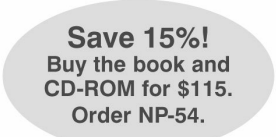


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This issue has a bumper number of articles. Because Christmas is coming there is a constructional article from Carl Lodström so that everyone can build something useful for the shack over the holidays. There is also a story from Gunthard Kraus to make you think twice about buying too much junk (I mean useful equipment!).

For those interested in the 12GHz prescaler, I have been warned by Eberhard Smolka of UKW Berichte that the first batch of divider ICs that arrived was faulty. This will mean a delay if you want to purchase a kit or ready built unit. PCBs, Kits and ready built units will be available so if you are interested, send me an email.

It only remains for me to wish you a Happy New Year and ask you not to forget to renew your subscription to VHF Communications for 2004.

73s - Andy



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Sigurd Werner, DL9MFV

New transmitter with higher performance for 76GHz

A simple concept for a 76GHz signal transmitter operating as a frequency octupler, was described in this magazine recently [1]. The new W band chip CHU3277 (from United Monolithic Semiconductors) is now being tested as part of an improved and more powerful transmitter module. Using only two MMICs, it requires a 5mW input signal at 9.5GHz to generate an output of 49.8mW (about 17dBm) at 76GHz. The 3dB output band covers approximately 3GHz. Options for increasing the signal strength are discussed.

1. Introduction

The octupler for 76GHz recently described [1] has proved its worth in practice, thanks to its uncritical behaviour. Several examples achieving an output of 13dBm, and in some cases considerably more (26mW). To enhance the 76GHz signal further, a circuit has been developed using two CHU2277 chips operated in parallel. This is no trivial matter. Moreover, in this MMIC the RF input has been moved through 90 degrees in relation to the RF output of the main port [2]. The equipment works as follows: The 9.5GHz signal is quadrupled using the CHX2092a chip (38GHz) and

fed through a Wilkinson splitter to two parallel mounted CHU2277 MMICs. The power levels obtained (76GHz) are fed on to two L form striplines (length approximately 3mm), combined with a $\lambda/4$ Wilkinson combiner (mounted on a 0.13mm quartz substrate) and finally fed to a WR12 waveguide (see Fig 1).

Although it was assembled very carefully, the results were very disappointing. The output obtained was only a mere 20% higher than the level attained by a standard module of the type described with only one CHU2277 chip.

This was clearly due to the four additional substrate to substrate transition areas. A large part of the additional power generated was lost at these discontinuities (and naturally in the Wilkinson combiner as well). But a more compact assembly is scarcely possible for technical reasons.

2. New concept

A few weeks ago, a new 76GHz IC developed by UMS was delivered, known as type CHU3277. This W band MMIC comprises a buffer stage, frequency doubler and two four stage amplifiers, which are combined into one output [3]. The input signal required for

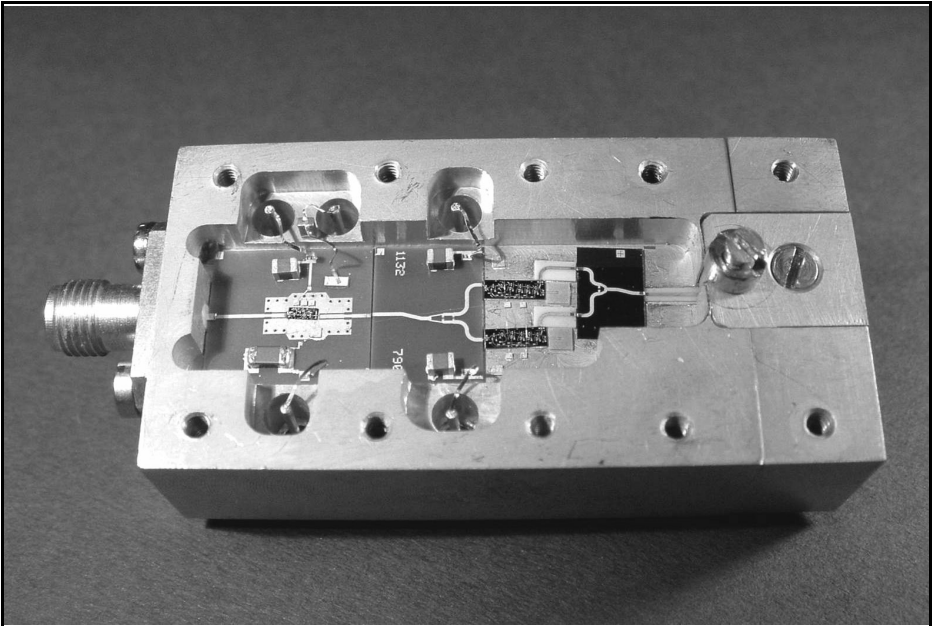


Fig 1: The first transmitter with two CHU2277 chips in parallel. The Wilkinson splitter and combiner can be seen clearly.

38GHz is fed in at 5dBm and the output is (typically!) 18dBm. One advantage here (especially in relation to parallel operation of the chip) is that the input and output ports are mounted in one plane (at the front of the IC in each case).

The design for the following frequency octupler is similar to the module already described [1]. The tried and trusted CHX2092a MMIC is again used as a quadrupler for the 9.5GHz input power, followed by the new CHU3277 chip. The output signal obtained is coupled into a WR12 waveguide through a stripline.

3. Mechanical and electrical assembly

The housing, which measured 38.1mm x 23mm x 15mm, was milled out of brass and then gold plated.

The quadrupler chip (CHX2092a) was mounted on a standard aluminium nitride substrate (0.23mm). The W band MMIC (CHU3277) was glued onto a 0.15mm high platform cut out from the base of the cavity (depth: 4.7mm), in order to balance out the difference in levels between the ceramic substrate and the 76GHz MMIC.

An SMA socket at 9.5GHz was used as the signal input. A stripline, projecting into the WR12 waveguide on aluminium oxide, acted as the 76GHz output. Details are shown in Fig 2.

In contrast to the module described earlier (with the CHU2277 chip), the feedthrough capacitors which lead to the DC circuits were not mounted on the housing base but were positioned on the side walls of the cavity. This made it possible to attach a heat sink to the underside of the housing (Fig 3). The DC lines are blocked by means of 100pF single layer capacitors, followed by 100nF ceramic capacitors. All connec-

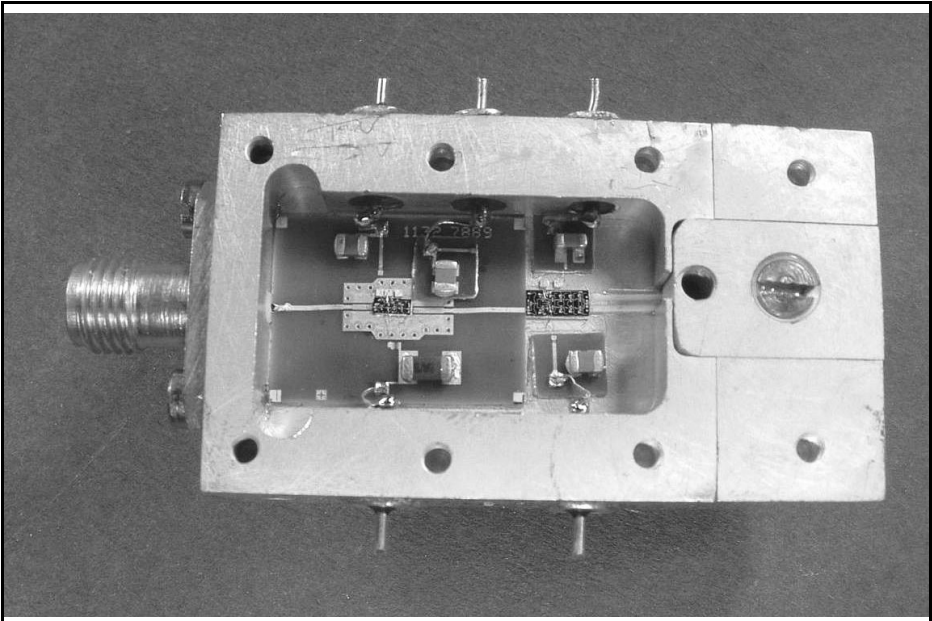


Fig 2: The new transmitter using the CHU3277 chip.

tions between the chips and the substrate were created using $18\mu\text{m}$ gold wire and bond technology (Hybond, Model 572).

4. Results

The drive power at 9.5GHz for the CHX2092a quadrupler chip was approximately 5mW. The gate voltages required were set as follows:

- Gate 1 and 2, minus 0.75V
- Gate 3, minus 0.15V
- Drain, 3.7V (60mA)

Additional details were given in [1].

The CHU3277 W band chip has an internal power supply network, i.e. it only requires two voltages (negative and positive in each case). The best results were obtained using symmetrical voltages (approximately equal values). The

first effect observed, as was to be expected, was proportional behaviour by the chip over a wide range, as between the current consumed and the power emitted:

At 280mA (corresponding to 4.6V in each case), the output was 34mW, at 310mA (corresponding to 5.0V in each case; specified as maximum value for voltage in data sheet), the output was a 42mW. (The current consumption of the negative section of the network in all cases was between 7.5 and 10mA).

The output of the MMIC was adjusted to 76032MHz by the positioning of tiny gold threads, and the coupling of the stripline into the waveguide was optimised. These miniature stubs were fixed in their positions using a UV activated adhesive (Vitalit 5603) [4].

The outputs observed at various frequencies with an output matched at 76GHz are shown in Fig 4. The 3dB output band extends from 74.6 to 78.7GHz (3.1GHz).



Fig 3: Side view of the completed transmitter showing the output feed and DC supply via feed through capacitors.

The 42mW output obtained (16.2dBm) is at the lower range of the specifications for the CHU3277 MMIC. The variation specified in the data sheet (measured on the wafer?) extends from 35mW to approximately 100mW (a difference of 5dB!), with a typical value of approximately 60mW (18dBm) [3]. That means that, in certain circumstances, it could be advantageous to pre-select the MMICs used, for example by using specimens with current consumption levels >300mA at the rated voltages.

An alternative to this procedure, which is not simple, could be to increase the voltages used to more than 5V, while simultaneously paying attention to the maximum permissible currents. These lie at 400mA at least (the absolute limit could even be 450mA). Thus the positive voltage increased to 5.45V (neg. voltage 5.35V), the current consumption increased by 330mA and the output increased by 46mW. A further increase in the current, to 355mA (corresponding to approximately 6V) yielded an output

signal of 49.8mW (just on 17dBm). This output was produced constantly for several hours in several tests, providing that the housing could be prevented from heating up (e.g. by attaching a small heat sink). During short time operation (<5 min) the solid brass body of the housing adequately dissipates the power loss of approximately 2W which occurs in the module. Cover effects could not be detected, which indicates careful matching work.

The two modules assembled to date display almost identical behaviour in both the DC and the RF performance. I am convinced that, by using somewhat higher operating voltages (which are clearly tolerated well), together with better matching of the CHU3277 chip, the 50mW output mark can soon be exceeded at 76GHz.

Using an experimental assembly, an attempt was also made to operate two of the CHU3277 semiconductors in parallel and to combine the power with a Wilkin-

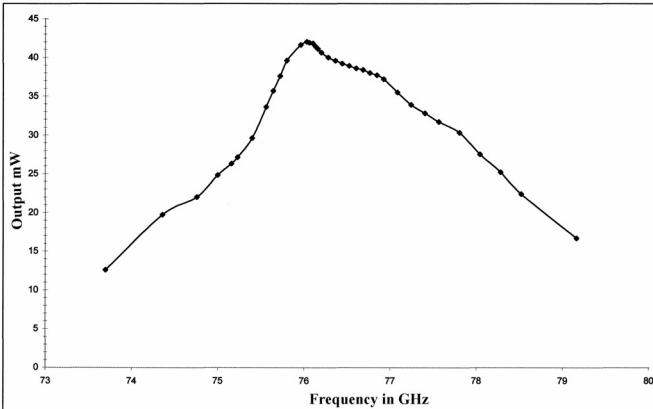


Fig 4: Graph showing the power output from the transmitter with the 76GHz chips drawing 310mA.

son combiner operating as $3\lambda/4$ (on a 0.23mm aluminium nitride substrate). However, the rather poor result demonstrates that some development work still needs to be carried out here (see below also).

5. Outlook and acknowledgements

The experiments demonstrate that highly efficient transmitter modules can be assembled with only a few MMICs. The power level of just 50mW obtained here could be doubled by means of a relatively low loss interconnection of two modules with the same construction. Instead of using a Wilkinson combiner with its high losses, I am thinking of arranging for power addition in the waveguide, using a magic T, though this would have to be well matched. A start has already been made on an assembly to implement this idea. The declared objective is to attain the magic level of 100mW power at 76GHz.

My thanks go once again to my colleagues (Dr. Jünemann, DK7AH and Dr. Hechtfisher, DG4MGR) from Rohde & Schwarz, Munich. The ceramic sub-

strates used come from the development stockpile, and were modified and adjusted by myself for the purposes described. I would like to single out Mr. W. Hohenester, who was a constant companion during this project. My thanks are also due to my friend Jochen Ehrlich (DK3CK), not only for providing mechanical assistance but also for his tireless constructive criticism.

6. Literature references

- [1] Sigurd Werner, DL9MFV, Frequency multiplier fro 76GHz with an integrated amplifier, VHF Communications 1/2003, pp. 35 - 41
- [2] Data sheet from United Monolithic Semiconductors, S.A.S. , Ref DS-CHU22770089, 29-Mar-00.
- [3] Data sheet from United Monolithic Semiconductors, S.A.S. , Ref. DS-CHU3277391, 01-Apr-03
- [4] Sigurd Werner, DL9MFV, A simple concept for an efficient 76GHz transverter, VHF Communications 2/2003, pp 77 - 83.



Alexander Meier, DG6RBP

12GHz divide by 1000 prescaler

Expanding a frequency counter by adding an external prescaler is a simple and inexpensive option if you wish to measure higher frequencies. This new divider module includes the 10GHz band and cuts out some tedious conversion by having a decimal division factor of 1000.

1. Introduction

A high quality, universal 5GHz frequency divider has already been presented in [1].

However, the requirements set for the divider module described below were rather different:

It was to cover the 10GHz band and was to have a decimal divider factor. The high division factor of 1000 means the module can be connected to almost any frequency counter. It is also a simple way of processing the divided signal (TTL level) for special tasks, using a micro controller. Thus, for example, a simple 12MHz frequency counter circuit could be created and combined with the divider module to display and monitor the output frequency of a transmitter.

2. Resolution and accuracy

Determining the frequency of a signal source is a fundamental measurement in high frequency measuring techniques, and it must, above all, be carried out very accurately. In this context, carrying out a measurement is nothing but a comparison between a known and an unknown value. The more accurately the known comparison value can be represented, the more accurate the measurement will be.

In frequency measurement, the frequency, f , is the reciprocal value of the period, t , which is a temporal value. Yet time itself is one of the most accurately reproducible experimental values. Thus frequency measurement becomes one of the most accurately performed measurements.

To measure a frequency, we need a counter, which counts the number of cycles in a specific period of time (gate time) and then standardises on a gate time of one second. The counter is usually a digital circuit, e.g. a micro controller, which is started and stopped by a quartz crystal timed gate.

The uncertainty, δf , of the frequency measurement is made up of the accuracy of the reference frequency of the time base, δf_{TB} , and the uncertainty of count-



ing. The uncertainty of counting amounts to ± 1 cycle, since the gate time is not synchronised with the signal frequency to be measured. Thus, depending on the position of the oscillations, one counting pulse more or less can be generated during the gate time. Depending on the length of the gate time and the level of the counting frequency, this error appears more or less pronounced. [2]

$$\delta f = \frac{\Delta f}{f} = \frac{f \cdot \delta f_{TB} + \frac{1}{f_{Tor}} \cdot \text{divisionfactor}}{f} = \delta f_{TB} + \frac{\text{divisionfactor}}{f \cdot t_{Tor}}$$

Time base + uncertainty of counting

where:

δf = relative uncertainty of frequency measurement,

Δf = absolute uncertainty of frequency measurement,

δf_{TB} = relative uncertainty of reference frequency (time base),

t_{Tor} = time gate,

f = frequency to be measured

In high frequency measurement ($>100\text{MHz}$), a frequency divider must first be used to split these frequencies into smaller frequencies which the counter circuit can process. This does not influence the part of the total error originating from the reference frequency. It is only the influence of the counting uncertainty (usually small) which increases.

The resolution of a frequency counter is just as important as the uncertainty. This is reciprocal to the gate time, which can be made clearer by an example:

A frequency counter has to count a frequency of 10MHz . If it counts the number of cycles in a gate time of 1 second, it will count 10 million. It has thus counted with a resolution of $10^6 \text{ Hz} / 10^6 \text{ oscillations} = 1\text{Hz}$.

If it counted with a gate time of 100ms , it could count only a million cycles in this

time. The resolution would thus be $10^6 \text{ Hz} / 10^5 \text{ oscillations}$, i.e. 10Hz .

It should be pointed out that frequency counters with reciprocal counting methods attain a higher resolution (usually by a factor of 10 or 100) for the same gate time, due to their special mode of functioning.

The resolution of a frequency counter with a prescaler connected in series deteriorates by the value of the division factor. This is because the actual counter can only count the lower frequency that is produced by the divider as an input signal.

$$\text{Resolution} = \frac{1}{T_{Tor}} \cdot \text{divisionfactor}$$

Thus the division factor for a prescaler should be selected as low as possible. This will give the highest resolution for a given gate time!

The division factor of 1000 selected here must now be analysed more closely:

Advantages:

- Conversion on the frequency counter can be carried out very simply. For example, a display of 10MHz corresponds to an input frequency of 10GHz .
- The existing frequency counter needs a frequency range of only 12MHz to be able to make full use of the frequency divider.
- Simple further processing for micro computer circuits based on the low output frequency with TTL level.

Disadvantages:

- The resolution is reduced by a factor of 1000. So, for a measurement time of one second, we obtain a resolution of 1kHz , and at 100ms gate time only 10kHz .
- This disadvantage, though, must be examined still more closely with

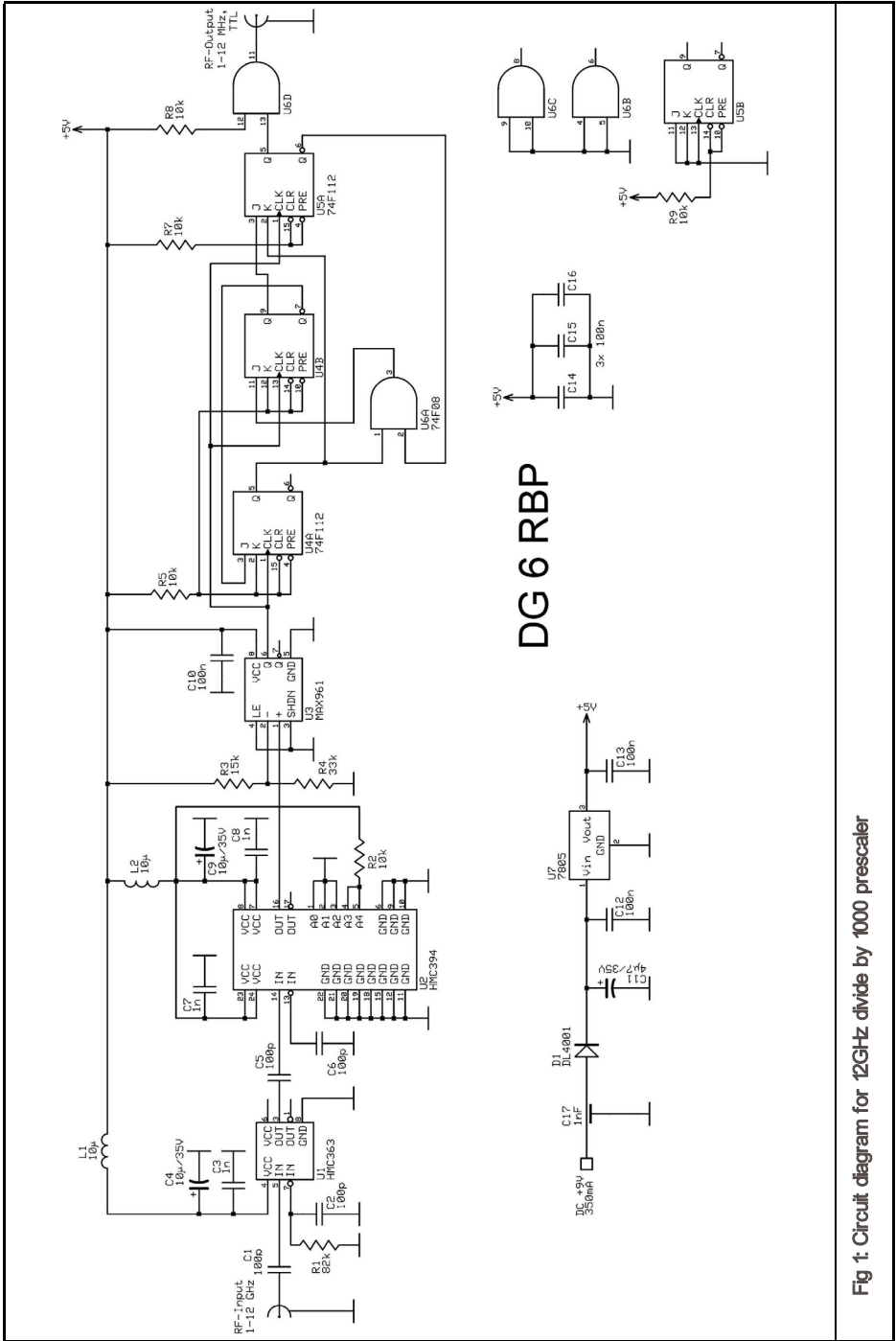


Fig 1: Circuit diagram for 12GHz divide by 1000 prescaler

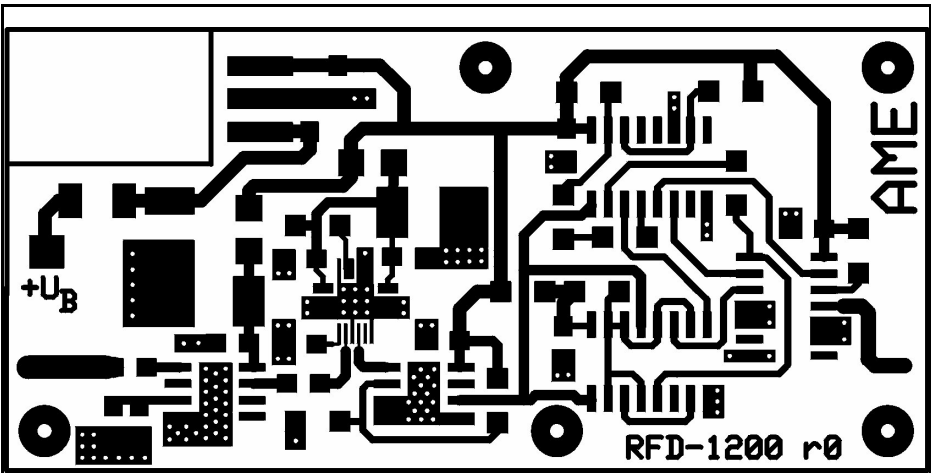


Fig 2: Printed circuit board layout for 12GHz divide by 1000 prescaler.

regard to the combination of a frequency counter and a prescaler! There are two important points to be taken into account here:

- 1) Practically all frequency counters in the medium price range can count directly only up to about 100MHz. All higher frequencies (e.g. 1.3GHz input) are divided down first, frequently by a factor of 64.
- 2) Does the uncertainty of the time base allow for meaningful measurements at high resolution?

