



*A Publication for the
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*Especially Covering VHF,
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VHF COMMUNICATIONS

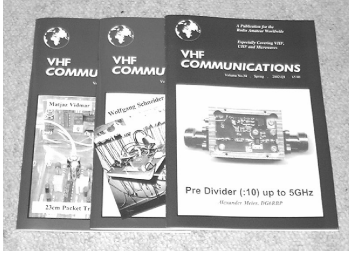
Volume No.37 . Spring . 2005-Q1 . £5.20



**Centre frequency counter for HP 8565A
and HP 8569 spectrum analysers**

Bernd Kaa, DG4RBF

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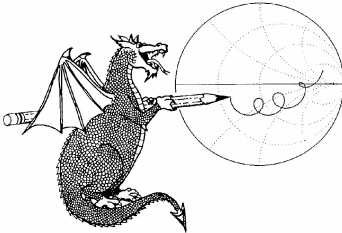
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Thank you to everyone who sent their comments about a CD version of the magazine. There was an overwhelming majority in favour of keeping the printed magazine. There were quite a few subscribers who wanted archive versions of articles on CD. I have been doing some research into the history of the magazine to give a talk at my local radio club and have found out that reprints of certain topic themes were available in the 1970s, so I will investigate this idea.

I have had some complaints about the readability of some circuit diagrams. The problem is that they are reduced in size to fit the A5 page size. I have tried some enhancements to the circuit diagrams in this issue and hope that they are improved. If for any reason you have a problem with the magazine, please let me know. I will always replace damaged magazines or send larger format diagrams.

I will be at the Ham Radio exhibition in Friedrichshafen again this year, see page 30 for details of the exhibition.

73s - Andy



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Carl G Lodström, SM6MOM & KQ6AX

The Noble Art of Optical Communication Part 2

The first part of this article, in issue 4/2004, dealt with the practical aspects of optical communications that can be used by radio amateurs. It showed how to construct a laser transmitter and receiver. The second part explains some of the optical theory including, how to select the lenses needed, how to make measurements using a diffraction grating and how to make a light chopper. There are some colour pictures mentioned in this article, unfortunately the magazine is black and white, but the colour pictures can be found on the VHF Communications web site.

1.0

Choosing a lens

1.1 Numerical aperture and f-numbers

Obviously, the range of our transmitter will be severely limited if we have no “antenna gain”! A lens can gather up the light and direct it in a nice, parallel, beam. We have “collimated” the light. The output beam may now have a much larger exit diameter to begin with, a few cm, equal to the aperture of the collimat-

ing lens or telescope, but the spread is now so small, much less than a degree, so that after a km it may still not be very large, and thus the signal is more easily detected due to the higher intensity.

Another thing we will benefit from knowing about is the concept of a “fast lens”, as photographers call it. A lens with a large aperture is a “fast” because it lets in more light, allowing the use of shorter exposure times.

A very common camera lens, that would be quite suitable for both the receiver and transmitter, described in part 1 of this article, has a Focal Length (FL) of 50 or 55mm and a f-number of 1.8, which means its aperture that would fit 1.8 times in the FL of the lens (Fig 1). So the aperture of this lens is $55/1.8 = 30.6\text{mm}$. These lenses are 13 to the dozen and can be had more or less for free. Check with the local used camera stuff store! Old Single Lens Reflex (SLR) cameras like Pentax, Yashika and the like used these lenses. They often have the “Pentax thread”, M42 x 0.75, adapters are available for this thread so you can make mounts to fix the lens on a chassis. The ability to focus the lens does not hurt either.

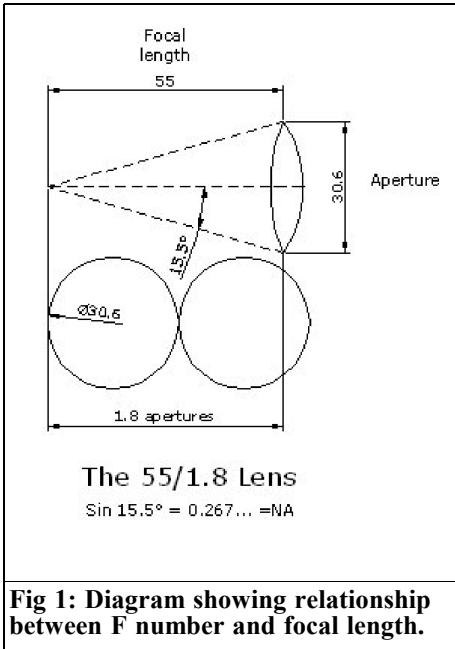


Fig 1: Diagram showing relationship between F number and focal length.

1.2 NA and *f*

For optical fibres, microscope objectives and some optical instruments, it is common to talk about the Numerical Aperture of the component. By this one means the Sine for the angle of half the light cone. Some microscope objectives approach a NA of 1! So they are looking at a bacterium not only from above, but almost from the sides as well! NA of 0.2 ~ 0.25 is not unusual for optical fibres. This indicates how much a light cone would come out of the end, or what maximum cone can be put in, thus what lens can be used for focusing light to or from the end of a fibre. To use an optic fibre with a particular NA together with a lens of a larger NA is a waste of lens. If NA is too small it will waste some of the light. A 20 ~ 40X microscope objective can be a very good choice for matching, getting parallel light in or out of, optic fibres. The objectives are usually marked with magnifications, length of the tube

they are designed for (usually 170 ~ 180mm) and NA.

A lens with an NA of almost 1 is the condenser in the table of a microscope. Parabolic mirrors can easily have an NA of more than 1.

Turn it the other way around: Our “Standard” 55/1.8 Lens (regarded ideally as a single element even though it is assembled from several glass elements, usually six in the Gaussian doublet) is then an aperture at a distance of 55mm from the film. The diameter of the aperture is 30.6mm. From the spot on the film, right behind the lens, one can imagine a cone to the edges of the aperture. The angle to the normal, the centre line, would be ~15.5°.

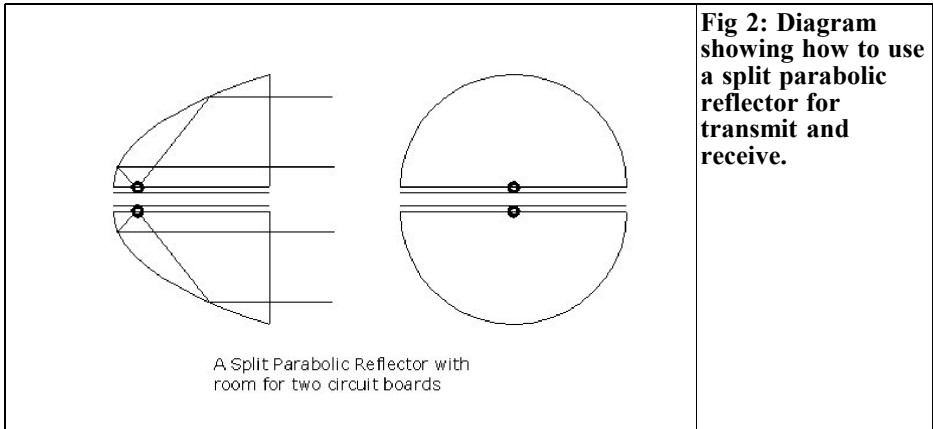
$$NA = \sin \left(ATAN \left(\frac{30.6}{2 \times 55} \right) \right) \quad (1)$$

2 x because the total angle is from side to side and the later part of the equation calculates ATAN for 30.6/2 divided by 55. In this case the half-angle is 15.5° for a total angle of ~31°. If used with a Laser Diode that spreads light ±16°, the major part of it would go into this lens, the desired result. Considering that these lenses ought to be almost for free, it is a good choice. For the Tx, projector lenses, that usually are even faster, may be even better choices and a FL of 15 ~ 25mm is enough, narrowing the field to lenses from small film projectors. With FL = 25mm I got as small a spot at 200m as I could handle!

SIN (15.5°) is 0.2672... so it is the NA of this lens. NA is useful for calculation of the maximum possible resolution of an objective. The smallest detail a lens can resolve (on the film plane) is:

$$\frac{\lambda/2}{NA} \text{ (metre)} \quad (2)$$

Where λ is in the same unit (metre). So, perhaps surprising, and contrary to experience, a fast lens has higher resolution



than a slower one! This is true, but it is difficult to make a fast lens without optical errors, so they “go soft”. For a normal lens, 55mm or so, this is the case at apertures larger than $\sim f/4$. For small apertures they go fuzzy again, just like the “Department store telescope” because of diffraction around the edges of the aperture. One can only resolve so much detail through a 1mm hole! So the “best” apertures for imaging with normal lenses are around $f/4$ to $f/8$. For telephoto lenses it may be $f/8$ to $f/16$. But this is a parenthesis; it is of little importance here.

But the NA of the lens is the most important factor, as we want to take care of all the light. This is true for both transmitter and receiver.

In case of the outgoing light, we like to have all the emitted light focused by the lens. With an elliptic output beam from the Laser Diode the lens must be large and close enough to absorb the light cone.

NA is the number that sums up this property and it is related to the f -number as:

$$f = \frac{1}{2 \times NA} \quad (3)$$

For various applications, one makes more sense than the other, even though they are two aspects of the same thing in the

same manner as frequency and wavelength.

With the particular Laser Diode used in part 1 of this article, where most of the light is inside a $16^\circ \times 6^\circ$ cone, the matching lens should have a corresponding angle. $\text{SIN } 16^\circ \sim 0.2756$, the needed NA for the lens. Converting it to an f -number, $\Rightarrow 1.813$ means that we can use these “regular” camera lenses. Some are even faster than this. Olympus had excellent lenses for their Pen-F series of cameras. 40mm/1.4 for example.

There are advantages with various lenses, and the philosophy may differ for the transmitter and the receiver. For the transmitter it is enough to take care of most of the light from the diode. We probably do not need a long focal length, or the beam becomes impossibly narrow even at a distance. For the receiver we want to capture as much light as possible, and get it onto the photo diode. So we want a lens with a large front element. If it is a long focal length, it will have to be aimed with care.

Parabolic mirrors are another possibility. They can easily have very small f -numbers, less than unity, or NA of more than 1. Some plastic reflectors for larger flashlights, lanterns, (torches) may have excellent mirrors. Of course, one can use them as they are, but it seems to me that

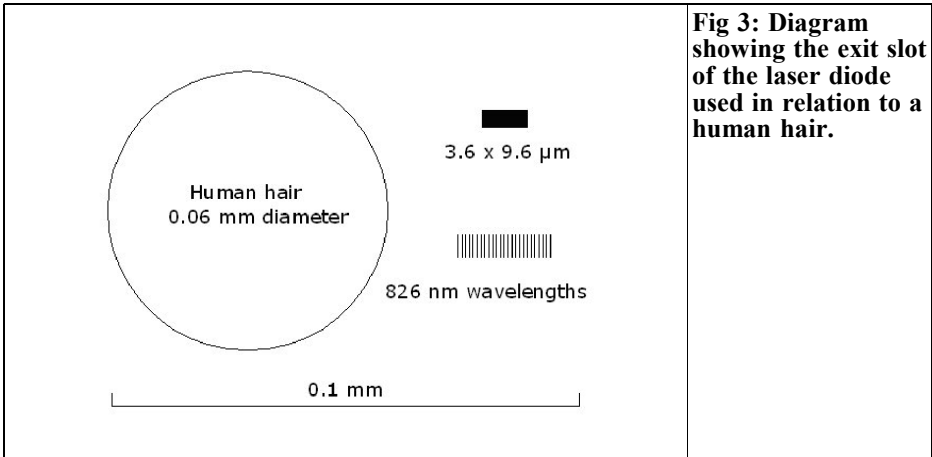


Fig 3: Diagram showing the exit slot of the laser diode used in relation to a human hair.

one could also cut one of them in half and use either half on a circuit board where the Laser Diode is mounted on one side and the detector on the other side. With the diodes “looking” sideways into the reflectors, from their focal points, the full angles of their emission and directivity are covered.

These reflectors cannot be used with the diode aimed towards the back anyway as they have a large hole there for the light bulb. Sideways is fine and it will handle all the light from/to the diodes!

I have not delved into this; although I have a nice reflector put aside for a rainy day... I assume that one will have to raise the reflector halves from the PCB by a small amount for to make up for lost material when cutting the reflector in half, as well as for the diodes sticking up a bit from the board. The focal point of the reflector is on its axis of symmetry. With two such reflector halves, one on each side, a nice instrument with an almost circular aperture can be built. Fig 2 shows how I visualize this, but it is up to you to decide what you want! It is probably also wise to build each circuit on a laminate with a ground plane. The ground planes can face each other and be connected.

2.0

Light measurement

2.1 Diffraction

Diffraction is something that matters for work with light in general and Laser Diodes in particular. It creates some problems (wider beams) but, as we will see later, we can also make diffraction work for us! Diffraction explains, for example, why it is physically impossible to get more useful magnification in a telescope than about 60x per inch of objective (or primary mirror) diameter. So, the “warehouse telescope” with 360x and a 2 inch objective diameter is for the birds, and I do not mean the ornithologists! The higher the magnification, the narrower is the “cone of light” available to the observer, as it originates in one distant point and ends with the objective diameter. Just like a highly directional antenna aimed at a distant source. A small aperture cannot separate two such cones, emanating from nearly the same point.

I remember it as Raleigh’s Diffraction Law (for round apertures).



Fig 4: Picture of light reflected in a taillight.

$$r = \frac{1.22 \times \lambda}{D} \quad (4)$$

where λ is the wavelength (in the same units as D) and D is the instrument aperture diameter (in the same units as λ) and r is the resolution in radians. Multiply with $180/\pi$ and get degrees.

The light from a Laser Diode follows the same rule. Since the aperture, the area from where the light exits from the chip, is so small, diffraction plays a major role for the shape of the exiting beam. The exit slot of the Laser Diode I am using is shown in Fig 3. The light spreads approximately by $16^\circ \times 6^\circ$ (each way from the central axis) so, for $\lambda = 827\text{nm}$, this is what we would expect from a $\sim 3.6 \times 9.6^\circ \mu\text{m}$ aperture.

2.1 The Diffraction Grating

This is a common component for determining the wavelength of radiation. In case of light, ruled flats are used. A very flat piece of glass is scribed by a diamond tip with some $100 \sim 2000$ lines (rulings) per mm! Special machines are designed and built for the purpose. For X-rays, crystals like Lithium Fluoride are used, where the atomic planes are the rulings. Gratings can be used either in a transmitting or reflecting mode. The reflective ones, for light, are usually coated with aluminium.

The grating acts like a prism, but for reflected light as well. We will see more about this soon. In a grating red light is deflected more than the blue. Prisms are the other way around. The angle of reflection for each colour is dependent on the wavelength of the light and the density of the lines, the rulings. The basic formula is:

$$d \sin \theta = m \lambda \quad (5)$$

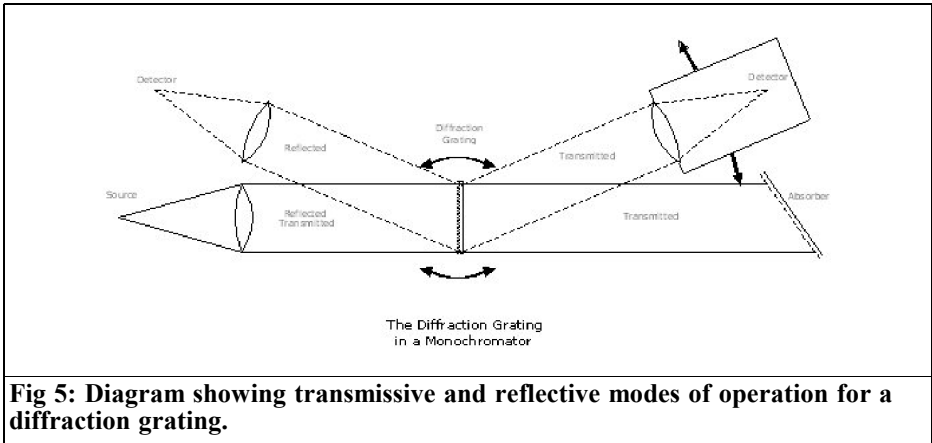
where:

- θ is the deflection angle
- d is the separation between each ruling line
- m is the order of the diffraction peak (1st, 2nd, 3rd ...)
- λ is the wavelength ($\sim 540\text{nm}$ for green light, $>700\text{nm}$ for IR)

Turning equation (5) around, one can calculate θ if all other parts are known.

$$\theta = \sin^{-1} \frac{m \lambda}{d} \quad (6)$$

Fig 4 shows a picture of the Sun reflecting in a taillight, photographed through a telescope and the insect screen in my window. We can see the diffraction peaks up to $m = 3$ on either side of the main spot. Each peak has the full spectrum of colours. The insect screen has 7.0



wires/cm giving a d of 1.486mm

Assume green light, “in the middle“ of the visible spectrum, $\lambda = 540 \cdot 10^{-9}$ metre (540nm). Using the equation (5) we can calculate the angle between the diffraction peaks in this picture!

$\theta = 0.02082^\circ$ or 1.25 minutes of arc, no problems for a telescope to resolve such a small angle, corresponding to 0.36m/km.

The visible range, ~390 to 700nm is contained within one octave. Had we been able to see more than one octave, the diffraction peaks above would have gone into each other and been smeared out. The IR from the first order will blend with the UV from the second. A diffraction spectrum from a monochromatic source, such as a laser or (almost monochromatic) a Sodium lamp, results in a sharp line for each order, m . Sodium has a double that are two lines very close together.

2.2 The monochromator, a little side project.

Sooner or later you will want to measure the sensitivity of various diodes, wavelengths of light from lasers or transmission or reflection from dielectric filters.

Or check if those “UV safe” sunglasses really are cutting below 375 nm!

You will need a diffraction grating! Fortunately, the Great Creator in Silicon Valley has come up with the CD so now everybody can have one! With a nominal 625 lines/mm it is a very good element for an optical grating! Obviously, it is reflective, but I found that the “silver” could be peeled off from the TDK CD-R that I used. Using tin snips, I cut two chords either side of the centre, resulting in a rectangular piece with parallel edges, about 1 inch wide for use as a grating. After a day the metallisation started to peel at the edges! Maybe humidity got in and did a job on it. I could peel off the metal layer with tweezers and got a clear, transparent, diffraction grating! It can still be used in a reflective mode, but this mode was better with the metal left on.

In Fig 5 shows two modes of operation pictured at once, one reflective and one transmissive mode. In either it is advantageous to have a lens on both source and detector, focused on infinity. Parallel rays then pass the grating and one gets more light and a better resolution.

In the reflective mode one can have the Source and Detector fixed and rotate the grating. At some angle one will get a plain reflection from the surface of the grating. This is zero degrees of angle.

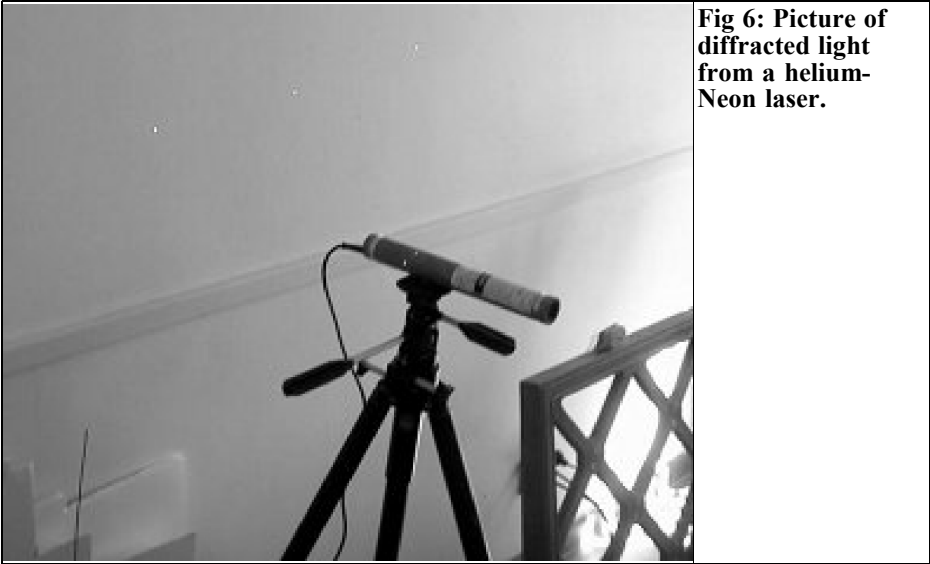


Fig 6: Picture of diffracted light from a helium-Neon laser.

