A Publication for the
Radio Amateur Worldwide

Especially Covering VHF, UHF
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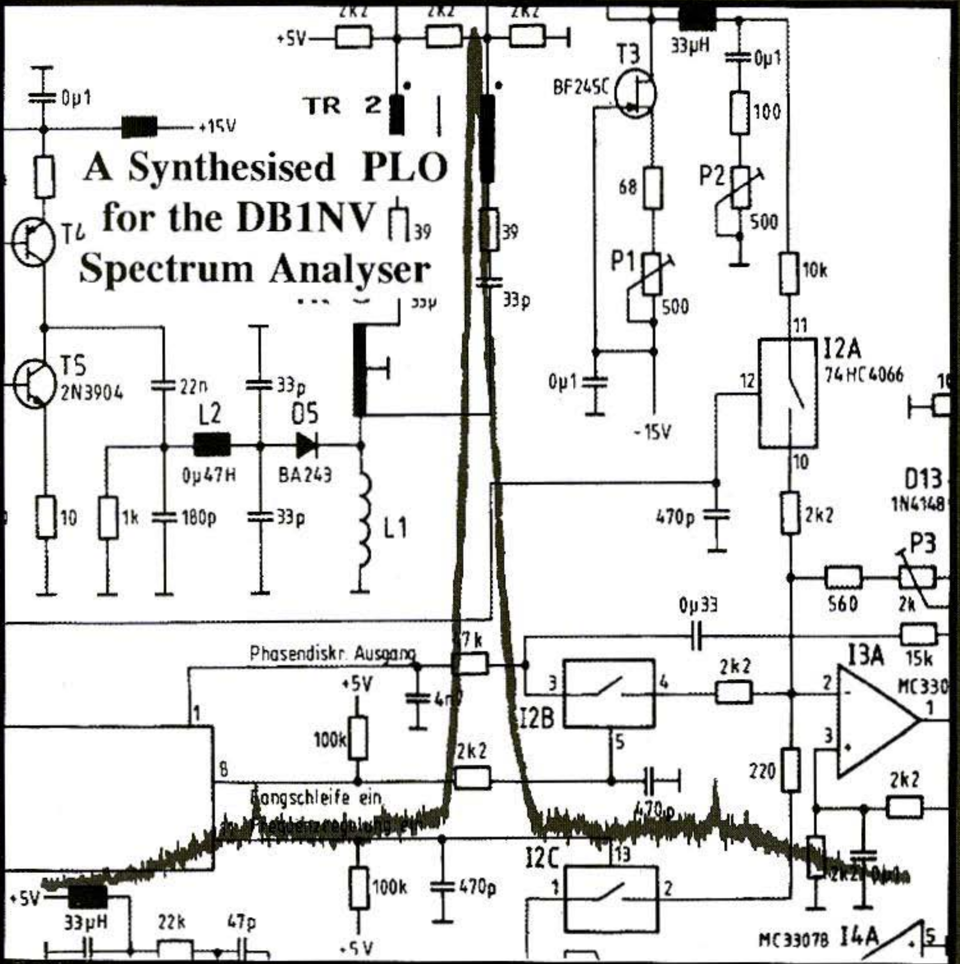
VHF COMMUNICATIONS

Volume No.26 . Winter . 4/1994 . £4.00

A Synthesised PLO for the DB1NV Spectrum Analyser



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Contents

We would like to wish you a belated Merry Christmas (if you celebrate it that is!) and wish you a Happy and Prosperous 1995. We hope that you have enjoyed VHF Communications, and thank you for supporting us, and hope that you will continue to do so in the future

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Dr. Jochen Jirmann, DB1NV

A Notch Filter for the S6 Special Channel

The occupancy of the S6 special channel in most cable networks is a terrible nuisance, at least for the radio amateur, since color and sound carriers are to be found in the 2m. exclusive band. Whether the interference has its origins in your own home or in that of understanding neighbors, the notch filter described below can be of assistance.

1. INTRODUCTION

In the early days of cable television technology, the origin of interference emissions from cable networks usually lay within the area of responsibility of the federal telecommunications authorities. Apparently the installation gangs had not been given sufficient training in how to deal with coaxial cables, with the result that badly mounted connectors and branches let high-frequency emissions escape. Since then, the level of experience here has improved, to such an extent that we can usually regard the telecommunications network as being "tight". If we are considering S6 interference, either the domestic distribution network (not renovated for reasons of cost) is emitting radiation or the equipment connected is insufficiently screened. In the one case, television and hi-fi equipment do not have the necessary screen damping, while in the other all cheap receiver connection cables act as slot antennae.

If, for example, a hi-fi receiver is the culprit, then the level of interference can be reduced by 2 - 3 S stages if the receiver input is provided with a rejector circuit. If the S6 program is not viewed throughout the entire domestic distribution system, then the best plan is to place the rejector circuit between the inter-connection point and the domestic connection amplifier.

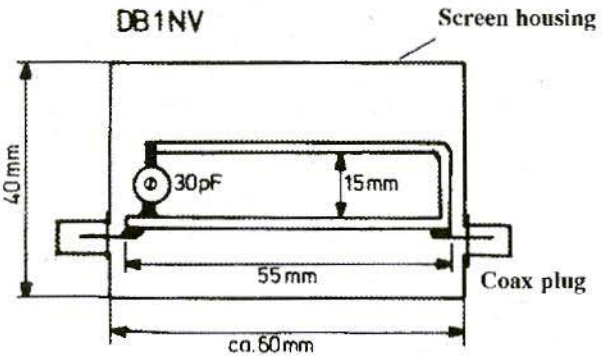


Fig.1: Assembly of the S6 Notch Filter

2. ASSEMBLY

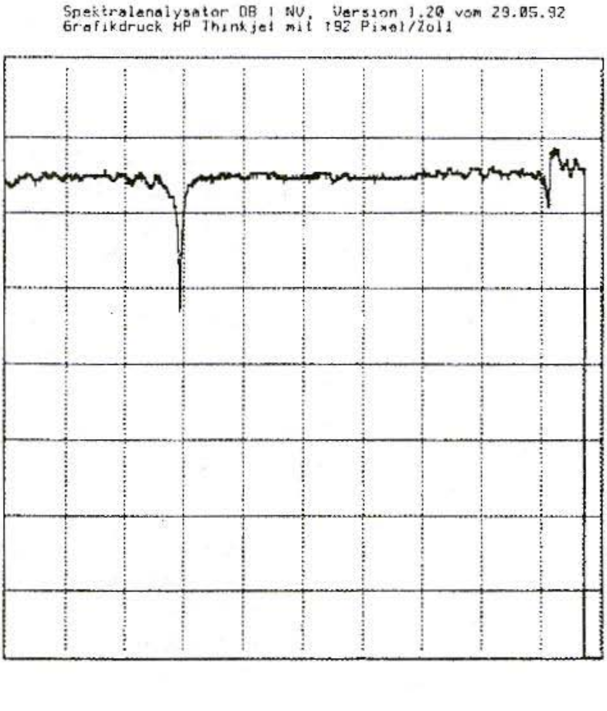
Assembling a notch filter is a simple matter. You need:

An HF-tight metal housing, dimensions 40 x 60 x 30 (mm.), which you can solder together from printed circuit board material if necessary

One 75Ω coaxial jack and one 75Ω coaxial plug each for soldering in, or corresponding cable plugs or jacks with short cable ends

A 30pF foil trimmer or ceramic trimmer

15 cm. of silver-plated 1-mm. wire



Test Object:	**S6 Notch filter**
Date:	**07.09.93**
Centre Freq:	**250 MHz**
Resolution:	**50 kHz**
Ref Level:	**+20dBm**
Mode:	**Normalised**
Horiz:	**50 MHz/div**
Video:	**100 kHz**
Vert:	**10dB/div**

Fig.2: Acceptance Curve of the Notch Filter

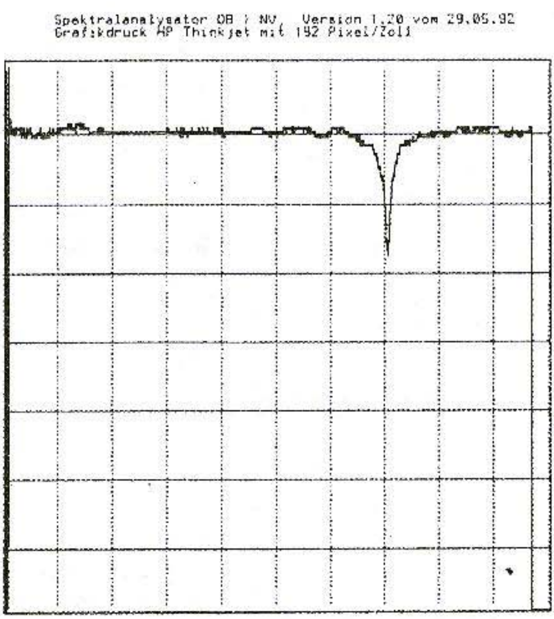
Bend the wire into a hairpin circuit, as per Fig.1, with dimensions of 55 x 15 (mm.). Solder it up to the trimmer and solder it unsupported between the internal conductors of coaxial jacks and coaxial plugs. Before closing the housing and making it HF-tight, drill a hole in the housing cover for calibration.