Re. point 1:

The present 12GHz frequency divider divides a frequency from 12GHz to 12MHz. A typical frequency counter can count these frequencies directly, with a resolution of 1000Hz at a gate time of 1 second.

If the 12GHz frequency divider divided by only 10, we would have a frequency of 1.2GHz. A typical frequency counter would internally divide this again by 64 before it could count it. The resolution would then be 640Hz for a gate time of 1 second. So we would gain nothing.

Re. point 2:

A normal frequency counter in the medium price range has a quartz crystal as reference, with a typical relative uncertainty of 10^{-6} . If we were to measure a frequency of 10GHz (10^{10} Hz) by direct counting, i.e. without a frequency divider or with a divider factor of 1, with a gate time of 1 second. At the end of the day the measurement would have an absolute uncertainty of $10^{10} \cdot 10^{-6} = 10^4$ Hz due to the reference frequency, plus 1Hz due to the counting uncertainty, which thus gives us 10.001kHz.

If we use the frequency divider, we obtain a resolution of 1kHz for the same input frequency and the same gate time. At the same time, we obtain an absolute uncertainty of 10kHz due to the reference frequency, plus 1kHz counting uncertainty. This gives a total error of 11kHz. As can be seen, the error in both cases, i.e. direct counting or with a frequency divider connected in series, exceeds the resolution with a divider. So it does not make sense to use the divider factor to increase the resolution.

Thus the division factor of 1000 is no disadvantage for an average frequency counter in the medium price range!

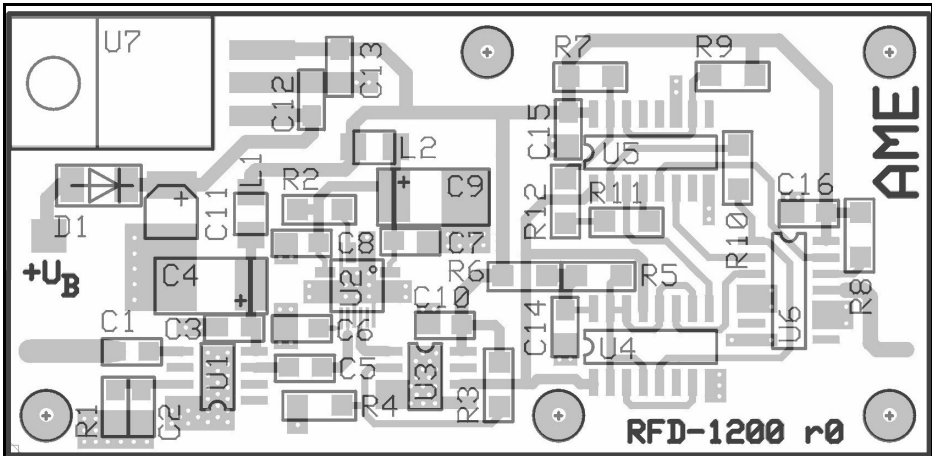


Fig 3: Component layout for 12GHz divide by 1000 prescaler.

Only very high quality equipment fitted with a precision ovened quartz crystal attains a relative reference frequency uncertainty in the region of 10^{-9} to 10^{-11} . The resolution could be meaningfully increased here with a smaller divider factor, but only if the counter can also count the input frequency divided by the divider directly (i.e. without an additional internal prescaler)!

3.

Circuit description

Fig. 1 shows the circuit diagram of the frequency divider. The input signal passes through the coupling capacitor, C1, to an HMC-363 12GHz divider (U1).

Unfortunately divider modules oscillate when there is no input signal, and this one is no exception. This behaviour is a problem when expanding a frequency counter. Without an input signal, the equipment registers an unstable frequency that is not really present. With many modules this can be prevented if the inverting input is earthed through a resistance in the $k\Omega$ range. The module

usually reacts to a reduction in resistance with a reduction in the tendency to oscillate, but it then also has reduced sensitivity. In the present case, the tendency to oscillate could be suppressed using the resistance R1, without having too much of a detrimental effect on the sensitivity.

The signal, divided by 8, then comes to the 5 bit counter, U2 (HMC-394), which is programmed with the connections A0-A4 for a division factor of 25. Here is a useful tip for anyone who is feeling a little irritated looking at the wiring diagram: according to the data sheet [3], you should select precisely the desired divider factor less 1.

The output signal from U2 is a type of square wave signal with a DC offset. In order to obtain a TTL signal from it, which is what the subsequent TTL divider needs, we use the U3 (MAX 961) rapid comparator from Maxim. It compares the signal from U2 with the threshold voltage from the resistance divider, R3 / R4, and thus generates a clean square wave signal at the output with an amplitude of 0V or 5Volts.

In the end, it has proved necessary to develop another synchronous counter

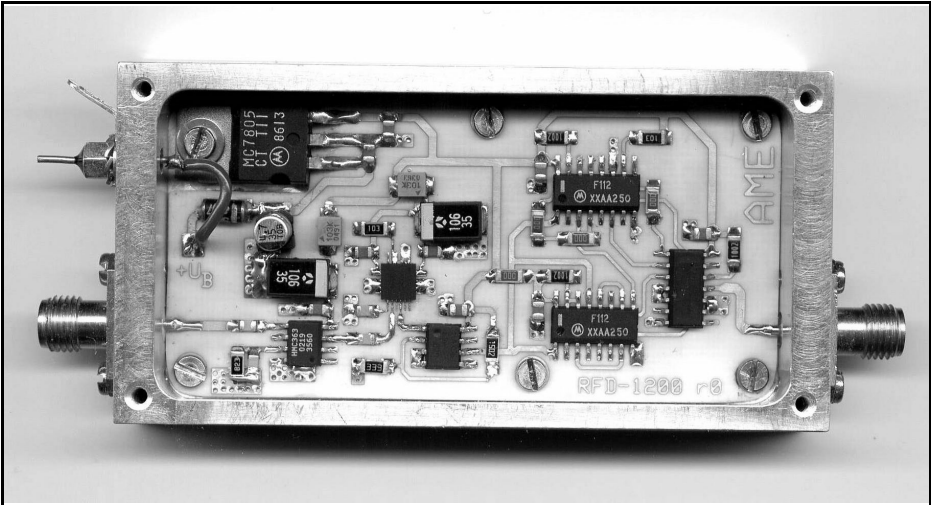


Fig 4: Completed prototype of the 12GHz divide by 1000 prescaler.

manually as a TTL divider by five. To be able to display the numbers 0 to 4, we need 3 Bits, and thus we also need the three JK flip-flops, U4/U5 (each 74F112).

The signal from the TTL divider is fed to the output via the free AND gate used as a buffer.

The operating voltage of the divider module (approximately 9 to 15V) is fed through the feedthrough capacitor, C17, and the reverse battery protection diode D1 to the voltage regulator, U7. This generates a regulated voltage of 5V for the divider modules. The supply voltages for U1 and U2 are fed through LC filters to reduce any RF residue.

4. Assembly of circuit

The circuit diagram was used to develop a board layout (Fig. 2) with the dimensions 71mm x 34mm. This size was selected because there were suitable tinplate housings available. However, these

are advantageous only if the price plays a significant role. A milled aluminium housings is definitely preferable. If a tinplate housing is used, a suitable heat sink should be provided to cool the voltage regulator and the divider module (especially U2). The prototype of the frequency divider was incorporated into a milled aluminium housing.

The base material used for the through hole plated board was 0.81mm thick RO4003C substrate from Rogers. This board material has outstanding high frequency characteristics, but it can be worked like standard epoxy (FR4) and can be obtained at budget prices.

The 50 Ohm track width is about 1.8mm. This width should be maintained, at least from the RF input to the first divider IC, U1. This track width can not always be adhered to as the RF signal moves over the board, since the connections to the divider ICs must be made very narrow. However, this is no great tragedy, provided the tracks remain very short.

The components are mounted by hand in accordance with the component drawing (Fig. 3). Problems arise only with the divider, U1, and the 5 bit counter, U2.

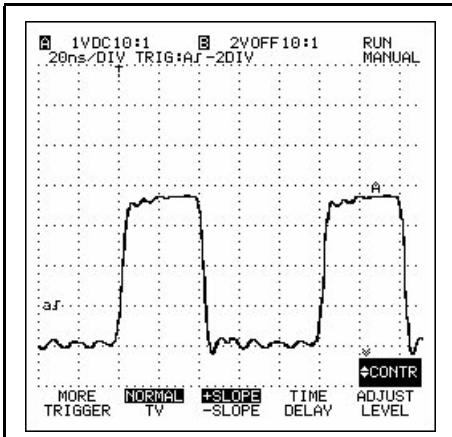


Fig 5: Output signal for 10GHz input.

The IC, U1, must be soldered to the earth area on its underside (!) This can be done through prior tin plating, or better using solder paste and hot air (or a hot plane). Things become more difficult with U2, the connections of which are, to a very large extent, on the underside of the chip, and where the pin distance, 0.5mm (!), is extremely small.

Anyone venturing to copy the divider should therefore have a lot of experience in soldering small SMD components.

By the way, the divider, U2, runs relatively warm, its current consumption being just on 200mA. It is cooled via its earth area and the through plating to the aluminium housing base.

Fig. 4 shows the prototype of the divider module.

5. Parts list

- SMD capacitors:
 C1,C2,C5,C6 100pF 0805
 C3,C7,C8 1nF 0805
 C10,C12-C16 100nF 0805
 C4,C9 10µF / 35V SMD tantalum

- C11 4.7µF / 3V SMD electrolytic capacitor
 C17 1nF feedthrough capacitor, Actipass FC-102P-10A
 L1,L2 10µH SIMID 1210
 R6,R10-R12 0Ω, 1206
 R2,R5,R7-R9 10kΩ, 1206
 R3 15kΩ, 1206
 R4 33kΩ, 1206
 R1 82kΩ, 1206
 D1 DL4001
 U1 HMC 363 S8G
 U2 HMC 394 LP4
 U3 MAX961CS8
 U4, U5 74F112 SMD
 U6 74F08, SMD
 U7 7805
 1 x LP DG6RBP-006
 1 x Aluminium milled housing
 or
 1 x Tinplate housing 71 x 34 x 30mm
 2 x SMA sockets with 4 hole flange

6. Measurement results

To test the frequency divider, a signal with a power of 1mW (0dBm) was fed in from a 20GHz HP 83732A signal generator, and the frequency was varied between 500MHz and 15GHz. The behaviour of the divider was monitored using frequency counters and an oscilloscope. No functional errors could be observed here. The output signal is shown in Fig. 5 for an input frequency of 10GHz.

If the divider is operated without an input signal, the output is either low or high level, depending on the counting step of the TTL divider. A tendency to oscillate without an input signal arose only where relatively long coaxial lines were connected without load.

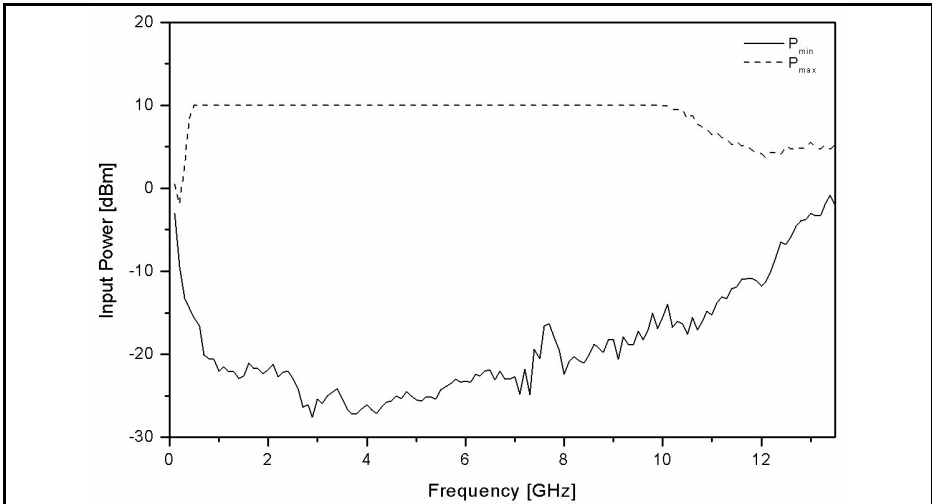


Fig 6: Minimum input power requirement for the 12GHz divide by 1000 prescaler.

In a second test, the power was varied at constant frequency (1GHz). In this test, it was established that at very low power levels false division factors arise! The precise input power level at which this behaviour begins depends on the input frequency and establishes the minimum level required for the frequency divider.

This minimum input power requirement for the divider module is shown in Fig. 6.

7.

Technical data

- Frequency range
 - Input min. 0.5 to 12GHz
 - Output 0.5 to 12MHz
- Divider factor 1000
- Input level approx. 10 to +5dBm
- Output level¹ 5V TTL
- SWR (HF input) typ. 2...6
- Tendency to oscillate with an open input - no²

- Connectors SMA sockets
- Operating voltage +15VDC
- Current consumption max. 350mA

¹ Without an input signal, there is either a low or a high level at the output.

² When relatively long, open coaxial cables are connected, or in similar situations, in certain circumstances a tendency to oscillate can arise when there is no input signal.

8.

Literature references

[1] Alexander Meier: Pre divider (:10) to 5GHz, VHF Communications 1/2002, pp 22 - 26

[2] Thumm/Wiesbeck/Kern: High-frequency measuring technique, Teubner Verlag 1997, Stuttgart

[3] Data sheet HMC-394, Hittite Microwave Corporation, 12 Elizabeth Drive, Chelmsford MA 01824



Dr Volker Grassmann, DF5AI

New 144 MHz transatlantic beacon

In December 2002 I received a dx report claiming reception of the Canadian 144MHz beacon VE1SMU in Germany. The Dubus magazine finally figured out that this beacon has been off the air for more than three years, i.e. the observed signal did not originate from Canada but somewhere else. Nevertheless, similar reports have existed for many years. In the VHF-DX email forum, the possibility of terrestrial 144MHz transatlantic propagation was recently discussed with passion. European radio amateurs miss the VE1SMU beacon because there is no alternative transmitter available on the opposite side of the big pond. Now there will be a new beacon in Canada. Please read the below email information which Andy, editor of the VHF Communications magazine, and I received from Frank, VO1HP:

“I have read with interest Dr. Volker Grassmann’s article in your VHF Magazine 2-2003 edition. In that article he states that it would be desirable to have a VHF beacon on the air from eastern Newfoundland. The Marconi Radio Club of Newfoundland (MRCN) and the Bacalieu Amateur Radio Klub (BARK) are currently in the process of placing such a beacon on the air. The beacon will transmit on 144.400MHz. The transmitter 250 W PA is the same one previously used by VE1SMU beacon. The VHF

exciter board and the CW ID board have been built by Joe Craig VO1NA. The 11 element Cushcraft yagi, 1/2 inch feeder and 35 amp power supply to run the beacon have been donated by Frank VO1HP. The yagi and feeder were installed by Boyd VO1CBS, Roy VO1XP and Joe VO1NA on August 30, 2003. The repeater site is in the middle of a large field at a turkey farm!!....and it has a clear view of the North Atlantic ocean through the mouth of Conception Bay on the South East coast of VO1 land. It is in the town of Carbonear at N47 45’24” W53 12’59” with the antenna at about 300’ (90m) ASL and beaming at 60 degrees true.

The beacon VO1ZA was placed on the air on October 19th 2003 at 1330UTC on 144.400Mhz by Frank VO1HP and Boyd VO1CBS. Any amateur station who claims to have heard this beacon are asked to send reports via email to either volhp@rac.ca or volna@rac.ca. Please describe what you have heard and if possible the complete information being transmitted by the beacon.

Frank Davis, VO1HP”

Up to date information can be found on Volker’s web site <http://www.df5ai.net>



Gunthard Kraus, DG8GB

Practical Project: Design and construction of a high quality 100MHz bandpass filter

The editors of VHF Communications and UKW Berichte recently issued a call for readers to describe successfully implemented projects and thus to give a new impetus to the activities of DIYers. Here is just such a project that required “intermediate development” at short notice. It did take longer than was expected to develop the circuit and to make sure that the targets were achieved.

1. Range of application

New projects are continually required for practical laboratory work. Various kinds of receivers or converters can be assembled from a set of microwave parts, some of these will be available and some will have to be designed and built, such as low noise amplifiers, mixers, stripline filters, VCOs etc.

A standard commercial car radio with synthesiser tuning and built in final stages (2 x 12 Watts) can be used as a 100MHz low frequency amplifier to reproduce FM or PSK signals, and for audio demonstrations to large groups of people.

The car radio was selected because such radios have high sensitivity and good screening built in as standard features,

and frequencies can also be easily adjusted or saved through a menu driven operation.

This bandpass filter can be connected between the intermediate frequency output of a microwave converter and the antenna input of the radio to ensure that in practice only signals converted to 100MHz are audible even if it is the image frequency.

2. Specifications

The following specifications were laid down for the development of the bandpass filter:

- System impedance: $Z = 50\Omega$
- Degree of filtration: $n = 5$
- Centre frequency f_0 : 100MHz
- 3dB bandwidth: $b = 5\text{MHz}$
- S21 Ripple in transmission range: 0.3dB
- Transmission loss in pass band: $< 10\text{dB}$
- Edge steepness: at 90MHz or 120MHz the attenuation should have risen to approximately 70dB with respect to the transmission range
- Filter attenuation in stop band in

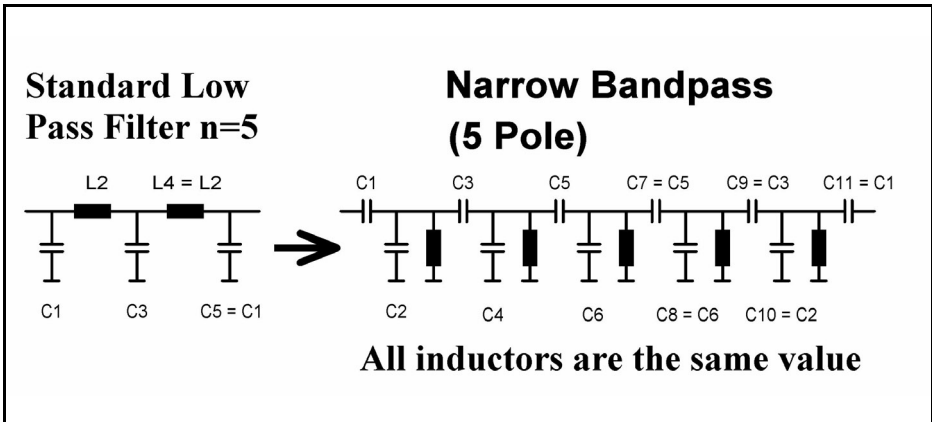


Fig 1: The standard low pass filter is transformed to the narrow bandpass filter with series capacitors and shut inductors.

frequency range from 0 to 1GHz
min. 70dB

- Filter type: Chebyshev Narrow Bandpass
- Mechanical: Printed circuit board, coated with 35µm copper on both sides, made from Rogers material R04003 with a thickness of 25 MIL = 0.813mm
- Board dimensions 30 x 130mm. Incorporated into a milled aluminium housing with SMA sockets as input and output
- Distance between board and cover, 13mm.

Following some test simulations, the so-called “Narrow Bandpass” was selected as the filter type, with $n = 5$, which can achieve very small bandwidths. It is created from the familiar, standard, low pass filter by suitable transformation work and is characterised by the fact that the inductances are the same value in all the resonant circuits.

As a result of experience, $L = 100\text{nH}$ was selected (for with reactive impedances exceeding 80 to 100 Ohms at 100MHz for coils or capacitors, problems rapidly arise with the broadband nature of the filter properties). This only works correctly if components of the highest possi-

ble quality are used.

This also applies to the circuit board, hard paper or FR4 material should not be used. Only genuine microwave materials are permissible, made from Teflon, ceramics or modern materials such as Rogers R04003 which has the combination of excellent mechanical and electrical characteristics which are not far below those of Teflon, even at 10GHz.

This type of filter is essentially high impedance, so it must be matched to the 50Ω source or load by means of an input or output capacitor.

The circuit diagram is shown in Fig. 1. The actual filter calculations are carried out using the free filter CAD program “fds” available from the Internet.