The diffraction peaks will appear on either side of this reflection.

In Fig 6 we can (hopefully) see a Helium-Neon (HeNe) laser at a known wavelength of 632.8nm and three spots on the wall behind it. The middle one is a pure reflection; the other two are the first order peaks on either side. The grating (a piece of a CD) sits on the back of a chair 100cm from the wall. The peaks are 45.2cm from the reflection. The angle is thus $\text{TAN}^{-1} 0.452 = 24.323^\circ = \theta$.

Re-writing equation (5) gives:

$$d = \frac{m\lambda}{\sin \theta} \quad (7)$$

For ($m = 1$), the primary peaks, $d = \lambda/\text{SIN } 24.323^\circ$ or $632.8\text{E-}9/0.412 = 1536.37 \text{ nm/line}$ or 650.88 lines/mm . Pretty close to the 625 that is supposed to be the CD standard. This measurement is probably not in error by so much and several standards may exist. I have noticed "Playing Times" marked on CD's. I remember some are 67 minutes and some 80 minutes, so maybe here the difference? You can always calibrate your instrument against some known lines in

Mercury and Sodium. If you intend to use the printed scales, calculated to be mounted as a straight line, enlarge them on a photocopier in proportion to the distance of your instrument. For example by 4x for 400mm distance. An example scale is shown in Fig 7.

In the transmissive mode, most of the light goes right through the grating. Call this angle 0° and find the diffraction peaks on either side, at an angle that corresponds to the wavelength and grating density. The detector must be swept through the angle; rotating the grating will not work in this mode.

Whatever type of monochromator you build, you can modulate the light source by an audio frequency (chopper!) and make a daylight blind detector, using a transformer, as described in part 1 of this article. You can now amplify the signal and measure it with an AC voltmeter! A gramophone pre-amplifier, for example, is very sensitive and can be used here. Their gain may very well be in the 60dB range below 1kHz. Notice that they are RIAA compensated, so the gain drops above this frequency! It is now easy to find, and measure, the peak from a

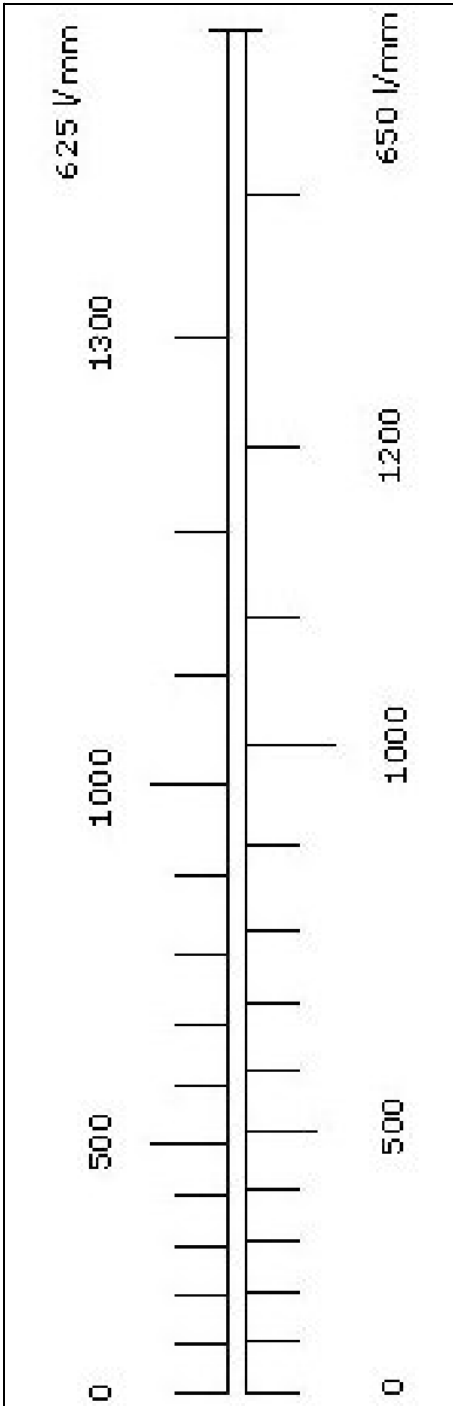


Fig 7: Scale for use with a diffraction grating. The scale length is 180mm and is for use at a distance of 100mm.

source, but to measure the sensitivity of a detector using a light bulb one may have to modulate or chop the light. The light from a bulb is flat enough over wavelength, to determine if a detector is more sensitive at 850 or 920nm. A small electric fan, or a CD player with holes in the CD, will do for an optical chopper. For the outmost in sensitivity, if it is needed, synchronous detection is unbeatable. It is an amplifier, maybe after a pre-amp, switching between gains of +1 and -1 as the light turns on and off. You can even detect signals in full daylight that are so weak that they cannot be seen in a dark room!

In Fig 8 Laser light, from a laser pointer hits the CD piece right on. We can see the main reflection towards the camera and the $m = 1$ and $m = 2$ order peaks on either side.

In Fig 9, a flashlight (torch) is photographed through the CD (transmissive mode) and we can of course see the light source in the middle. We can also see the two first order peaks, one on each side. Each of them has the blue closest to the lamp and the red at the outer edge. IR will of course fall just outside the red. Zooming in on the left spectra gives a nice continuum from red to violet (left to right), Fig 10.

Fig 11 shows a Sodium street light, we can see that there is some Mercury vapour in it too. The intense yellow is really a double line, 588.9950 and 589.5924nm, of Na, the standard test for spectrographs... Hg has strong green, cyan, blue and violet among other lines. With lenses, as suggested, the monochromator would be able to separate the Na doublet, 0.6nm apart!

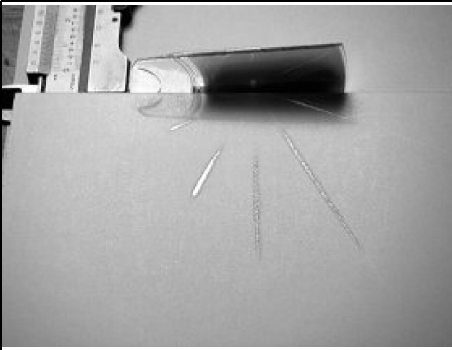


Fig 8: Light from a laser pointer hitting CD.



Fig 9: Flashlight photographed through the CD.

2.3 An Actual Monochromator

Inspiration for the homebrewed Monochromator may come from taking a look at a simpler commercial instrument, one that easily resolves the two lines in Na. It is not that hard to put something together!

This particular instrument, shown in Fig 12, is a Spectrometer (if it had a power supply and a meter), but that is because it is a monochromator with a light source and a detector! A light bulb on the right side and a Photo Multiplier Tube (PMT) on the upper left. On the upper right, a mechanism, here showing 446nm, with a mechanical counter that keeps track of how much the Diffraction Grating is rotated. Fig 13 shows the same instrument from below, lid removed.

Although we are not using a PMT in our applications, it is still one of the most

sensitive detectors known to man. It will detect about every third photon that reaches its window and all (detected photons) results in a pulse, the height of which corresponds to its energy and the gain (adjustable) of the tube! The blue sensitivity is great, the IR not so great. A PMT requires a stable, and variable (for setting of the gain) HV power supply, usually in the 800 – 1200V range with almost no current. The rise/fall time is in the ns, or a few, range. The lifetime is virtually infinite, as long as the tube is not abused. The output current can well be 0.1 ~ 1mA so it can drive a meter directly. The gain is dependent of the HV and can be as high as 10^8 but is certainly above 10^5 . All the pins on the socket are just connected to a ladder of high value (~1M Ω) resistors, dividing the HV to the dynodes, as the electrodes are called. that's all!

A fortunate circumstance, where geometry and trigonometry coincide

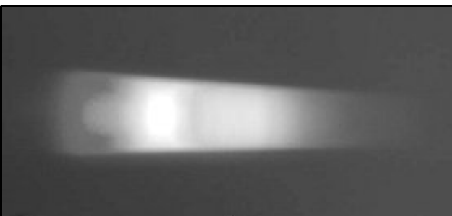


Fig 10: A zoomed view of the left spectra of Fig 9.

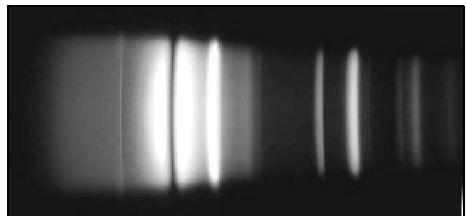


Fig 11: Looking at a Sodium street light.

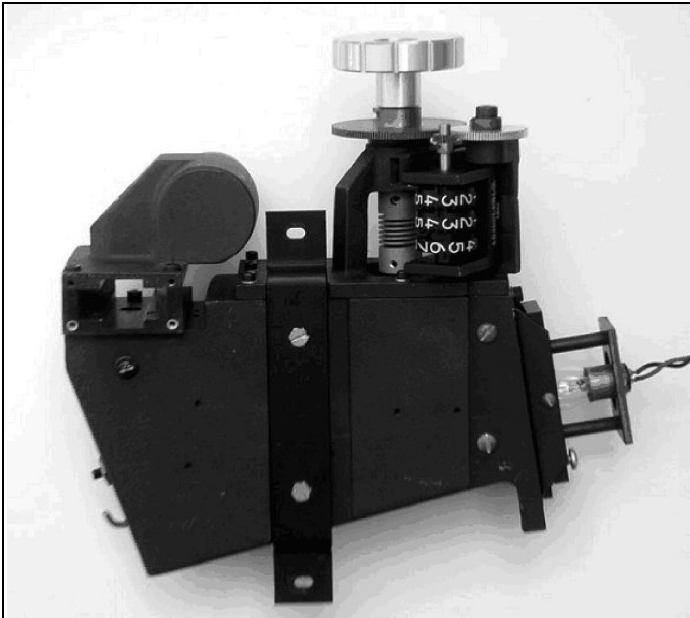


Fig 12: Picture of an actual monochromator.

(hyperbolic cosecant functions, or something the like, that is above the head of this author) allows us to use a simple

mechanical device for setting and indication of wavelength! But let us first see how the light bounces around in this

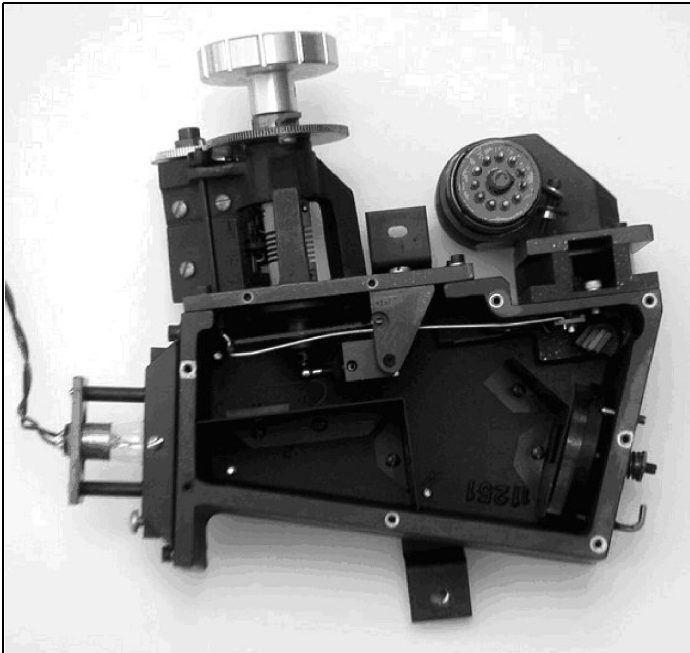


Fig 13: Picture of monochromator with the lid removed.

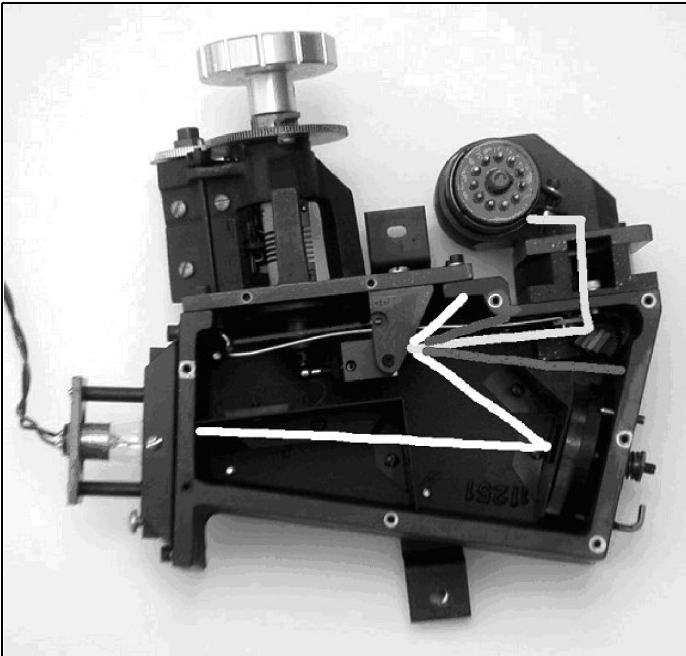


Fig 14: Light path inside the monochromator.

instrument (Fig 14):

Light from the lamp enters a slit in the left wall. It proceeds (and spreads) until it reaches a spherical mirror at the right. Here it is reflected and partially focused as it reaches the Diffraction Grating (upper side, middle) from where the direct reflection, and the diffracted colours of that side, are absorbed in the wall above. This, and the lower "rainbow" are swept by rotation of the Diffraction Grating. Below the total reflection, from top down, a blue, a yellow and a red beam are added to the picture. As pictured, only the yellow reaches the detector, first via a 45° mirror, then through a slit (the focus of the spherical mirror), through an area where a cell with a sample can be placed, off a second 45° mirror and to the PMT. By rotating the Diffraction Grating, any part of the spectrum, colour, may reach the PMT, as the sample allows. In this instrument only about $\frac{1}{2}$ ~ 1nm at a time will pass through the slit.

The mechanism for rotating is a fine screw with a dial. The dial can be

calibrated directly in nm! With another placement of one edge, the dial can indicate the inverse, the "wave number" (waves/cm). Here is where my brains short out, but I know it is true. I once rebuilt a large SPEX spectrometer from wave numbers to wavelength, but it was long ago but I do not remember how I figured it out... The pivot point for the grating must be selected correctly for this to work. A geometrical problem that maybe some reader is familiar with.

With a CD-Grating (625 1/mm) the angular difference in the ~700nm range is about 4° per 100nm. I have three monochromators, but if I were to build yet a fourth one, using a CD as a Diffraction Grating, I would use a 15-turn knob with a counter. Using the 4 ~ 11 turn counter numbers. One turn would have to represent 100nm. One cannot go much below 400nm or above some 1100nm with a Si diode, and it may not be interesting anyway. I would use a lead screw of steel, as large as possible and with as fine thread pitch as possible. The nut I would

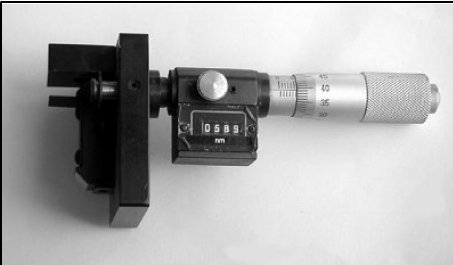


Fig 15: Picture of micrometer screw.

make of three brass nuts, spring loaded to each other (thereby cancelling the play of the thread). A micrometer screw (Fig 15) is even more appealing, especially if it has a digital readout. With 0.5mm per turn, to generate 4° of rotation of the arm, onto which it acts, must be $TAN(4^\circ)/0.5 = 7.15$ mm long.

(As I check out eBay on the Internet for real Diffraction Gratings I see a 110kg Perkin Elmer Spectrometer, complete with motors and amplifiers, for \$100. The advantage of building your own is that you would avoid the shipping charges from Northern California...)

In the pictured spectrometer the Diffraction grating is mounted in a holder, and this holder is hinged in two steel ball bearings. The arm has a ball on the end, resulting in a well-defined point of con-

tact with the flat end of the micrometer screw. A U-clamp/spring makes sure the arm is in contact with this micrometer end even while backing off the screw.

In Fig 16 and 17: MS is the Micrometer Screw that makes contact with a ball at the end of the arm. The length of this arm is from a line between the pivot points P to the contact point on the ball. The surface of the Grating is on this line. Below you can see a knurled lock-nut as the length of the arm is adjustable for calibration of the instruments dispersion (nm/turn). The Diffraction Grating Constant in this instrument may be other than 625/mm => different length of the arm. G denotes the Diffraction Grating and P the pivoting points.

As this instrument is intended for detecting weak absorption lines in a transparent sample, it uses surface mirrors throughout. Thus, the light does not have to go through anything but the air in the instrument and the glass in the lamp and the PMT.

Fig 18 shows a common configuration for a monochromator. The mirror is spherical, easy to make large and with a large NA (= small f-number) so a lot of light can pass through. This design would probably be your best bet. Although it would not be a surface mirror, you can adopt this design using a spherical shaving mirror and for the small mirrors, the ones in the lid of women's pancake makeup are pretty good. Scribe and break one in half for the small folding mirrors. The spherical, main, mirror can be

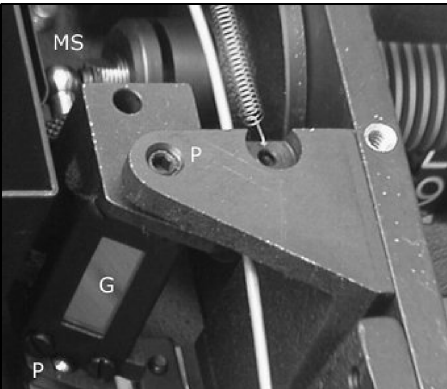


Fig 16: Picture showing detail of the monochromator.

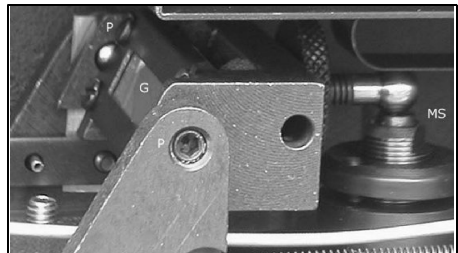


Fig 17: Picture showing detail of the monochromator.

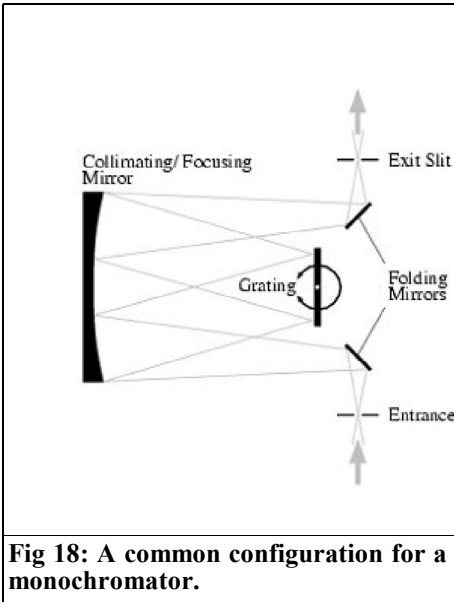


Fig 18: A common configuration for a monochromator.

masked with black paper so that two rectangular areas, side by side, are available to the passing light. It is all that is needed anyway, and keeping unused surfaces black, or sooted, cuts down on the stray light that otherwise can pass the instrument. The main mirror focuses the input slit onto the output slit (and the other way around) so the distance between the mirror to a slit (via a small mirror) is equal to the mirror radius. The grating can be rotated with a screw and a cam as in the machine in the earlier pictures.