3. CALIBRATION

Anyone who has no wobble measuring position for calibration can also get by using a 2m. FM unit, which must have an S-meter. Link one of the filter leads (the filter is symmetrical) to the cable TV network and the other lead to the radio equipment. Set the S6 sound carrier frequency at 145.740 MHz and use an attenuator to bring the S-meter reading precisely to the end-scale deflection point. Then calibrate the notch filter trimmer to minimum. Since most S-meters are not worthy of the name and are more in the nature of linear HF-voltmeters, you should now reduce the damping until you obtain a new reading and re-balance the trimmer. And that's all the calibration you need.

4. MEASUREMENT CURVES

Fig.2 shows the acceptance curve for the notch filter in the frequency range between 0 and 450 MHz. We can see a reduction of about 20 dB at the interference frequencies in the 2-m. band. Fig.3 shows the frequency range from 0 to 200 MHz, and we can recognize the form of the filter curve better.



Test Object:	**S6 Notch filter**
Date:	**07.09.93**
Centre Freq:	**100 MHz**
Resolution:	**50 kHz**
Ref Level:	**+20dBm**
Mode:	**Normalised**
Horiz:	**20 MHz/div**
Video:	**100 kHz**
Vert:	**10dB/div**

Fig.3: The Filter Curve Form is easily recognizable

Matjaz Vidmar, S53MV

A DIY Receiver for GPS and GLONASS Satellites

Part-3b

Quadrifilar Backfire Helix Antenna

Although long range, precision navigation systems like GPS or GLONASS were designed to be independent as much as possible of the performance of either transmitting or receiving antennas, the antennas used still have some influence on the system performance.

1. ANTENNA REQUIREMENTS

The transmitting antennas installed on the spacecraft have a shaped beam to supply any Earth-located users with the same signal strength and use the on-board transmitter power more efficiently.

Maintaining the same signal strength is especially important in CDMA, since the GPS C/A-codes are too short to offer a very good crosstalk performance. The ideal receiving antenna should have a hemispherical radiation pattern, offering the same signal strength from a satellite at zenith and from another satellite just above horizon. Further, the receiving antenna should match the transmitter polarisation (RHCP) in all valid directions.

Finally, the receiving antenna should attenuate any signals coming from undesired directions, like signals coming from negative elevations, since these are certainly reflected waves and the latter are a major source of measurement errors due to their unknown propagation path.

2. THE TURNSTILE ANTENNA

Although a turnstile antenna (two crossed dipoles fed in quadrature) with or without a reflector is frequently used for satellite reception, this antenna is-not very suitable for satellite navigation for several reasons. The polarisation of a turnstile antenna is circular only in the zenith direction and is completely linear in the horizon plane. Therefore, a turnstile antenna offers no discrimination between the desired RHCP direct wave and the unwanted LHCP reflected wave, since circularly polarised waves changed their sense of polarisation on each reflection. Reflected waves cause severe measurement errors and a relatively slow and deep signal fading, so that the receiver even looses lock on the signal.

3. MICROSTRIP PATCH ANTENNA

A better alternative is a microstrip patch antenna. A single microstrip patch reso-

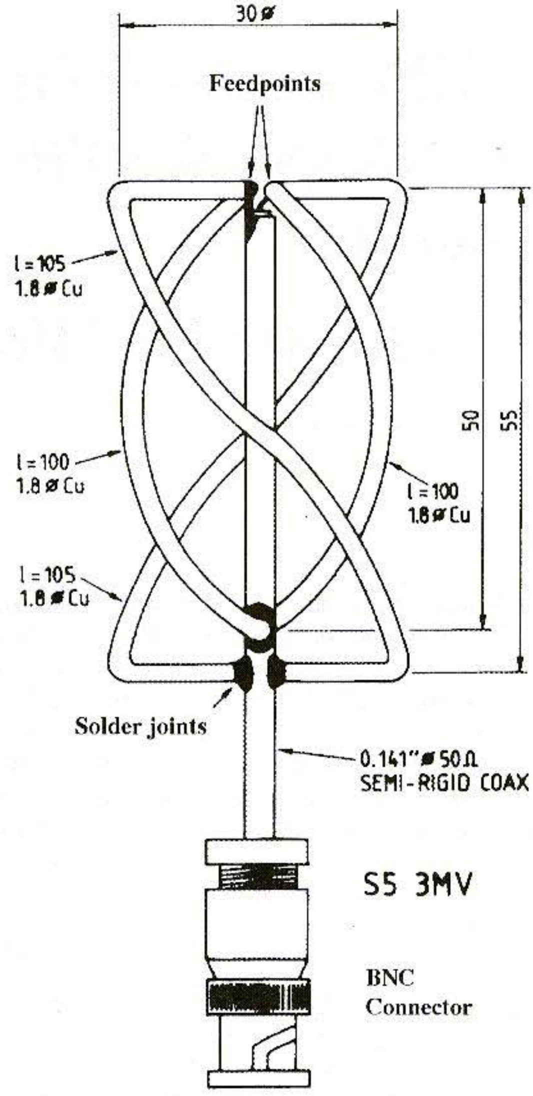


Fig.1: The Quadrifilar Backfire Antenna with Right-Hand Circular Polarisation

nator provides a useful radiation pattern with a reasonably circular polarisation over a wide range of elevations. Unfortunately the radiation pattern of a microstrip antenna falls down to zero in the horizon plane. Microstrip antennas are usually used when a simple, low-profile antenna is required, usually to be installed on a vehicle roof. Since low-elevation satellites can not be received, a microstrip antenna usually limits the available GDOP.

4. THE QUADRIFILAR BACKFIRE ANTENNA

The best antenna for satellite navigation and other applications requiring hemispherical coverage seems to be the quadrifilar backfire helix (also called a "volute" antenna). Such an antenna provides a shaped conical beam. The beam shaping and cone aperture can be controlled by adjusting the helix radius, turns pitch distance and number of turns as described in [11]. By the way, the same type of antenna is frequently used on low-Earth orbit satellites, like the NOAA weather satellites.

As the GPS and GLONASS satellites already provide a constant signal strength for Earth-located users regardless of the satellite elevation, no particular beam shaping is required for the receiving antenna. The optimum number of turns of a quadrifilar backfire helix used as a GPS or GLONASS receiving antenna seems to be between 1.5 and 3. Making a quadrifilar backfire helix longer by increasing the number of turns does not have much effect on the gain or the beam-cone aperture, but it improves the beam shaping and further attenuates the undesired lobe in the opposite direction (downwards).

Although the best GPS receivers use such a quadrifilar helix with 1.5 or 2 turns, such an antenna is difficult to manufacture and test. In particular, the four helical wires have to be fed in quadrature and there is very little space on top of such an antenna to install the feeding network. Further, a 2-turn backfire helix is rather large (20cm high) for a portable receiver. If its improved pattern performance is to be fully exploited, the direction of its axis should not deviate too much from vertical and this is not a very practical requirement for a portable receiver.

Most GPS/GLONASS receivers therefore use a simpler antenna, usually a short one-half turn backfire helix like shown on Fig. 16. Making the quadrifilar helix shorter resonance effects can be used to feed the four helical wires with the proper signal phases. In particular, one pair of wires is made shorter to make its impedance capacitive at the operating frequency and the other pair of wires is made longer to make its impedance inductive at the operating frequency.

To obtain RHCP a conventional endfire helix has to be wound like a right-hand screw. The backfire helix is just opposite: to obtain RHCP the backfire helix has to be wound as a left-hand screw, besides the proper phasing of the four helical wires, of course! Further, the backfire helix requires no reflector. The four helical wires are fed at one end of the helix and shorted together at the other end of the helix. Since the main (desired) radiation lobe is directed towards the feedpoint and away from the shorted end, such an antenna is called a backfire antenna.