3. Circuit diagram of filter

Entering the design information into the input mask of the filter program (fds), we obtain a picture like Fig. 2. You should take the trouble to draw the associated circuit immediately, because this is actually where we start to think about the practical implementation.



```

FilDes--Filter Design System--Ver 1.20
Narrow BandPass Filter
What is the value for the inductors? (in nH)
100
Coupling caps are:
C(1,2) = 0.842 pF
C(2,3) = 0.675 pF
C(3,4) = 0.675 pF
C(4,5) = 0.842 pF
Resonating caps are:
C(1) = 24.49 pF
C(2) = 23.81 pF
C(3) = 23.98 pF
C(4) = 23.81 pF
C(5) = 24.49 pF
The characteristic impedance of the filter is 2017.42 ohms
Match to a lower rs/r1? (Y/any key)
Default impedance is 50.00 ohms. Type 'D' for different src and load
or hit any key to match to default.
Add 5.074 pf to the source end, and 5.074 pf to the load end.
Then change C(1) to 19.539 pf and C(5) to 19.539 pf.
(S)ave to file (P)rint or (C)ont?
-

Type: NBP Poles: 5 Ripple: 0.30 Zsrc 50.00 Fo 100.00 MHz
BW design (ripple or 3 dB): 3 dB
    
```

Fig 2: The results produced by the filter design program "fds".

As already mentioned we will need coils of extremely high quality and stability, so here we use industrially manufactured helical components, e.g. from Neosid, with $L = 100\text{nH}$ and a basic Q of 130.

The circuit capacitors, like the input and output capacitors, are a combination of two or three SMD size 0805 standard values made from "NP0" material or better. The tiny coupling capacitors are under 1pF and must be accurate so only printed inter-digital capacitors can be

considered. Fig. 3 shows the complete circuit design.

As a check, two simulations of the filter characteristics for the ideal case were carried out using PUFF. Fig. 4 shows the frequency range from 90 to 110MHz, it is easy to recognise that the requirement for "at least 70dB filter attenuation at 90MHz" is adequately met. Fig. 5 shows the permissible Chebyshev ripple of 0.3dB in the pass band and confirms the efficiency of the free filter program used.

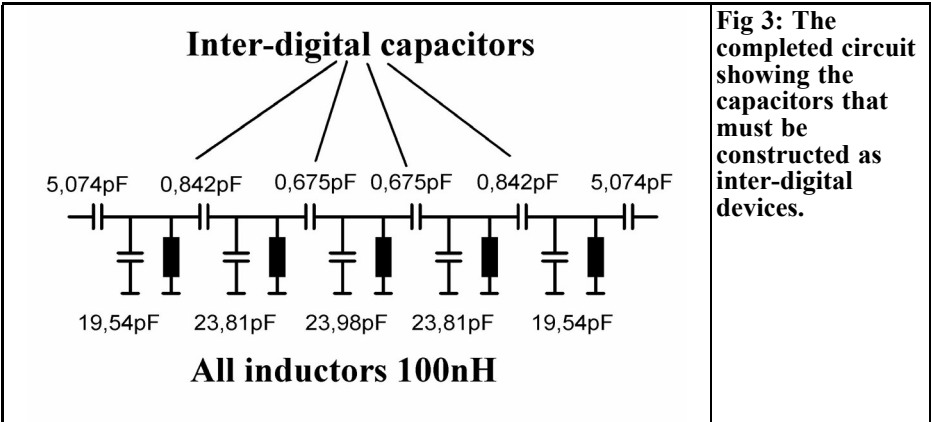


Fig 3: The completed circuit showing the capacitors that must be constructed as inter-digital devices.

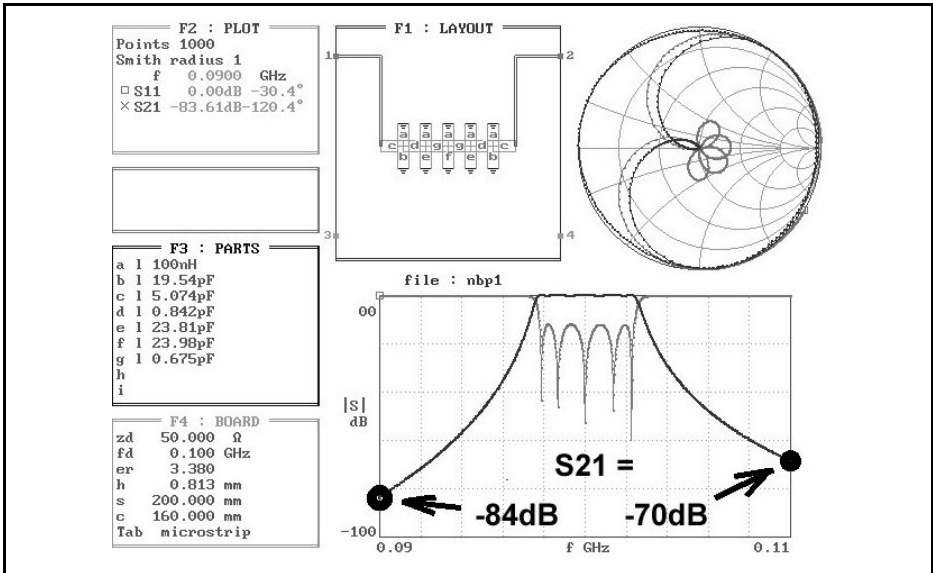


Fig 4: PUFF simulation showing adequate attenuation at 90 and 120MHz.

4. Realisation of the inter-digital coupling capacitors

The values required for the coupling capacitors are very low and lie between 0.5 and 0.9pF, but must be produced with a high degree of precision. For this reason, discrete solutions are excluded, and we are left with a “printed format”.

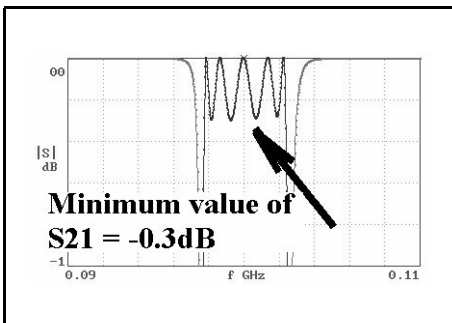


Fig 5: Chebyshev ripple in the pass band.

The design of such capacitors, using nomograms, has already been described in VHF Communications articles. [1] [2].

However, when the first attempts were made to assemble the printed circuit on the board, it became clear that the divergences of 10% to 15% between theory and practise arising using these procedures can not be tolerated and another solution is required. Fortunately, the free student version of APLAC allows for the simulation of an individual capacitor of this type with really high precision, although it then refuses to examine the complete filter circuit saying “Memory restricted in this version...”.

We therefore propose the following path:

- The two inter-digital capacitors required (0.675pF and 0.842pF) are individually developed and optimised one after the other. In addition, an S parameter file is generated for each capacitor that can be entered into the PUFF simulation.
- The complete filter circuit is first simulated using PUFF and then

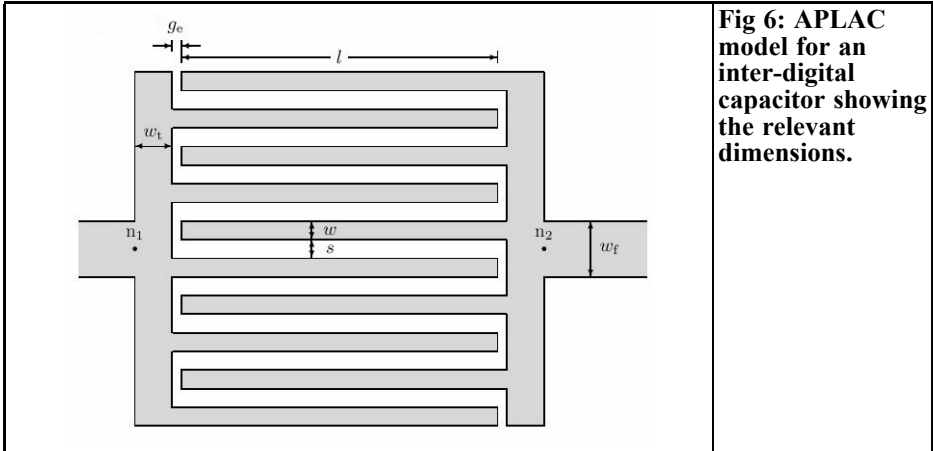


Fig 6: APLAC model for an inter-digital capacitor showing the relevant dimensions.

manually optimised (the inter-digital capacitors not only affect coupling but also add extra inherent capacitance to the circuit capacitors, the latter must be correspondingly reduced).

- Only now is the coil Q taken from the data sheet also inserted into the circuit as a loss resistance for any inductance and thus simulates the transmission curve to be expected from the structure of the printed circuit board.

We began with the design of the small capacitor of 0.675pF. The first step, according to the start of the APLAC program, is to open a new file and then right click on the empty screen. Under “Basics”, we find the components “Port” and “Ground”, each one is needed twice (but in the student version they have to be called up individually one after the other). The inter-digital capacitor can be inserted by right clicking on mouse / microwave / microstrip / Micap. Following positioning of these five components, double click on the left hand mouse button on a free space, which allows a roll of wire to come out of the cursor and the circuit can be wired up (see also the APLAC presentation in [3]).

The S parameter file for the simulated capacitor is not generated until you have

clicked on the output port and, under “Attributes”, entered the line

```
STORE C0675.s2p GHZ MA
```

The two system resistance ports have already been automatically selected at 50Ω, as the entry “50” is automatically posted. However, to make things clearer, you should always make all attributes visible. Now the crucial question arises: how are the technical details of an inter-digital capacitor programmed?

Fig. 6 can help us here, since it originates from the corresponding manual (RF Components / Microstrip Parts / Micap) and tells us what APLAC needs for correct simulation, where:

- W = Finger width
- L = Finger length
- N = Number of fingers
- S = Finger interval
- Ge = Gap at end
- WT = Width of connecting line for individual fingers on each side, which is indicated as Transmission Line.
- WF = Width of microstrip supply line leading to capacitor connections

0.25mm was selected for the finger widths and the various intervals. Firstly, this can be controlled satisfactorily by the board manufacturer, and secondly

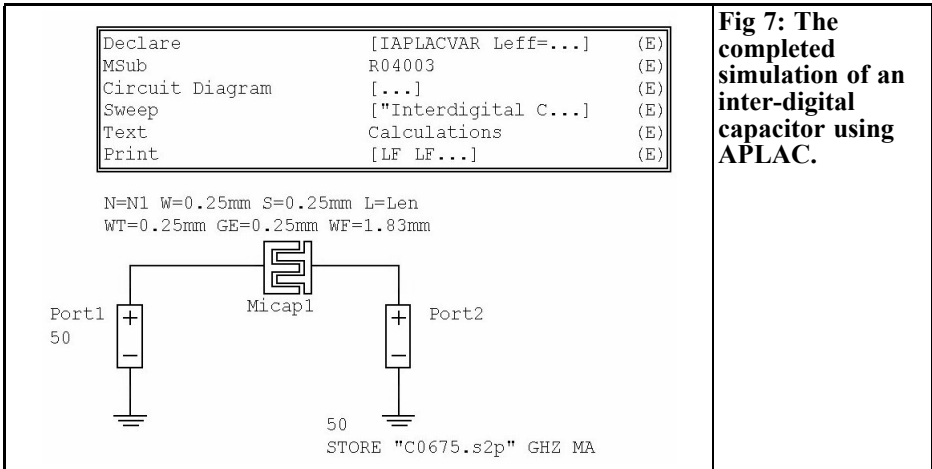


Fig 7: The completed simulation of an inter-digital capacitor using APLAC.

this value is not so high that it brings the desired coupling capacity down too far.

The selection of six fingers was based on the older designs already mentioned [1], [2], and the start value for the finger lengths was likewise taken from them and subsequently optimised. These values must now be correctly transferred into the attribute field of the capacitor (after double clicking on the symbol) In Fig. 7 all the moves discussed can be seen. Just make sure that an additional variable, "N1", has been introduced for the number N, and that a new variable, "Len", has likewise been introduced for the finger length. This is not absolutely necessary, but it makes things clearer for outputting the calculated values later. Moreover, it is now high time this project was saved under a suitable name.

In addition, the "Object Box List" was also mixed in. It keeps a record of the necessary simulation files and can always be called up onto the screen using "Presentation / Show Object Box List". The entry "Isweep" (which can not be used here!) is deleted.

Five new files have to be added to existing "Circuit File". These are described in detail below. Please note that the sequence of the following files in the Object Box List must be adhered to...

The "Declare" file

In this file, all "Initialised APLAC variables" (IAPLAC- VAR) required later are declared. They must always have a start value of some sort. In the simplest case, just enter a "1". We then have:

```
IAPLACVAR Leff=1.0
IAPLACVAR Cg=1.0
IAPLACVAR Cm=1.0
IAPLACVAR Cgtot=1.0
IAPLACVAR Cmtot=1.0
IAPLACVAR Len=7.30mm
IAPLACVAR N1=6
```

Comments:

Leff - represents the actual effective finger length, calculated by the program.

Cg - expresses the grounded capacitance per unit length, i.e. the additional relative inherent capacitance per unit length arising. It will detune the resonant circuit to the left and right of the capacitor.

Cm - corresponds to the coupling capacity desired, though once again expressed per unit length.

Cgtot - is the actual effective inherent capacitance for each individual capacitor connection, by which the associated capacitor must be reduced. It is calculated as the product of Cg and the effective length, Leff.

**Table 1: Contents of Sweep file.**

```
"Interdigital Capacitor 0.675pF"
LOOP 1001  FREQ  LIN    0.000001GHz  3GHz
WINDOW=0  Y      ""      "dB"   -40    0
SHOW W=0   Y      MagdB(S(1,1))
          +Y     MagdB(S(2,1))
```

Cmtot - is the true objective: the actual coupling capacitance, calculated as the product of C_m and Le_{ff}

Len represents the finger length verifiable in the layout

N_1 - is the number of fingers used.

The "Msub" file

All the data concerning the board material (Rogers R04003) and the circuit board are contained in this file. The following entries are needed:

```
ER=3.38
H=0.813mm
T=0.035mm
RHO=0.75
RGH=2um
TAND=0.0005
COVER=13mm
LEVEL=2
```

The following details are included, in this order:

```
Relative permittivity: 3.38
Board thickness: 0.813 mm
Copper coating: 35 μm
Specific resistance of copper in comparison with gold: 0.75
Surface roughness: 2μm
Board material dielectric loss factor: 0.005 where  $f = 100\text{MHz}$ 
Distance between board and cover: 13 mm
LEVEL = 2 means simulation must be extremely precise, regardless of the amount of calculation
```

The "Circuit" file

It is automatically set up when the wiring diagram is drawn and no additional en-

tries are required.

The "Sweep" file

This is not only used to generate the S parameter file for the desired capacitor, but also allows S_{11} and S_{21} to be represented in the frequency range from 1kHz to 3GHz. A start value of 1kHz is used because AppCad 3.0 gives an error message if the frequency in an S parameter file begins with zero.

The "Sweep" file required is shown in Table 1.

Significance of individual lines of "Sweep" file:

- 1st line: Designation of simulation project, included in desired printouts
- 2nd line: Simulate 1001 points in frequency range from 1kHz to 3GHz and divide this range up linearly.
- 3rd line: Provide diagram for outputting results with vertical axis divided up from 40dB to zero dB.
- 4th line: It must always be empty, must NEVER be forgotten and separates the simulation section from the display section
- 5th line : Display magnitude cycle of S_{11} in zero window
- 6th line: Display magnitude cycle of S_{21} in same window.

The "Text" file

This is used in calculation and saved separately as a public declared variable, and subsequently determines the precise capacitor data (see entries and comments in "Declare" file):

```
Call  $Le_{ff} = \text{Ref}(\text{Micap}1, L)$ 
```


**Table 2: Contents of Print file.**

```

LF LF
S "Calculation of an Interdigital Capacitor of 0.675pF"
LF LF
S "Number of Fingers = "REAL N1
LF
S "Length of a finger = "REAL Len
LF
S "Total grounded capacitance on every side = "REAL Cgtot
LF
S "Total coupling capacitance = "REAL Cmtot
LF

```

```

Call Cg=Ref(Micap1,C1)
Call Cm=Ref(Micap1,C)
Call Cgtot=Cg*Leff
Call Cmtot=Cm*Leff

```

The "Print" file

It provides for the outputting of the calculated parallel and coupling capacities on a page of their own. To make things clearer, "LF" stands for line feed. And "S" identifies a string. This is the associated explanatory text (Table 2).

If this work is correctly done (and the new files are listed in the specified order), the simulation can be started

using <Control + S>.

After the necessary calculation time, something appears on the screen. But only the printout is important (Fig. 8), whilst the representation of the S parameter is more for information. To make things clearer, it should be explained that the calculated values are always output in the basic units. The entries must therefore be indicated in the following way:

- Number of fingers = 6
- Finger length = 7.30mm
- Total effective inherent capacitance on each side = 535.489 Femtofarads = 0.535pF

```

APLAC 7.51 Thu Dec 26 2002 at 07:50:15
Copyright (c) APLAC Solutions Corporation, Finland, 1998-1999

```

```

APLAC 7.51 WARNING: Mlin T less than three skindepths
(TLine_VCCS2<-Mlin_TLine<-tim2<-Micap1)

```

```

Calculation of an Interdigital Capacitor with 0,675pF

```

```

Number of Fingers = 6.000
Length of a finger = 7.300m
Total grounded Capacitance on every side = 535.489f
Total coupling capacitance = 675.010f

```

```

End of APLAC 7.51 Thu Dec 26 2002 at 07:50:18 (CPU-time = 2.89 s)

```

Fig 8: The APLAC printout for the 0.675pF inter-digital capacitor.

**Table 3: Setup file for PUFF showing important parameters in bold.**

<code>\b{oard}</code>	<code>{.puf file for PUFF, version 2.1d}</code>
<code>d</code>	<code>0 {display: 0 VGA or PUFF chooses, 1 EGA}</code>
<code>o</code>	<code>1 {artwork output format: 0 dot-matrix, 1 Laser Jet, 2 HPGL file}</code>
<code>t</code>	<code>0 {type: 0 for microstrip, 1 for stripline, 2 for Manhattan}</code>
<code>zd</code>	<code>50.000 Ohms {normalizing impedance. 0<zd}</code>
fd	0.100GHz {design frequency. 0<fd}
er	3.380 {dielectric constant. er>0}
h	0.813mm {dielectric thickness. h>0}
<code>s</code>	<code>200.000mm {circuit-board side length. s>0}</code>
<code>c</code>	<code>160.000mm {connector separation. c>=0}</code>
<code>r</code>	<code>0.010mm {circuit resolution, r>0, use Um for micrometers}</code>
<code>a</code>	<code>0.000mm {artwork width correction.}</code>
mt	0.035mm {metal thickness, use Um for micrometers.}
sr	5.000Um {metal surface roughness, use Um for micrometers.}
lt	5E-0004 [dielectric loss tangent.]

- Total coupling capacitance = 675.01 Femtofarads = 0.675pF.

The 0.842pF capacitor is also calculated in this way. In the “Declare” file, the number of fingers is altered to $N1 = 8$ and the finger length is changed to $Len = 6.44\text{mm}$. And dont forget to double click on the output port on the screen and change the designation of the S parameter file in “C0842.s2p”. Now save, under a new name, and then check the sequence of files, carry out a simulation and incorporate the result:

- Number of fingers = 8
- Finger length = 6.44mm
- Total effective inherent capacitance on each side = 601.940 Femtofarad = 0.602pF
- Total coupling capacitance = 842.387 Femtofarad = 0.842pF.

So APLAC has completed its task and can be shut down again.

5.

Preparation of circuit with the help of PUFF

Preliminary note on PUFF:

To simplify working with PUFF, we create a new folder for a new project and unzip or copy PUFF2.1 into it. This folder is then given the name of the corresponding development and an Icon with the same name on the Windows screen. You should only work with the better Protected Mode version of PUFF and should therefore set up a path to the puffp.exe file! Right click on your PUFF icon and select the properties. Select the memory tab and use the settings shown as per [4] and [5]. Activate “Auto” under the “Conventional memory” heading in both windows. Under EMS, XMS and DPMS the entry required is 4096 Kilobyte. You can now experiment with PUFF without problems. Once development work has been completed, only the results data are saved, and then the operations folder is deleted again.

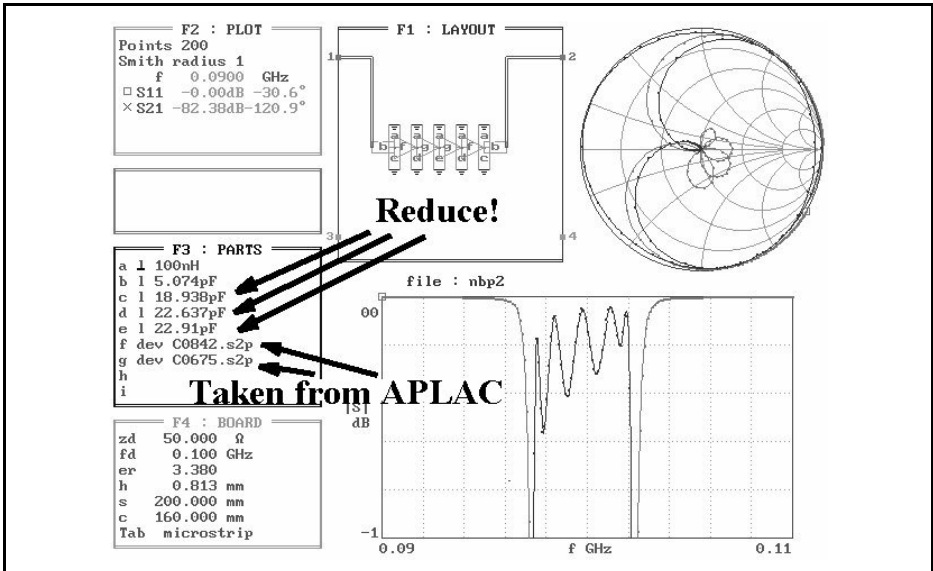


Fig 9: PUFF simulation using inter-digital capacitors showing results that are not ideal.

After that preliminary note, we can get started. Open Windows Explorer and copy the two S parameter files generated by ACLAP (C0675.s2p and C0842.s2p) into this new PUFF folder. Then use the text editor to open the setup.puf file, to enter the correct material and circuit board data.

Table 3 lists the data from the file for the Rogers R04003 circuit board used (the important points are in bold type!).