If you build this one, make the slits (and the small mirrors) tall enough to handle one upper and one lower light source (on one end) and detector (on the other). Maybe 10-15mm total. There will soon be another article in which something will be described that will allow you to do some really nifty things with this machine! You could even include a small baffle (perpendicular, horizontal, black card, just between the small mirrors and the slits) separating the upper and lower parts of both apertures. Together with a clean main mirror and grating it may give

enough separation. Both upper and lower beams will come out the opposite on the other end, and both will use the entire grating and sectors of the main mirror.

It is very easy to calibrate your Monochromator! If you use an optical chopper, especially with a synchronous detector, you can use faint amounts of light from streetlights. They are usually discharge lamps with Sodium and Mercury vapour. They have characteristic emission lines. Regular fluorescent tubes contain Mercury too. To find out the wavelengths of lines from these elements, go to the web site http://physics.nist.gov/cgi-bin/AtData/lines_form? and enter Hg or Na, or any other element, in the Spectra window!

3.0

Modulation of constant light sources

A standard way of doing this is to “chop the light”, with a “chopper” of course. In this country (the U.S.) AOL is kindly supplying CDs at the rate of one every few months. Sometimes they come in a nice box too. One can go to the post office and get more of them, for free, if there is a need. How come music CDs cost \$18.00? Anyhow, I got the idea to build a chopper of one recent delivery, using the motor from a dead CD player.

In Fig 19, the small black rectangle below the motor, is an opto-coupler. It is coupled by the reflections in the CD, when it is not facing a hole. It has one IR emitting LED and an open collector photo transistor. A k or a few from each to 5 V works fine. This way one can synchronize a synchronous detector, or a lock-in amplifier, that is about the same thing. See my article about a NF meter [1] for a circuit! This way you can detect very weak signals from any light source of your choosing, through your home-



Fig 19: Light chopper using a CD and CD player motor.

built monochromator or some dense filter. You can get the result direct on a meter, ready to read. Other sources, even if they have a lot of 120Hz modulation on them, does not matter the least. You can get the light from a distant streetlight, chop it alone, focus it into the monochromator, and do synchronous detection on it. In this application, a Transconductance Amplifier is beneficial after the detector diode. It is not an RF application. One may still use an audio transformer if there is a lot of ambient light to get rid of.

The motor seemed to be of a good quality, it kept a very steady speed. There should be no problem to find an old CD player to junk the motor from. They have a chuck that holds CD's also! The measured current consumption was almost entirely from the wind resistance of the CD, it probably would be good to round the edges of the holes. With various voltages I got these current consumptions, speeds and chopped frequencies (for my 13 hole CD):

Motor voltage	Current (mA)	RPM	Hz
0.2	~10	85.4	18.5
1	40	921	199
2	90	1630	353
3	140	2221	481
4	200	2696	584
5	270	3113	674

A few hundred Hz will do very well for the application. It must not be a multiple of the line frequency.

4.0

References

[1] A Simple Noise Figure Meter, Carl G Lodström, SM6MOM, KQ6AX, VHF Communications 2/2002, pp. 105 - 110



Gunthard Kraus, DG8GB

Practical Project: Stripline low pass filters

Our “Practical Projects” series continues with this article on how to design and make stripline low pass filters as individual universal modules.

1.0 Foreword

Anyone who used to play with building blocks as a child will easily understand this “long-term project” by the author. What a joy it is, for a real “RF Freak”, to be able to open a drawer and assemble a circuit or a receiver using existing modules, or to try out a new principle, or to check something that has been published or suggested by someone else. Also to design a new “building block” and bring it to life, for another frequency range can be considered as a real challenge.

Fig. 1 shows what can be expected in the course of a developer’s practical work on stripline filters, the various design ideas are collected together and waiting to be housed in a project. Since every frequency range has its own perils, it is good to be able to pass on the results obtained to other DIYers. After all, we don’t want everybody to have to re-invent the wheel.

2.0 Technical data and general standards for module system

All circuit boards are designed in accordance with the same prototype:

- The board size is 30mm x 50mm, or 30mm x 130mm for exceptional cases.
- “RO4003” from ROGERS is used as the printed circuit board material. It is very low-loss up to 10GHz, advantageously priced, and makes a splendid processing material from the mechanical point of view. It is copper-coated on both sides (useable up to a few GHz with a layer thickness of 35µm) and 0.813mm thick (25 MIL) and all connections to earth are created from the bottom surface, using 0.8mm feedthroughs made from full tubular rivets to the continuous earth layer.
- The circuit boards are built into milled aluminium housings with SMA connectors. 50Ω striplines on the boards lead to these SMA connectors.

If supply voltages etc. are also required in active circuits, then SMA connectors

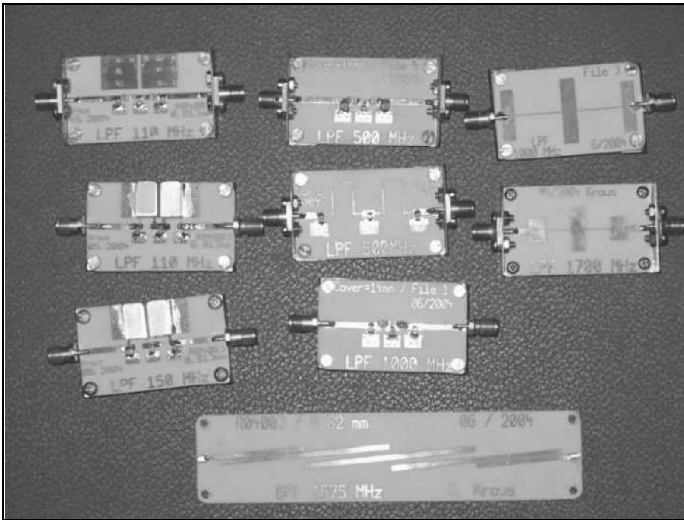


Fig 1: Some examples of the author's stripline filter development boards.

are positioned on the top side for this purpose. A third RF connection (as, for example, in mixers) is located on the bottom side, once again taking the form of an SMA connector. Fig 2 shows what this looks like for a double-balanced mixer module.

So now for the filters:

The Chebyshev type, class $N = 5$ and with a maximum ripple of 0.1dB in the transmission range, has been demonstrated to be a good compromise, with sufficient edge steepness in the stop band. This ripple effect is associated with

a reflection attenuation value of 16.4dB, and the equipment operates at the “ripple cutoff frequency” when using the filter calculation program.

3.0 Various 110MHz low pass filters

3.1. Component values and printed circuit board designs

For anyone who’s surprised by the choice of cutoff frequency, here’s the secret: it does a very good job of reducing the distortion factor of a signal generator from 50MHz at up to 100MHz. It can also be used with receivers or converters with an input frequency up to 100MHz, without this even making the noise factor worse through a marked increase in the attenuation.

So now for the circuit:

Here we can work, for example, with the “fds.zip” program (from the Internet). Determining the components required is

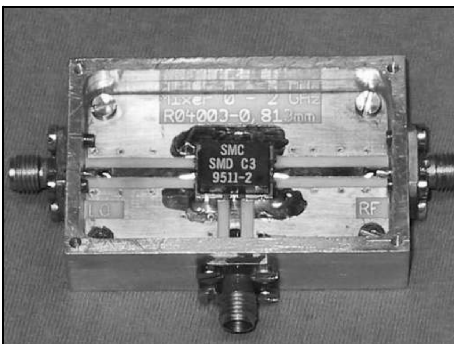


Fig 2: Picture of a double balanced mixer in the standard format.

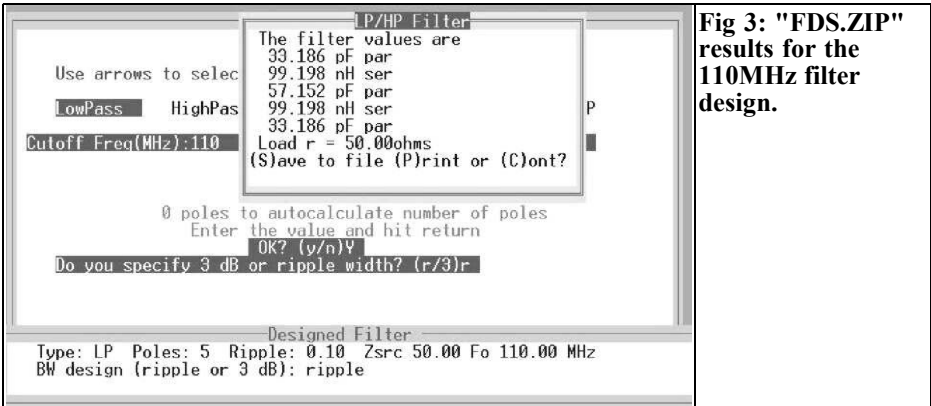


Fig 3: "FDS.ZIP" results for the 110MHz filter design.

a small matter if you just select the following options one after another from the menu and enter the appropriate values:

- Chebyshev type
- N = 5
- Ripple limiting frequency = 110MHz
- Z = 50Ω at input and output (symmetrical filter)
- Maximum ripple = 0.1dB

Fig 3 shows the results screen. If we take a close look, we get a nice surprise: the two inductance values are almost exactly 100nH. As a result, we can work with ready-made coils and we don't have to

do any winding ourselves. There is no problem making the capacitor values, they are approximated using parallel connection of up to three SMD capacitors.

The printed circuit board shown in Fig 4 has been created so that we can test out various types of coil. It is easy to recognise the 50Ω stripline running from left to right in the middle of the board, which is 1.83mm wide (it was calculated using the "TRL85" program, while a cross-check was carried out using the most recent version of APLAC, the stripline calculator, which can be found under "Tools").

The feedthrough pads for the three capacitors in the filter can be seen below the centre, whereas the two big earth pads above the centre are there to earth the coil housings. If only small modern 1210 or 0805 SMD RF choke coils are used, these can be soldered directly using the breaks in the stripline.

3.2. Simulation of filter circuit

The author has obtained the following results and adopted the following habits while working with this equipment:

The quickest method is still to use the PUFF simulation software, provided all the models and functions required are available. Should this not be the case, use

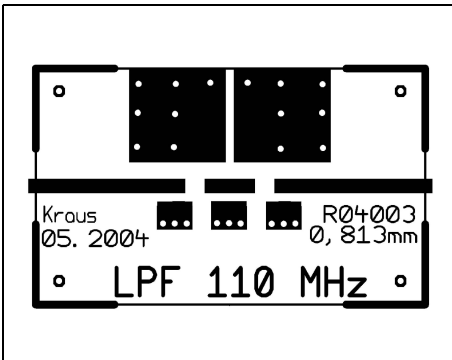


Fig 4: The test PCB, the 50Ω stripline can be seen in the centre.

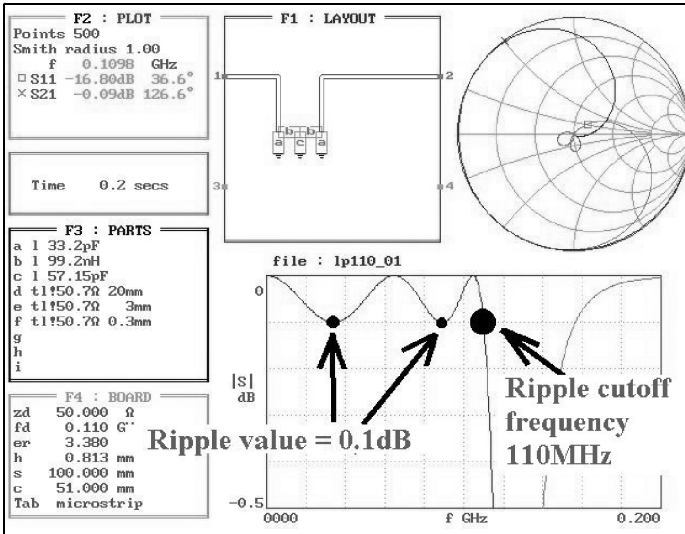


Fig 5: Results of PUFF simulation. Check the ripple in the pass band and the cutoff frequency.

the free student version of APLAC and generate the missing model (e.g. the open-end extension for a stripline at no-load or the transition from a narrow stripline to a wide one, etc.) as an S-parameter file and then process it further using PUFF.

If the project is too demanding for PUFF or the APLAC student version, then the only thing to do is to go to the section with the full version of APLAC. It's true that there is also a "dodgy" older version of APLAC haunting the Internet; but we won't go there!

First, carry out a simulation using PUFF and the components taken over from the filter program, and check that the proposal is correct. To do this, just check, in the transmission range, whether the value for the S21 ripple is exactly 0.1dB (Fig. 5). While you're doing this, just take a glance at the ripple cutoff frequency. With a display range converted to 0 to -50dB, you should then test to make sure that the reflection attenuation does not exceed the maximum value of -16.4dB for the ripple value of 0.1dB (Fig. 6).

A reminder: the ripple and reflection values here are inter-connected by the following formula:

$$Ripple = 10 \log \left(\frac{1}{1-r^2} \right) \text{ in dB}$$

where r is the reflection factor.

If, on the other hand, we use the reflection factor associated with the ripple value, then, following the conversion, we obtain this relationship:

$$r = \sqrt{1 - \frac{1}{10^{\frac{Ripple}{10}}}}$$

And since the S11 parameter is usually printed out in dB in the simulation program, here's the formula to use for this:

$$S11 = 20 \log(r) = 20 \log \left(\sqrt{1 - \frac{1}{10^{\frac{Ripple}{10}}}} \right) \text{ in dB}$$

And now for the first circuit:

For inductances, we use high-quality helical coils manufactured by the NE-

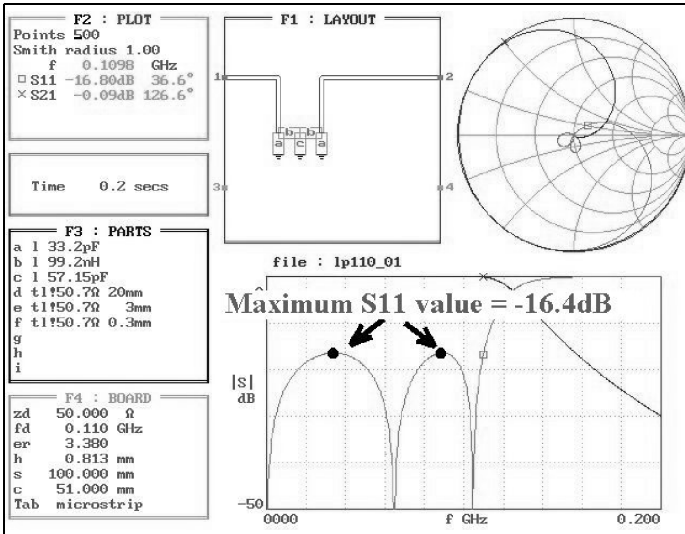


Fig 6: With the range set to 0 to 50dB, check the maximum ripple is -16.4dB.

OSID company, in a silver-plated copper housing, with $L = 100\text{nH}$ and with a quality rating of $Q = 130$. At 100MHz , this quality factor gives a series loss resistance of approximately 0.5Ω , which can be entered, directly opposite the coil, in the F3 list in PUFF (required entry: "1 100nH+0.5"). We thus have no difficulty in determining the filter behaviour in this frequency range, because the capacitors and the printed circuit board material do not pose any problems here.

The set-up file from PUFF is now edited and supplemented by the following printed circuit board material data for

R04003:

- Relative permittivity $\epsilon_r = 3.38$
- Loss factor $lt = 0.0005$
- Board thickness $h = 0.813\text{mm}$
- Copper coating $mt = 0.035\text{mm}$
- Surface roughness $sr = 0.002\text{mm}$

The complete simulation circuit required can be seen in Fig 7. It contains, not only the stripline sections provided for on the printed circuit board, but also the four "open-end extensions" required for the

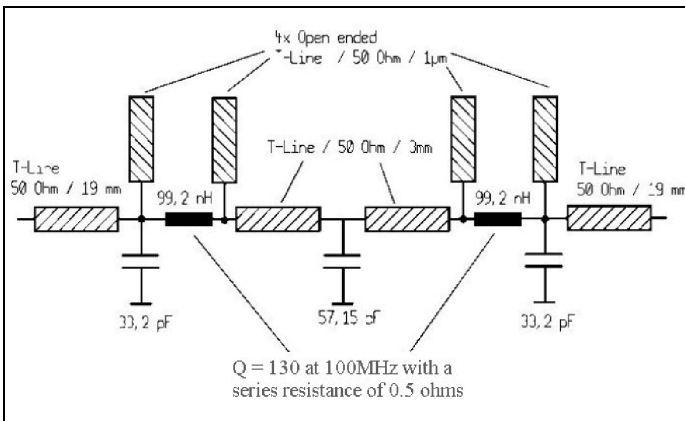
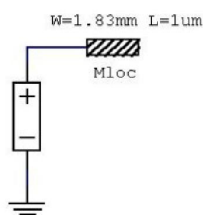


Fig 7: Circuit used for simulation.





W=1.83mm L=1um
Mloc

```
Sweep
"open-end-fringing"
loop 501 freq lin 0 5GHz
Window=0 Smith
Window=1 Y "" "dB" -40 0

Show
+W=0 RI S(1,1)
+W=1 DB S(1,1)
EndSweep
```

```
50
STORE "fring01.slp" GHZ MA
```

```
MSub R04003
ER=3.38 H=0.813mm T=0.035mm RHO=0.75 RGH=0.002mm TAND=0.001
COVER=500mm LEVEL=2
```

MSub	R04003	(E)
Circuit Diagram	[...]	(E)
Sweep	["open-end-fring..."]	(E)

Fig 8: Input parameters used for APLAC simulation of 50Ω open circuit stripline.

two breaks in the stripline (to build in the two coils). It is well known that PUFF cannot model these things, so you can get some help from the student version of APLAC as follows:

We simulate a 50Ω stripline at no-load (“Mloc” = MicrostripLine with open connection) with a length amounting to 1 micrometer, using the R04003 printed circuit board data in the frequency range of 0 to 5GHz. The S-parameter file generated is saved under a suitable name and then copied into the PUFF directory.

However, for future use you should set up a directory somewhere with attachments of this type, and collect everything there which comes up in the course of a developer’s life.

With APLAC you have to go through three operations before you reach your objective (see also [1]): apart from the circuit diagram, you need an “Msub” file with the material data, and naturally you need the “Sweep file”. This contains the data relating to the frequency range, the simulation step size, the listing of the sizes to be simulated, the type of screen output, etc.

The Msub file looks like this:

```
ER=3.38
H=0.813mm
T=0.035mm
RHO=0.75
RGH=0.002mm
TAND=0.001
COVER=500mm
LEVEL=2
```

And the sweep file for the frequency range from 0 to 5GHz is formulated like this:

```
“open-end-fringing”
loop 501 freq lin 0 5GHz
Window=0 Smith
Window=1 Y "" "dB" -40 0

Show
+W=0 RI S(1,1)
+W=1 DB S(1,1)
```

The complete APLAC entry required can be seen in Fig 8. The simulation circuit diagram consists of a 50Ω port, to which the circuit section at no-load is connected. The S-parameter file “fring01.slp” is saved using the attribute entry:

```
STORE „fring01.slp GHZ MA
```

After the simulation, this S-parameter file is immediately copied into PUFF and also saved somewhere.