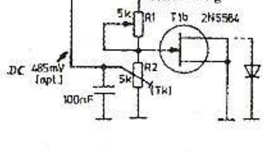
The feedpoint impedance is in the 50Ω range, symmetrical. A good match to 50Ω is usually sacrificed for the radiation pattern which is much more important. Usually one of the four helical wires is replaced by a semi-rigid coaxial cable of the same outer diameter to form an "infinite balun". on the other hand, the performance of the antenna is not degraded much if no balun is used as shown on Fig. 1.

In the practical construction of a half-turn quadrifilar helix it is especially important to respect the exact lengths of the helical wires, since the antenna uses resonance effects and is rather narrowband. The dimensions shown on Fig. 16 are for the GPS L1 frequency (1575.42 MHz). A GLONASS L1 antenna should be approximately 3% smaller. Finally, an antenna for both GPS and GLONASS L1 channels can be built by designing it for the average of the two frequency bands.

IMPROVEMENTS - CHANGES

Suppression of interference in 70-cm ATV mode using highly selective notch filter, by E.Berberich; 1/94 pp.45-55.

Some errors crept into Fig.12 on p. 52, so here's the circuit again.



Improvements and additions to the Spectrum Analyser by Dr.J.Jirmann, DB1NV

Some points were not clear regarding the structure of the spectrum analyser and need correcting:

1. Printed circuit board 007 (LO/PLL): Circuit diagram and components diagram gave different values for resistance of 17, pin-2: the version with a 56k resistance to earth is correct.

The capacitor at pin-4 of 13 (NE 5534) has a purely blocking function. It could be given a value of, for example, 0.1uF.

2. Printed circuit board 009 (run-off control): The tendency of the emitter follower to oscillate did not become apparent until the layout had been completed. It can be remedied by means of a 1nF (not 1uF) ceramic capacitor on the foil side.

Circuit diagram and components diagram gave different values for resistance of 12, pin 2 to earth. The correct value here is 150k; at 39 k, the tuning diode in the second LO would have a bias voltage in the conducting direction.

Carl G. Lodström SM6MOM/W6

An RF Power Meter with a Linear Scale

During the work with the Complex Impedance Meter (1) I got the idea to try to see how low a signal level could be detected with a straightforward diode and a good DC amplifier.

As I worked on this, I was struck by the accuracy of the meter as long as the diode operated in the square law region. A simple power meter with a linear dial!

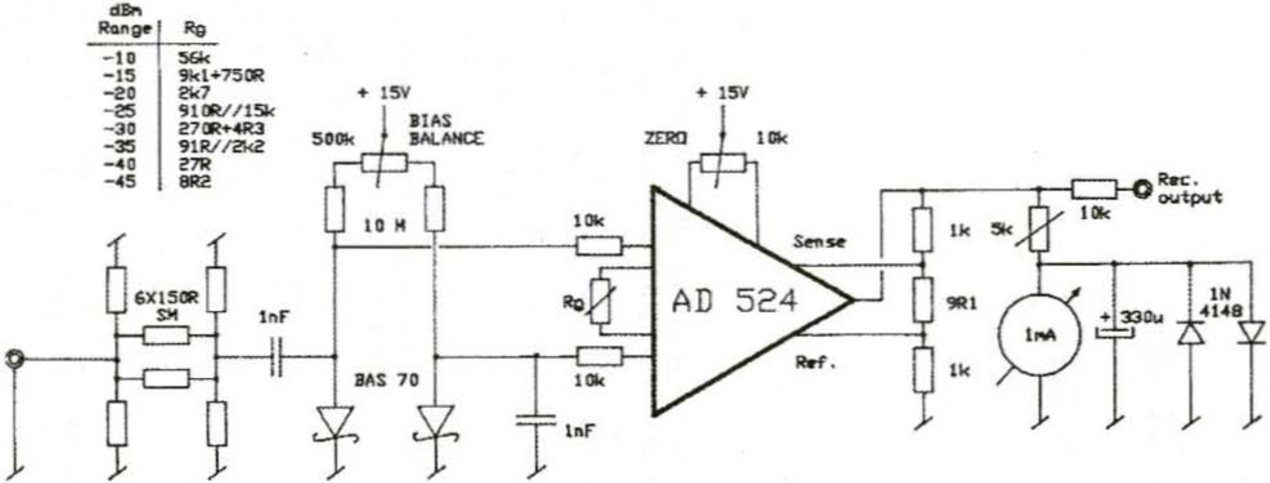
1. DESIGN CONSIDERATIONS

Clearly a detector diode does not have 50Ω impedance (kΩ at this bias) and that attempts to match it would not only complicate the instrument, but lessen the flat frequency response as well. Another insight was that temperature variations would create drift problems.

With this in mind, I decided to try two diodes, one pumped by RF signal and the other not. Both biased in the forward direction by about 1µA of DC current. My favourite instrumentation amplifier is AD 524 by Analogue Devices. It works very well in this application.

To provide a reasonable good 50Ω input impedance for the instrument the input circuitry is an attenuator/termination, providing good match and some 6dB of voltage attenuation. The detector diode impedance is assumed to be >50Ω.

The necessary close thermal coupling between the detector diodes is achieved by using the dual Schottky BAS 70-05 or -06. Common anode should work as



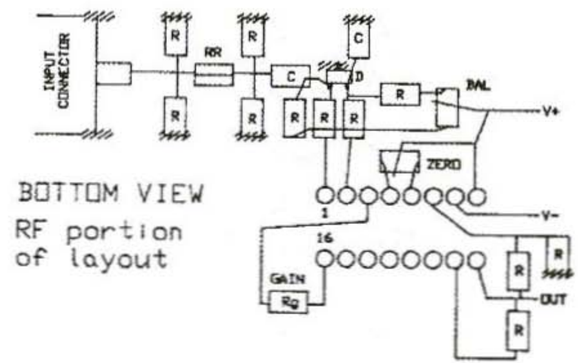


Fig.1: Circuit Diagram and Component Layout for the Power meter

well as common cathode. Just feed the 1μA bias supply from -15 V. Thereafter everything is pretty much straightforward.

2. CIRCUIT DESCRIPTION

The circuit and component overlay for the power meter are shown in Fig.1. The AD 524 input zeroing is used to null the instrument, and the balance between currents in the two diodes can also be varied by a small amount, making it possible to find a point of optimal thermal tracking. The output amplifier is connected for a gain of approximately 100x by the 1k-9.1-1k resistors. The gain at the different ranges is selected by selecting a different resistor value for each range by means of a rotary switch.

The values are most easily determined if you pick the lowest (-45 dBm range) and adjust the resistor in series with the meter movement for an accurate reading. Go then up one range at a time, substituting with a decade resistor box to find the accurate values. I have listed the ones that worked for me.

The gain formula is:

$$|G| = (40k/R_g) + 1$$

but with 20% inaccuracy.

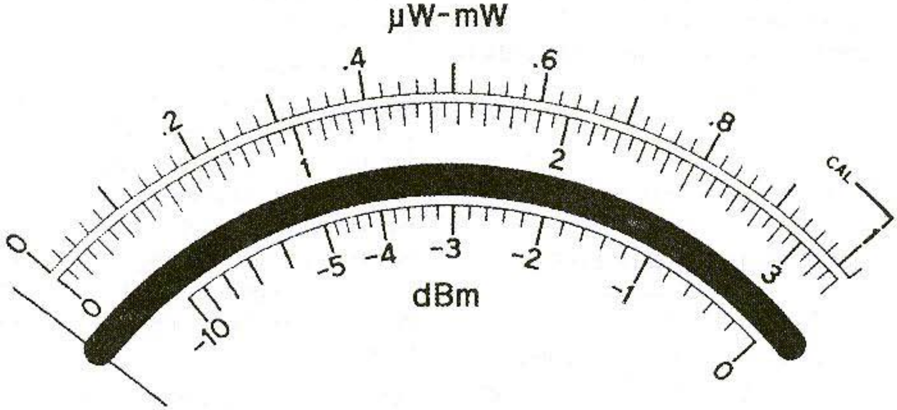
Without the gain in the output stage, the gain selection resistor for the -45 dBm range would have to be 82mΩ, an impractically low value. If one connected the meter direct to the amplifier output, reasonable values could still be used, but read on!

It is advisable to pay attention not to burn out the meter, or bend the needle, with a possible transient of full output from the amplifier! The simplest method of protection is to put a few cross-connected diodes across it. Together with an electrolytic capacitor and a limiting resistor in series, the meter can handle a slowed down overload.