Once this is complete, start PUFF and enter one component after another in list F3. However, the entire circuit must be modified compared with Fig. 3 and Fig. 4, at the following points:

Instead of the discrete coupling capacitors with 0.675pF or 0.842pF, the two S parameter files generated by APLAC, C0675.s2p and C0842.s2p, are now used. This happens in F3 with the two lines:

```
device C0675.s2p
device C0842.s2p
```

In addition, in all five resonant circuits the circuit capacitors must be reduced by

the additional inherent capacitances introduced by the adjacent inter-digital structures. This produces the following changes:

- First and fifth circuits:
 $C_{circuit} = 19.540pF \ 0.602pF = 18.938pF$
- Second and fourth circuit:
 $C_{circuit} = 23.810pF \ 0.602pF \ 0.535pF = 22.637pF$
- Middle circuit:
 $C_{circuit} = 23.980pF \ 0.535pF \ 0.535pF = 22.91pF$

If we now repeat the simulation as per Fig. 4 with this circuit changed in this way, the result seems rather a long way from the desired ideal (Fig. 9). Fortunately, the solution is relatively simple, since the reason for this situation should be sought in the (simple!) relationships between the frequencies and the inter-digital capacitance values! So you keep making careful corrections to the values of the resonant circuit capacitors until you have achieved the best compromise. Fig. 10 shows this objective and the

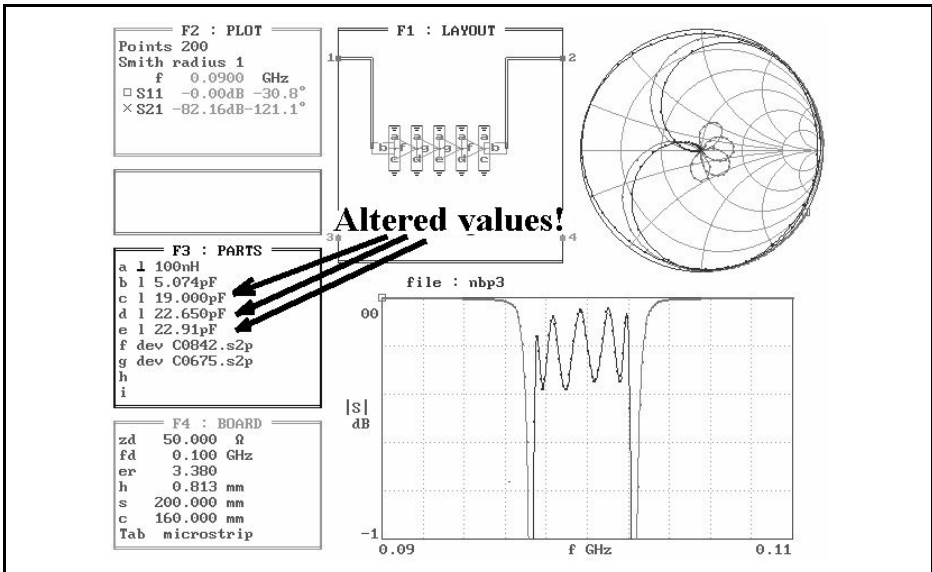


Fig 10: PUFF simulation using inter-digital capacitors showing results after adjusting the capacitor values.

capacitor modifications required for it.

Things don't get interesting until you bring the actual coil losses present into the simulation. With the Neosid helical filters used, the data sheet specifies a basic Q of $Q = 130$ for the frequency $f = 100\text{MHz}$. However, if you rotate the core inward during adjustments in order to reduce the inductance, then this quality is reduced! We are actually dealing with brass cores, into which eddy currents are induced via the coils magnetic field, thus reducing the inductance. But this is equivalent to an increase in losses and it is more advantageous to work with a quality of only $Q = 100$. If we consider these losses as a series resistor for each coil, then we must set 0.63Ω on each side. Fig. 11 shows the effects of these losses. They cause the transmission loss to increase to approximately 7.3dB, whilst the stop band remains practically unchanged.

It becomes interesting later on to compare the measured results with this prediction, for so far, very optimistically, we have assumed that all circuit capacitors

are ideal and loss free and we must therefore undoubtedly count on another slight deterioration.

6. Practical construction of the bandpass filter

The board dimensions are defined by the home made milled aluminium housing, with a screwed on screening cover, used by the author. These dimensions are used for all microwave circuits and, on grounds of rationalisation, they now come in only the two board sizes, 30mm x 50mm and 30mm x 130mm. The four fixing screws are then arranged in a grid, 24mm x 44mm or 24mm x 124mm.

If you take a closer look at the bandpass filter circuit board (Fig. 12), you will very quickly recognise the principle used, which is a tried and trusted one for microwave circuits:

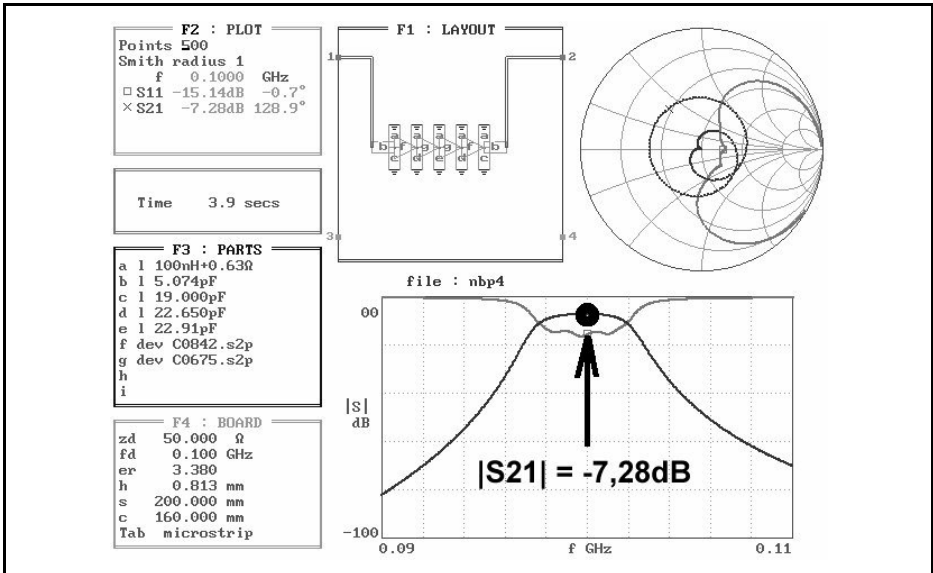


Fig 11: PUFF simulation showing the effect of losses.

The underside of the circuit board is a continuous earthing area. The two SMD coupling capacitors are inserted at the input and output in the continuous 50Ω microstrip line, which has a width of 1.83mm, on the top face. In addition, this line is interrupted by the four inter-digital capacitors required. The five insulated earthing areas, with numerous feed through links, are designed for the five resonant circuits. Do not use any other earthing areas for frequencies from 100MHz upwards, since these through hole plated islands ensure low inherent instability and satisfactory broadband behaviour!

The helical filters do pose one small problem due to their overall height, which with the connecting feet are approximately 15mm, and thus exceeds the clearance height of 13mm between the circuit board and the cover. This problem is solved by flush mounting with the silver plated housing soldered directly to the associated earth islands.

The circuit diagram and the components required can be seen in Fig. 13, whilst Fig. 14 shows the final structure.

If you add up the capacitances in each individual circuit and compare them with the values calculated above and then add

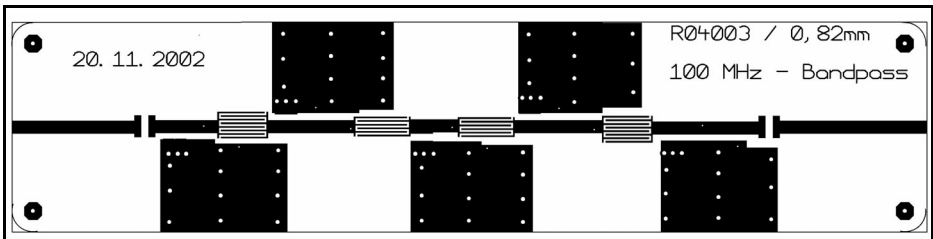


Fig 12: Printed circuit board layout for the 100MHz bandpass filter.

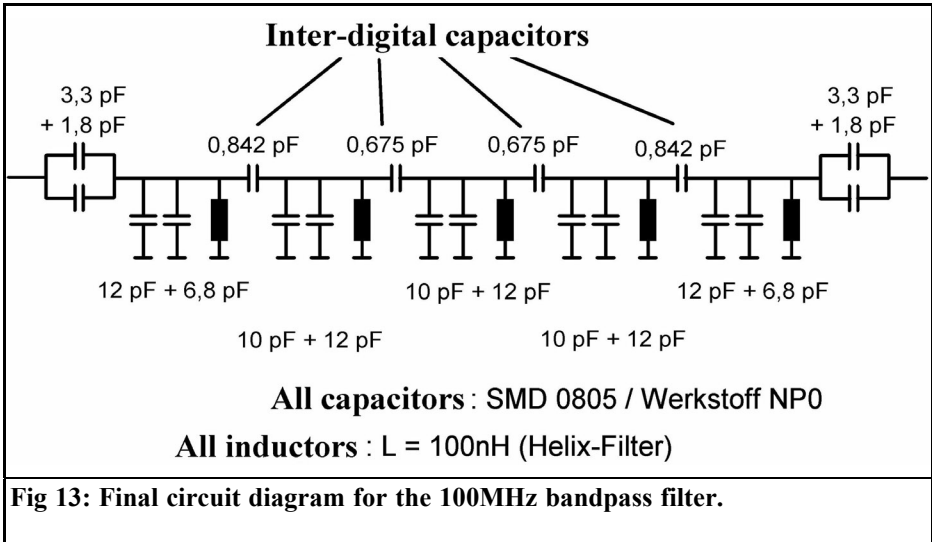


Fig 13: Final circuit diagram for the 100MHz bandpass filter.

in the capacitance of the continuous stripline (the circuit board data selected gives an inherent capacitance of approximately $0/5\text{pF}$ for a line length of 5mm), you come to the conclusion that the soldered component values are too high!. This is true, but when the core is fully rotated out the helical filters have a maximum inductance of 100nH , which can be reduced right down to 93nH when the core is rotated in (see above!). Thus it is preferable to add a little extra to the capacitance and tune the filter correctly. The following measurements indicate that these observations are correct.

One problem remains: after the filter is built how do you get at the inductor cores? One solution would be suitably positioned tapped holes in the housing, which are closed using grub screws. But a second method works better on this filter, with minimal stray fields. Use two small copper plates as carriers for the two SMA sockets at the input and output. The central conductors of the sockets are soldered to the associated microstrip lines and the plate squares are simply screwed to the circuit board. Fig. 15 shows this arrangement. Now the circuit board can be connected to the network analyser, measured and correctly tuned.

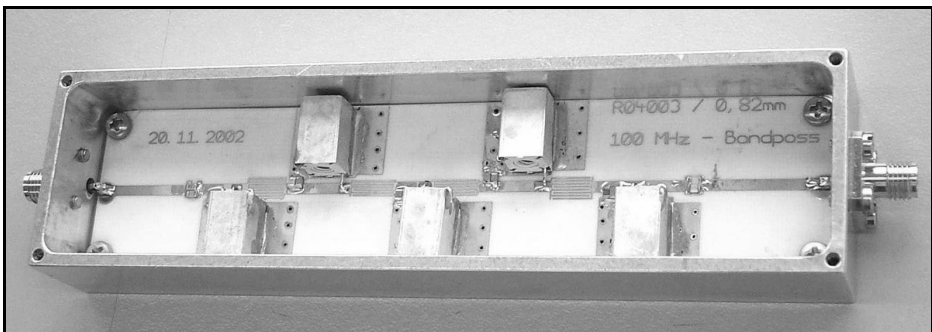


Fig 14: The completed 100MHz bandpass filter in its aluminium housing.

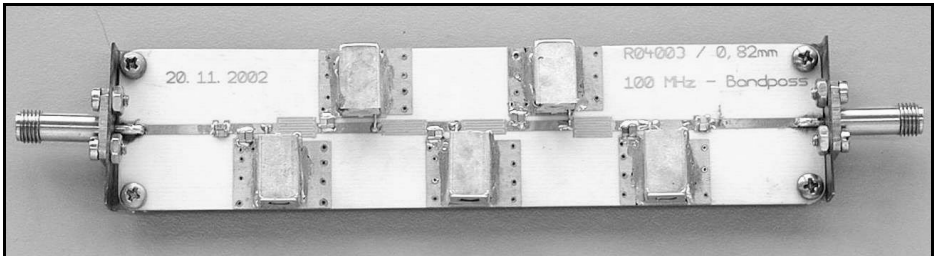


Fig 15: The bandpass filter with SMA connectors fitted to copper plates.

A subsequent measurement check following incorporation into the screening housing, with the cover screwed on, shows that the transmission curve is not noticeably altered as a result.

Incidentally: this method is also extremely useful for rapid, basic functional testing of all microwave circuits in the range up to 10GHz. When equipment is in a housing the first thing you are usually struggling against is the fact that the layout acts as a waveguide or a cavity resonator, with its natural resonances and other effects. It usually needs a bit of effort, e. g. using conducting foamed material glued to the inside of the cover or additional dividing walls, to get con-

trol of these “bad habits”.

7. Measurement results

Fig. 16 shows the transmission curve in the frequency range from 90 to 110MHz, so that we can make a direct comparison with the simulation as per Fig. 11. Naturally, there are difficulties with the rise in the transmission loss to 10dB (as we feared). If we carry out a few more specimen simulations with varying loss resistance values, then this gives us a series resistance of 0.88Ω for each coil,

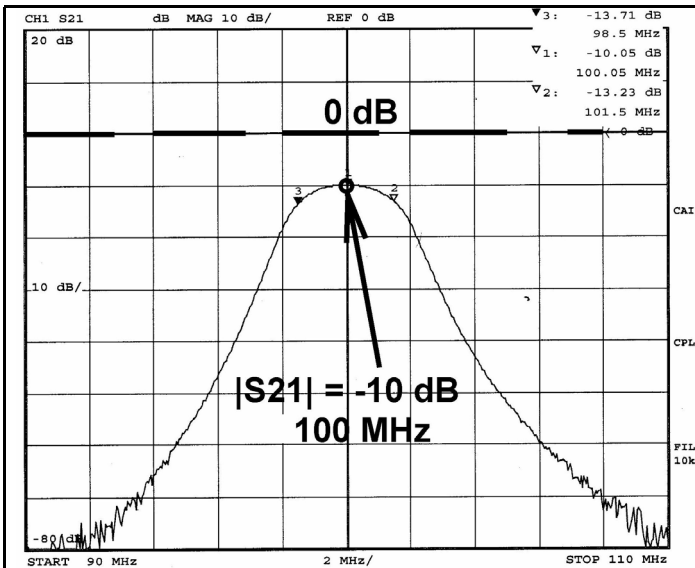


Fig 16: Transmission curve from 90 to 120MHz.

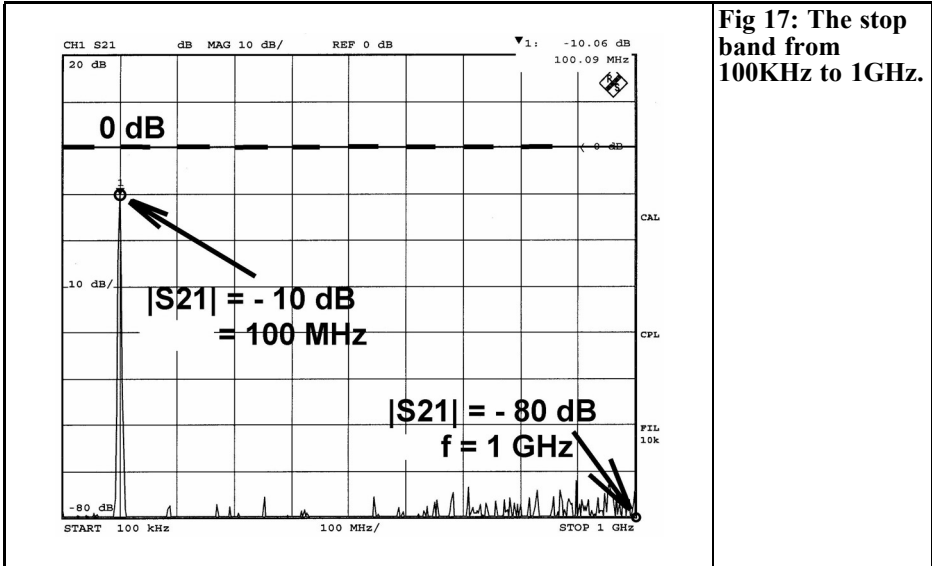


Fig 17: The stop band from 100kHz to 1GHz.

or $Q = 72$. But since not only the coils but also the capacitors play a part in these losses, it would be rewarding to do another experiment here using very expensive but high quality microwave ceramic capacitors (as incorporated used in the Satellite LNBs).

Finally, Fig. 17 shows the stop band in the range from 100kHz to 1GHz. In principle, no unpleasant surprises should be expected here as, through the use of several SMD capacitors in each resonant circuit, the inherent resonance is pushed sufficiently far upwards. To recap: the inherent inductance of these capacitors does not depend on the capacitance value but on the size of the housing, and for “0805” is about 0.5nH. If we use several capacitors in parallel the inductances is in parallel and there is a corresponding fall in the total inductance. So it is only above this inherent resonance that the filter attenuation collapses, because the resonant circuit no longer displays capacitive behaviour, being replaced by inductances. And unfortunately their reactive impedances then diminish as the frequency rises.

The reduction in the filter attenuation as

the frequency rises can be prevented. Since we are dealing with a direct coupling of the output to the input the effects of these capacitors own inherent inductance, which we have just been discussing, also have a role to play here, we must use three capacitors in each resonant circuit, instead of just two.

One other comment: as a precaution, the inside of the cover was laminated with a conducting foamed material 2mm thick, in order to suppress the expected housing resonances in the GHz range. However, the effect of all these measures for the range exceeding 1GHz was not investigated further.

8. Summary and outlook

The combination of modern design and simulation software, including undesirable “parasitics”, together with a correct circuit board layout and correct housing construction, means that nowadays even a DIYer of limited means can assemble high quality circuits with precisely pre-



```

FilDes--Filter Design System--Ver 1.20
Narrow BandPass Filter
what is the value for the inductors? (in nH)
72                               Resonating caps are:
Coupling caps are:              C(1) = 16.27 pF
C(1,2) = 0.461 pF              C(2) = 15.90 pF
C(2,3) = 0.369 pF              C(3) = 16.00 pF
C(3,4) = 0.369 pF              C(4) = 15.90 pF
C(4,5) = 0.461 pF              C(5) = 16.27 pF
The characteristic impedance of the filter is 2544.98 ohms
Match to a lower rs/rl? (Y/any key)
Default impedance is 50.00 ohms. Type 'D' for different src and load
or hit any key to match to default.
Add 3.108 pf to the source end, and 3.108 pf to the load end.
Then change C(1) to 13.226 pf and C(5) to 13.226 pf.
(S)ave to file (P)rint or (C)ont?
-

Type: NBP Poles: 5 Ripple: 0.30 Zsrc 50.00 Fo 145.00 MHZ
BW design (ripple or 3 dB): 3 dB

```

Fig 18: A design for a 145MHz filter using "fds".

dictable characteristics. Converting to other frequencies is not a problem either, a pocket calculator and version 2.1 of PUFF are all you need!

In conclusion, I'd like to make a wish come true for a radio ham friend of mine: During a technical discussion on the problems arising in relation to the development of the bandpass filter, he said: "Now, do something aimed directly at radio amateurs. I could make good use of a filter like that if I could use your converter idea, but with a 2m receiver instead of the car radio".

He's right, and so Fig. 18 shows just such a design. For this project, the circuit inductance has been reduced to 72nH (for the next smallest filter obtainable can be altered only between 76nH and 67nH) and the band width has been increased to 6MHz due to the higher mean frequency. The simulation in the range of 120 to 160MHz is shown in Fig. 19, where the work was carried out with $Q = 100$ (this shows the enormous values for the stop band at 120 or 160MHz!).

Complete development requires a little

more work because the microstrip line in the centre of the board shows up parallel capacitances for each resonant circuits. True, these are only percentage changes, but this influence unfortunately does slowly make itself felt.

A series of such line sections with various lengths can be listed in PUFF list F3 and they can be introduced between the individual filter components in accordance with the board layout. Then, of course, comes the laborious precision work to correct the circuit capacitors.

Due to the higher operating frequency, the direct capacitive coupling between the input and output sockets within the housing can now have a greater effect on the filter attenuation and can make it worse. So presumably dividing walls are required and it looks as if this could mean a new, independent project.

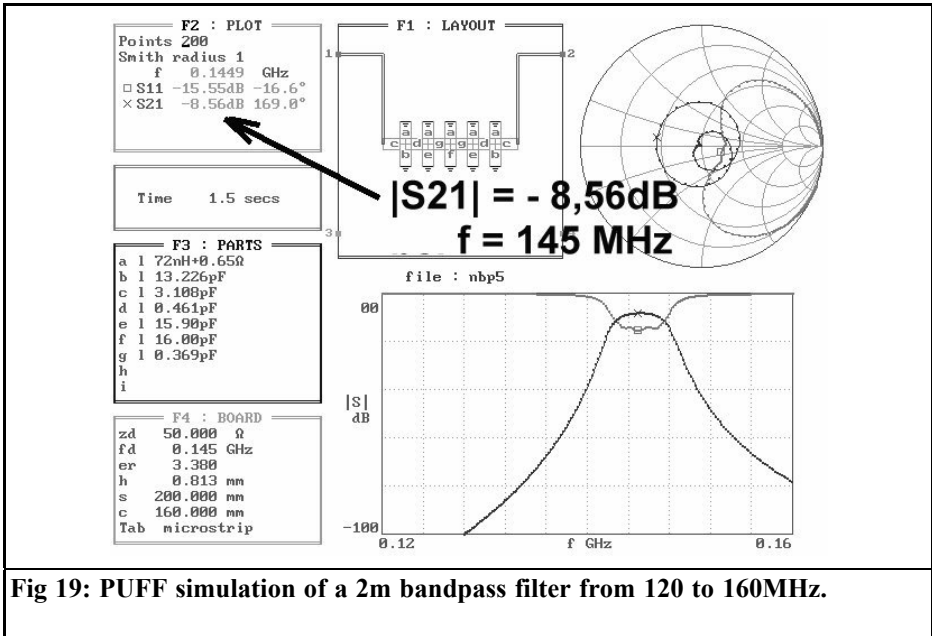


Fig 19: PUFF simulation of a 2m bandpass filter from 120 to 160MHz.