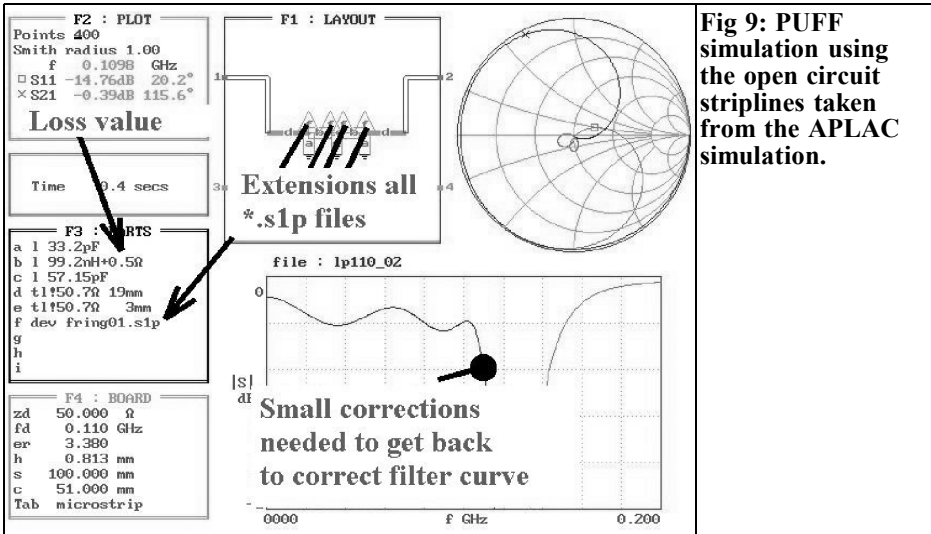


Fig 9: PUFF simulation using the open circuit striplines taken from the APLAC simulation.

The PUFF simulation, with all the modules constructed in accordance with Fig 7, can be seen in Fig 9, with the four open line ends included using the imported S-parameter files. Together with the two central line sections, they require slight corrections to the component values as opposed to the initial circuit, especially at higher frequencies, to get back to the correct filter curve and the original limiting frequency desired. The losses in the transmission range can be seen easily. Naturally, the simulated attenuation at very low frequencies should not be taken seriously: the actual loss resistance rises with the frequency and has considerably smaller values below this, so that the complete circuit will also behave considerably better here.

By the way, for anyone who's surprised that we can't just go on working on these not too complicated filter simulations using APLAC, here are three reasons why. Firstly, PUFF and APLAC give exactly the same results for this circuit using S-parameter simulations. Secondly, the PUFF entry for the circuit for something routine is somewhat quicker than setting up the APLAC screen and the two files required. And thirdly, it's very

annoying when the APLAC student version is frequently and unforeseeably suddenly closes with the message "Memory restricted in this version!" as soon as anything in the circuit is modified (in our case, see the addition of open-end extensions to the filter). It is a real pity that this happens with lots of Windows software.

3.3. Measurement results using prototype with helical coils

The printed circuit board can be seen in Fig 10. The measured S21 response in the transmission range can be seen in Fig 11 – with the best will in the world, it doesn't conform to Fig 9 and the basic attenuation (that increases with the frequency) is considerably higher than expected. The search for the "why" rapidly brings the cause to light:

The author has often preached in favour of the exclusive use of SMD capacitors made from the low-loss materials "COG" or "NP0" in filters. This material can easily be identified by its violet colour. A standard capacitor with an indefinable grey-white appearance was used in the prototype. However, who would have thought that this would have had so many

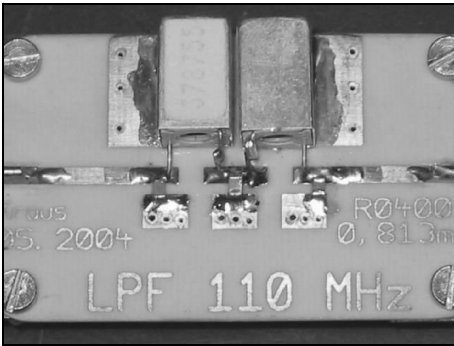


Fig 10: Picture of the PCB fitted with helical coils.

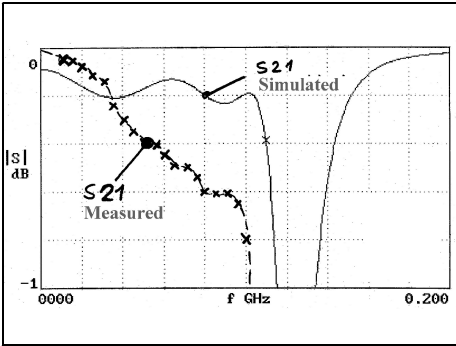


Fig 11: Measured response of filter using helical coils.

consequences? Well anyway, it's simply, one more warning to be more careful.

Once the culprit had been removed, the transmission range looked as seen in Fig 12; apart from the slightly increased ripple (caused by the fact that the capacitor combinations were not absolutely perfectly put together and the coils were not entirely correctly tuned), it's a thoroughly encouraging picture and about what we would have expected.

It was only when the stop band up to 2GHz was measured (Fig 13) that I really did start to frown a lot, for just above 1GHz and around approximately 2GHz the stop band attenuation briefly collapsed.

Anyone who has already amassed some experience with stripline filters knows that very thin circuit sections also act as inductances. Only if the circuit is not very short, as opposed to the wavelength, does the coil simply become a circuit again which and behaves in a peculiar way at all multiples of $\lambda/4$. The helical coil behaves in the same manner. At some point it changes from a coil into a circuit and we therefore experience this unattractive behaviour of the filter. Consequently, this design format is not used except when the behaviour above 1GHz is of no importance but the attenuation in the transmission range has to be as low as possible.

Perhaps some of you have already been asking yourselves how the very low attenuation values for the transmission

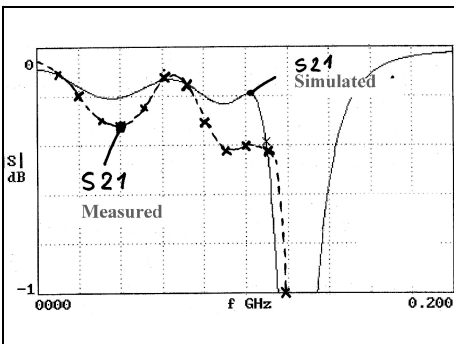


Fig 12: Measured response of the filter after the capacitors changed.

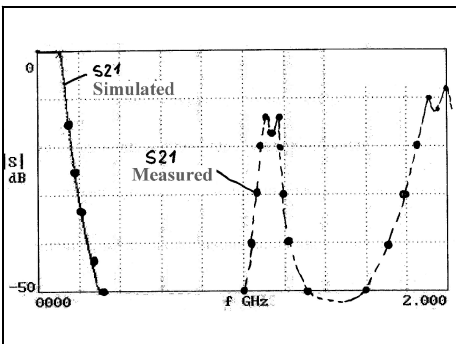


Fig 13: Filter stopband measurement up to 1GHz.

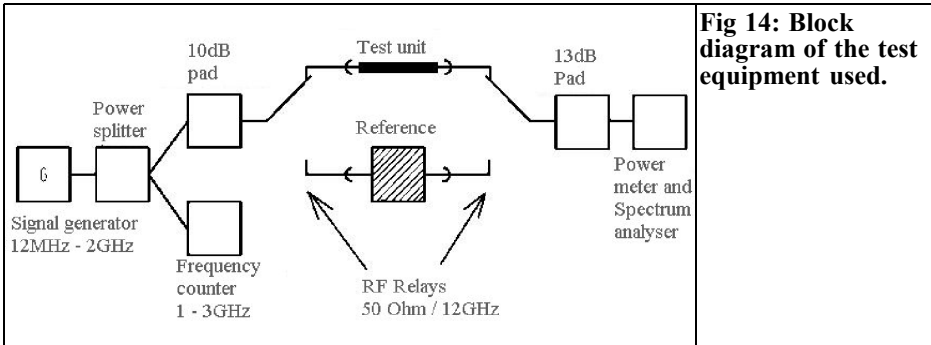


Fig 14: Block diagram of the test equipment used.

range can be measured in a “hobby workshop”. After all, there’s no expensive modern network analyser available on the premises, but “only” the HP8410. Here we can obtain some assistance from Fig 14, in which we can see the measurement equipment assembled for this work. Unfortunately, it looks just as unruly as that when assembled in practice (Fig 15). However, if everything is correct and is neatly calibrated, you quickly get used to it. The trick with this layout is the same as with the most expensive measuring machines: calibrate the layout with a “reference” (here this is simply a short piece of Teflon cable with SMA plugs) and then use two RF relays to convert to “Measure”. Instead of the reference, the module to be inspected is now inserted into the signal path and the difference in the values displayed on the output measurement equipment really is due only to the object being investigated.

3.4. Conversion to SM choke coils

The broadband behaviour up to levels exceeding 2GHz cannot be improved unless minute coils with correspondingly high self resonances are used. Here modern SMD chokes in sizes 0805 to 1210 spring to mind straightaway; their only disadvantage is a quality level which is markedly worse than that of the helical coils, usually not exceeding $Q = 50$.

The filter data obtained can be seen in Fig 16. As expected, the attenuation level has risen somewhat in the transmission range (see Fig 12). But the stop band attenuation now follows the simulated path precisely, and up to 3GHz is so high (and without any interruptions) that it is no longer displayed by the HP8410 network analyser (exceeding 60dB). We can see from Fig 17 how wonderfully small a circuit of this kind can be.

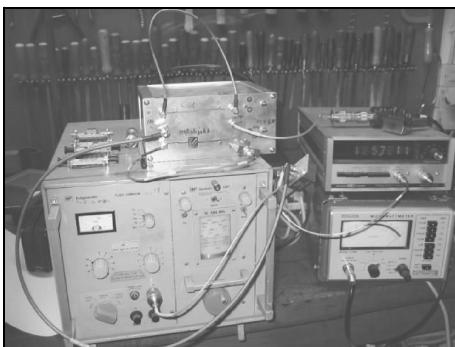


Fig 15: Picture of test equipment used.

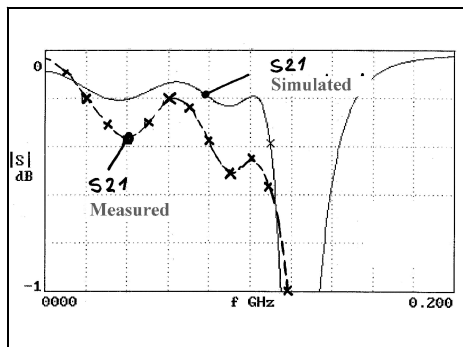


Fig 16: Measured response of filter using SMD coils.

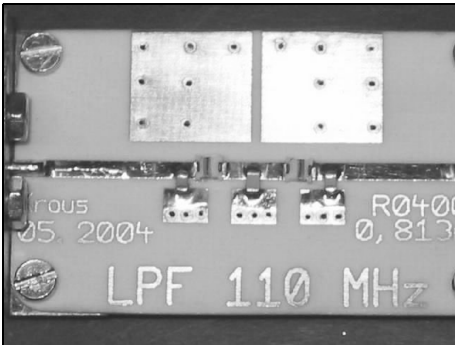


Fig 17: Picture of PCB fitted with SMD coils.

3.5. Helical version with 150MHz limiting frequency

Following my good experiences with helical coils in obtaining very low values of transmission attenuation, I wanted to construct a low pass filter of a similar nature, to carry out measurements in the 2m band (144 to 146MHz). The limiting frequency was therefore increased to 150MHz and, following my previous experiences, the assembly proceeded at a spanking pace.

Stage 1:

I started by calling up the “fds.zip” program and making appropriate changes to the values; the results can be seen in Fig 18.

Stage 2:

Next came a simulation check of the filter results using PUFF. Everything went according to plan, so no diagram is needed here.

Stage 3:

The simulation circuit used, in accordance with Fig 7, was converted to 150MHz. Since the same circuit board is used, it is necessary only to adjust the coil and capacitance values (Fig 19).

Stage 4:

Now we carry out a simulation with PUFF. The PUFF screen used for this and the predictions for the transmission range can be seen in Fig 20. It certainly looks quite perfect, but please just take a look at the parts list: as the frequency rises, the circuit sections and the open-end extensions become more and more noticeable. So the component values had to be varied until:

- the maximum value of S11 for both “bumps” in the transmission range was once again approximately -16.4dB
- the ripple limiting frequency had risen again to exactly 150MHz. It had actually dropped by approximately 3 %

Stage 5:

The circuit board was now assembled.

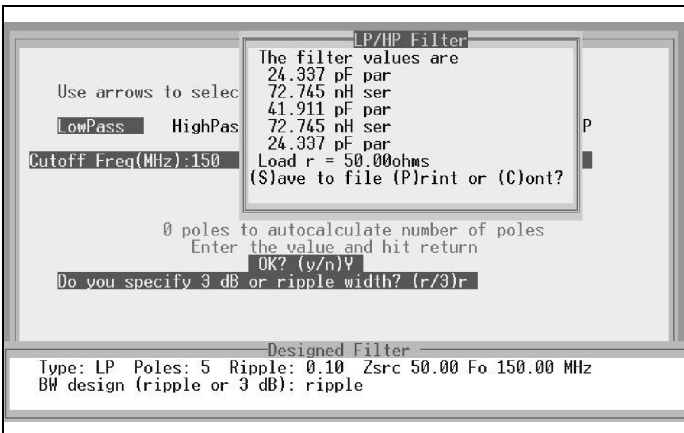


Fig 18: Parameters for 150MHz filter.

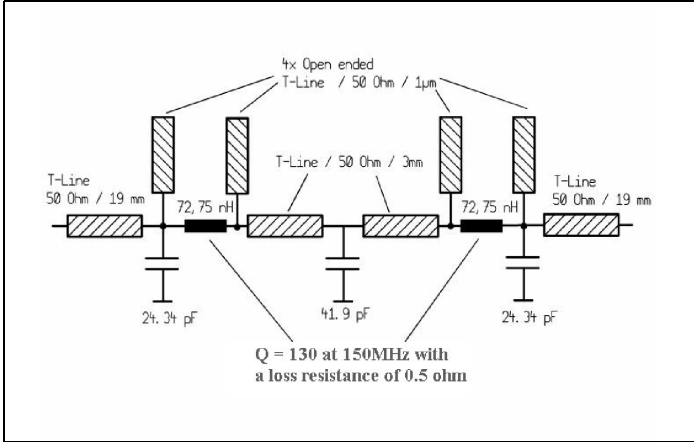


Fig 19: Circuit used to simulate 150MHz filter.

Instead of the 100nH helical coils, the smallest version obtainable was used (76nH) in the same type of housing – with the hope that the coils could be trimmed down to about 70nH by screwing the brass balancing core in hard. Guaranteed genuine NPO type components were soldered onto the SMD capacitors. The necessary capacitance values were obtained in the following manner:

- 24.18pF consisting of 22pF parallel 2.2pF

- 40.78pF consisting of 39pF parallel 1.8pF.

Stage 6:

Now, of course, the anticipation was high before the measurement.

Fig 21 shows the behaviour in the transmission range. It can easily be seen that the hope that the inductances could be “trimmed down” was in vain as, firstly, the limiting frequency is visibly too low, and, secondly, the ripple value has increased. And yet the design concept was absolutely correct. Anyone who com-

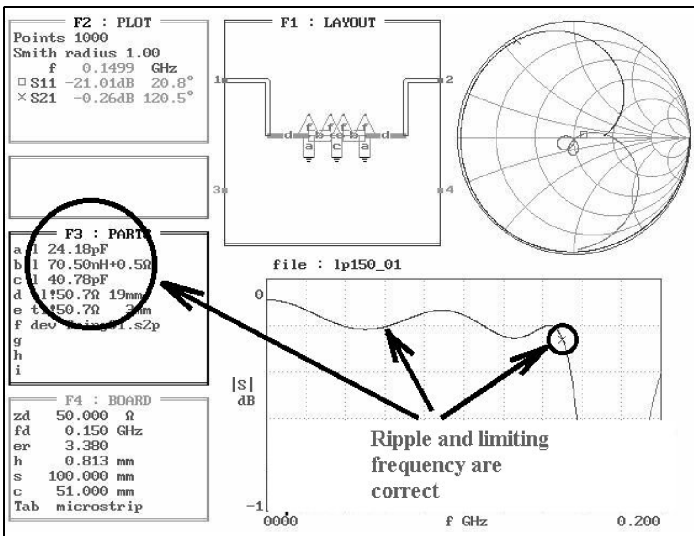


Fig 20: PUFF simulation of the 150MHz filter.

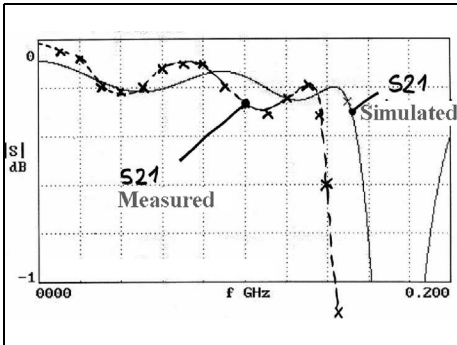


Fig 21: Measure response of the 150MHz filter.

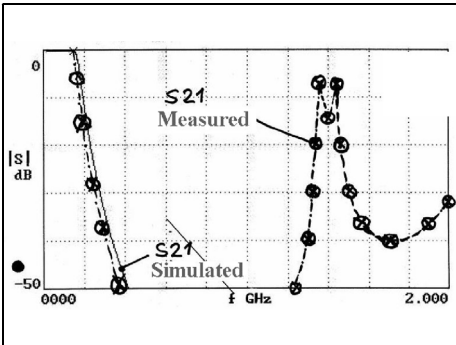


Fig 22: 150MHz filter stopband measurement.

compares the basic attenuation paths of the simulation and the measurement and carries out the calculations will quickly recognise that they are almost identical. And if precisely matching coils had also been used and correctly balanced, then we would have obtained the correct limiting frequency and the ripple value measured would have matched the predicted level.

We might prefer to say nothing about the stop band range, as up to 1GHz there are a few problems. But then come the tiresome line effects of the coils, as we already know from the 100MHz helical version (Fig 22). A photo of the board is

not required, since it looks exactly the same as in Fig 10. For that very reason you should make a written note of the new limiting frequency on the PCB.

(To be continued)

X. Literature references for Part 1

- [1] An interesting program APLAC, Gunthard Kraus, VHF Communications 2/2003, pp 90 - 105

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Gunthard Kraus, DG8GB

Internet Treasure Trove

All about the Smith chart

You often need to print out a Smith diagram quickly in order to indicate or monitor something. On this homepage you can find everything (but everything!) on this subject. It doesn't matter whether you're looking for a tutorial or printed blanks or tools or books or other web-sites: it's all here somewhere.

Address:
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The MINT Project

Here you can find information on the technology behind a modern radio telescope or radio interferometer for measuring the CMB (Cosmic Microwave Background). The receiver and evaluation technology is very interesting, but then the links to appropriate CMB tutorials etc. are good fun as well.

Addresses:
<http://background.uchicago.edu/>
<http://cosmology.princeton.edu/mintweb/instrument/>

Collection of antenna design software

Just another little collection of CAD programs for antennas. Things keep turning up here that you didn't know before.

Address:
http://www.dxzone.com/catalog/Software/Antenna_analysis/

The Designer's Guide

This link's title doesn't tell you all there is to be found there: the best plan is to search in every nook and cranny and open every file. Some very nice application notes!

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Summitek

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And the publications available for downloading are on a suitable generous scale. And of course, we can't do without an IM calculator for doing calculations at home.

Address:
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Millitech

Here's something for people who work with the higher frequencies in the millimetre wave range. The link in question leads to some interesting technical articles and application notes.

Address:
<http://www.millitech.com/cadiv/technotes.htm>

LC Filter Design

Anyone who has to develop an LC filter in a big hurry usually turns to our tried and tested "fds.zip" or "Faisyn". For those who are in a real hurry but don't have these programs installed on their computers, there's this online calculator. You just key in the data and press <Enter> - and you've already got the component values, the circuit and the simulated attenuation path on the screen. Slightly simplified, it's true, but the readings are absolutely correct. Do please try it out.

Address:
<http://www-users.cs.york.ac.uk/~fisher/lcfilter/>

Besser

Besser ("better") – it sounds like an incentive for someone's own work ethic, and in fact it does have something to do with that. What's actually behind the name is a company that is mainly active in the field of further education in the RF and communications technology area. Not only are these people kind enough to let you download quite a lot of good calculators for HF engineering, but there are also first-class articles on offer as "BesserBits" ("Better Bits") covering various subjects (DSP / noise measurement in mixers, etc.).