Full deflection ranges from -10 dBm to -45 dBm are meaningful. Signals of -50 to -55 dBm are about the lowest that can be measured. Using 50% AM modulation and a selective audio amplifier (HP 302A) levels of -65 dBm have been detected with this same device. Given the 5 Hz bandwidth in the HP 302A, a synchronous detector in combination with switching of the RF signal alternatingly between the diodes should allow for another 10 - 20 dB, setting the limit at some -80dBm. More complicated circuitry that certainly will affect the flatness of the response, though.

These days, with inexpensive wideband amplifiers readily available, it may be just as wise to use one or two of them. Another 20 - 40 dB is then easy to get, and the flatness may be just as good as with RF switching between the diodes.

The input attenuator is realised by six 150Ω surface mount resistors. Needless to say, they should be built in accordance with good UHF practice. Before I replaced two individual Schottky diodes with the single BAS 70 device the response was very flat (within a dB) to 1.3 GHz, where it peaked briefly and



died. If anything, it ought to be better now, but I have no "flat" source to try it. It seems to work past 1.8 GHz though.

The meter I use (from a flea market) has an individually calibrated dial that is not entirely linear. It is provided here anyway, for reproduction, since it "is better than nothing." An old, non working HP 431 power meter, has just this kind of meter and can usually be had for less than a new µA meter.

Note that the -10 dBm is at 10% of full scale, whereas a voltmeter reads -20 dB here.

3. LITERATURE

(1) Complex Impedance Meter, C.G.Lodström, VHF Communications 2/92

Dr. Ing. Jochen Jirmann, DB1NV

A Synthesised Local Oscillator for the DB1NV Spectrum Analyser

The following article describes a synthesiser which replaces the previous first oscillator of DB1NV spectrum analysers 006 to 011 and which, as well as having better frequency precision, improves the usable dynamics for signals lying close together by at least 5 dB. It is also easier to operate than the previous PLL's.

The synthesiser has a tuning range of 450 - 1,300 MHz with a 50 kHz step size. Without the use of any special components, it was possible to attain a phase jitter interval of 105 dBc/Hz for a carrier interval of 20 kHz, at moderate expense.

The synthesiser has actually been conceived primarily as a local oscillator for the spectrum analyser, but can also be used as a simple synthesised test oscillator.

1. THE ADVANTAGES OF SYNTHESISER - LOCAL - OSCILLATOR

Five years ago, the author introduced his first design for a spectrum analyser created using resources available to an amateur. The basic concept is still proving serviceable, but it has also become clear that the first local oscillator is the weak point in the circuit. With respect to resolution and volume range, the remainder of the circuit can generate considerably more than can be used by the relatively poor oscillator. This is documented in Figs. 1 to 3. Here, a test oscillator (HP8640A) acts as a signal source at 140 MHz, which is modulated at 20 kHz and with a modulation factor of 2 per thousand,

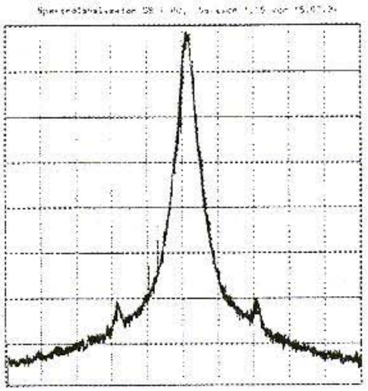


Fig.1: **Previous Spectrum Analyser LO with powerful Noise Bar**

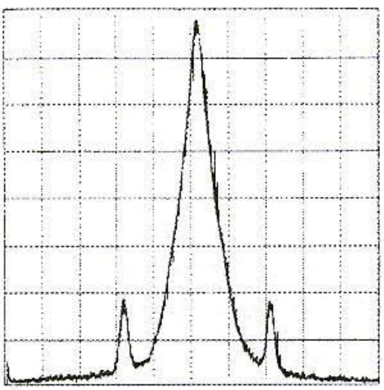


Fig.2: **Use of a Singer SG 1000 as Local Oscillator**

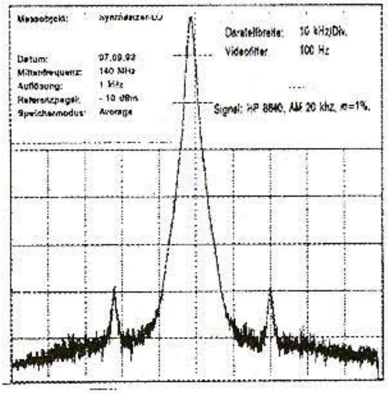


Fig.3: **Synthesiser LO markedly improves SA Dynamics**

which corresponds to a side-band amplitude of 60 dB below the carrier.

In Fig.1, the original first oscillator was used, with a display width of 10 kHz/section. The intermediate-frequency band width is 1 kHz and the video band width is 100 Hz.

If the image storage is used to give an average value, the modulation at ± 20 kHz is still recognisably within the noise base of the oscillator.

In Fig.2, a second standard signal generator, a Singer SG1000, acted as a local oscillator producing less noise. With no other change in the settings, the modulation stands out more than 10 dB above the noise.

We can thus recognise that the intermediate-frequency and video sections of the analyser have a considerably greater dynamic, which can not be made use of through the poor first oscillator.

Now, Fig. 3 shows the same signal, now with the synthesiser / local oscillator. Here too, we can observe the significant dynamic gain, although it is not as great as with a low-noise standard signal generator.

Fig.3 was plotted without digital averaging, so that the noise displayed is somewhat greater than in the previous diagrams.

In addition to the improved measurement dynamic, the synthesiser offers better frequency accuracy and is easier to operate, as the rather tiresome search for scanning points separated by 2 MHz in the previous PLL is no longer necessary.

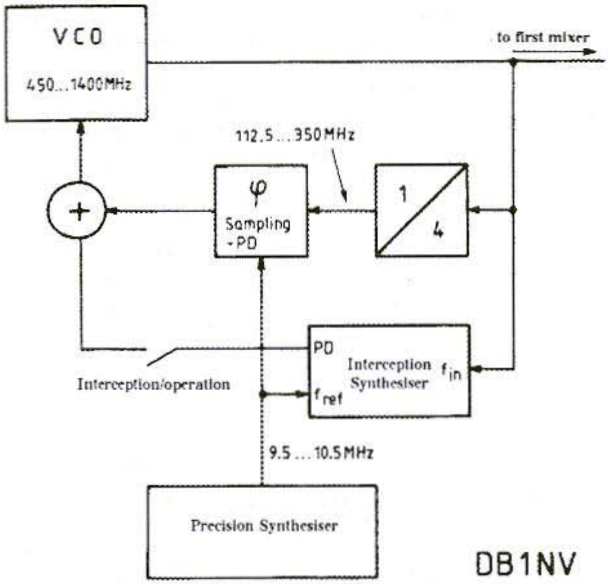


Fig.4: Synthesiser with Sampling Phase Discriminator

DB1NV

2. POSSIBLE METHODS OF ASSEMBLING THE SYNTHESISER

We can estimate the minimum phase jitter interval required for the first oscillator in the following way. At a band width equal to 1 kHz, the intermediate-frequency filter of the analyser still allows signals to be resolved with a level differential of 75 dB, at a frequency interval of 20 kHz. So the phase jitter of the first oscillator must be at least 75 dB below the carrier level when the interval between the oscillator and the carrier is equal to 20 kHz. Since the filter band width is 1,000 Hz, the standardised phase jitter interval then becomes 105 dBc/Hz. This value is not exactly outstanding. The Hewlett-Packard test oscillator, which is widely used, even by amateurs, can attain 130 dBc/Hz with the same carrier interval. Many first-generation synthesiser/standard signal generators and simpler spectrum analysers are of course not much better than the 105 dBc/Hz desired with regard to phase jitter. It should thus be possible, even using amateur equipment, to attain the required phase jitter interval using modern components. Yig oscillators, the new price of which is still around DM 1,000, should certainly not be used.

2.1. The Options

There are two possible ways of building a low-noise synthesiser of this type.