9.

A disadvantage which unfortunately could not be avoided

Everything is working excellently, the filter is doing its duty, the articles written and now it has to be sent off. What is there left to do? Well, something does come creeping up now from an unexpected quarter. In this case from Finland - the CD with the newest test version 7.80 of APLAC turned up on my desk.

So you run the update and, just for interest, test the improvements, using the new version on existing completed projects. And what does this lead to? True, you can certainly calculate the inter-digital capacitor again, with its dimensions and fractional capacities, but generating the S parameter file (and thus the sweep) is barred, the reason given being "Memory restricted in this version". Following several emails and some overheating among the specialist

brains working things out in Helsinki, this is the situation:

- The two tasks (i.e. S parameter sweep AND capacitor calculation) can function together only if you use version 7.51, which is indeed what the author has been using up to now.
- With the newer versions, only the capacitor calculations can be carried out since, due to the recently improved model, the simulation cost for sweeps has increased, and has therefore exceeded the limits of the student version.
- APLAC is rather embarrassed, and in such circumstances may well wish they'd been a bit less nit-picking in issuing cheap-rate "University and Education Licenses". So they are being very generous about issuing free 45 day test complete versions (and you can apply more than once....). But please contact APLAC direct with your queries!

The author can also suggest another solution, which is to carry on using the



previous version 7.51. His email address is: krausg@elektronikschule.de

10.

Literature references

[1] Puff beginners project: Low-noise amplifier for the 137MHz NOAA weather satellite range or for the 145MHz 2m amateur radio band. Part 3, Section 12.2.: Building an inter-digital capacitor by Gunthard Kraus. UKW Berichte 3/1999, Page 170 (This article did not appear in VHF Communications but can be translated if there is sufficient demand - Andy)

[2] Design and building of microwave

circuit .Part 10: Development of a 2GHz oscillator. Section 13.7.: Dimensioning of an inter-digital capacitor by Gunthard Kraus. UKW Bericte 2/1999, Page 117 (This article did not appear in VHF Communications but can be translated if there is sufficient demand - Andy)

[3] An Interesting Program APLAC by Gunthard Kraus. VHF Communications 2/2003, pp 90 - 105.

[4] Problem with running Puff on Window 98, Windows ME and Windows XP, and how to eliminate them by Gunthard Kraus. VHF Communications 4/2002, pp 229 - 231.

[5] Lets try again with PUFF by Gunthard Kraus. VHF Communications 3/2003, pp 170 - 180.

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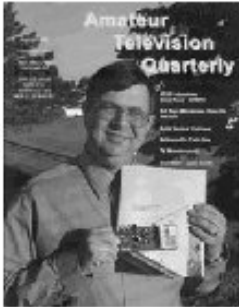

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Wolfgang Schneider, DJ8ES

1 to 65MHz DDS signal generator with a switchable output level

One extremely important aid for the active radio amateur is a signal generator. Whether you need a test signal to check the short wave receiver, or you want to re-calibrate the S meter or you'd like to measure the quartz filter you've assembled, the areas of application are numerous. Since a large part of DIY activities is involved with the short wave area, the signal generator described below is designed for the range of 1 to 65MHz. The signal level can be adjusted in 1dB steps between -117dBm and +10dBm and the frequency can be selected to 1Hz resolution.

1. General

In recent years, new DDS (Direct Digital Synthesiser) components have been continuously coming onto the market. Modern DDS components, with their simple external wiring supply sinusoidal output signals going far up into the VHF range. One possible area of application for this kind of frequency generator is, for example, as a signal generator. For general control of the DDS component, it makes sense to use a micro controller, which makes various control options possible, together with the displays showing the frequency and other parameters.

The system determines that the output signal shall be filtered through a high quality low pass filter. In addition to the desired frequency (the fundamental), a DDS also generates alias products such as the clock frequency + the fundamental and clock frequency fundamental. Also the clock frequency multiplied by two, three, etc. This also gives an endless number of mixed products. These can be removed using a fixed low pass filter at 30 to 40 % of the clock frequency.

You can find additional details about DDS ICs in [1] and [2].

The DDS signal generator with switchable output level, consists of the two function blocks:

- DDS with control and display
- Controlled amplifier with adjustable attenuator.

The controlled amplifier followed by an adjustable attenuator gives an output signal level range of -117dBm to +10dBm. The usable frequency range is 1 to 65MHz. All adjustments are carried out by the micro controller.

2. Description of circuit

The core of the circuit is the AD9851

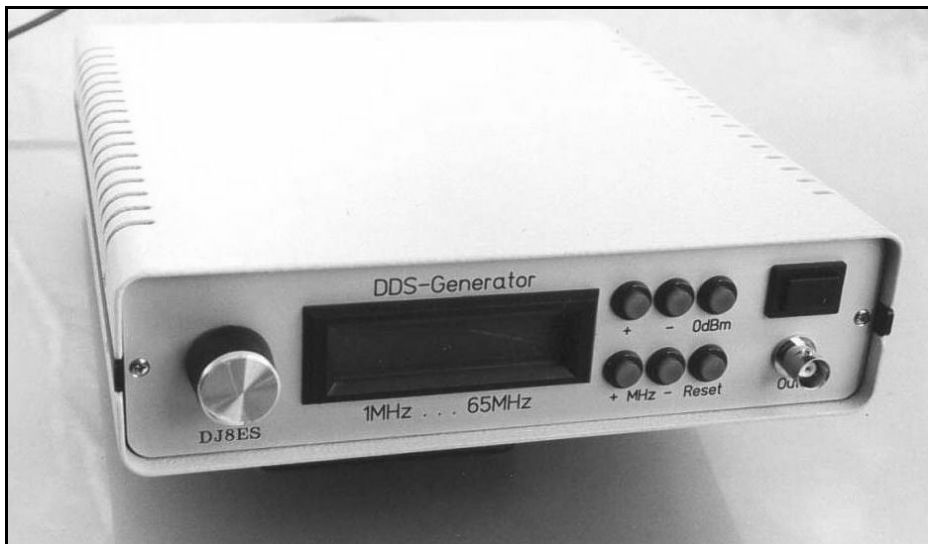


Fig 1: The completed 1 - 65MHz DDS signal generator with switched output level.

integrated DDS IC (IC2) from Analog Devices. It supplies a sinusoidal output voltage from 0Hz to approximately 33% of its clock frequency into 50 Ohms. Thus if we use a clock oscillator at 30MHz, and activate the sextupler in the DDS IC by means of the software, we obtain a clock frequency of 180MHz, which then gives a maximum usable frequency of approximately 65MHz.

The required ripple and harmonic suppression are obtained using a 17 pole Butterworth low pass filter. It is designed for 50 Ohms and the limiting frequency is approximately 65MHz. All values are brought back to standard values.

The amplifier described below using the transistor T1 (BFR96S) is based on tried and tested 50Ω broadband technology, increases the output signal by maximum of 23dB.

A small part of the output is rectified, using the Schottky diode D1 (BAT 45) for control purposes (Fig. 2). The DC voltage is compared, in the operational amplifier IC3a (LM358N), with the reference voltage obtained from the LED,

using the trimmer R12 (50 kΩ). The operational amplifier thus controls the output level of the DDS, using the FET stage T2 (2N7000) connected as an adjustable resistor, via the RSET input on IC2. The output from the broadband amplifier stage, set to +10dBm (10mW), does not vary by more than +/- 0.3dB over the entire frequency range. This interesting method for stabilising the power comes from Bernd Kaa (DG4RBF) and is described in detail in [1].

The entire signal generator is controlled by an AT90S8515 (IC1) micro controller, manufactured by ATMEL (Fig. 3). Its 40 pin housing contains enough ports to control the DDS module (IC2, AD9851) via a shaft encoder, the LC display (2 x 16 characters) and the switchable attenuator for setting the output. Specific functions can be directly selected by pressing one of the six control keys, connected through the ports PA0 to PA7.

The LC display is connected via the plug connection K8 and wired 1 to 1. The pin configuration of the plug connections for

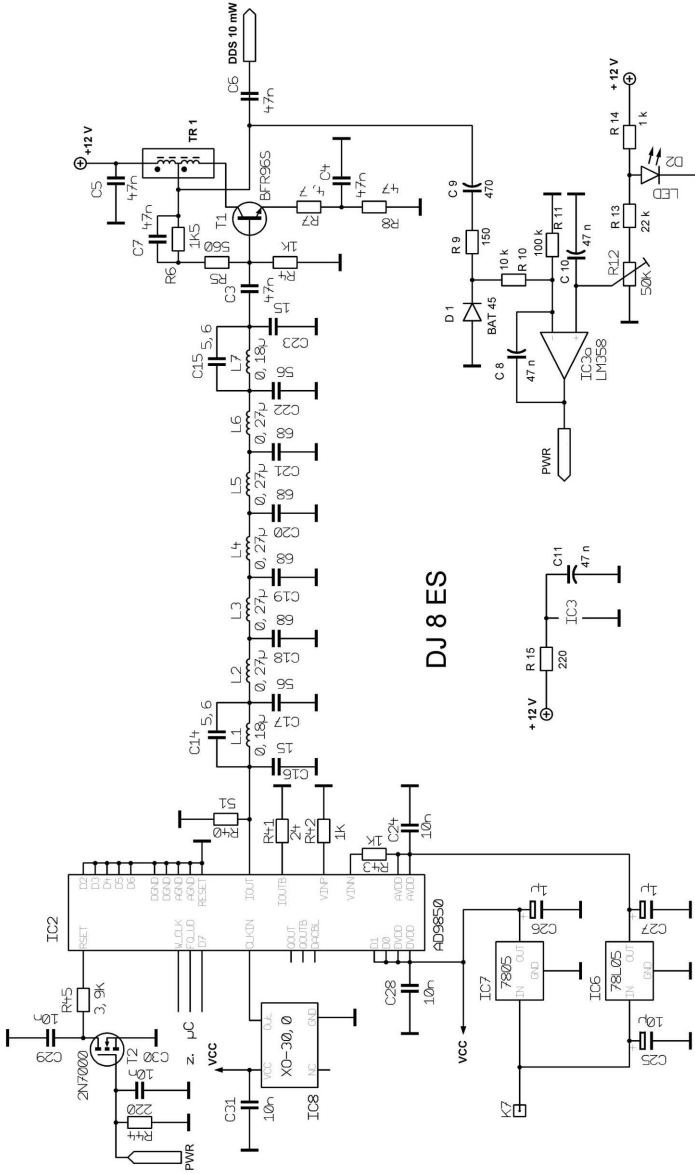


Fig 2: Circuit diagram of the signal source and low pass filter.

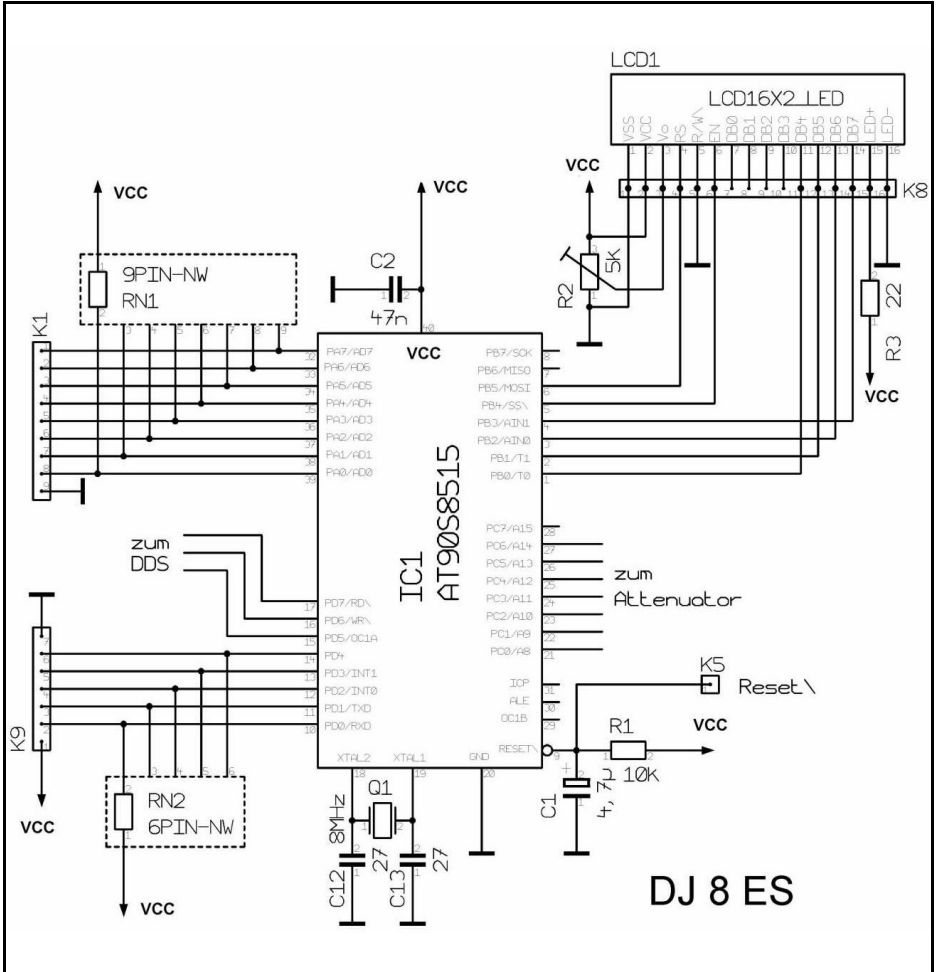


Fig 3: Circuit diagram of the micro controller.

control keys and shaft encoders is as follows:

K1 μC Function

- 1 PA7 not wired up
- 2 PA6 Frequency 1MHz
- 3 PA5 Frequency + 1MHz
- 4 PA4 Tuning step values
- 5 PA3 Attenuator + 1dB
- 6 PA2 Attenuator 1dB
- 7 PA1 Output level 0dBm
- 8 PA0 Output level + 10dBm

9 GND

K9 μC Function

- 1 +5V (Vcc)
- 2 PD0 not wired up
- 3 PD1 not wired up
- 4 PD2 not wired up
- 5 PD3 Shaft encoder channel B
- 6 PD4 Shaft encoder channel A
- 7 GND

As can be seen from the table, not all the

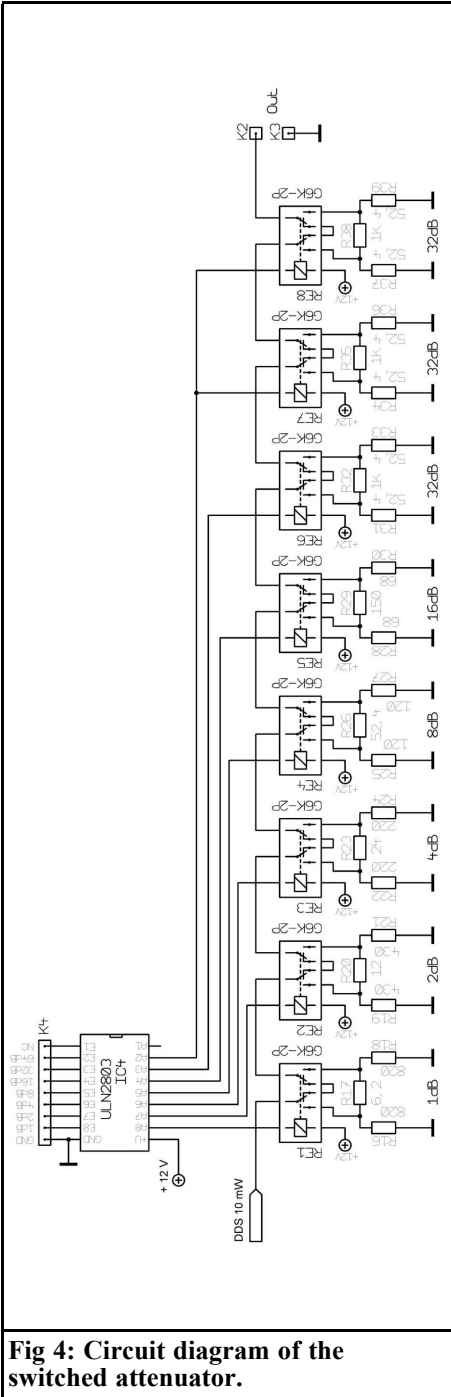


Fig 4: Circuit diagram of the switched attenuator.

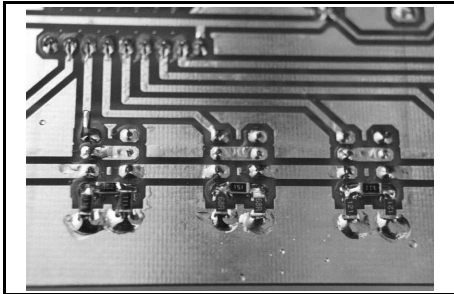


Fig 5: Detailed view of SMD components.

connection ports are used. There is still space left for future software versions and/or possible individual adjustments.

When the micro controller is switched on it is reset using “Power on Reset” or the “Reset” connection (low active) thus the signal generator is always set to 14.000MHz with an output level of 0dBm (1mW) as default values. The tuning step width is initially defined as 100Hz/step. This can be switched between 1Hz, 10Hz, 100Hz and 1kHz by pressing the relevant key.

The most effective method of assembling switchable attenuators is to arrange them in binary coded stages. The structure 1dB, 2dB, 4dB, 8dB, 16dB, allows all possible combinations to be created with a minimum of expenditure. The adjustable attenuator in the signal generator is constructed in this way. It has 8 stages thus there is an output adjustment range of +10dBm (10mW) through to 117dBm.

Special sub miniature RF relays, controlled by the micro controller, switch the individual PI attenuators. The resistors used are SMD format 1206 types. The detailed photo (Fig. 5) gives a clear view of a suitable assembly for RF.

The 64dB attenuator is in 2 x 32dB format. Note that:

- High values of attenuation can be only be obtained with precise values of resistances. The 32dB attenuator cannot be assembled using normal resistance values from the E24

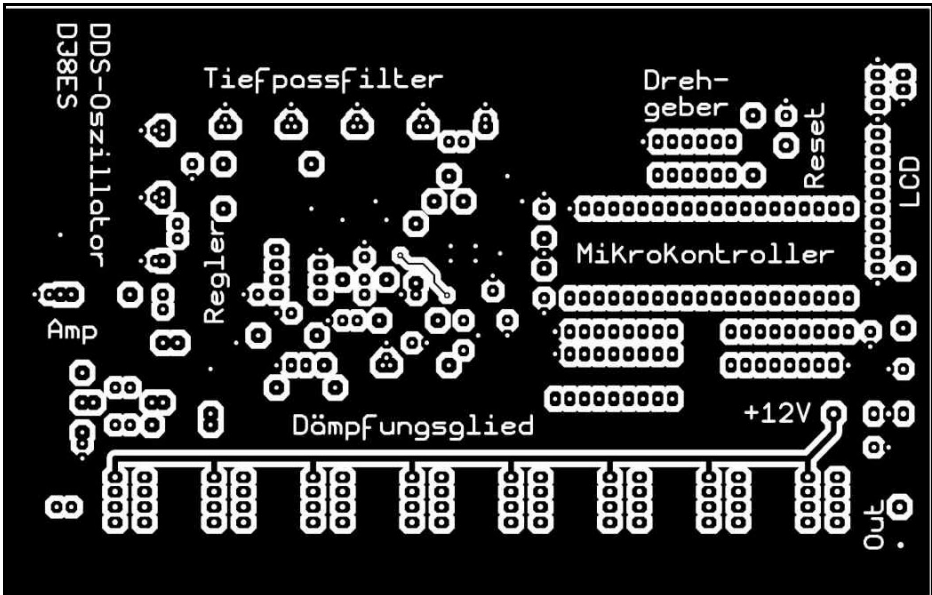


Fig 6: Top side of PCB layout.

range. The 52.4Ω resistance can be obtained fairly precisely from a parallel circuit involving 100Ω and 110Ω .

- We must also expect cross-talk.

The ULN2803 driver IC (IC4) is used to drive the relays so that the micro controller is not loaded with the switching current of the relay. The recovery diode normally used for relay coils is integrated in the driver IC.

3.

Assembly instructions

The complete signal generator circuit is mounted on a $100\text{mm} \times 160\text{mm}$ (Europa board) double sided copper coated printed circuit board. The assembly thus fits into a standard commercially available $102\text{mm} \times 162\text{mm} \times 30\text{mm}$ tinplate housing.

As the tracks in the area of the DDS IC

can sometimes be very delicate, and a lot of through plating is required, it is recommended that you use a professionally manufactured circuit board. These boards can be obtained from VHF Communications.

The components are not mounted on the DJ8ES-065 printed circuit board in any particular order, except that the AD9851 DDS IC (IC2) and the AT90S8515 micro controller (IC1) are fitted last. Before these two components are fitted, since they are SMD parts, we should first check the function of the two 5V fixed voltage regulators (IC6, IC7).

To make the LC display easier to read in background lighting, a 16pin multiple plug is used for the connections. The connections are 1:1 through connected. The external controls (shaft encoders and control keys for setting frequency and/or level) are connected through similar multiple plugs.