Address:
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1995	Y	Y	Y	Y
1994	Y	Y	Y	Y
1993	Y	P	Y	P
1992	Y	Y	Y	Y
1991	P	P	P	P
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Bernd Kaa, DG 4 RBF

Centre frequency counter for HP 8565A and HP 8569 spectrum analysers

Spectrum analysers are a subject of perpetual interest for DIY radio amateurs, but then they do make it possible to take a good look at what would otherwise be the concealed world of high frequencies. Moreover, even the popular spectrum analysers HP 8565A, HP 8569A and HP 8569B are slowly becoming available at prices that are within the means of radio amateurs.

1. Introduction

In general, HP 8565 and HP 8569 spectrum analysers from Hewlett Packard display data easily usable by radio amateurs. The centre frequency display for these units is a digital display that indicates the frequency in MHz. However, this display is decidedly imprecise and, according to the manufacturer's specification, deviations of several MHz are normal (± 25 MHz at 12.5 GHz!). Nowadays, you really do expect a more accurate frequency display from high quality measuring instruments. In order to overcome this problem, the author has developed a small additional circuit that can be added on without interfering with the analyser and displays the precise centre

frequency with a resolution of 0.1MHz.

This circuit uses genuine frequency counting. In addition, a frequency marker (luminous spot) is generated on the screen, that represents the exact point for the frequency shown in the display. Not only can this luminous spot be used to determine the precise centre frequency, but you can even highlight every spectrum line on the display and then read off the associated frequency on the LC display.

The block diagram of the circuit for the centre frequency counter can be seen in Fig 2.

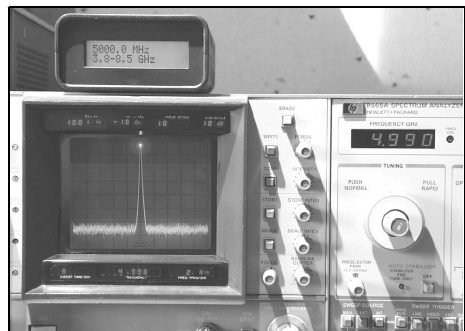


Fig 1: Picture of Centre frequency counter fitted to an HP 8565A spectrum analyser.

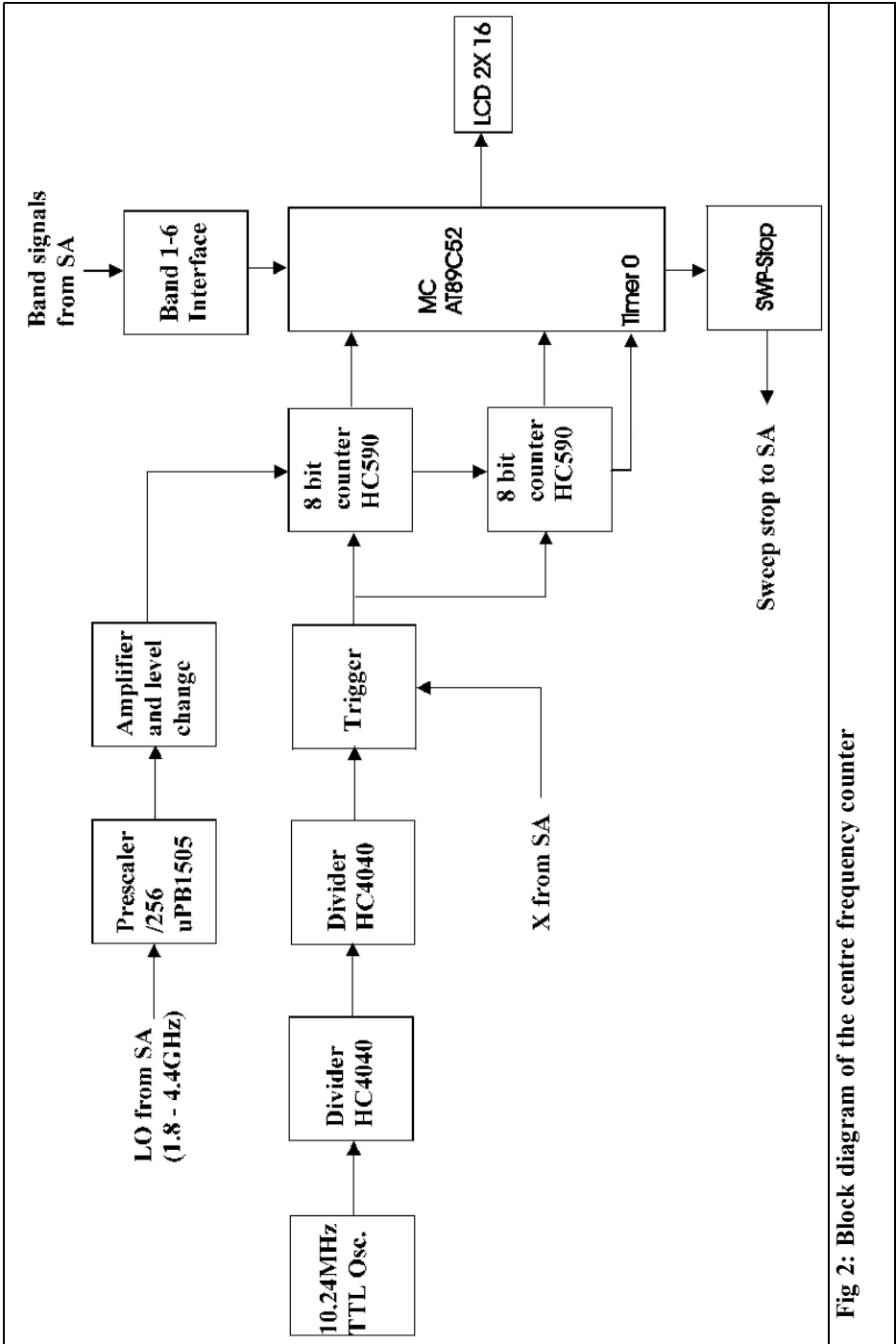


Fig 2: Block diagram of the centre frequency counter



2.

Circuit description

In principle, we are dealing here with a frequency counter with a fixed gate time, triggered by the sweep. The gate time is generated from a 10.24MHz quartz crystal by two binary dividers, IC3 and IC4 (74HC4040). The LM358 operational amplifier is wired up in such a way that it is always triggered at a specific point of the sweep, depending on the setting of the R7 trimmer (+ the R15 potentiometer). Connecting this to the two IC's, IC7 and IC5, means the gate of the downstream counter IC1, IC2 (74HC590) stays open from the trigger point for the gate time. At the same time an interrupt is triggered which starts additional processing by IC9, an AT89C52 microcontroller with internal program memory.

To be able to count the frequency of the local oscillator (1.8GHz to 4.4GHz), the μ PB1505 prescaler (IC11) is needed, it divides the frequency by 128. The maximum frequency of the LO, 4.4GHz, is at the upper limit of the μ PB1505, so rather more input power is required (approximately +3dBm) to ensure that it operates correctly. The downstream 74HC00 gate (IC12) is used to change the level and controls the clock input of the first counter IC (IC1).

These spectrum analysers from HP are equipped with 3 button operation, when the frequency is counted there is a problem because, depending on the bandwidth selected, the sweep speed of the analyser is automatically matched. That would lead to a difference in the counter reading based on the sweep speed. In order to solve this problem, the sweep of the analyser is continued for as long as the gate of the frequency counter is open. This gives a correct counter reading and simultaneously displays the very useful frequency marker on the screen.

The HP spectrum analysers are designed for stopping the sweep. To this end, they have a special connection at the rear (AUX B / pin 7), which continues the sweep as long as it receives a voltage of approximately 20 - 25V. This is controlled through the transistor, T1. It is a Mini PROFET (2), which makes it very simple to switch a voltage of approximately 20V with a control voltage of only 5V. The microcontroller makes sure that the display is not updated too often at very low sweep speeds since several sweep passes are required for a measurement.

3.

The microcontroller

The software is stored in the microcontroller, so the system does not need an EPROM for program memory. Only a serial EEPROM is provided, to store the individual set-up data. Here the first and second intermediate frequencies of the analyser can be stored, so that the correct centre frequency is displayed.

The HP spectrum analysers, HP 8565 and HP 8569, have a first intermediate frequency of 2,050MHz, which can, however, deviate by about 100kHz, and a second intermediate frequency of 321.4MHz. In band 1 (0.01 - 1.8GHz) the first intermediate frequency is used. All additional bands use only the second intermediate frequency.

The spectrum analyser has switching outputs for the individual bands, so that the microcontroller can display the correct centre frequency, up to 22GHz, depending on the band. The combination of the two resistance arrays (R14, R17) and the 6 individual resistances forms the interface for the signal voltages of the analyser (-15V and +15V).



Table 1: Inter-relationships of intermediate frequencies.

Band	Frequency (GHz)	IF (MHz)	Harmonic Mode
1	0.01 - 1.8	2050	-1
2	1.7 - 4.1	321.4	-1
3	3.8 - 8.5	321.4	-2
4	5.8 - 12.9	321.4	-3
5	8.5 - 18	321.4	+4
6	10.5 - 22	321.4	+5

4. Set-up

Temporary key switches can be linked to connections K9, K10 and K11, in order to do the once-off set-up adjustments of the first and second intermediate frequencies. The data then remains permanently saved in the EEPROM (IC8). The key switch on K9 acts as the enter (“Return”) key; the key switches on K10 and K11 act as the UP and DOWN keys to set the two values.

When the “Return” key (K9) is activated, the set-up program is called up and the

first intermediate frequency can be set using [UP/DOWN]. The selected value is then confirmed using “Return”. The second intermediate frequency then appears, and is set and confirmed in the same way. A third, corrective value can then be entered, in order to match the circuit to any possible deviations of the analyser.

The key switches need not necessarily be removed, they are needed only once at the beginning and subsequently only very rarely, in order to balance out, for example, temperature drift or long-time drift. One advantageous solution is to position the keys at the rear of the equipment.

Table 1 shows the inter-relationships of the intermediate frequency, the band and the harmonic used when the frequency is synthesised in the analyser.

5. Assembly

The complete circuit is accommodated

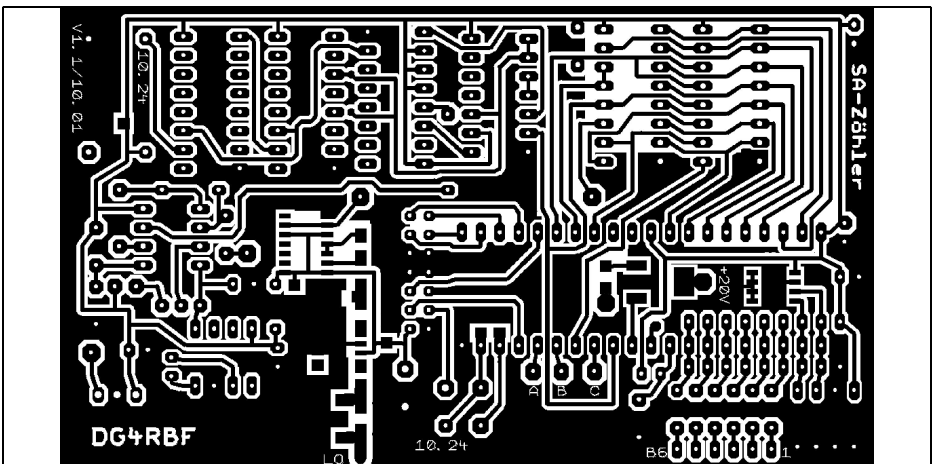


Fig 4: PCB layout for centre frequency counter. Can be downloaded from [1].

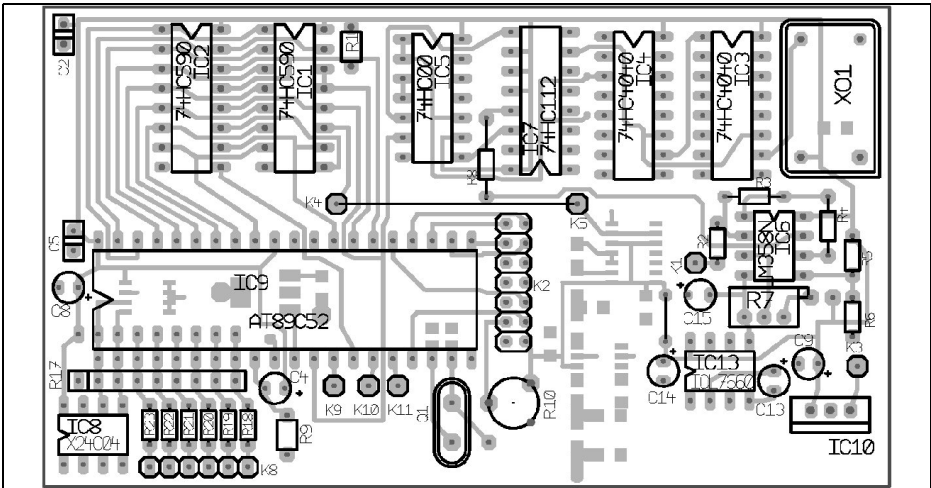


Fig 5: Top side component layout of centre frequency counter.

on a single sided circuit board (109mm x 62mm). The circuit board layout is shown in Fig 4 or can be downloaded from the author's homepage [1] as a PDF file. The component diagram for the top side can be seen in Fig 5.

The EEPROM (IC8) can be mounted on the top side as a normal (DIP) IC or, optionally, on the bottom side as an SMD IC. The XO1 TTL oscillator can be dispensed with if a 10.24MHz quartz crystal is used for Q1 and the signal of XTAL1 (pin19 at IC9) is fed to CLK (pin10 of IC3). It would also be possible to use only the 10.24MHz TTL oscillator and then provide the microcontroller with this frequency at pin 19.

A 40 pin precision socket should be fitted for the microcontroller. The two bridges on the top side are created using a piece of insulated jumper wire. The component drawing for the bottom side can be seen in Fig 6. The SMD components and the R14 resistance array are mounted here.

When mounting the components, take precautions against static charging.

The display is connected to K2. This connection is wired in such a manner that a two row connector using the IDC method of termination can be used for a simple connection.

6. Power supply

The circuit requires approximately +9V (+12V) for operation and approximately +20V to +25V control voltage for the SWP-STOP function. The voltages can be supplied by an external power source (which means it is not necessary to interfere with the analyser), or they can be extracted from the analyser. The two cables soldered on to provide a power supply directly from the analyser (+10V und +20V) can be seen in Fig 7.

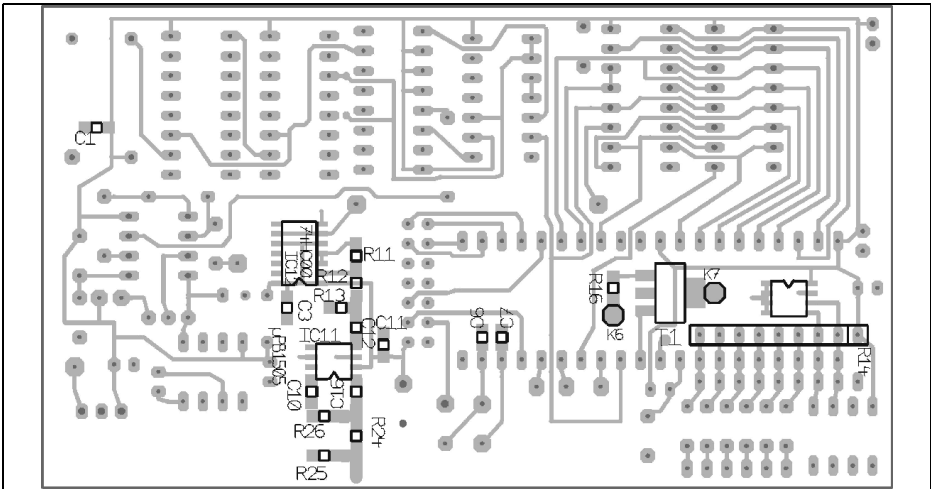


Fig 6: Bottom side component layout of centre frequency counter.

7.

Connection to analyser

First LO:

(1.8 – 4.4 GHz) At the rear of the analyser, there is a BNC or SMA connection for the first LO. About +5 to +8dBm are available here. An attenuator with a value of approximately 3 dB is provided at the input of the prescaler (μ PB1505) and can also be appropriately adjusted.

Band switching and sweep stop:

All connections required for this are likewise at the rear of the equipment and are available at the terminals AUX A and AUX B. The configuration of these connections can be seen in Fig 8. The switching outputs for the individual bands, pins 1-6 at AUX B, are connected to K8 (bands 1-6). The control voltage which is required to continue the sweep is available at K6 and is connected to the analyser at pin 7 / AUX B.

If no original plug is available for these connections, a standard SUB-D computer plug can be converted. To do this the tin shield around the edges of a 15 pin

connector is removed and the superfluous pins are squeezed out. A plug converted in this way for the connections to AUX B can be seen in Fig 9.

8.

Putting into operation and calibration

If the circuit is to be used only for precise centre frequency displays, then the R15 potentiometer can be dispensed with and the connections are bridged on the circuit board. If everything has been connected up correctly, a start message should appear on the LC display when the equipment is switched on.

The frequency is then displayed in the first column and the selected band is shown in the second column. A luminous spot (marker) appears on the screen of the spectrum analyser and is positioned in the centre of the screen with the help of the R7 trimmer. If you use an HP8569, switch off the digital image memory to

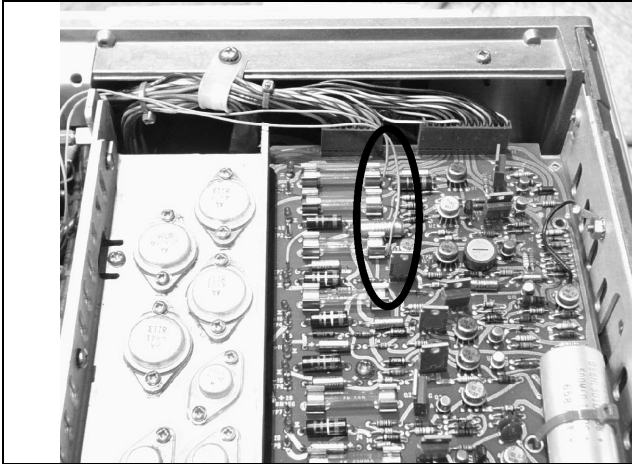


Fig 7: Power supply connections to power the supply in the spectrum analyser.

do this, as this makes the marker easier to see.

For the centre frequency to be displayed correctly, the first and second intermediate frequencies still have to be entered as described under SET-UP. The first intermediate frequency can be measured at the LO output of the analyser if you set it to the zero line, using the ZERO-SPAN mode. Use the internal 100MHz test signal for fine calibration.

We use band 2 (1.8 to 4.1GHz) to check the second intermediate frequency. A suitable signal in this range can be used for this, or we take a harmonic of the 100MHz test signal.

A correction factor is also available for discrepancies in the upper bands. To calculate the frequency, multiply this correction factor by the value of the harmonic used e.g. band 6: Frq * correction factor *5

9. Parts list

IC1, IC2	74 HC 590
IC3, IC4	74 HC 4040
IC5,	74 HC 00
IC6	LM 358N
IC7	74 HC 112
IC8	X24C04
IC9	AT 89C52
IC10	7805
IC11	PB 1505
IC12	74 HC 00
IC13	ICL 7660
T1	BSP452
R1,3	100kΩ
R2,4	22kΩ
R5-7	100kΩ
R8	47kΩ
R9,11	10kΩ
R10	5kΩ
R12	270Ω
R13	68kΩ
R14	8x6.8kΩ

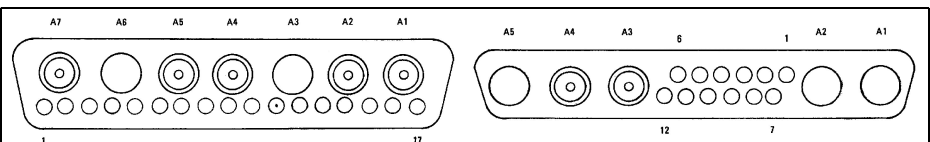


Fig 8: Details of the AUX A (left) and AUX B (right) connectors fitted at the rear of the spectrum analyser.