- You can divide the tuning range from 450 to 1,300 MHz into many sub-ranges, so that the sub-oscillators automatically supply a sufficiently low-noise sig-

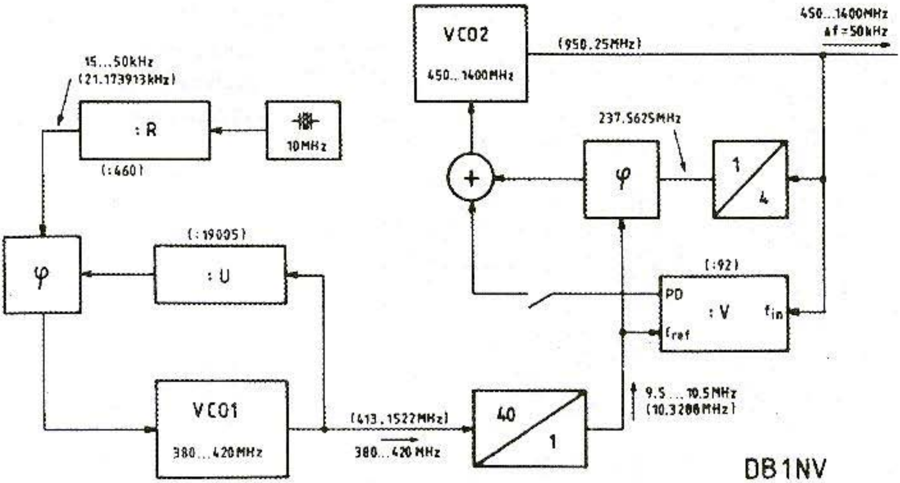


Fig.5: Block Diagram of the 450 to 1300 MHz Synthesiser

nal, due to their resonator quality. A slow PLL loop compensates only for the frequency changes arising from fluctuations of temperature and low-band interference, e.g. hum.

- You can retain the existing oscillator, which is tuneable over an extremely wide range, and stabilise it using an extremely rapid PLL with a control band width of several 100 kilohertz. An extremely low-noise reference signal must be generated at the same time.

The commutation of several sub-oscillators may be a solution for a pure synthesiser. For the spectrum analyser, it must also be possible to wobble the oscillator signal over the entire tuning range. We can therefore expect some transient problems to arise from the periodic commutation of the sub-oscillators within milliseconds.

Analogue wobbling of the oscillator using a frequency control loop, on the other hand, was a tried and tested technique. To keep the operation of the analyser simple, the step size of the synthesiser had to be less than 100 kHz.

The author's first attempts, using conventional divided PLL's with phase monitoring frequencies between 100 kHz and 1 MHz proved a disappointment. The interference dispersion of the oscillator, divided down by a factor of 1,000 or more, was too transient for normal phase detectors to generate a control signal which could be evaluated. By contrast, it was possible to attain the desired signal quality right away with the help of an oscillator which could be tuned over a narrow band (40 MHz tuning range at app. 400 MHz) and a modern single-chip synthesiser, an SP8858 from Plessey. Now all that was needed was the resolution to transfer the signal quality of this synthesiser to the actual first oscillator of the spectrum analyser. To control even small phase deviations from the oscillator, frequency dividers in the control loop should be dimensioned to give the lowest possible divider ratio.

Fig. 4 shows the approach. The output signal from the VCO is initially divided by four and thus displaced into a frequency level which is easier to control. A sampling phase discriminator compares the divided VCO signal with an auxiliary frequency of about 10 MHz, i.e. the divided VCO signal can be synchronised on all harmonic waves of 10 MHz.

Dividing by four represents a compromise between the undesirable dividing down of the phase jitter from the first oscillator and the degree to which a sampling bridge can be created using amateur resources. This procedure gives VCO scanning points at 40 MHz intervals. So that the sampling PLL can find the correct scanning point, a simple single-chip synthesiser, of a type used in television engineering, works in parallel with it as an interception stage. Its only task is to tune the VCO into the vicinity of the correct sampling PLL scanning point following a frequency change, and it is then switched off.

The sampling PLL is designed for a control band width of approximately 200 kHz, and is thus rapid enough to control the phase jitter of the VCO. If we now wish to be able to vary the frequency of the VCO in stages, the 10 MHz auxiliary frequency has to have fine-stage tuning. At the same time, an extremely good phase jitter interval must be attained, since all interference at the auxiliary frequency will be

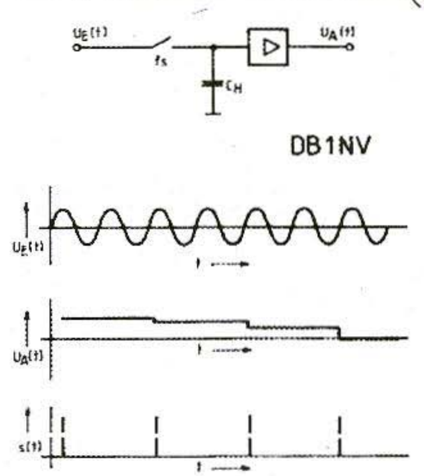
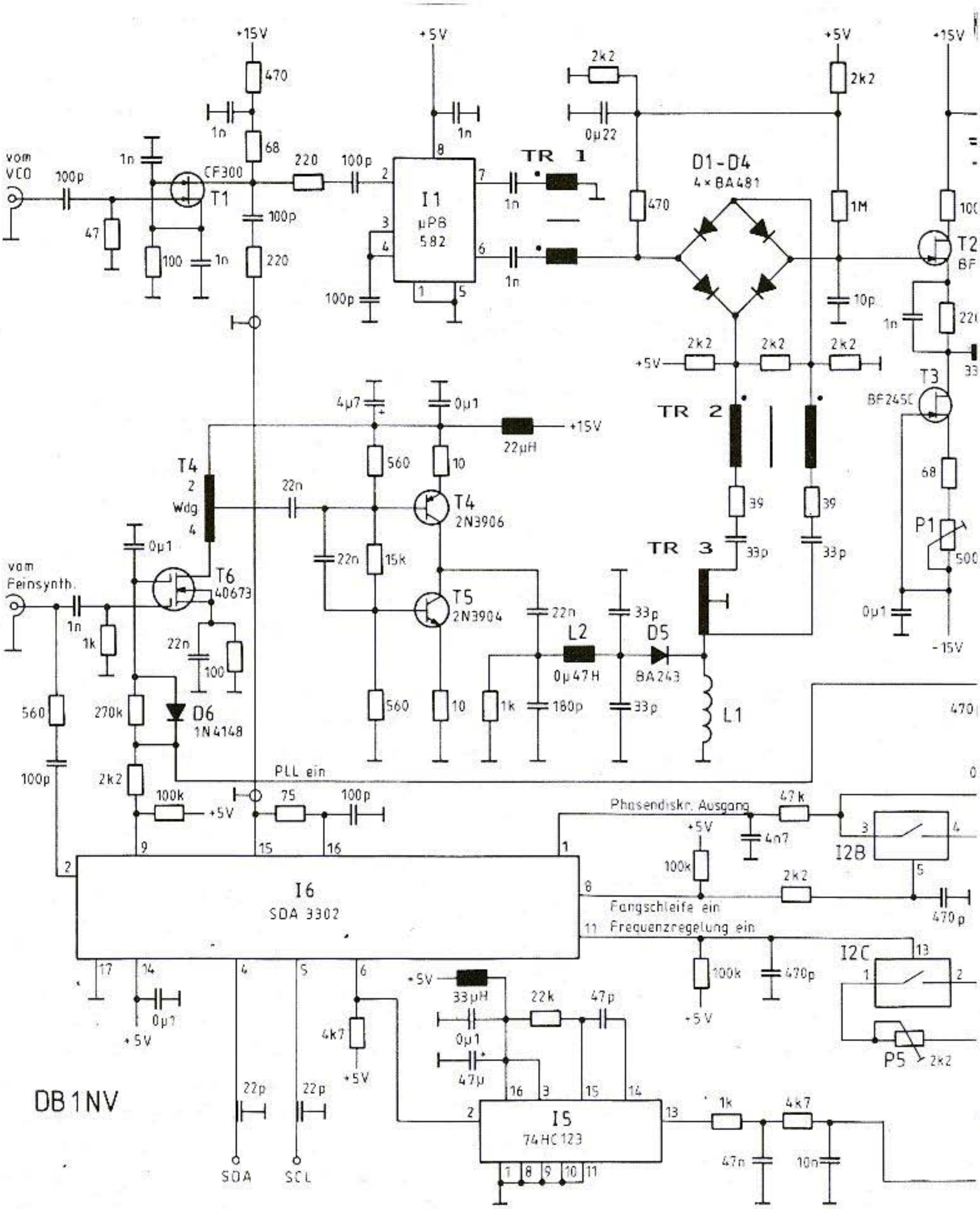


Fig.6: Structure and Operation of a Sampling Phase Discriminator

multiplied at the final frequency to a corresponding degree. As mentioned above, a narrow-band synthesiser with the low noise level required is relatively easy to make. So the 10 MHz reference is provided by dividing up a synthesiser which is tuneable from 380 to 420 MHz.