The equipment can not be put into the tinplate housing until the entire printed

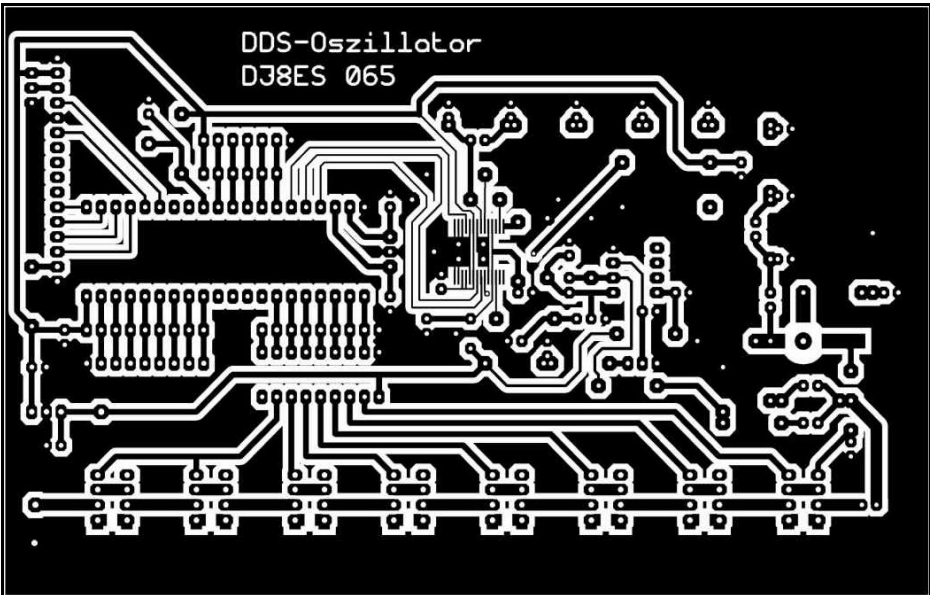


Fig 7: Bottom side of PCB layout.

circuit has been assembled. However, the SMA output socket and the 1nF feed through capacitor for the 12V power supply should be inserted into the sides of the housing first. For good RF practice the earth areas of the printed circuit board are soldered to the tinplate housing all around on both sides.

3.1. Parts list

IC1	AT90S8515, Micro controller		
IC2	AD9851, DDS component		
IC3	LM358 Operational amplifier		
IC4	ULN2803, Driver IC		
IC6	78L05, 5V, Voltage regulator		
IC7	7805, 5V, Voltage regulator		
IC8	TTL/CMOS Oscillator 30MHz		
T1	BFR96S, Transistor		
T2	2N7000, FET		
D1	Schottky diode BAT 45		
D2	LED, red		
Q1	Quartz 8.0MHz		
TR1	Coil 2 x 12 wdg. 0.2mm Cu enamelled wire on ferrite core FT 23-43, double wound		
C1	4.7µF, 16V, Tantalum		
			electrolytic capacitor
C25	10µF, 16V, Tantalum		electrolytic capacitor
C26, C27	1µF, 16V, Tantalum		electrolytic capacitor
L1, L7	0.18µH, Axial, RM 10		
L2 - L6	0.27µH, Axial, RM 10mm		
R2	Precision spindle trimmer 5kΩ, Format 64W		
R12	Precision spindle trimmer 50kΩ, Format 64W		
RN1	Resistor array 8 x 10kΩ, SIL 9-8 10k, RM 2.54mm		
RN2	Resistor array 5 x 10kΩ, SIL 6-5 10k, RM 2.54mm		
RE1-RE8	Sub-miniature relay 2 x Um, 12V, Type G6K-2P		
K1, K4	Contact socket board 9 pin, RM 2.54 mm		
K8	Contact socket board 16 pin, RM 2.54 mm		
K9	Contact socket board 7 pin, RM 2.54 mm		

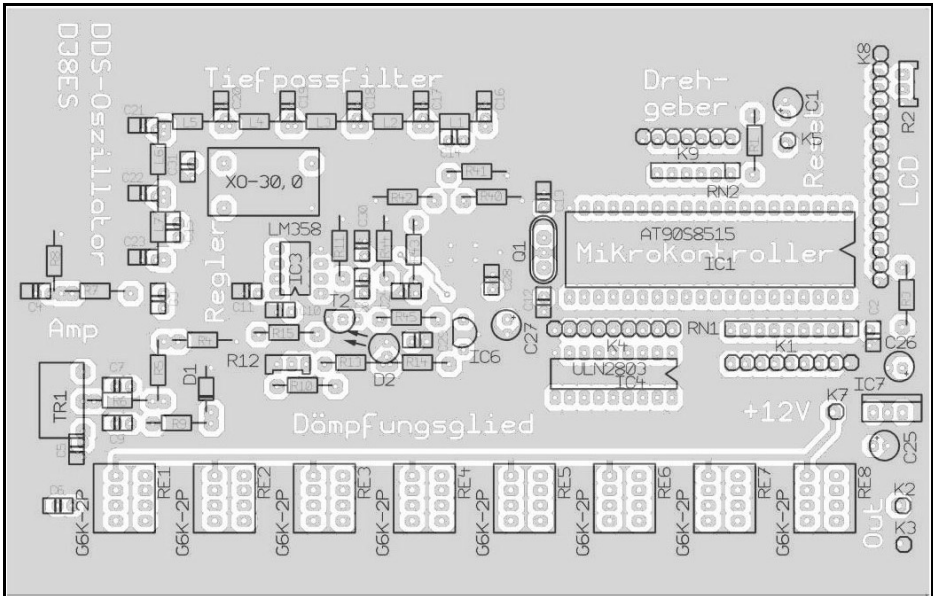


Fig 8: Component layout for top side of the PCB.

- K2, K3, K5, K7 Terminal pin 1mm
- 1 x Tinplate housing 102mm x 162mm x 30mm
- 1 x LC display 2 x 16 characters with back light
- 1 x Opto-electronic shaft encoder (32 pulses/revolution)
- 6 x Miniature pressure keys
- 1 x SMA flanged bush
- 1 x 1nF Feedthrough capacitor, solderable
- 1 x Precision IC base, 8 pin
- 1 x Precision IC base, 18 pin
- 1 x Precision IC base, 40 pin
- 1 x Printed circuit board DJ8ES-065

Resistors 1/4W, RM 10mm

- R7 4.7Ω
- R3 22Ω
- R41 24Ω
- R8 47Ω
- R40 51Ω
- R9 150Ω
- R15, R44 220Ω

- R5 560Ω
- R4, R14, R42-R43 1kΩ
- R6 1.5kΩ
- R45 3.9kΩ
- R1, R10 10kΩ
- R13 22kΩ
- R11 100kΩ

SMD resistors, format 1206:

- R17 6.2Ω
- R20 12Ω
- R23 24Ω
- R28, R30 68Ω
- R25, R27 120Ω
- R29 150Ω
- R22, R24 220Ω
- R19, R21 430Ω
- R16, R18 820Ω
- R32, R35, R38 1kΩ
- 7 x 52.4Ω (100Ω and 110Ω parallel)

Ceramic capacitors EGPU, RM 2.54mm:

- C12, C13 27pF

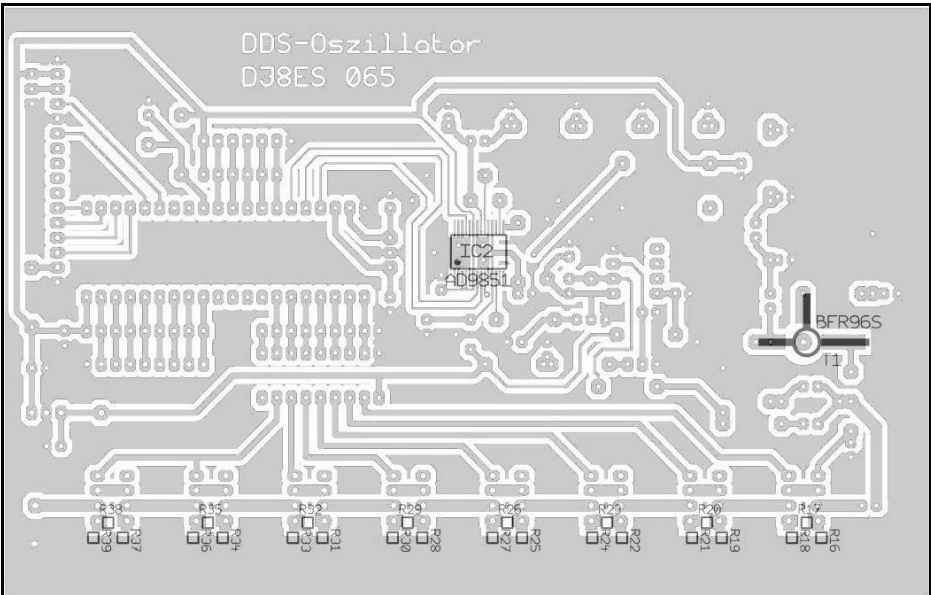


Fig 9: Component layout for bottom side of the PCB.

C14, C15	5.6pF
C16, C23	15pF
C17, C22	56pF
C18 - C21	68pF
C9	470pF
5 x	10nF
9 x	47nF

4. Putting into operation

When the circuit board is almost complete (excluding micro controller and DDS) it should be given a visual check. Provided no errors are found the supply voltage of +12 V should be applied for the first time.

Now check the voltages of both 5V fixed voltage regulators with a digital voltmeter. Only then are the two missing components mounted.

The LC display and the shaft encoder are connected, using suitable pin and socket

connectors, via the multiple plugs provided. A power meter and a frequency counter with a high-impedance input are connected to the RF output.

The current consumption for the entire circuit (including shaft encoder and illuminated display) should be approximately 300mA when the equipment is switched on. The signal generator should supply a frequency of 14,000,000MHz with an output level of 0dBm (1mW). The desired output is set precisely using the precision trimming potentiometer R12 (50kΩ). The variation range is approximately ±3dB.

Deviations of a few hertz in the output frequency result from the standard inaccuracy of the 30MHz clock oscillator. After a running-in phase of at least a week, its frequency can be measured precisely with a suitable frequency counter and programmed into the micro controller. In the prototype, the deviation was precisely -130Hz.

In practical operation, the signal genera-

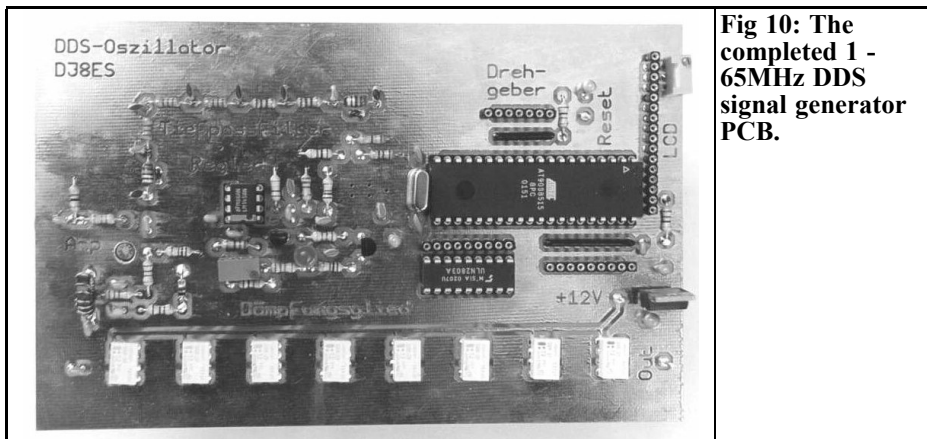


Fig 10: The completed 1 - 65MHz DDS signal generator PCB.

tor and DDS assembly's current consumption varies depending on the output level selected. An additional current consumption of approximately 10mA should be calculated for each relay used in the 8 stage attenuator. Thus a maximum of just on 400mA for +12V operating voltage is required.

Apart from setting the output level, no more calibration work is necessary. The signal generator can now be tuned in steps of 1Hz, 10Hz, 100Hz or 1kHz. The default setting is 100Hz. Operating keys for +1MHz or 1MHz are provided for rapid frequency changes.

The output level can be selected in 1dB steps between maximum +10dBm (10mW) and minimum -117dBm. The

default setting when the equipment is switched on is 0dBm (1mW). Here too operating keys are provided for the settings 0dBm and +10dBm, which are frequently required.

5. Literature references

- [1] Bernd Kaa, DG4RBF, 1 65MHz short-wave synthesiser with DDS and wobble function, UKE Berichte 4/99, pp. 205 - 222
- [2] Wolfgang Schneider, DJ8ES, Direct digital synthesis using AD9850 IC, UKW Berichte 2/98, pp. 67 - 75



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Eugen Berberich, DL8ZX

A new space saving omnidirectional antenna - the H O Loop

Well known horizontal polarised omnidirectional antennas such as the Halo, the corner reflector, and the Big Wheel, have considerably larger physical dimensions than the very compact H O Loop. This form of antenna, from M² INC. USA (M Quadrat), is efficient, easy to install, and space saving; it comes in 6m, 2m or 70cm versions. The antenna is described below using the 2m version as an example.

1. Antenna construction

The antenna consists of a dipole bent into a square (Fig. 1), the ends of which face each other. These two ends act as if they represented the shorting capacitor where a high RF voltage forms during transmission.



Fig 1: Photograph of the 2m H O Loop antenna from M²-Inc.

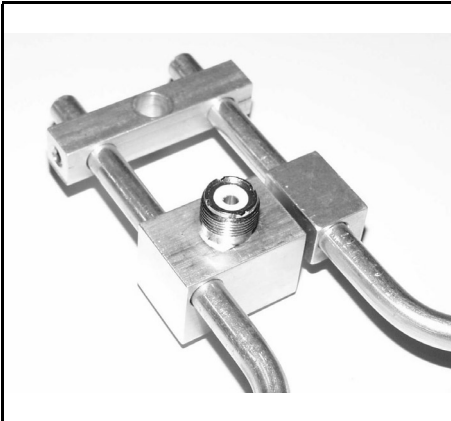


Fig 2: The weatherproof connection block.

The distance between these two dipole ends can be used for tuning to the desired frequency.

In the 2m version, the dipoles are shortened at both ends by 3mm, which increases the actual frequency by approximately 1MHz, depending on the installation instructions.

The antenna is connected using a coaxial socket, which is incorporated in a weatherproof milled aluminium block (Fig. 2). The interior of this terminal block is filled with a gel suitable for use with RF as a protection against moisture and corrosion.

The installation screws are made of stainless steel and take the form of Allen screws, which should be considered as a very positive factor!

2. Characteristics of H O Loop

2.1. Gain specification from manufacturer

The gain for the 2m model is specified at 4dBd (10°), 8dBd (9° Elevation) with a double stack.

The installation height above ground also enters into the gain calculation, because although ground absorption takes place, it largely disappears if stacking is introduced, due to the flatter radiation (Fig. 3 and 4).

If a 2m H O Loop is installed on a metallic surface (e.g. a car roof) at a height of 46cm, a low side upward echo beam is produced with a lower gain [2] in the horizontal plane.

This article also has examples relating to satellite operation for 70cm/2m.

Similar values are obtained for the gain of a 70cm H O Loop.

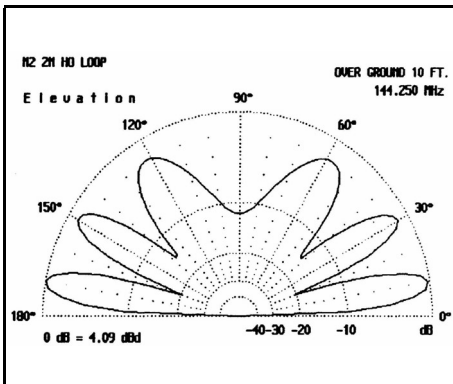


Fig 3: Polar diagram for one 2m H O Loop antenna.

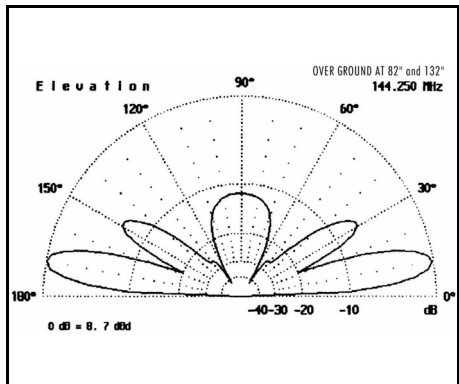


Fig 4: Polar diagram for two stacked 2m H O Loop antennas.

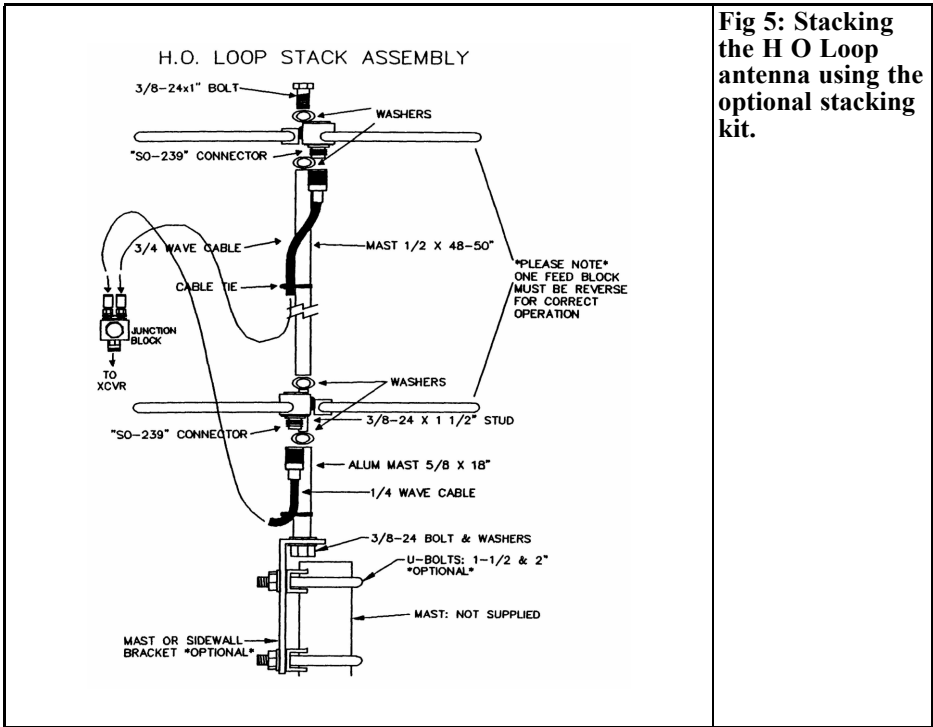


Fig 5: Stacking the H O Loop antenna using the optional stacking kit.

2.2. Antenna stacking

Stacking H O Loops is a very effective option for increasing the gain.

The stacking can be carried out either using the stacking kit supplied by the manufacturer or else using a standard matching harness (Fig. 5). The stacking intervals for 2m H O Loops should be as per the installation instructions, e.g. between 36" and 50" (91cm to 127cm).

2.3. Omnidirectional characteristics

In principle, what we are dealing with here is a horizontal omnidirectional mono-band antenna.

The omnidirectional characteristics are unfortunately not made clear by the manufacturer's documentation, but empirical investigations revealed an approximately round beam, with two sag points diagonally opposite each other of a few dB.

2.4. Rating

According to the manufacturer's specifications, the model for the 2m band is rated at 800Watts, or 1.5kW if 2 antennas are stacked.

The socket for the 2m model is a Teflon UHF socket (SO239 socket) (Fig. 2).

2.5. Siting antennas

If you have a site where it is not possible to use a rotator, you can use the H O Loop versions for 6m, 2m or 70cm to create a structure like a Christmas tree.

An antenna for the next highest frequency range can be installed between the stacking intervals of the antenna for one band (Fig. 6).

2.6. Installation options

The H O Loop has a central fixing point in the short circuit clip of the antenna; the 6m version has an additional support

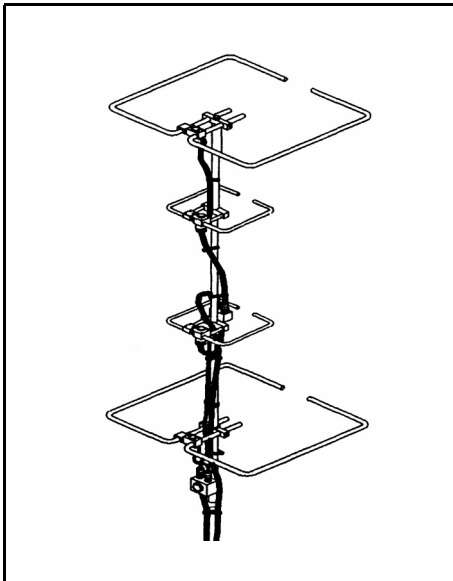


Fig 6: Stacking a 2m antenna with a 70cm antenna.

as well. The manufacturer provides for the antenna to be fastened to a thin rod (approximately 18mm). This rod can be mounted on a boom or at the top of the mast (Fig. 7).

If the antenna is not fastened using the central fastening hole but instead fastened directly to the mast using an angle clamp (Fig. 8), we can expect the fre-

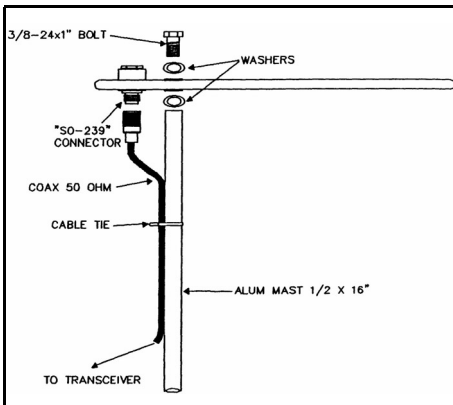


Fig 7: Mounting the 2m H O Loop antenna.

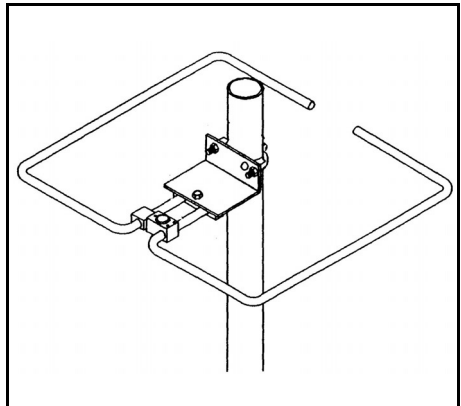


Fig 8: Mounting the H O Loop antenna on a standard mast.

quency of resonance to be displaced upwards by approximately 800kHz (in the 2m version), this should be taken into account during calibration.

A magnetic mounting can be obtained from the manufacturer for vehicle mounted operation (Figs. 9 and 10).

2.7. Effects of the weather

The DF0ANN beacon is successfully operating in the 2m band using this antenna, it is sheltered by a tiled roof. Humidity and/or rain and ice noticeably de-tune the antenna downwards. Fig. 11 shows the two curves for a dry and a wet

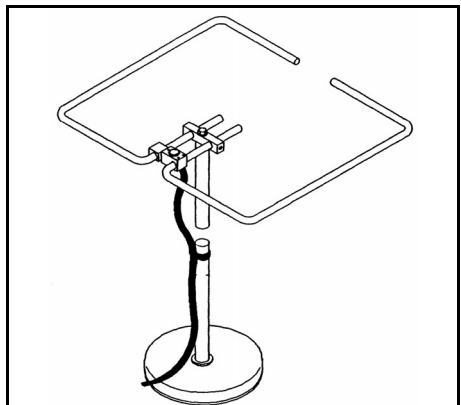


Fig 9: Mag mount for mobile use of the H O Loop antenna.