Table 2: Pin connections for AUX A (J1).

Pin No.	Wire Code	Signal
1-6	-	NC
7	925	Normaliser, Control In
8 - 10	-	NC
11	968	Horizontal Sweep Out
12 - 17	-	NC
A1	978	Vertical Output
A2	957	Pen Lift
A3	-	NC
A4	923	Normaliser X In
A5	924	Normaliser Y In
A6	-	NC
A7	926	Normaliser Blank In

- R15 POTI
- R16 22kΩ
- R17 8x4.7kΩ
- R18-23 10kΩ
- R24 18Ω
- R25,26 270Ω
- C1-3,5 100nF
- C4 3.3μF
- C6,7 27pF
- C8,9 10μF
- C10,11 1nF
- C12 22nF
- C13-15 10μF
- C16 10pF
- BU1 SMA or SMC
- LCD1 LCD 16X2
- Q1 10,24 MHz
- XO1 TTL-Osc.10.24 MHz
- K1 K1x1

Table 3: Pin connections for AUX B (J2)

Pin No.	Wire Code	Signal
1	913	B1
2	914	B2
3	915	B3
4	916	B4
5	917	B5
6	918	B6
7	95	Stop Swp
8	904	Ytf Mod
9	94	Marker
10 - 12	-	NC
A1 - A2	-	NC
A3	2 (coax)	Yto Freq Analog
A4	3 (coax)	Ttf Freq Analog
A5	-	NC

- K2 K2x7
- K3-7 K1x1
- K8 K1x6
- K9-11 K1x1

10. References

[1] Homepage for Bernd Kaa, DG4RBF: www.dg4rbf.de

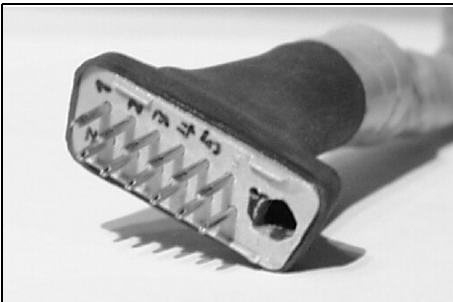


Fig 9: Modified SUB-D connector.

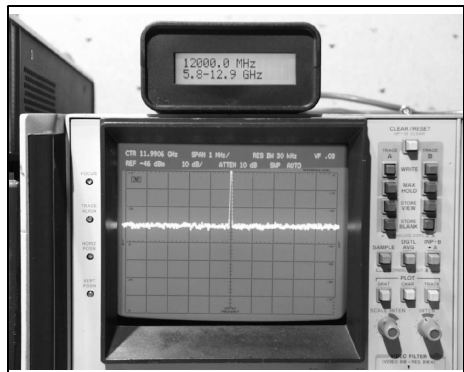


Fig 10: Centre frequency counter in use with an HP 8569.



Wolfgang Schneider, DJ8ES

DDS oscillator for QRP experiments on short wave and 6m band

A large part of radio amateur DIY activities are related to the short wave range. So a universal, programmable DDS oscillator assembly is described below. A freely programmable oscillator can be especially helpful for experiments on these frequencies, whether it be for a receiver, a transmitter or even a complete transceiver. All ranges required are covered, right up to the 6m band. The standard centre frequency offsets for the intermediate frequency and the SSB and CW modes are programmable. Depending on the mixer and circuit concept used, the oscillator frequency can be selected in minimum steps of 1Hz, and the oscillator level can be selected within the range from +13dBm to -6dBm.

1. General

Modern DDS components (Direct Digital Synthesiser) can supply sinusoidal output signals extending far into the VHF range using simple external components. One possible use for this type of frequency generator is, for example, as an oscillator for a transmitter or receiver in the HF and VHF ranges. It is advantageous to use a

microcontroller to control a DDS; this makes it easy to obtain the required frequency tuning and the display of the frequency on an LC display.

It is necessary to filter the output signal by means of a high quality low pass filter. In addition to the desired frequency, a DDS module also generates alias products, such as: clock frequency + carrier frequency and clock frequency - carrier frequency. This also applies to frequencies that are double, treble... the clock speed. This results in an endless number of mixed products. This cannot be prevented except by having a low pass filter which is as sharp as possible at 30 to 40% of the clock frequency.

The alias products generated by the DDS module cannot be completely suppressed, even using a low pass filter. Spurious effects generated by the system still remain, with a maximum level of up to -50dBc. One elegant combination pairs a fine tuned DDS oscillator with a PLL oscillator, although the circuit complexity is increased, especially when the complete short wave range and the 6m band have to be covered by a single module, which has to include various intermediate frequencies as well. It is however an extremely interesting solution if only one frequency range or amateur radio band is to be covered, with a fixed intermediate frequency offset.

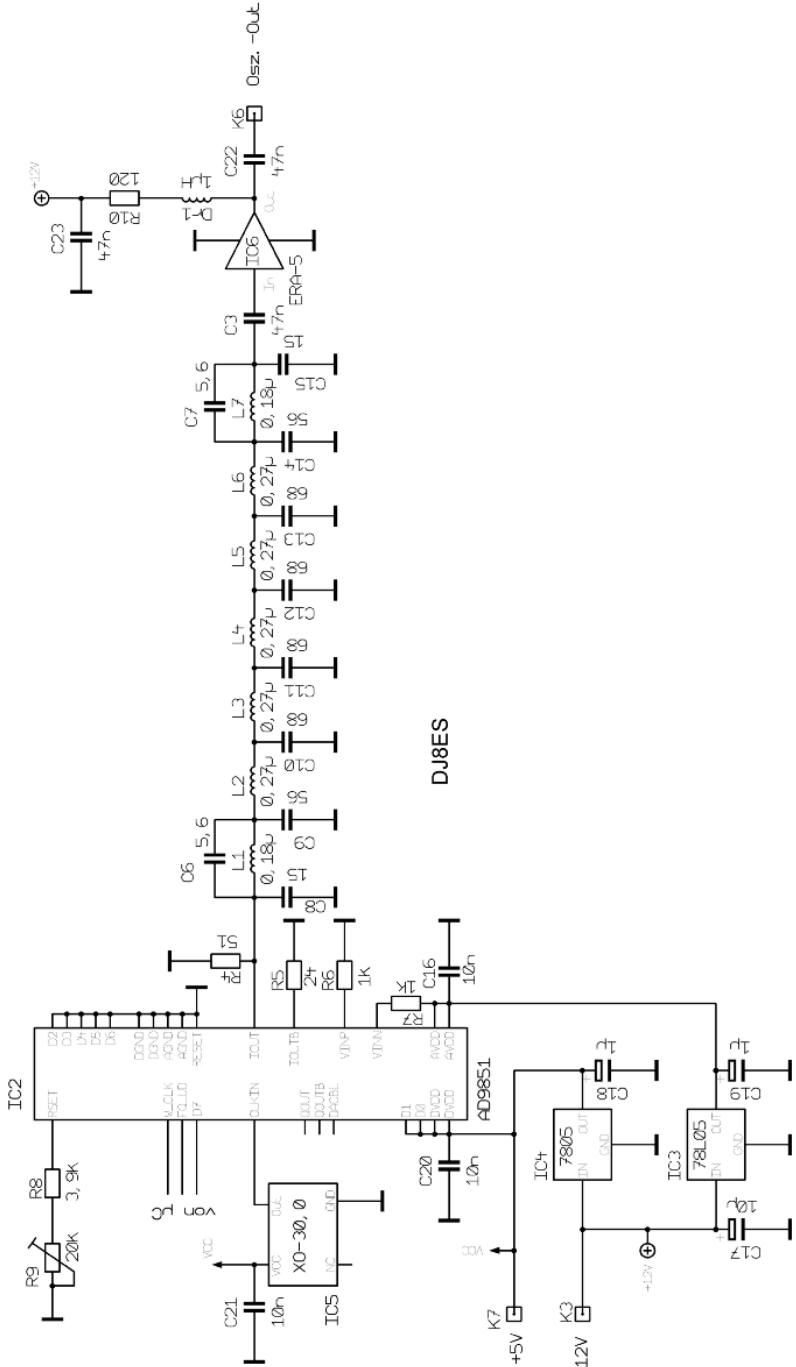


Fig 1: Circuit diagram of DDS oscillator and low pass filter.

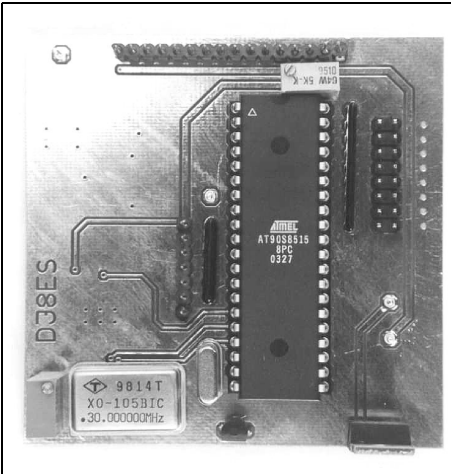


Fig 2: Picture of the assembled DDS oscillator.

2.

Circuit description

The core of the circuit is the integrated DDS module AD9851 (IC2) from Analog Devices. It supplies a sinusoidal output voltage between 0Hz and approximately 33% of its clock frequency at 50 Ohms. So with a 30MHz clock oscillator, with an internal sextupler activated by software to 180MHz, the maximum usable frequency is approximately 65MHz.

A 17 pole, 50 ohm Butterworth type low pass filter is used to give the required ripple and harmonic suppression and is shown in Fig 1. All values are taken back to standard values. The cutoff frequency is approximately 65MHz.

The output of the oscillator assembly can be set by means of a precision trimming potentiometer (20kOhm, R9) at the RSET input of the DDS (IC2). The available level range runs from +13dBm (20mW) down to -6dBm (250µW). Thus commercially available mixers, both

passive and active, can be controlled at the required power.

The whole DDS oscillator is controlled by a AT9058515 microcontroller (IC1) from ATMEL (Fig. 2). Its 40pin package contains enough ports to control the DDS module (IC2, AD9851) through a shaft encoder and the LC display (2 x 16 characters). The desired frequency band, the intermediate offset and the amount of SSB/ CW deviation are set at ports PA0 to PA7 (Fig. 3).

The LC display is connected up through the plug and socket connection K4 and is 1:1 wired. The following pin configuration applies to the shaft encoder, even if it is connected through a plug and socket connection:

K5	µC	Function
1	-----	+5V (Vcc)
2	PD0	not wired up
3	PD1	not wired up
4	PD2	shaft encoder channel A
5	PD3	shaft encoder channel B
6	PD4	key for frequency step
7	-----	GND

After resetting the microcontrollers via "Power on Reset" or the (low active) reset port, the microcontroller always sets the DDS oscillator to the frequency band programmed at the port, PA0 to PA7, taking into account the intermediate frequency and the SSB/CW deviation. If anything changes at this port, the microcontroller reacts at once and changes the output frequency. Thus, for example, various deviations for TX/RX mode can be implemented.

The tuning step width for the shaft encoder is defined, following switching on as 100Hz per step. The interval can be switched to 1Hz, 10Hz, 100Hz or 1kHz

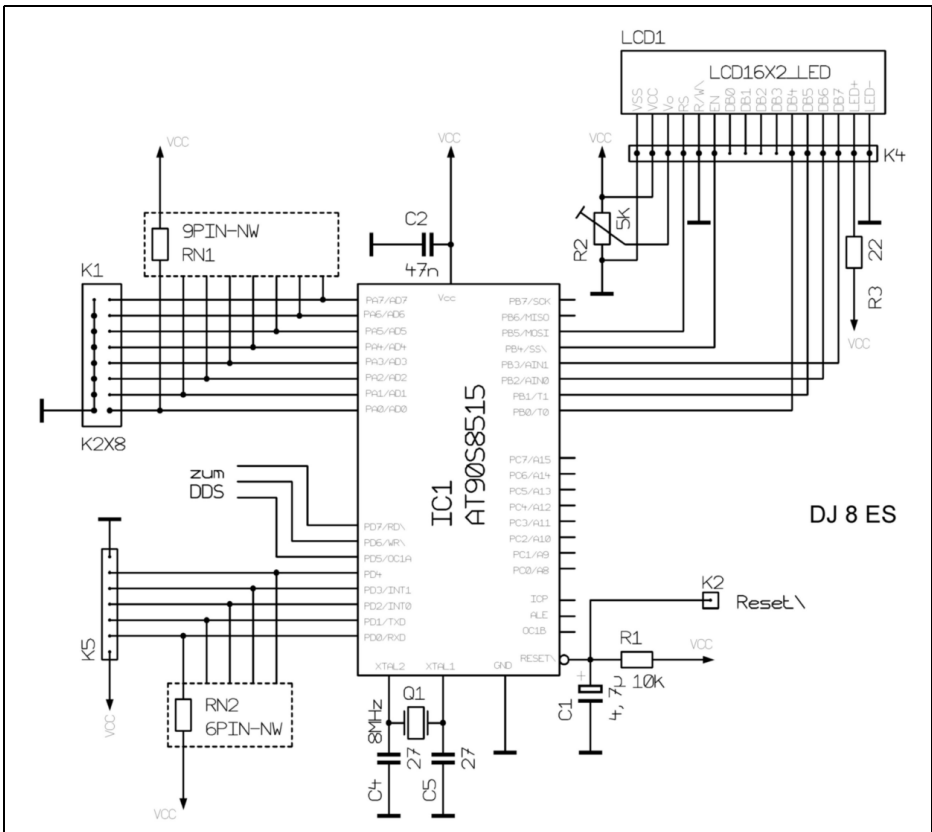


Fig 3: Circuit diagram of the microcontroller for the DDS oscillator.

by pressing a key.

3. The operational software

Frank Peter Richter (DL5HAT), has done a splendid job of writing the software for the microcontroller (IC1, AT90S8515) from ATMEL in Basic. AVR and BASCOM-AVR are offering a special Basic compiler tailor made for this controller family that can be used with the Windows operating systems.

All data relevant to operation are represented on a 2 line LC display, controlled by means of the software. Here we should select the current frequency band, the present operating frequency and the selected mode, USB/LSB or CW (see also Fig. 10).

These operating modes are programmed by means of jumpers or DIP switches through the connector block K1. The individual options can be found in Table 1 (programming table).

If the DDS oscillator is to be used, for example, in a 20m transceiver with the intermediate frequency 10.7MHz, then

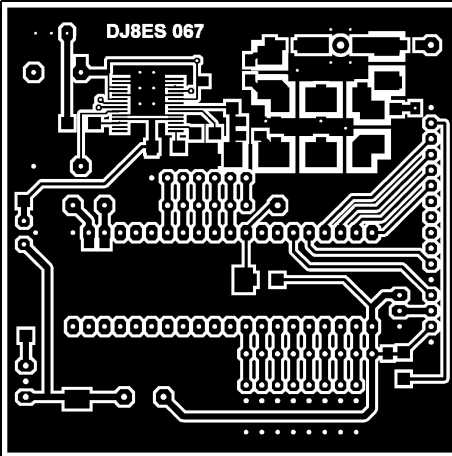


Fig 4: Bottom side of DDS oscillator PCB.

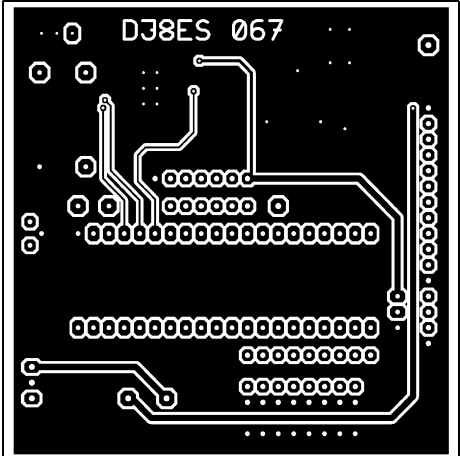


Fig 5: Top side of DDS oscillator PCB.

the following setting is necessary at the port for radio mode in USB:

PA7	PA6	PA5	PA4	PA3	PA2	PA1	PA0
0	1	0	0	1	1	1	0

where:

- 0 = Jumper present, low level at port PA(n)
- 1 = No jumper, high level at port PA(n)
- X = Redundant, no influence

Table 1: Programming details for DDS oscillator.

PA7	PA6	PA5	PA4	PA3	PA2	PA1	PA0	Band or IF	Comment
0	0	0	0	X	X	X	X	1.800 - 2.000MHz	160m
0	0	0	1	X	X	X	X	3.500 - 3.800MHz	80m
0	0	1	0	X	X	X	X	7.000 - 7.100MHz	40m
0	0	1	1	X	X	X	X	10.000 - 10.150MHz	30m
0	1	0	0	X	X	X	X	14.000 - 14.350MHz	20m
0	1	0	1	X	X	X	X	18.000 - 18.500MHz	17m
0	1	1	0	X	X	X	X	21.000 - 21.500MHz	15m
0	1	1	1	X	X	X	X	24.000 - 24.500MHz	12m
1	0	0	0	X	X	X	X	28.000 - 28.500MHz	10m
1	0	0	1	X	X	X	X	28.500 - 29.000MHz	10m
1	0	1	0	X	X	X	X	29.000 - 29.500MHz	10m
1	0	1	1	X	X	X	X	29.500 - 30.000MHz	10m
1	1	0	0	X	X	X	X	50.000 - 50.500MHz	6m
1	1	0	1	X	X	X	X	50.500 - 51.000MHz	6m
1	1	1	0	X	X	X	X	51.000 - 51.500MHz	6m
1	1	1	1	X	X	X	X	51.500 - 52.000MHz	6m
X	X	X	X	X	X	0	0	f = f - 800Hz	SSB / CW *
X	X	X	X	X	X	0	1	f = f + 800Hz	SSB / CW *
X	X	X	X	X	X	1	0	f = f + 1500Hz	SSB / CW *
X	X	X	X	X	X	1	1	f = f - 1500Hz	SSB / CW *
X	X	X	X	0	0	X	X	no IF, SSB / CW offset	IF definition
X	X	X	X	0	1	X	X	n IF, any SSB / CW offset	IF definition
X	X	X	X	1	0	X	X	f = f + 9.0MHz ± SSB/CW	IF definition
X	X	X	X	1	1	X	X	f = f + 10.7MHz ± SSB/CW	IF definition

* Together with IF offset of 9 or 10.7MHz will mix to give desired sideband. Will be shown on display.

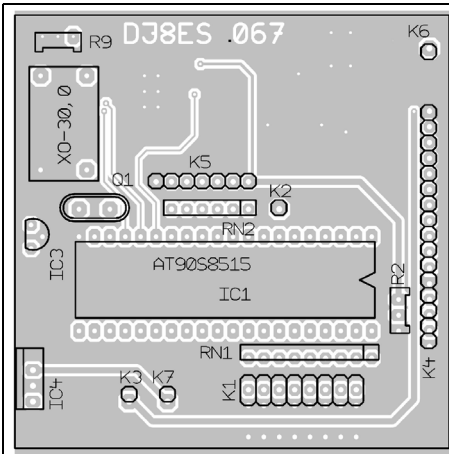


Fig 6: Component layout for DDS oscillator.

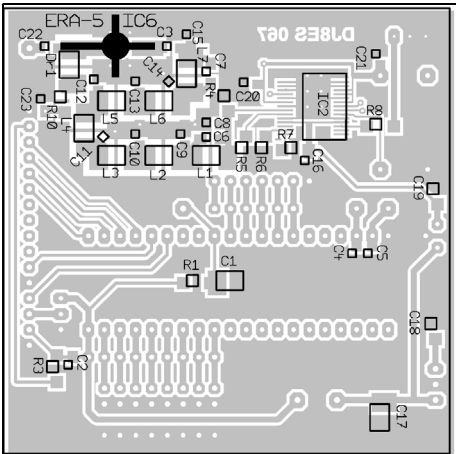


Fig 7: SMD component layout for DDS oscillator.

It is important to know that the DDS assembly is designed for operation with “high value” intermediate frequencies. That means that the oscillator always oscillates above the wanted frequency, offset by the intermediate frequency. This results in a sideband reversal (USB/-LSB) for the wanted signal!