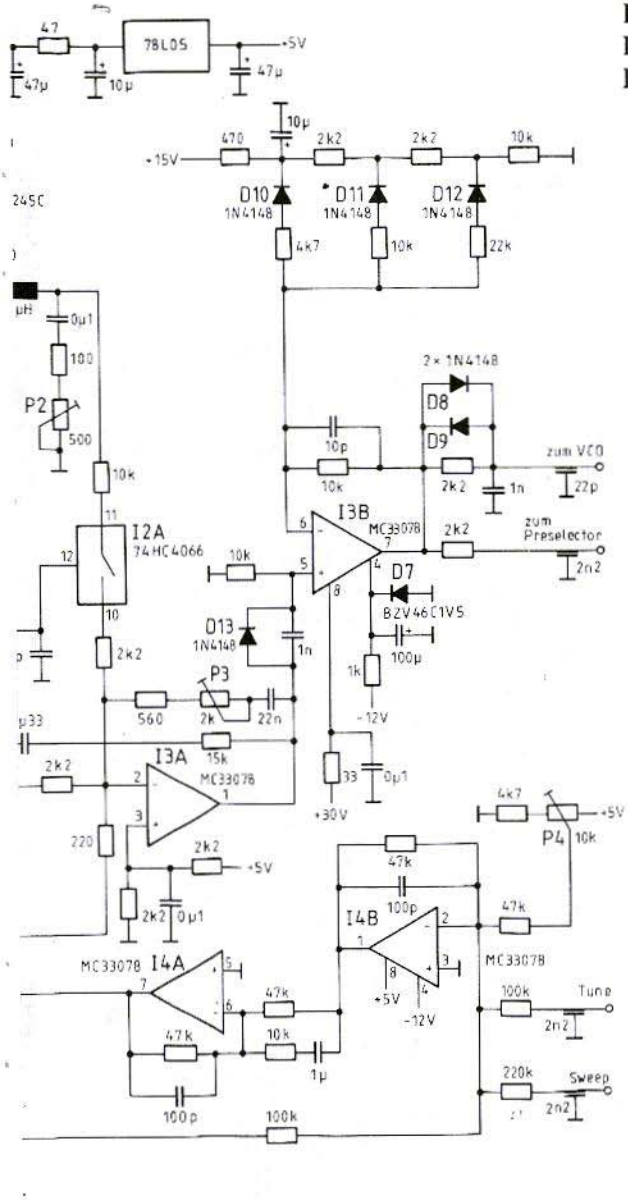
To obtain a uniform frequency grid at the end frequency, the step size of the "precision synthesiser" should be switched in accordance with the grid divisions of the harmonic wave used in each case. Since almost all synthesiser IC's have freely programmable reference dividers, this can easily be done. Strictly speaking, changing the step size calls for a change in the dimensioning of the loop filter in the precision synthesiser in order to optimise the spurious interval and the transient behaviour. Since the SP8853 proposed has current sources in the phase monitor which can be programmed through the interface, these elements can be used to



+15V
+5V
470
1n
68
vom VCO
100p
47
CF300
T1
100
220
100p
I1
μPB 582
TR 1
0μ22
2k2
470
D1 - D4
4 × BA481
1M
10p
T2
T3
BF245C
68
P1
500
0μ1
-15V
TR 2
39
33p
TR 3
4μ7
0μ1
22μH
+15V
560
10
T4
2 Wdg. 4
22n
15k
T4
2N3906
T5
2N3904
vom Feinsynth.
T6
40673
1k
D6
1N4148
270k
2k2
100k
L2
0μ47H
D5
BA243
180p
L1
PLL ein
75
Phasendiskr. Ausgang
47k
4n7
I2B
I6
SDA 3302
Fangschleife ein
Frequenzregelung ein
470p
I2C
P5
2k2
33μH
22k
47p
47μ
4k7
22p
SDA
SCL
I5
74HC123
47n
10n
DB1NV

Fig.7: Entire Main Loop Circuit

Feinsynth = Precision Synthesiser
Phasendiskr = Phase Discriminator
Frequenzregelung = Freq. Control
Fangschleife = Interceptor Loop



Winding Data:

TR1: 2 x 1.5 turns AW630 on FB101-43 Ferrite bead

TR2: approx. 20cm twisted AW630 on T50-43 toroidal core (yellow-white) (app. 20 turns)

RR3: 2 x 15 turns AW630 on FB101-43 ferrite bead

TR4: 4 x 2 turns 0.28 CuL on FB101-43 ferrite bead

L1: 2.5 turns 0.5 CuL, 4mm diam

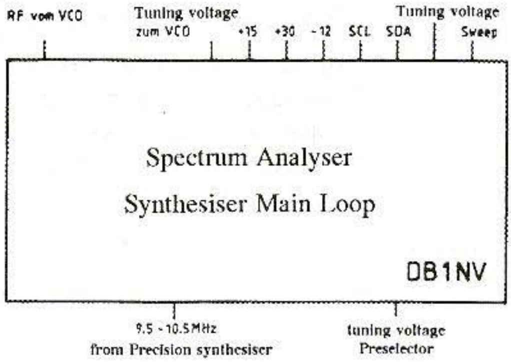


Fig.7a: Main Loop Connections

optimise the control behaviour of the loop (in approximate stages) in accordance with the step procedure selected.

2.2. Synthesiser Control

The relatively complex control of such a synthesiser can be handled only by means of a single-chip micro-controller (if a reasonable number of units are to be assembled). In this context, it fulfils the following functions:

- Evaluation of pulses from an incremental transmitter for tuning, with a progressive tuning characteristic line
- Controlling the frequency display - an 8-character LED dot matrix display
- Determining the appropriate harmonic wave and setting the interception synthesiser
- Setting the precision synthesiser (frequency and control behaviour)
- Time-controlled transfer from interception synthesiser to sampling PLL
- Control of a digital/analogue converter to generate the tuning voltage of the VCO if the VCO has to be wobbled. In this case, a conventional frequency control loop is wired in.

The proposed divider/multiplier procedure looks very elegant at first sight, but it has a crucial weakness, i.e. the 400 - 10 MHz frequency divider.

This assembly decides the total attainable phase jitter interval, since divider stages also generate noise side bands. Even with an ideal input signal, a good ECL divider has a broad-band noise base of about - 145 dBm/Hz. This value represents the limit for the purity of the 10-MHz signal. Each time the 10-MHz signal is doubled, the phase jitter interval is reduced by 6 dB. For example, to attain an output frequency of 1,280 MHz, it is necessary to multiply by a factor of 128, which corresponds to a rise of 42 dB in the phase jitter. The desired phase jitter interval of - 105 dBc/Hz can thus not be maintained at the higher end of the band.

The situation could be improved by mixing the VCO down by segments, using low-noise permanent oscillators, and only then feeding the signal to the sampling phase discriminator. For reasons of expense, it was decided not to do this, since permanent oscillators (e.g. with coaxial ceramic resonators), for their part, have to be connected to the master crystal using slow PLL loops.

When the frequency divider was investigated further, it became apparent that the signal noise interval could be reduced down as far as 10 dB using low-current drain dividers - e.g. the SP8716. A divider consisting of a CA3199 and an SP8695 created considerably less broad-band noise, but on the other hand needed more current.

3. DETAILED OPERATION

Fig.5 shows a detailed block diagram of the synthesiser. To appreciate the tuning procedure, let us assume that the synthesiser is to be tuned to f_o = 950.250 MHz, and that the frequency grid gives Δf = 50 kHz. Let a crystal oscillator with f_{ref} = 10 MHz be available as the reference frequency of the precision synthesiser. Every change of frequency involves the following procedures:

1. The total divider ratio required is calculated as:

   $$N = f_o/\Delta f = 950.250 \text{ MHz}/50 \text{ kHz}$$
   $$= 19{,}005$$

2. The micro-controller determines which harmonic wave of the precision synthesiser (tuneable from 9.5 to 10.5 MHz) falls on the final frequency. The range involved runs from the 91st harmonic wave to the 100th. Since, owing to the pre-division ÷4, only harmonic waves divisible by 4 can be selected, we are left with harmonic waves 92, 96 and 100. In this ambiguous case, the lowest-order harmonic wave is selected. The harmonic wave to be selected is V = 92.

3. The interception synthesiser receives the precision synthesiser signal as a reference frequency. It is pre-set to tune the auxiliary VCO to V times its reference frequency. In the example, V = 92 is programmed.