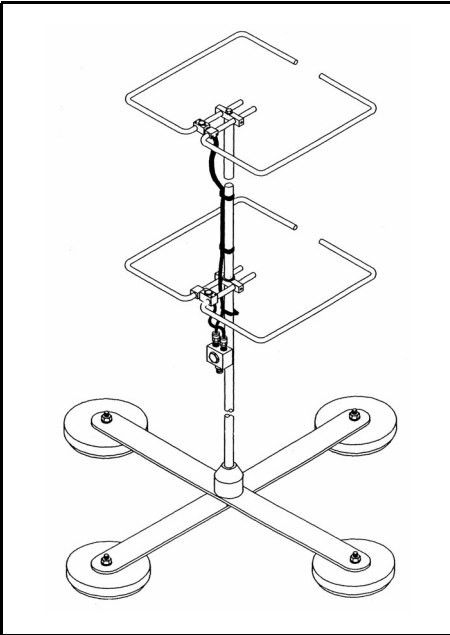


Fig 10: Heavy duty mag mount with 4 mount points.

antenna, plotted against the frequency.

A plastic hood suitable for use with RF, of the kind normally used for microwave ovens, can be placed over the 70cm version of the H O Loop, to keep snow

and rain off of the antenna. Unfortunately this hood is not very UV resistant, as it is a milky colour and is semi-transparent.

Black UV resistant plastics usually contain graphite and are thus RF attenuators.

3. Practical experiences

Operating the H O Loop is not complicated. The small size of this type of antenna can have many advantages!

It should not be forgotten that the antenna is relatively insensitive to interference (switching pulses) from the mains, and essentially absorbs fewer of these than other types of antenna.

The open dipole ends have proved to be a disadvantage because they ice up, and this eventually leads the ends to breaking open.

The two opposite ends should be sealed in a watertight manner with plugs made from a UV resistant material.

Connecting the two ends with a Teflon rod that fits into the open ends would

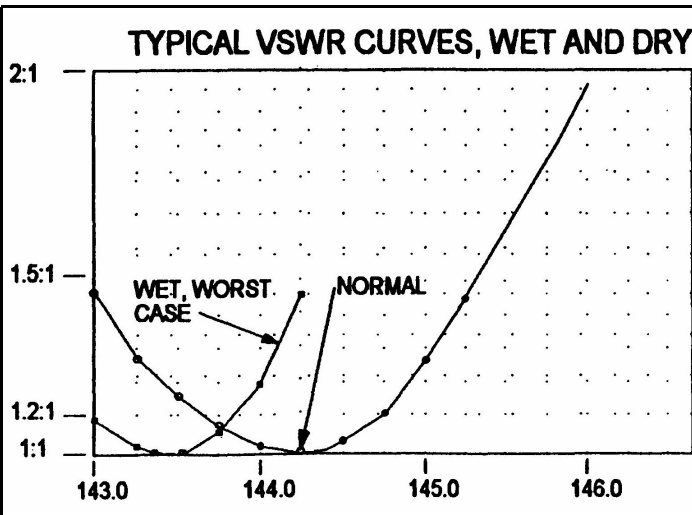


Fig 11: Performance curves for wet and dry antennas.

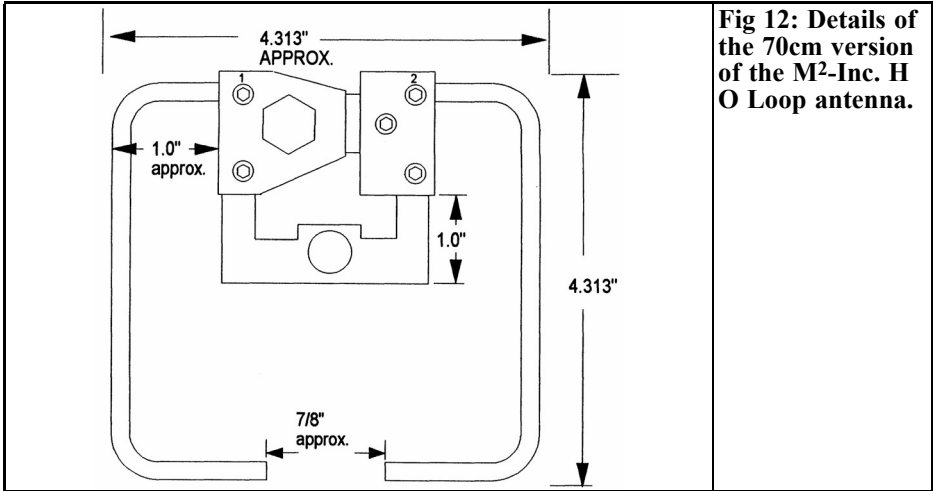


Fig 12: Details of the 70cm version of the M²-Inc. H O Loop antenna.

have the same effect, and would also improve the mechanical stability (problems with birds). You should expect more de-tuning of the actual frequency during wet weather though, since the Teflon rod retains more moisture.

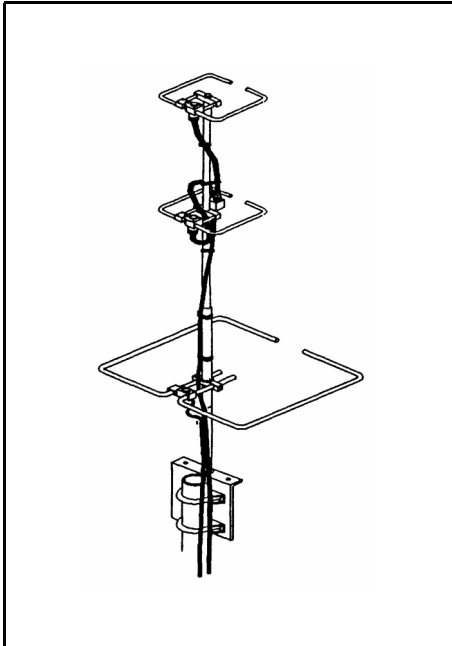


Fig 13: Stacked 70cm H O Loops above a 2m H O Loop antenna.

3.1 Calibration to actual frequency

Following calibration to the actual frequency, the antenna can be adjusted to give optimal matching, using a non contact plunger.

The band width, 500kHz in the 2m band, is relatively narrow (SWR < 1.2), but the range for a usable matching (SWR < 1.5) should be specified at 2MHz.

3.2 DIY

Unfortunately I've not found any instructions so far, but radio hams who enjoy experimenting can use the illustrations, which will suggest many possibilities for DIY work.

4. Literature

- [1] 2m H O Loop data sheet from manufacturer, M² INC. U.S.A www.M2inc.com
- [2] 2meter S Q Loop data sheet (forerunner of the H O Loop)
- [3] Article on H O Loop in CQ Ham Radio, Hicksville, NY, USA



Carl G. Lodström, KQ6AX & SM6MOM

The noble art of the Rheostat

1.0

Introduction

Rhe'os is Greek for "flow" and *stato*'s is Greek for "set" or "to set". In Swedish, flow is *ström*, part of my name. Maybe that is what inspired the article...

The word "Rheostat" is sometimes used a bit freely. Although I am sure most readers of VHF Communications knows exactly what it is, maybe some do not. Let us make sure:

- The Rheostat is a Resistor with one end adjustable, or it is a Potentiometer where one end is left unused (or connected to the wiper) see Fig 1.

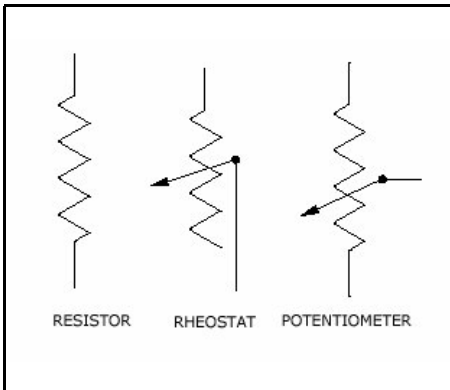


Fig 1: Resistor, Rheostat and Potentiometer.

At one time or another we have all burnt up one end of a potentiometer as we connected it as a Rheostat and tried to vary something drawing too much current! In Fig 2 we can see a Rheostat (one end of the winding is not available), compared to a "normal" size potentiometer, able of handling maybe 50W *over its full length*. Clearly, it has got a bit too much, while only part of it was in circuit. This illustrates the nature of the problem. The Rheostat, described in this article, dissipates *all* the power in the *entire* device *all* the time! Be it 1 μ A or 10A. The Rheostat in the picture has a total resistance of 24 Ω . Its Max current would then be 1.4A and it is clearly not of much use for currents less than some 0.1A or in excess of a few amps.

The problem is well known and obvious: A dial lamp for 6V 0.3A can be run from a 12V battery with a 20 Ω resistor in series. I may want to dim it as well, so I may pick a 100 Ω Potentiometer (connected as a Rheostat, not as a Potentiometer). When it is set for normal operation, 6V x 0.3A = 1.8W is developed in the lamp and the same in the 1/5th of the potentiometer in use (Fig 3). So, the entire potentiometer must be able to handle 5 times as much, or 10W. This certainly excludes any small Rheostat, maybe therefore the limited popularity of the variable resistor as a Rheostat. If the currents involved are yet higher, either the available resistance must be lesser or

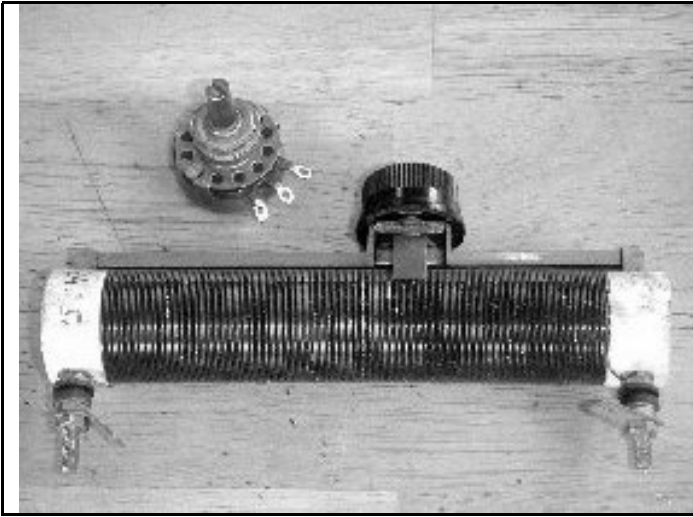


Fig 2:
Photograph of a
normal
potentiometer
and Rheostat.

the power rating must be higher. A Rheostat truly must be hand picked for the application and selected with some skill. Neither is it a true Rheostat. If the available voltage, or the load resistance, varies, so will the set current.

2.0

The true rheostat!

My solution to the problem can be set to control currents in the μA range, as well as in the tens of amperes by the turn of a

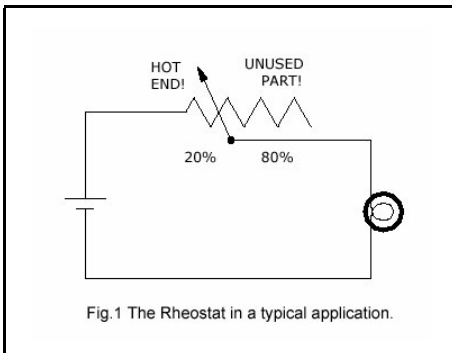


Fig 3: Controlling a dial lamp.

knob! This would normally call for a resistance range of some $1\text{M}\Omega$ to 0.1Ω (or even less), and of course many totally different devices! Perhaps as many as the seven magnitudes of current range.

The Rheostat presented can also handle hundreds of volts and it can dissipate hundreds of watts (with a sufficient heat sink of course). As such it is superior to all normal variable resistors! The control of this one is by means of a regular knob, on a regular potentiometer, easily turned. Have you ever tried to turn a husky rheostat, rated 100W, with your fingertips on a small knob?

2.1 Some Applications

One very useful application is to charge batteries with constant current (or to discharge them with the same to measure their capacity).

To charge some NiCd batteries with a constant 0.3A at up to 15V. Find a source of 15V, good for at least 0.3A. A wall plug adapter maybe. Connect the Rheostat in series and adjust for 0.3A (in short circuit or with the batteries in circuit)! There you are! Your batteries will be charged with 0.3A until the supplied voltage limits further increase.

You have a diode that you want to use

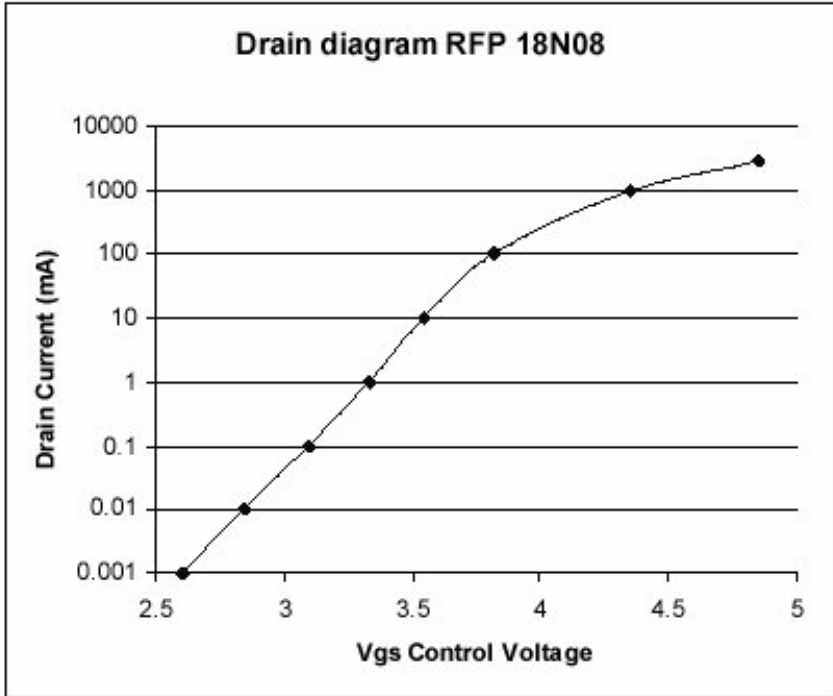


Fig 4: Characteristics of the MOS FET

for an outside thermometer by reading the forward voltage drop at a small current. You are not planning to include the conductor resistances in the measurement! The diode drop is about 0.65V but with a constant current it changes very linearly with temperature at the rate of some -2.1mV/K . Often a diode coupled transistor, with base and collector joined together, is used but it is not needed. A 1N4148 or 1N914 will do fine. Connect the Rheostat (turned down!) in series with the diode! Turn it up to 1mA (if that is what you want, the temperature constant is about the same anyway, but you do not want this current to heat up the diode) and read the voltage drop across the diode. As the diode forward voltage drop changes with temperature, the current remains constant.

2.2 Making the Rheostat

The heart in this Rheostat is a Power MOS FET. An inexpensive transistor that can handle tremendous power levels (with the proper heat sink!) and it is a resistor (with a diode in parallel) that is purely voltage controlled! It does not even consume any gate current, so the source of the variable gate voltage can well be a battery, and it will last as long as if you left it on the shelf! In my case it will probably last the rest of my life. So just solder it in!

My Junk Box provided a robust potentiometer marked "2.5 MEG" but I measured 3.3 M Ω . The Junk Box MOS FET happen to be a TO220 package, "RFP 18N08" by RCA. I have not even bothered to look up the data sheet. With 5V

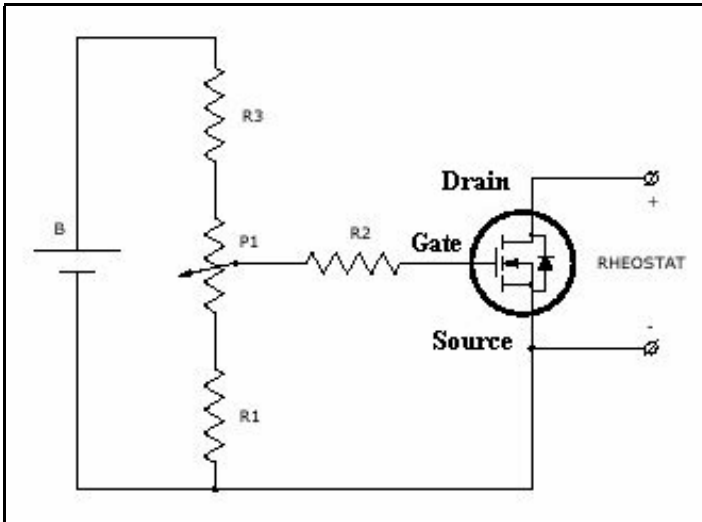


Fig 5: The circuit diagram of the Rheostat.

on the gate, the resistance was pretty much bottomed out at a few tenths of an ohm. This tells me two things: It is not a Very High Voltage device, or the resistance would have been higher. Neither is it a Very High Current device (50A or so) or the resistance would have been lower, below 0.1Ω . The part number probably means 18A and 80V and that is fine with me. It is all I need for a general purpose Rheostat. There are other MOS FETs in the Junk Box should I need more!

I started out this project by taking a look at the general character of the control properties of this transistor. Such a big one would normally not be considered in the μA range, but if it works there, why not make a note of it? The Graph in Fig 4 shows what Drain currents I got with a 5V on the Source and for various Gate-Source voltages. The current was practically constant, independent of the applied Source voltage, for all currents up to 1A or so, varying at most with 1% from less than 1V to over 20V.

Clearly, 2V is sufficient to turn this transistor completely off, and 5V enough to turn it fully on. Maybe more than you would want. This means that we want 2V at the bottom of the Potentiometer (if we

plan to control such small currents). Across the pot, P1 (Fig 5), that in this case is $3.3\text{M}\Omega$, we would like to see 3V. This makes $R1 = 2.2\text{M}\Omega$. Neither do we have 5V batteries. Maybe we have a 9V battery? Or a 6V? Suffice it to say that, in this case, we can add $1.1\text{M}\Omega$ per battery voltage above 5V on top of the potentiometer (R3). This helps keeping the drain on the battery to lowest possible current as well as making the “range of interest” equal the mechanical range of the potentiometer. You do not want all the action at one end of the dial!

We can see that for currents up to 100mA it adheres beautifully to a logarithmic function. Above there I suspect that the resistance in the Source leg (pin and bond wires) adds a voltage drop counteracting the applied control voltage on the Gate. At 1A we can see that it should have been about 4V, but it is 4.35V. A resistance of 0.35Ω would explain this, and this may very well be the actual leg resistance + bond wire. This value will of course be different for different transistors. By no means do I suggest that this particular transistor is better than any other, I just happened to have it laying around.

If you use the newer MOS FETs with

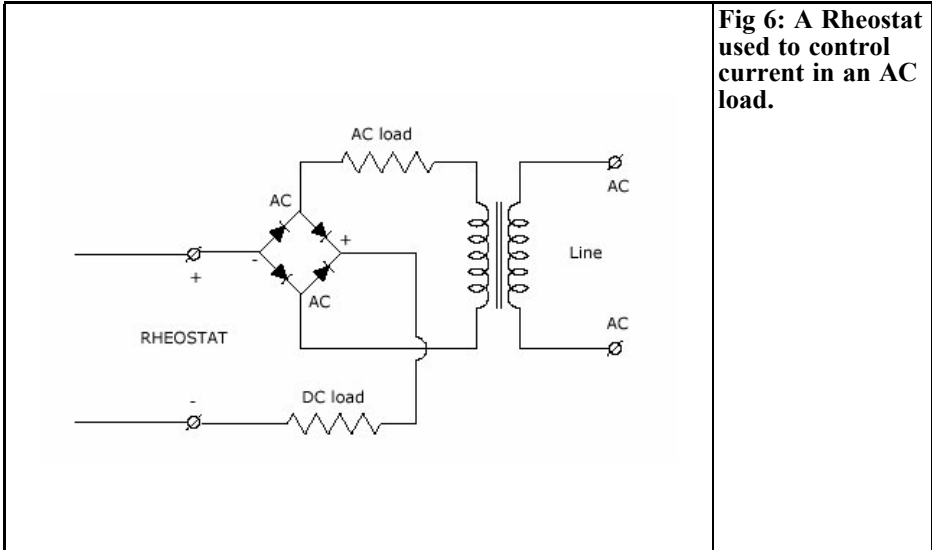


Fig 6: A Rheostat to control current in an AC load.

very low on resistances in the $10\text{m}\Omega$ range, remember that this is the resistance of a few inches of connecting wire! So, if you want the use of this exceptionally low resistance, you need appropriate connections and wires. A good Rheostat for $<1\Omega$ would be neither simple or inexpensive to make by conventional means!

The Rheostat described here has three unusual, and very useful, properties

- Power handling, it can handle the power that the transistor can handle, no matter at “what end” of the resistance it is! 80V and 0.1A is 8W, as is 8A and 1V. With conventional Rheostats you would need totally different ones for these two cases, here one single device will do!
- Constant current, few other devices will control your currents this beautifully! In this respect it is a true Rheostat; it controls a set current, as long as the load and supply allows, but in spite of their variations.
- The adjustment of the limiting current is logarithmic in nature. This makes it possible to set some μA of constant current with the same ease as a few mA, or a few A, using a

linear potentiometer for the gate voltage.

3.0

Special applications

For very small overhead voltages ($< 0.5\text{V}$ or so) it works in both directions! At some 0.6V a parasitic internal diode takes over in the reverse direction. This diode is a result of the way the transistor is made and we just have to live with it.