The frequency setting (or frequency tuning) is carried out by means of a shaft encoder. The prototype used a

mechanical unit with 32 steps per revolution. However, other optical shaft encoders can also be used. Tuning is controlled via “Interrupt”. The step width (which can be switched) can vary between 1Hz, 10Hz, 100Hz and even 1kHz per step.

The individual start frequencies for each band are determined at the start of each band (e.g. 14,000,000Hz). This also applies to the frequency of the clock oscillator of the DDS module and, of course, to the text on the LC display. There is space for individual adjustments here, if applicable.

4. Assembly instructions

The entire DDS oscillator circuit is mounted on a double sided copper coated printed circuit board (DJ8ES 067) with the dimensions 72mm x 72mm. The assembly thus fits into a standard 74mm x 74mm x 30mm tinplate housing.

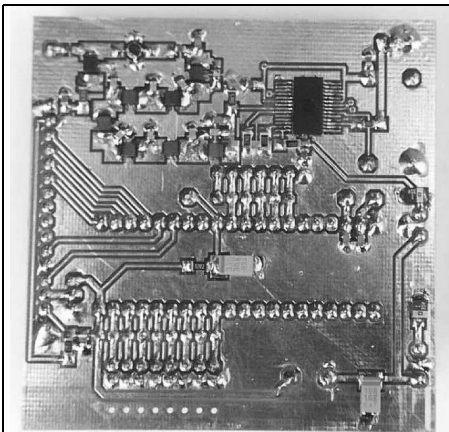


Fig 8: Picture showing SMD components.

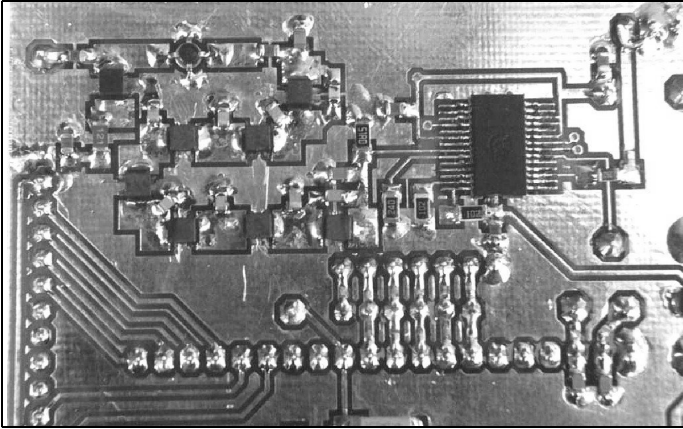


Fig 9: Closeup view showing DDS IC and ERA-5 MMIC.

Figs 4 and 5 show the layout of the circuit board. As some of the tracks in the vicinity of the DDS module are very delicate, and a lot of feedthroughs are needed, it is recommended that you use a professionally made circuit board.

The DJ8ES 067 board is assembled in no particular order, but the AD9851 DDS module (IC2) and the AT9058515 microcontroller (IC1) should not be fitted until the two 5V fixed voltage regulators (IC3, IC4) are fitted and functionally tested.

Be careful when mounting the AD9851 DDS module (IC2). This IC is upside down. The IC should be positioned on the board and the “legs” compressed and soldered on to the board individually, one after the other, using a fine tip.



Fig 10: Picture of display.

To make it easier to read the LC display has a backlight. A 16 pin plug is used to connect the display. The connections are connected straight through. The shaft encoder is also connected to a corresponding plug. The frequency setting, intermediate frequency offset, etc. can be programmed either, as in the prototype, using jumpers, or using an 8 pin DIP switch, or using a ribbon cable with external switches.

The board should not be fitted into the housing until it has been fully built. First the SMA connectors and the 1nF feedthrough capacitor have to be inserted in the side walls for the 12V power supply. Appropriate recesses can be filed in the edge of the cover for the ribbon cable on the shaft encoder and the display.

The board is inserted in such a way that the 5V volt regulator (7805, IC4) lies flush with the top edge of the housing. They should be screwed together. A little heat conducting paste should be applied first. For good RF performance, the earth surfaces of the circuit board are soldered to the tinplate housing all round on both sides.



4.1. Parts list

IC1	AT90S8515 microcontroller
IC2	AD9851BRS DDS module
IC3	78L05 voltage regulator
IC4	7805 voltage regulator
IC5	30MHz TTL/CMOS oscillator
IC6	ERA-5 MMIC
Q1	Quartz, 8.0MHz, HC18-U
Dr1	Choke, 1μH, SMD
C1	Tantalum electrolytic capacitor, 4.7μF, 16V, SMD
C17	Tantalum electrolytic capacitor, 10μF, 16V, SMD
C18, C19	Tantalum electrolytic capacitor, 1μF, 16V, SMD
L1, L7	Fixed inductance, 0.18μH, SMD
L2 - L6	Fixed inductance, 0.2μH, SMD
R2	Precision spindle trimmer, 5kΩ, 64W model
R9	Precision spindle trimmer, 20kΩ, 64W model
RN1	Resistance array, 8x10kΩ, SIL 9-8 10kΩ, RM 2.54mm
RN2	Resistance array 5 x 10kΩ, SIL 6-5 10k, RM 2.54mm
K1	Contact socket terminal strip 2 x 8pin, RM 2.54mm
K4	Contact socket terminal strip, 16-pin, RM 2.54mm
K5	Contact socket terminal strip, 7-pin, RM 2.54mm
K2,3,6,7	Terminal pin, 1mm
1 x	Tinplate housing 74mm x 74mm x 30mm
1 x	LC display, 2 x 16 characters with back light
1 x	Mechanical shaft encoder (32 pulses/revolution)
1 x	SMA connector
1 x	1nF feedthrough capacitor, solderable
1 x	Precision IC base, 40-pin
1 x	Printed circuit board, DJ8ES 067

SMD resistances, type 1206:

R1	10kΩ
R3	22Ω
R4	51Ω
R5	24Ω
R6, R7	1kΩ
R8	3.9kΩ
R10	120Ω

SMD capacitors, type 1206:

C4, C5	27pF
C6, C7	5.6pF
C8, C15	1 pF
C9, C14	56pF
C10 - C13	68pF
C16, C20, C21	10nF
C3, C22, C23	47nF

5.

Putting into operation

After a visual check of the printed circuit board before the microcontroller and the DDS are fitted, the operating voltage of +12V is applied for the first time. Now the two 5V fixed voltage regulators can be tested with a digital voltmeter. Only then are the two remaining components mounted.

For the full function test, the LC display and the shaft encoder are connected up, together with appropriate plug and socket connections, using the socket terminal strips provided for the purpose. The RF output is connected to a mW meter. A frequency counter with a high input impedance should also be connected.

The current consumption for the entire circuit (including the shaft encoder and display) amounts to approximately 230mA. When the operating voltage is applied, the oscillator module immediately supplies the programmed frequency of, for example, 14.000 000MHz +/- any frequency offset

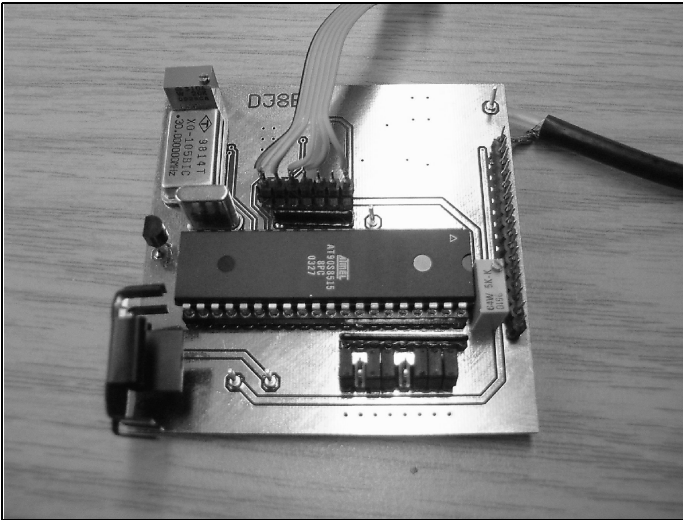


Fig 11: Completed DDS oscillator.

present.

Details on the programming of the frequency bands, the intermediate frequency offset and the possible SSB/CW deviations are given in Section 3. The output is set to the desired value using the precision trimming potentiometer, R9 (20k). The output variation range extends from max. +13dBm (20mW) to -6dBm (250 μ W).

Possible deviations from the output frequency of a few Hertz result from the normal imprecision of the 30MHz clock oscillator. Following a running in phase of at least a week, the module's frequency can be precisely measured using a suitable frequency counter and programmed in the microcontroller. In prototype the deviation amounted to precisely +90Hz.

No further calibration work is required, apart from setting the output level.

6. References

-
- [1] Wolfgang Schneider, DJ8ES, 1 to 65MHz DDS signal generator with switchable output level, VHF Communications 4/2003, pp. 226 - 235
 - [2] Wolfgang Schneider, DJ8ES, Direct digital synthesis using IC AD9850, UKW Berichte 2/1998, pp. 67 - 75



Alexander Meier, DG6RBP

A simple detector to estimate the immission from GSM mobile phone base stations: Part 2

Continued from issue 3/2004

The first part dealt with the principles of measurement and conditions. The following section describes the practical construction of a suitable measurement instrument.

2.

A broadband detector for GSM base stations

The broadband detector developed for the frequency range of GSM base stations is a compromise between feasibility, cost and precision. In addition to power density measurements, it is also extremely well equipped for tracking down concealed base stations using its analogue display. The technical data of the detector are shown in Table 1.

Thanks to a switchable filter at the input, the power density of the two GSM bands can be determined separately. Moreover, signals from other wireless communication services (e.g. radio, mobile phones) are suppressed to the greatest possible extent.

The measurement range covers three decades, from $1\mu\text{W}/\text{m}^2$ to $1\text{mW}/\text{m}^2$ and therefore covers most measurement re-

quirements. The measurement result is displayed on a moving coil meter. The accuracy of reading is sufficient for estimated measurements. An external attenuator between the antenna and the detector provides the option of expanding the measurement range to higher power densities. The detector is calibrated for a dipole with approximately 2dBi gain.

2.1. Circuit description

The block diagram of the GSM detector can be seen in Fig 1. The signal from the dipole passes through a 5 pole high pass filter (850MHz) to suppress unwanted signals, e.g. from very powerful radio stations. The desired frequency band (935 to 960MHz / 1805 to 1880MHz) is then selected using a switchable band pass filter. An amplifier balances the losses in the filters and raises the signal level a little before the signal is converted into a DC voltage using a logarithmic true power detector. The peak value of the power density, which varies with time, is determined using a peak detector.

To display the mean value, the time constant of the logarithmic detector can be switched. To display the measured value, a moving coil meter is used, controlled by an operational amplifier with an adjustable offset. The offset adjustment thus determines the lower

**Table 1: Technical data for broadband GSM detector.**

Frequency range	935 - 960MHz and 1805 - 1880MHz
Measurement range	$1\mu\text{W}/\text{m}^2$ to $1\text{mW}/\text{m}^2$
Display	Moving coil meter
Crosstalk attenuation between GSM bands	20dB
Typical accuracy	3dB
Measurement antenna	Dipole with 2.15dBi gain
Detector	True RMS
Power supply	9v battery
Current consumption	65mA
Battery indicator	LED for low battery
Housing	165mm x 102mm x 41mm

limit of the display ($1\mu\text{W}/\text{m}^2$).

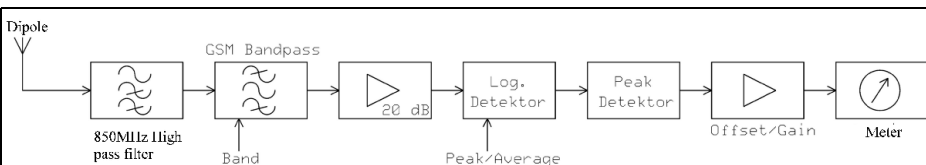
The circuit diagram of the GSM detector can be seen in Figs 2 & 3. The antenna signal is fed through a 5 pole high pass filter (850MHz), consisting of C1 to C3 and L1-L2, to the switch, U1. In combination with switch (U2), the switch, U1, is used to feed the signal through one of the two SAW band pass filters. This enables the immission on both GSM bands to be measured separately, and moreover the response to mobile phones suppressed due to their long duplex interval. There is no practical way to suppress the frequencies of cordless telephones using CT1+ (930 to 932MHz) or DECT standard (1880 to 1900MHz), as they are too close to the GSM frequency band. However, this is also an advantage, since the detector can be used on the many DECT base stations. The mean power density of the beacon signal from such a telephone is very small but the peak value can be determined. However, owing to the attenuation used in the frequency band above 1,880MHz, the absolute power density cannot be measured

precisely.

The integrated amplifier, type ERA-3 (U3) from Mini-Circuits, is used to balance the losses in the filters and raise the signal level a little. Its operating voltage is stabilised at 5V, so the amplification factor is independent of the battery voltage.

The asymmetrical output signal is then balanced by the transformer, TR1, at an impedance of 200 Ohms and fed to the logarithmic, modulation independent, (true power) detector (U4). Its output voltage (PIN 12) is proportional to the input level (dBm) at approximately 50mV/dB. The peak value of the output signal is determined using a peak value detector (U5A). The time constant of the detector can be extended to form a mean value, using the capacitor, C19.

The frequency response of the GSM detector was determined via the output (PIN 12) of the logarithmic detector and is shown in Fig 4. The ripple in the

**Fig 1: Block diagram of broadband GSM detector.**

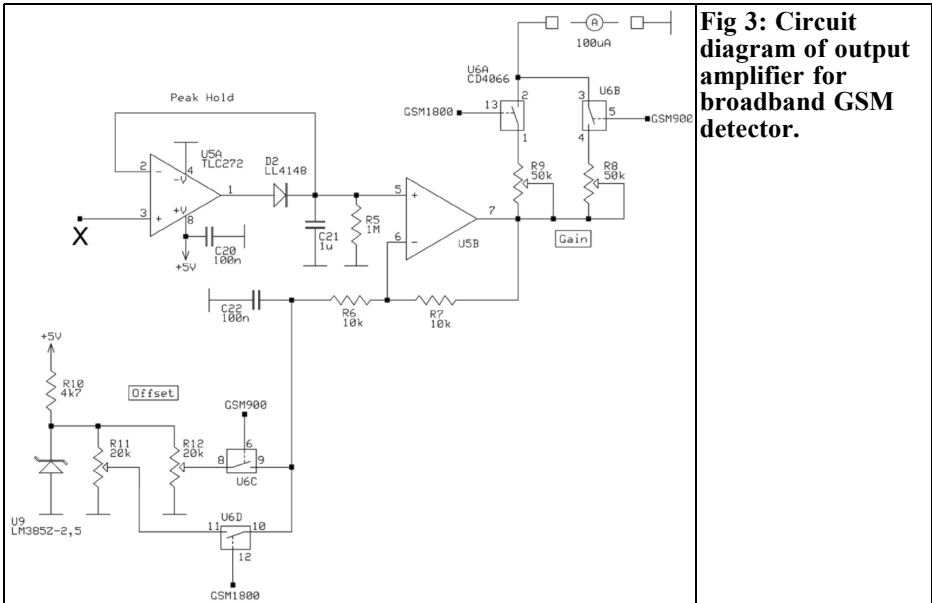


Fig 3: Circuit diagram of output amplifier for broadband GSM detector.

transmission range is approximately 2dB for both GSM bands and is predominantly due to the filter. The crosstalk attenuation between the two GSM bands is approximately 15 to 20dB and can be further improved by the selection of the dipole antenna.

The lower limit of the display is set using a DC voltage fed to amplifier U5B, this can be separately adjusted for each of the two bands using R11 or R12. Thus the output voltage from U5B (PIN 7) is 0V if the power density is below 1µW/m². The range for the moving coil meter is set by means of the trimmers, R8 and R9. The full scale deflection corresponds to a power density of 1mW/m². The scale for the moving coil meter is illustrated in Fig 5.

The power source is a 9V battery, so the equipment is also portable. The battery voltage (>5.2V) is stabilised at 5V using the low drop out voltage regulator (U8). The resistance splitter, R13/R14, is used to compare the battery voltage with a reference voltage in the voltage regulator. If the battery is run down, the level at

PIN 1 of U8 is taken to earth and the LED goes out as a battery indicator light. At 65mA, the current consumption for the circuit is unfortunately not particularly economical, so that the service life of the battery is only a few hours.

2.2. Assembly and calibration

First, the 74mm x 55mm x 30mm tinplate housing is drilled for the BNC connector and the feedthrough capacitors. Then all SMD components are mounted on the printed circuit board and the board is soldered into the housing. The layout of the circuit board is shown in Fig 6, and component layout in Fig 7. Next the wired components are added and the connections to the BNC connector and the feedthrough capacitors are made.

The entire assembly is easy to construct but SMD soldering experience is required for the logarithmic detector and the band pass filter.

Following an initial functional test, the circuit can be built into an aluminium housing and wired up to the display. The

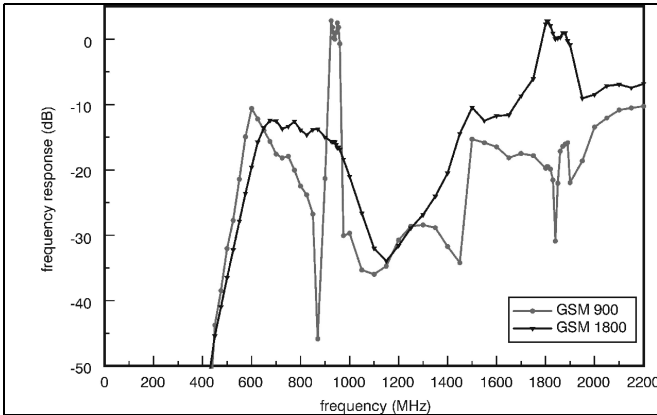


Fig 4: Frequency response of the detector for GSM 900 and GSM 1800.

equipment is then calibrated.

To calibrate the moving coil meter for a display in $\mu\text{W}/\text{m}^2$, there must be an inter-relationship between the power density and the received power of the measurement antenna. An antenna with a gain, G (dBi) and a power density, S (W/m^2), with a wavelength, λ (m), gives a received power, P_E of:

$$P_E = S \cdot 10^{\frac{G(\text{dBi})}{10}} \frac{\lambda^2}{4\pi} [W]$$

The unmodulated carrier of a standard signal generator is used for calibration, simulating a dipole antenna in a radiation field, S . The signal generator is directly connected to the GSM detector and a received power corresponding to the

power density is selected. For a dipole with approximately 2.15dBi gain, the inter-relationship is as shown in Table 2. In each case, the frequency is adjusted to the band centre (947.5MHz or 1,842.5MHz).

First the GSM 900 frequency band is calibrated. An output of -48.8dBm is selected from the signal generator. The trimmer, R12, is used to set the zero the output of U5B (PIN7), it is adjusted until the offset voltage for the operating amplifier is reached (a few mV). Thus, if a dipole antenna is connected and exposed to a power density of less than $1\mu\text{W}/\text{m}^2$, the output voltage of U5B corresponds to 0V.

Next an output of -18.8dBm is selected from the signal generator and a display of $1\text{mW}/\text{m}^2$ set. This balancing is repeated in the corresponding manner for the GSM 1800 frequency band.

If you want to track down concealed base stations using the GSM detector, a directional antenna must be used instead of the omni-directional dipole. The AS-LOG [3] from UKW Berichte can be used (Fig 8), this covers both GSM frequency bands. So that the power density displayed on the GSM detector is correct, an attenuator must be added. This compensates for any antenna gain,

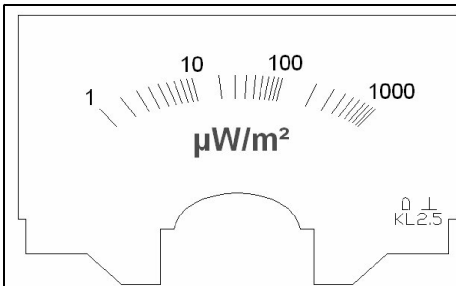


Fig 5: Scale of output meter.