4. Next, the reference divider factor, R, and the setting divider factor, U, of the precision synthesiser are determined. The auxiliary frequency generated by the precision synthesiser is equal to:

   $f_h = (f_{ref} \cdot U/R)/40$; moreover,

   $f_o = V \cdot f_h$ should also be true;

   this gives $f_o = V \cdot (f_{ref} \cdot U/R)40$;

   We can then freely select one of the divider factors, U and N. Suppose we pre-set U. R becomes:

   $$R = V \cdot (f_{ref} \cdot U/f_o)/40$$

   If we select U = N, the reference divider becomes;

   $$R = (V \cdot f_{ref})/40\Delta f$$

   The surprising thing is that only the expression $f_{ref}/40\ \Delta f$ has to be a whole number to make it possible to set the reference divider value. This can easily be arranged by selecting a suitable crystal frequency. You can actually go a step further and make the expression equal to a power of two. The multiplication is then reduced a displacement operation which can be handled by a micro-controller.

5. The internal phase monitoring frequency of the precision synthesiser is equal to:

$$f_p = f_{ref}/R = (40\ \Delta f)V;$$

In our case, f_p = 21.74 kHz

The micro-controller scans a table to find the control gradient which is appropriate for this standard frequency, and programmes the precision synthesiser's phase discriminator accordingly.

6. The precision synthesiser is programmed to the final frequency and the interception synthesiser is switched on.

7. After a response time of app. 200 msec., the interception synthesiser is switched off.

As can be seen from the derivation procedure (admittedly somewhat mathematical), controlling the synthesiser is not so complicated as it seems to be at first, and requires only a few lines of Assembler or C program.

The attractive point about this solution is that, with moderate phase jitter demands, we obtain a circuit concept which turns out not to need mixer/filter equipment in the control loop or inter-stage loops, and can therefore be added to and tested using amateur resources. And the risk of spurious effects is also relatively slight.

The key assembly for synthesisers in the process described is the sampling phase discriminator. As already mentioned in earlier articles, the sampling phase discriminator functions on the basis that, when a switch (in the form of an FET or a diode bridge) is briefly activated, the transient value of the input voltage is transmitted to a storage capacitor and retained there. If the switch pulses are synchronised with the input voltage, then a DC voltage arises in the storage capacitor, the level of which depends on the form of the input signal and the phase position. In the non-synchronised case, the mixed product of the input voltage and the switch pulses arises in the storage capacitor, and harmonic waves may be mixed into this up to high orders of magnitude. Fig. 6 shows the principal mode of operation. We can also interpret the circuit as a harmonic mixer at an intermediate frequency of zero.

4. DESCRIPTION OF CIRCUIT

4.1. Assemblies

The synthesiser consists of 3 assemblies, the VCO DB1NV 012, the main loop assembly (sampling PLL, interception stage and frequency control loop) DB1NV 014, and the precision synthesiser and cycle control assembly (computer section), DB1NV 015. Any other oscillator with a suitable frequency range can also be used as VCO as long as the tuning voltage it requires does not exceed 30 V.

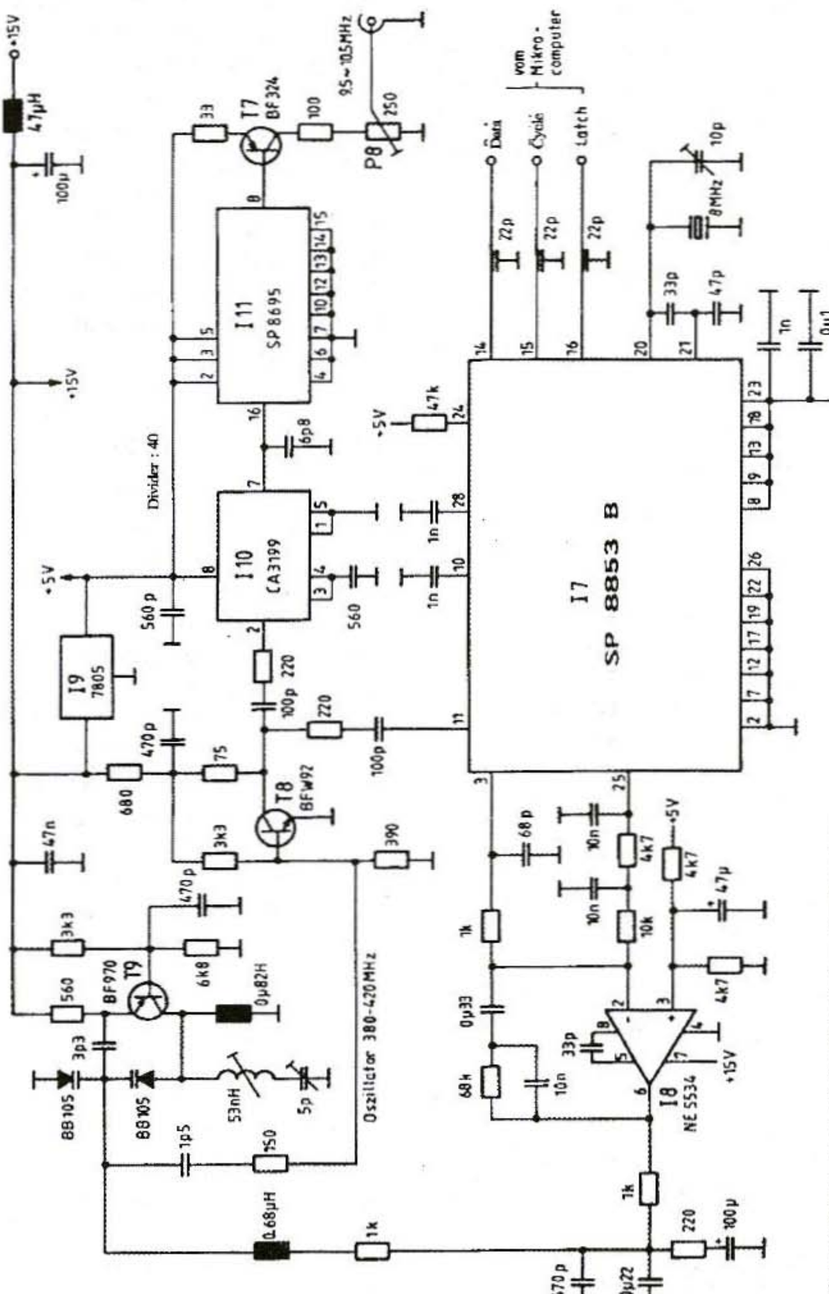


Fig.8: Circuit Diagram of the Precision Synthesiser

4.2. The Main Loop

Fig. 7 shows the entire main loop circuit. With a frequency stroke exceeding 200 kHz/component, the assembly acts as a conventional frequency control loop, as it did in the previous oscillator. With a frequency shift of 200 kHz/component or less, the second oscillator of the spectrum analyser is wobbled, and the first oscillator acts as a frequency synthesiser with a step size of 50 kHz. To make tuning simpler, the precision tuning range of the second oscillator is then brought down from approximately 3 -4 MHz to about 100 kHz, and the tuning resolution is thus improved.

The 450...1,300 MHz signal originating from output 2 of the VCO comes into the assembly and initially reaches a buffer stage, with a CF300, as per (1). Without the buffer stage, internal signals from the dividers, 1 and 16, go back into the VCO and from there into the intermediate-frequency section of the analyser, generating ghost signals. I1, a uPB582, divides the VCO frequency by four and energises the sampling phase discriminator with the diodes, D1 - D4. The Guanella transformer, TR1, de-symmetrises the symmetrical outputs of the uPB582, thus doubling the voltage shift at the sampling bridge input. The bridge circuit with the four Schottky diodes forms the actual sampling switch. In the normal case, the diodes are blocked, and are briefly switched on for about 300 psec. by means of a pulse applied through TR2, so that the transient value of the input voltage in the output-side 10 pF capacitor is "frozen". The BA481 Philips diode used can be replaced by any other low-capacity Schottky diode, e.g. the Hewlett-Packard 5082-2835. The voltage sampled is removed through a temperature-compensated source follower using T2 and T3. This source follower circuit is used in FET keys for oscilloscopes. The excitation stems from (2). Since double FET's can usually not be obtained by amateurs, two similar (best measured) BF245C's are used here.