AC line regulation is actually possible! If you can accept the drop ($\sim 1.5\text{V}$) in a rectifier bridge. The transistor must be able to stand the voltage being turned off, i.e. it must be able to stand the full peak of the line voltage. For 120V this is 167V and for 230V it is 325V. The rectifier bridge must also be able to handle this! A 200V and a 400V (respective) bridge and FET should do it, but 600V, or even 800V, components cost very little more, and you are safer from spikes and transients. The control potentiometer, battery, resistors and possibly the heat sink, will now be dangerous to

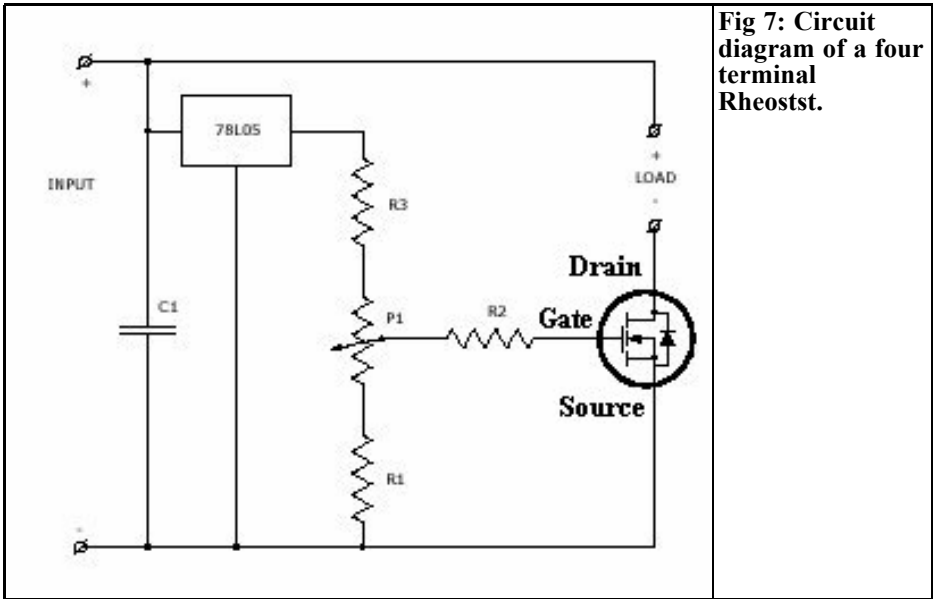


Fig 7: Circuit diagram of a four terminal Rheostat.

touch. If this is what you want to build, enclose it all in a metal box, use a safety ground, insulate the transistor from the heat sink with a mica washer and use only in grounded outlets.

In this application the transistor is connected with the Drain (usually also the tab on a TO220 or the body of a TO3 package) to the + corner of the rectifier bridge and Source as well as the - side of the battery to the - corner. The two AC corners are placed in series with the AC

source and the load (Fig 6).

Beware of the power in this case, it can get very high very quickly! You may have 1A in the blink of an eye and with 230V you also have 230W! Either your heat sink is of a very good size or you will not have a working Rheostat for very long! For high powers several MOS FETs of the same kind can be paralleled without matching. This will also distribute the heat more evenly over a heat sink.



Fig 8: Photograph of the four terminal Rheostat.

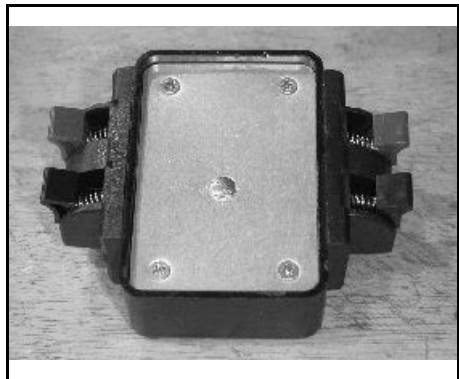


Fig 9: Photograph of the four terminal Rheostat.



3.1 RF concerns.

Since the transistor has gain, it can be made to oscillate. Probably not more than at a few MHz, if even that, but it would result in poor regulation. Since no current flows in the gate, a series resistor there will not matter. A few $k\Omega$ in series with the gate lead, located near the transistor body will ensure that no oscillation occur. Since the gate draws no current, this has no other drawback.

Another RF concern is the possible sensitivity to RFI from your transmitters. The Rheostat is not likely to be sensitive to it, but test! Especially for a case like using a diode for outdoor thermometer. The long conductors will act as an antenna and may pick up enough for the diode to rectify it, or for to carry it back to the MOS FET (which should not react, but one never knows!). Put an extra few 100Ω in series with each conductor at the Rheostat end, maybe a small capacitor across the diode, and twist the long wires lightly, minimising pickup. The extra resistors will not affect the voltage drop across the diode, as long as there is enough voltage available. If it is important that your equipment is insensitive to RF, as always, test it thoroughly before you rely on it!

3.2 A four terminal Rheostat

A four terminal Rheostat can be built, which may be more practical for some applications (Figs 7,8 & 9). I found the components at Radio Shack. Loud-speaker terminals are good for universal use. Here I visualised the availability of a power supply (of more than 6.5V) so one would not need the battery, although one can keep that part of the design. The bottom plate is also the heat sink for the MOS FET so it can only handle a few watts.

In this one I used a 5V regulator, LM78L05 for the gate voltage. It needs some 6.5V in for regulation, and can handle some 32V. If 5V is not enough

for your particular FET, use the LML08 or 12V regulators! Do not omit C1, 10 - 100nF or so, because the regulator may then well become an oscillator at a few MHz! Not good for the performance!

4.0

Conclusion

You will soon find a lot of applications for this Rheostat, especially if your power supply does not have a current limiter.

NiCd battery charging can be done with a transformer, a rectifier and the Rheostat as well! Assuming it is a full transformer (with a low voltage output), not an auto transformer, then safety is taken care of. If it has a 12V secondary, the peak voltage is 17V, and for a light load, such at the end of the charging cycle, it may well be 5-10% more, 18V maybe. The rectifier bridge will take $2 \times 0.7V$, so 16.6V is left, more than the 15V we wanted, so watch out! The current is constant so the batteries will be charged, but you should have a voltmeter on them and remove them when the correct voltage is reached. Naturally, the current will be regulated but only when the rectified voltage in is high enough! As the batteries charge, the conducting fraction of the period time will be less and less. So the average current becomes lower, possibly only flowing 30% of the time, and thus it may appear as less than the set current on the average, taking take more time than you thought to charge the battery. An easy way out of this, if it is important, is to put an electrolytic capacitor across the DC corners of the bridge. In our example we will now have an almost constant 18V and the the set current will flow 100% of the time! Even more reason to keep an eye on the voltage, or the batteries will be destroyed by overcharging.

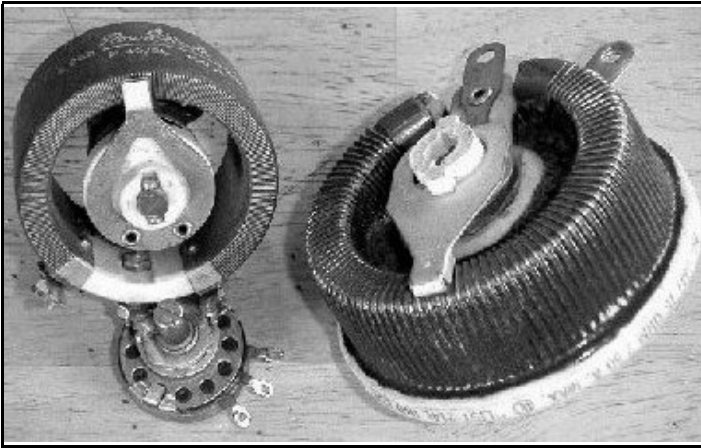


Fig 10: Examples of Potentiometers and Rheostats.

I have noticed one thing with the Rheostat that is a potential problem for small stuff. When the Rheostat "turned down", and one applies voltage to the circuit, there is a spike of higher current for a ms or so until it regulates down the set current. I assume this is due to the D-G capacitance, but have not delved into it. So, for small, sensitive loads, make sure the power source cannot deliver transients! Somebody may be sitting and playing around with crystal diodes or very small light bulbs.... It works fine if the power is already applied to the Rheostat to begin with, and it is turned it up.

Real Potentiometers, or Rheostats can be very large, Fig 10 shows "normal" size Potentiometers. The Rheostat described

in this article (Figs 11 &12) can handle as much power, or more, than the larger one pictured.

This article has described a profoundly simple and versatile Rheostat with excellent performance and a low component count/cost. The power handling sets it apart as it can handle what the MOS FET and its heat sink allows over the entire range. There is no "last few turns" of resistance body in this Rheostat! The circuit makes the Rheostat a practical possibility and a rugged universal component capable of *constant power dissipation regardless of resistance setting*. There are a lot of uses for it in the well equipped Shack!

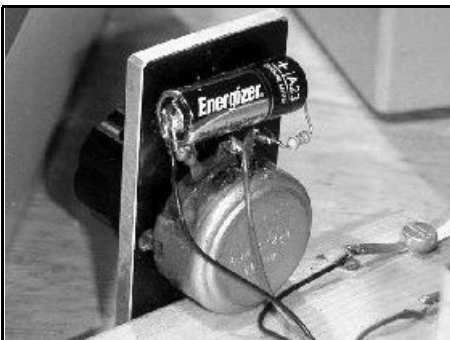


Fig 11: The author's Rheostat.

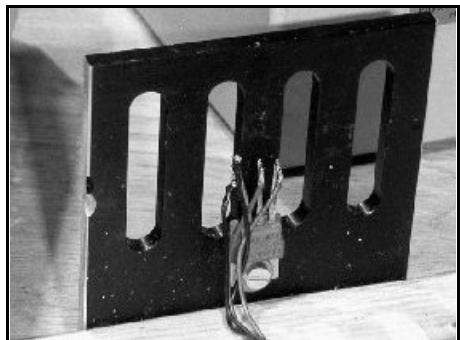


Fig 12: The author's Rheostat.



Gunthard Kraus, DG8GB

Funny electronic stories

(**Translators note:** The German story contained some dialogue in the Swabian dialect, so I have exercised some poetic licence and set the scene in Birmingham, with the Swabian character becoming a Brummie.)

The following story is the result of a friendly agreement between the author and the editor of UKW Berichte on the subject of the “electronic essay”. The question was whether there was room in a technical periodical for anything other than just technology in articles kept suitably technical. So what follows is a sample, an incident from the everyday life of a man with a passion for electronics.

Today:

The treasure trove of electronic scrap

It is a well known fact that we all have many unalterable characteristics. However, even today little is known, and little scientific research has been done, into the characteristics and genetic traits of the incorrigible tinkerer, who never walks past a scrap yard or a flea market without investigating.

In the electronics field, both the elec-

tronic scrap recycling companies and the miscellaneous amateur flea markets exert a magnetic attraction which no “real enthusiast” can withstand.

According to my wife, Im not much better. I neednt, and indeed cant seriously deny it! It was no accident that my students invented a new nickname for me (DIY Danny) a name my family somehow found out about (I still have no idea how).

But to my tale:

Now and then you have to pause awhile and leave off the work, work, work, to give ones faithful spouse an unexpected treat. In our family this can mean, for example, a day out to the big city, maybe 100km away. For one thing, its my wifes home town, and for another its where I completed my studies and then worked for a few years afterwards (and yes, youre right, its where our lifelong partnership began). When we get there, we practice a new version of the marital break-up: I leave my wife in the pedestrian precinct for a couple of hours so that she can have a good look round the shops without being hampered by an impatient husband. True, this is frequently expensive, but in these two hours Im free to have a thorough rummage around the 10 to 20 cubic metres of



circuit boards available in the electronic scrap recycling centre operated by a former fellow student of mine. There I am free to do what I like, and I always find something, since this recycling centre takes in supplies from a radius of at least 150km.

When the two hours are up, I link up with my spouse again and we spend the rest of the day doing things together. This is a tried and tested system. Each of us is usually extremely satisfied with his or her finds, and I can certainly recommend this as a risk free technique to others.

But there was one occasion..... It was sometime last year, and the weather was very warm. It hadn't rained for a long time and so for once the scrap yard didn't look like the normal bog. So there I was rooting around in a box, when I noticed something was a bit out of place. Whole areas of the ground were gleaming here and there under the circuit boards! It definitely wasn't water, and I went to try and spin out the suspense. After about one hour of digging, I was the proud possessor of about 3kg of small, modern 10MHz crystals, which had just been chucked onto the big heap. Let's draw a veil over the foolish antics involved in reaching this position, since I had to:

- shift vast quantities of scrap circuit boards
- use my fingers as if I was panning for gold and plough through all the dust, and then
- find a suitable container so that I could carry the stuff away.

The final volume filled a 2 litre gherkin jar, I know this because I ended up using the said jar as a means of transport. The first problem, of course, was the proprietor, my old fellow student, who asked me (as he was entitled to do): "What are you doing with that then?" (actually he asked me in broad Brummie, so it sounded something like: "Warra yo doin wi

that?"). I mumbled some reply which might have been: "Dunno really" or "I've got a few ideas" or even: "I know lots of people who're looking for this kind of thing". The scrap price to be paid was minimal, about what you'd pay if you ordered two new crystals like this, VAT and postage included.

What might appear as the next problem, the wife, was not a problem at all since the first thing you do in this kind of situation is to wash your hands somewhere and the next is to be absolutely on time for the rendezvous. So her question: "Well, did you find anything nice or useful?" could be answered in the affirmative without any problems, and with great satisfaction.

The scene now changes to the domestic home, a few hours later. What real DIY enthusiast could resist the attractions of a gherkin jar like this? Not me! Tragically, this was where I forgot to pay attention, and suffered from an excess of enthusiasm. For this led to a practical confirmation of something from my old physics textbook:

"The pain threshold for the human ear lies at approximately 110 dBa".

Would you like to know the details? OK just carelessly use the sacrosanct, extremely new pasta strainer to wash about 3 kilograms of filthy crystals in the sacrosanct stainless steel rinsing sink in the sacrosanct kitchen....

Fortunately, even in Nature every thunderstorm and every hailstorm comes to an end and you're free to try out all the possibilities of your new toys: from the quartz oscillator through the ladder quartz filter to the multiplier, you can simulate and assemble the lot. But eventually, there are no more games to play - and even your friends tend to start when you happen to meet them, afraid you're going to ask them *again* whether they need any 10MHz crystals. The gherkin jar and its contents have a melancholy look, but something must be done with



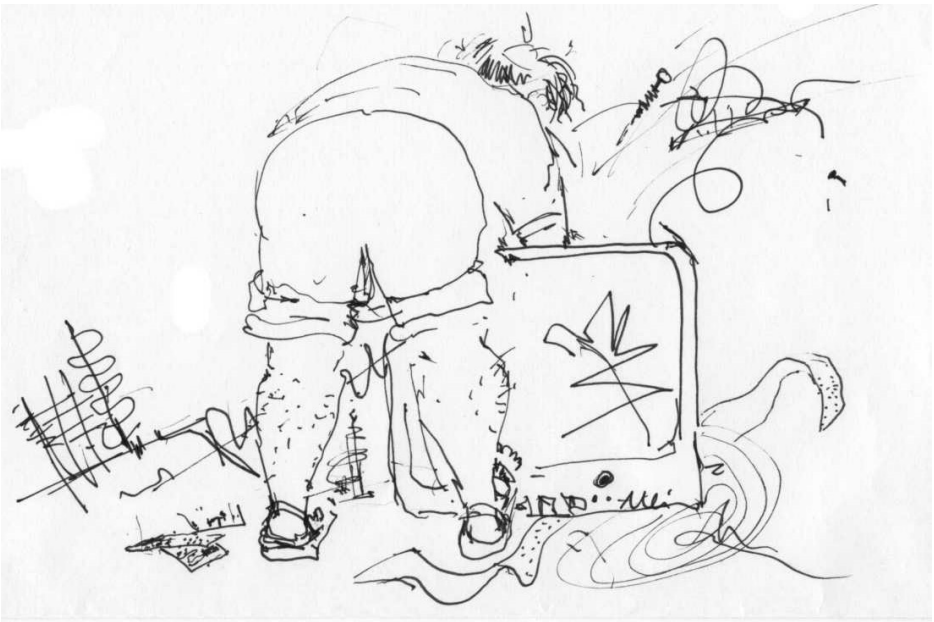
them. When everyone has been provided with 10MHz crystals, Ill perhaps solder up most of them in a lead coffin and bury them somewhere as hidden treasure for our descendants to find. Perhaps in 200 years time they wont know what crystals are any more, and that will give the archaeologists something to puzzle their heads about.

The question remains, dare one take the risk of going digging for gold in the recycling centre again and of finding new treasures? Im reminded of another experience which serves as a terrible warning.

It was high Summer last year, and I was scrabbling around on the circuit board mountain 2 metres high (wearing birchwood sandals and shorts), when the beastly thing began to slide. Naturally, I lost my sandals in the fall (that wasnt a problem, I found them again!). But cir-

cuit boards fitted with components look like beds of nails for Indian fakirs when seen from below and I had to clamber over them to save myself. Dont ask me how I explained the sudden and inexplicable disappearance of my socks to my wife (by the end, they were ripped to pieces and smeared with blood), for I dont remember clearly what I said myself any more. At all events, since then, to be on the safe side, Ive resisted the gold fever for quite some time, and confined myself to pottering about with components Ive already got handy at home.

But now the itch is slowly getting stronger again. For who knows what you might find in the circuit board boxes this very day, right now?. And in the end thats a feeling that could wake a real DIYer even if he was hibernating for the Winter!





Gunthard Kraus, DG8GB

Internet Treasure Trove

Theory of Antennas

That's the heading above the lists of links and downloads on this homepage. But we're dealing with tutorials on the basic principles here, the list is never ending, and it's impossible to itemise everything there is here on this subject. So just have fun rummaging around!

Address:

<http://members.fortunecity.com/xelbef/theory-antennas.htm>

Radiometrix

A company which specialises, in particular, on data transmission and modems, including those operating over radio routes. But there are also some neat application notes for the RF freak, quite apart from the many data sheets and other information regarding interfaces.

Address:

<http://www.radiometrix.com/>

TX Lines, Antenna Systems, PC Joystick

talking about antennas, but a homepage with such good tutorials and with so many different applications deserves a mention.

Address:

<http://www.tmeg.com/tutorials/tutorials.htm>

Radio Waves

The site's name gives you a pretty good clue: it deals with developments and products relating to electromagnetic waves. The main theme is antenna development and the production of complete antenna arrays. And because in this kind of work you have to pay very close attention to the customer's wishes, there's also an extensive planning section. One splendidly attractive gift from these people to all RF freaks is the "RW calculator" for calculating and planning every last little thing relating to antennas and receivers, which you can download free of charge. This is really something great, with many different applications of all kinds!

Address:

http://www.radiowavesinc.com/html/414_rf_system_planning_tools.cfm

We don't want to spend all our time



Longwave links

Just for a change, here's something in German, and something for the community of longwave fans. This homepage is a must for anyone wanting to know more about this field and this technology.

Address: <http://home.t-online.de/home/h-hcuno/afu/longlink.htm>

Amateur Television and high speed data signalling

Even though this homepage is hosted from France, you don't need any knowledge of French, as there is an English version. We are certainly grateful to the owner for the opportunity to download information on well documented DIY projects, going up to 2.3GHz, which will gladden the heart of any do-it-yourselfer.

Address:
http://perso.wanadoo.fr/jf.fourcadier/accueil/accueil_e.htm

G4FGQ's Software Site

An interesting page, put together in a way that's just right for the amateur and for anyone who likes fiddling about with RF projects. You can spend a lot of time here searching, investigating and downloading. Most of this site is about programs, all "home made" and all similar in appearance. So you know at once what it's all about, and you don't have to spend a long time in experimental operations. Very rewarding and highly recommended!

Address:
<http://www.btinternet.com/~g4fgq.regp/#S102>

W2BRI's Magnetic Loop Site and PA3CQR magnetic loop antenna page

As always on such sites, the more you nose around and the more stones you turn over, the more treasures you unearth. Very enjoyable and informative to see everything concerning loop antennas summarised here.

Addresses:
<http://www.standpipe.com/w2bri/index.htm>
<http://www.iri.tudelft.nl/~geurink/magn-loop.htm>

Alphabetically arranged radio technology links

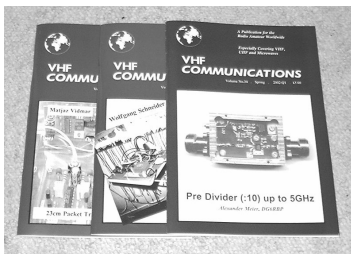
A site like this makes it considerably easier and more comfortable to search for a program, an article or a homepage to solve a problem.

Address: <http://home.wanadoo.nl/~a.schipper/fuurlinks.htm>

Myer's Engineering International

23 years of experience with radio communication systems and antennas... that's what it says in the introduction to this site. Another homepage which can reveal masses of new or fascinating facts for those with an interest in such matters. Seek, and ye shall find...

Address:
<http://www.myerseng.com/index.html>



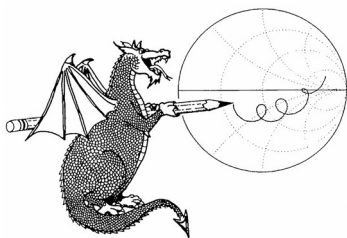
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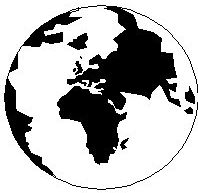
PUFF Version 2.1 Microwave CAD Software

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



VHF Communications Web Site www.vhfcomm.co.uk

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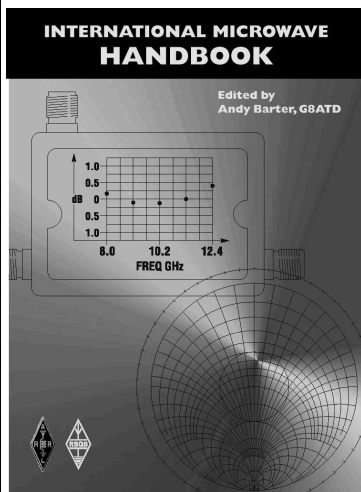
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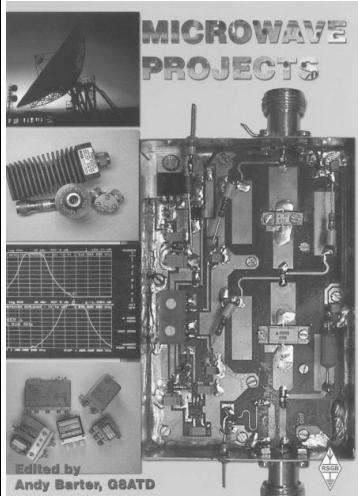
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