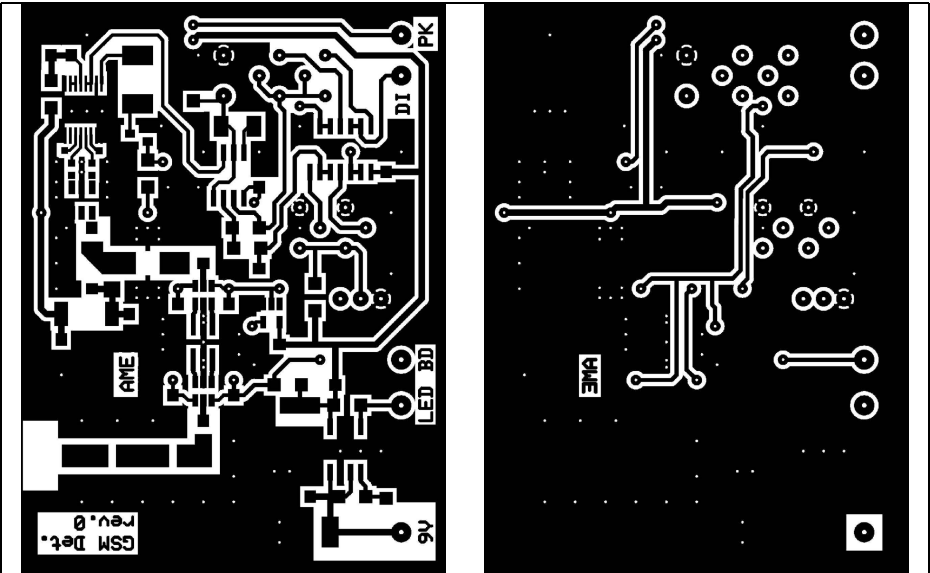


Fig 6: Printed circuit board layout for broadband GSM detector. Full size 71mm x 53mm x 1.5mm.

the antenna gain for the AS-LOG is approximately 7 to 10dBi.

A completed printed circuit board in a tinplate housing can be seen in Fig 9. In this case, the wired components have been mounted on the other side of the circuit board, since the board for the prototype was not through hole plated. With a fully through hole plated printed circuit board all components can be mounted on the components side.

Fig 10 shows the prototype of the GSM detector in an aluminium housing, together with a dipole antenna for the GSM 900 frequency band.

Table 2: Relationship between power density and input from dipole.

Power density	Equivalent input from dipole GSM 900	GSM 1800
1µW/m ²	-48.8dBm	-54.6dBm
10µW/m ²	-38.8dBm	-44.6dBm
100µW/m ²	-28.8dBm	-34.6dBm
1mW/m ²	-18.8dBm	-24.6dBm

2.3. Parts list

- C2 2.2pF 0805
- C1, C3 3.9pF 0805
- C8-9, C10, C12 100pF 0805
- C13-C16 150pF 0603
- C29 330pF 0805
- C18 1nF 0805
- C4-C7, C11, C17, C20, C22, C24, C26-C28 100nF 0805
- C21 1µF MKS
- C23 4.7µF / 35V SMD electrolytic capacitor
- C19, C25 10µF / 35V SMD electrolytic capacitor
- L1, L2 8.2nH
- L3 10µH
- R4 3.3Ω 1206
- R2 39Ω 2512 (1 W)
- R15 1.2kΩ 1206
- R10 4.7kΩ 1206
- R1, R3, R6, R7 10kΩ 1206
- R11, R12 20kΩ 64Y
- R8, R9 50kΩ 64Y

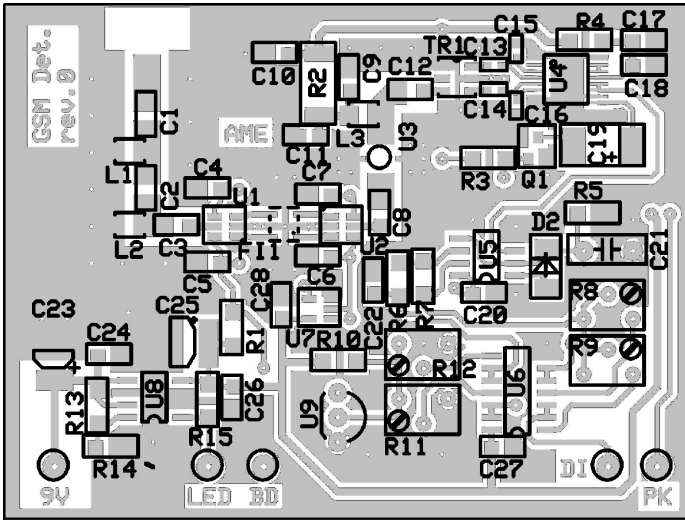


Fig 7: Component layout for broadband GSM detector.

- R14 470kΩ 1206
- R5 1MΩ 1206
- R13 1.6MΩ 1206
- U1, U2 HMC 221
- U3 ERA-3
- U4 AD8362 BRM
- U5 TLC 272
- U8 MAX 883 CSA
- U6 CD 4066
- U9 LM 385Z-2.5
- U7 TC4SU69F
- FI1 GSM dual band filter
SAWCD942MLB0T05
- TR1 TCM4-19
- Q1 BSS 138
- 1 x LED 3mm Low current with
insert socket
- 3 x rocker switch 1xUM
- 1 x battery compartment for 9V unit
- 1 x moving coil meter 100µA + scale
- 1 x tinplate housing 74 x 55 x 30mm

- 1 x BNC connector (e.g Radiall R
141 404 000)
- 5 x 1nF DF capacitor, solderable;
- 1 x printed circuit board, two-layer,
through hole plated
- 1 x aluminium housing with front
plate

3. Literature references

- [1] 26th Federal Immission Protection Directive BImSchV (Bundesimmissionsschutzverordnung), Directive Governing Electro-magnetic Fields
- [2] Electromog Information Page - Dipl. Ing. Ralf Dieter Wölfe

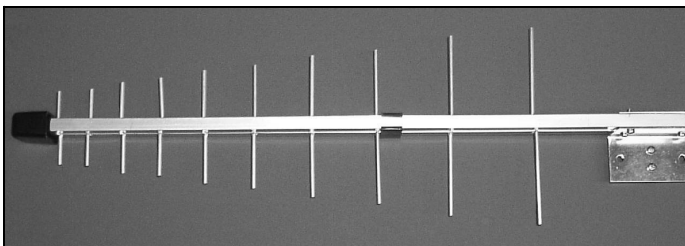


Fig 8: Picture of the AS-LOG antenna.

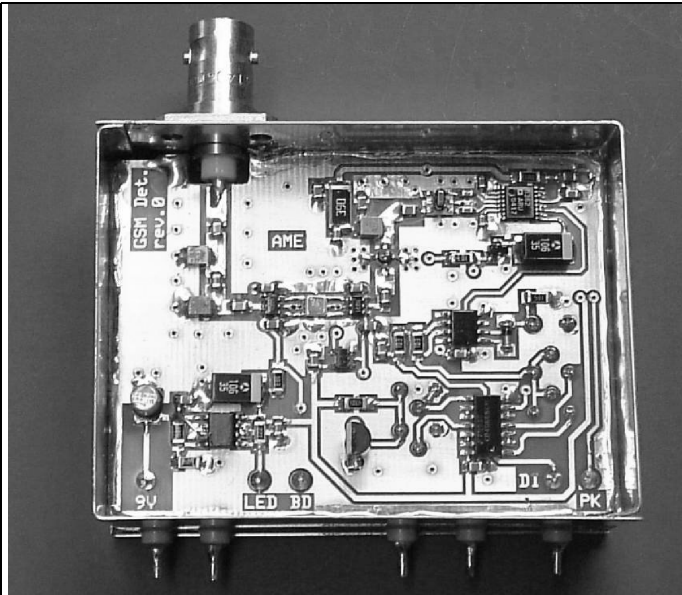


Fig 9: Picture of the completed broadband GSM detector.

<http://www.ralf-woelfle.de/elektrosmog/index.htm>

800MHz - 2GHz; Art.No. 00247; obtainable from Technical Dispatch Department for Radio Accessories, UKW Berichte; 7, Jahnstr., 91083 Baiersdorf

[3] AS-LOG, log- per. antenna for

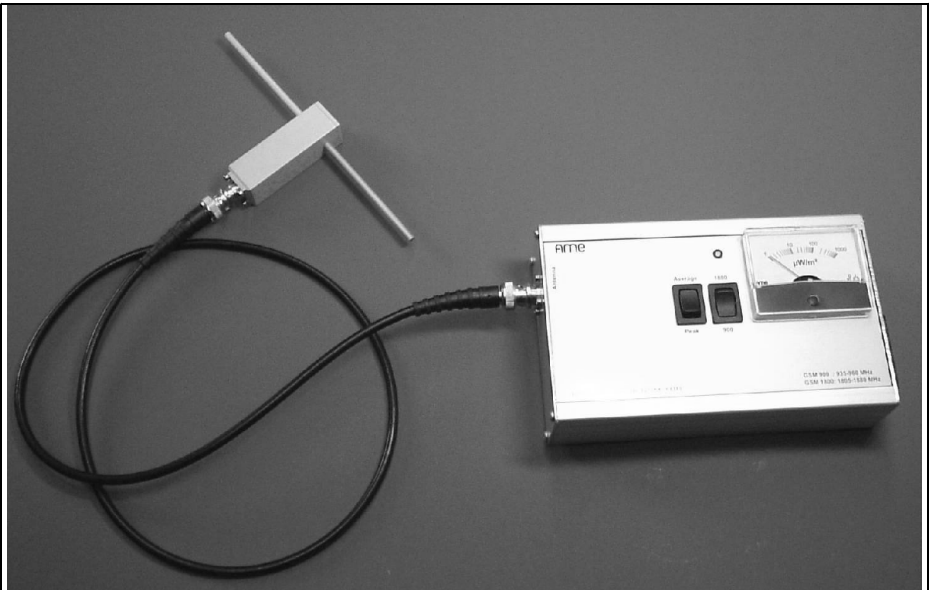


Fig 10: Picture of the broadband GSM detector in its aluminium case and fitted with the dipole antenna.



Gerrard Galve, F6CXO

Franco's Finest

Inexpensive 12 to 24GHz doubler described by F6CXO

Franco Rota runs an RF component supply company in Italy called R F Elettronica. His main objective is to sell bulk components such as SMD parts to the electronics industry. He attends some radio rallies in Europe and often has interesting items for sale that can be used or adapted by radio amateurs for use on the amateur bands. This is the third article of a regular series that will describe one of Franco's products with details of its use by radio amateurs. If you require more details about the products you can contact VHF Communications or Franco – rf.elettronica@tiscalinet.it

1.0

Introduction

Franco sells the circuit board shown in Fig 1 for €3, it was featured in issue 3/2004 when its use as a 10GHz preamplifier was described. The circuit board was initially purchased to salvage the 4 x NE32584's. It was then examined and it is quite rare to find satellite boards with pre-amplifiers so well aligned and so suitable for modifications. After some initial operations using them as inexpensive 10GHz pre-amplifiers, I had the idea of testing these circuits in a 12/24GHz doubler.

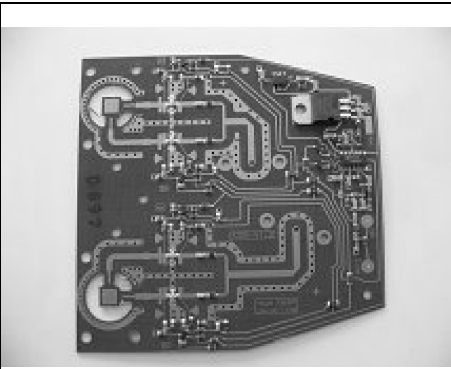


Fig 1: Picture of circuit board used.

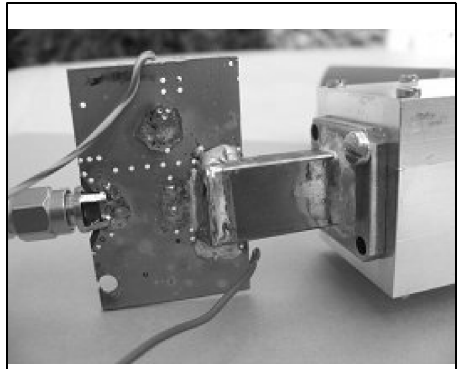


Fig 2: Picture of waveguide fitted to PCB.

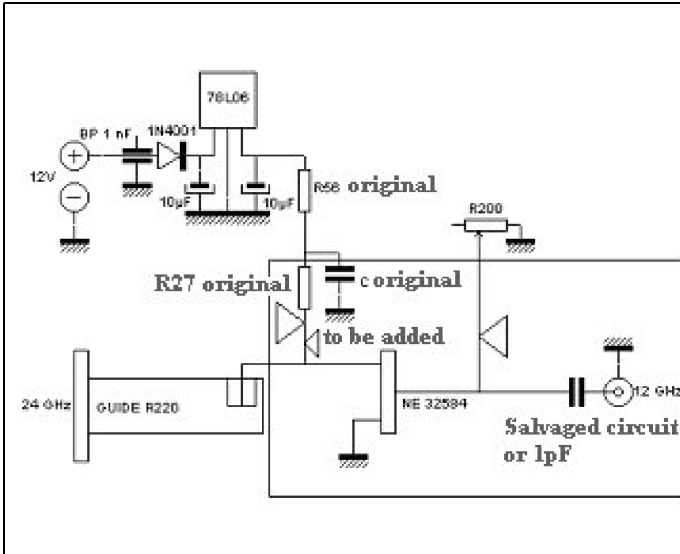


Fig 3: Circuit diagram of the modified circuit used.

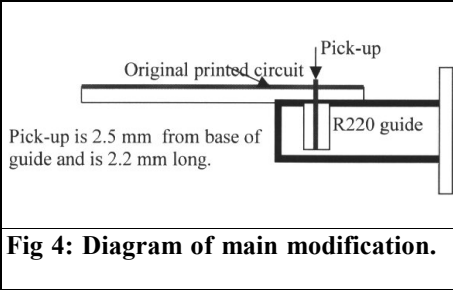


Fig 4: Diagram of main modification.

2.0 Modification

At the output, a short waveguide is used to recover the signal through a little piece of 1.7mm coax which has had its outer sheathing removed. Fig 2 show the waveguide fitted to the PCB, Fig 3 shows the modified section of the circuit used and Fig 4 is a diagram of the main

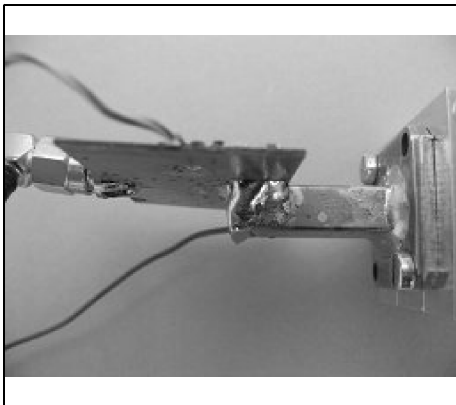


Fig 5: Side view of waveguide.

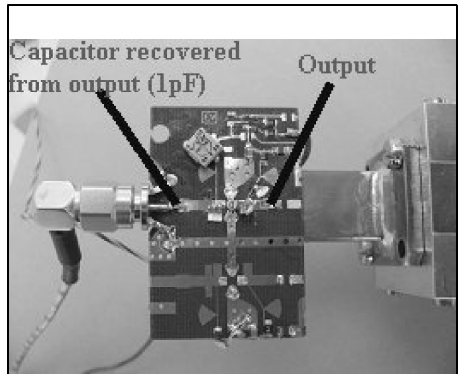


Fig 6. View of underside of PCB showing input and output.

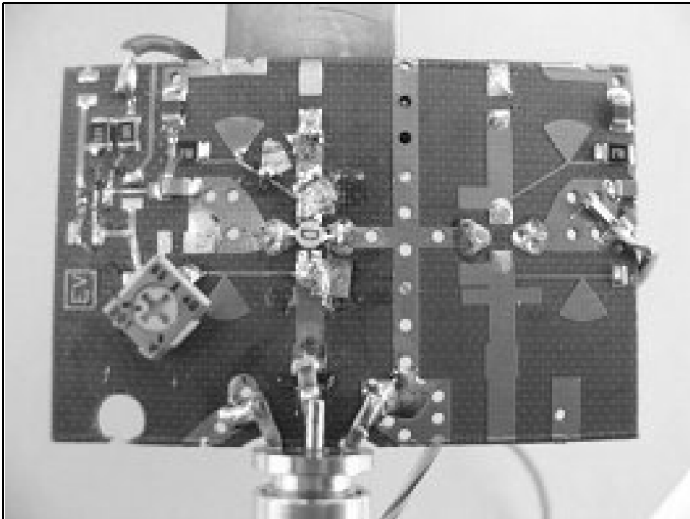


Fig 7: View of PCB showing optimisation foil flags.

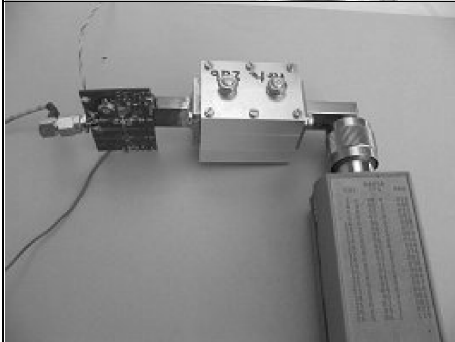


Fig 8: Test assembly using filter by OE9PMJ.

modification. Fig 6 shows the underside of the PCB and Fig 7 shows a close-up where you can see the foil flags used to optimise the circuit. To ensure that the output power measurements that I made were correct I used a filter designed by OE9PMJ to filter the output signal (Fig 8). The output measurements are show in Fig 9.

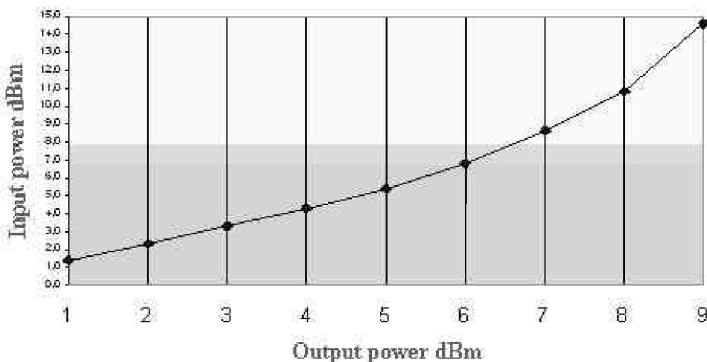


Fig 9: Graph showing test results of 12 to 24GHz doubler. Test conditions: V drain = 2.315v, I drain = 40mA.



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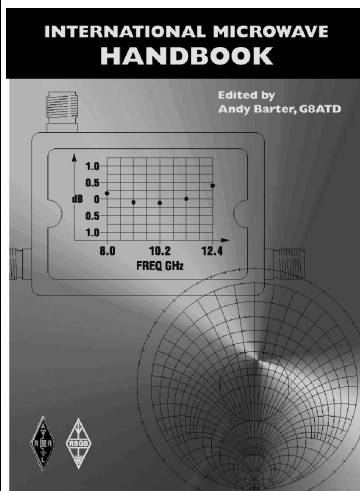
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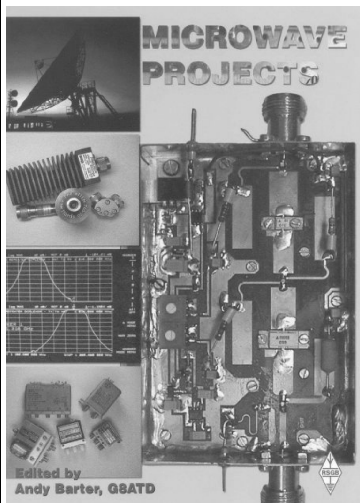
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