4.3. The Trigger Pulse

Generating the narrow trigger pulse for the sampling bridge called for the use of the little grey cells. At first, the author used a circuit which had been utilised in some of the older Hewlett-Packard measuring equipment, but this was very expensive. The most important thing to come out of this was that you do not need an expensive step recovery diode to generate the pulse, because ordinary tuner switching diodes such as the BA243 and BA283 also display this effect. Remember the outline wiring diagram - the precision synthesiser controls the sampling phase discriminator. Its signal - 9.5 - 10.5 MHz and a level of about 0.5 Vss - arrives at the buffer and control stage via T6. Almost any dual gate MOSFET can be used here - e.g. the ancient type 40673. A smaller ring core transformer, with a 3:1 transmission ratio, provides the connection to the power stage through T4 and T5. The two transistors act as B-type push-pull power stages in the transmitter circuit and supply about 15 Vss to the step recovery diode, D5. The special characteristic of a step recovery diode is that following a current reversal from the conducting state, it initially remains conductive and is then "abruptly" blocked. The current drop within a few hundred picoseconds generates a half-sine voltage pulse in the resonant circuit consisting of L1 and the diode capacity, with an amplitude of 5 V and a width of approximately 300 psec.. The transformer, TR3, converts this into two zero-symmetrical pulses. The attempt to use these pulses directly to trigger the diodes did not succeed, as the diode bridge represents a non-linear resistor and produced serious pulse distortion effects. When the author investigated professional sampling stages from HP, Philips and Tektronix, he noticed that the symmetrical stepping lines between the step recovery diode and the diode bridge were unusually long and were often wound round ferrite ring cores. An attempt to use 20 cm. of twisted wrap-around wire with telephone insulation led immediately to a breakthrough. When the pulse is generated, the step recovery diode "sees" a symmetrical circuit with an impedance level of about 120 Ohms, since the reflection at the diode bridge has not yet run back through the circuit. If the pulse then switches the diode bridge on, it again sees a source impedance of 120 Ohms. This works as long as the doubled transit time in the circuit exceeds the pulse width. As TR2, the circuit is wound around a highly permeable ferrite or iron powder core. If, instead of the Teflon wire, we use enamelled wire, then the transmission loss for the circuit greatly increases. With normal wire enamel, losses are clearly too high in the gigaHertz range.

The connection to gate 2 of T6 has a special task. It provides for a slow increase in the sampling pulse amplitude, and thus a gradual release of the sampling PLL. If the equipment is switched on abruptly, it can happen that the sampling PLL jumps away from the pre-set scanning point and synchronises with an adjacent scanning point.

The output voltage of the sampling discriminator at the drain of T3 is about 1 V max., and the DC level is set at 2.5 V through P1 in the source line of T3. P1 should be a multi-speed trimmer, as the setting must be carried out very precisely. After the source follower, the control voltage passes through an RLC filter which can be set using P2 and arrives at the 13 A gain control amplifier through a 74HC4066 CMOS switch (12 A). The MC33078 dual operation amplifier from Motorola represents almost the ideal operational amplifier. It produces almost as little noise as special low-frequency opamps, and has a high amplifier band width product of 16 MHz. The input for the two operating voltages is synchronised to within 1 V, and the operating voltage outputs also oscillate less than 1 V. The control circuit proportional fraction can be set using the trimming potentiometer, P3. The VCO could be controlled directly through the output of the gain control amplifier. The experiments demonstrated that because of the curved tuning characteristic of the diode-tuned VCO the transient response and noise behaviour of the PLL at the lower and upper limit frequencies is markedly different. The tuning characteristic is therefore roughly equalised, using I3B and the diode network D10...D12. The non-linear low-pass filter at the output of I3B reduces the response time of the circuit and optimises the noise behaviour.

4.4. The Interception Circuit

As explained above, a phase control circuit with a sampling phase discriminator can engage with any harmonic wave of the sampling frequency. So an interception circuit, which leads the VCO to the correct scanning point, is needed for secure operation. Since no special requirements are laid down for this "interception synthesiser", it can take the form of a simple single-chip synthesiser, as used in television engineering.

In the present circuit, I6, an SDA3302 from Siemens, acts as the interception synthesiser. According to (3), it is suitable for typical input frequencies of 1,400 MHz, has an I^2C bus interface, and possesses 7 switching outputs, which here are used for control functions. The entire assembly can thus be controlled using the I^2C bus. Anyone wanting to know more about the I^2C bus, the standard bus for entertainment electronics, can find the necessary details, for example, in (4).

The SDA3302 has the input signal from the VCO at pin 15 and the precision synthesiser reference signal at pin 2. Pins 2 and 3 are actually intended for the connection of a 4-MHz crystal as a reference, but if the circuit is controlled from outside it still operates satisfactorily at reference frequencies of over 50 MHz. According to (5), Telefunken electronic can provide a pin-compatible IC with largely equivalent data, with the type description U6204B. The internal gain control amplifier of the SDA3302 is not used. The control signal from the phase discriminator travels through pin 1 to the CMOS switch, I2B, and then on to the gain control amplifier, I3A. Pins 8, 9 and 11 are the switching outputs referred to, which are used to control the CMOS switch. All the elements in the assembly control make use of the bus circuits SDA (pin 4) and SCL (pin 5) of I6.

The block-off capacitors at the control inputs of the CMOS switches, I2A to I2C, are important. They attenuate capacitive interference from digital interference factors entering the control loop.

The SDA3302 is used as a frequency divider in assembling the frequency control loop. The divided input signal can be switched through an I2C bus command to the pin 6 control output, and I6 can thus be operated as an adjustable frequency divider. A 74HC123 monoflop (I5) acts as a frequency/voltage converter, and the tuning voltage and the wobble sawtooth voltage are added together in I4B, as in previous frequency control loops. A basic fixed bias, which can be adjusted using P4, establishes the frequency closed circuit frequency for the minimum tuning voltage. After inversion in I4A, the aggregate signal reaches the gain control amplifier, I3A, via the CMOS switch. The control behaviour of the frequency control loop can be set using P5.

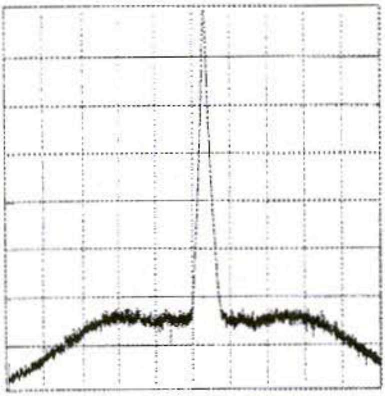


Fig.9: Noise Base using SP 8716 Divider

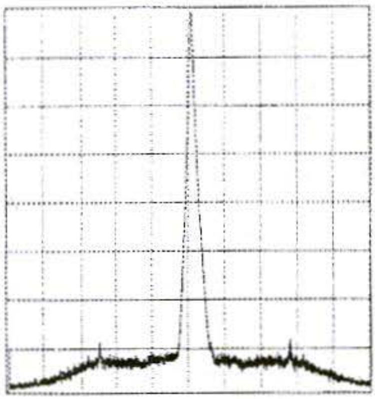


Fig.10: Smaller Noise Base, Divider Chain - CA3199 & SP8695

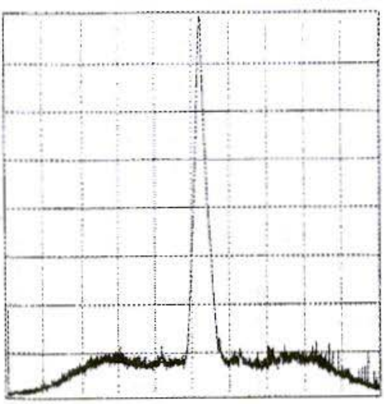


Fig.11: Signal from HP 8406A Standard Signal Generator

4.5. The Tuneable Reference Frequency

The precision synthesiser generates a reference frequency for the main loop which is tuneable between 9.5 and 10.5 MHz. Fig. 8 shows the detailed circuit. The synthesiser essentially consists of the SP8853 or SP8861. The two modules are identical except for the temperature range and the housing. The SP8853 is obtainable only with a ceramic housing, whereas the SP6681 is supplied in a 28-pin SMD housing (PLCC 28). Both circuits operate at input frequencies which can exceed 1.3 GHz and can be programmed through a serial three-wire bus. They offer a range of useful functions, only some of which are needed here:

- The phase discriminator can be programmed for control direction and control gradient
- The built-in dual modulus divider can be switched to :8/9 or :16/17
- The entire circuit can be switched off through the "power down" input
- Two frequency registers are available (e.g. transmission and reception frequencies), which can be selected using a switch pin

The precision synthesiser consists of the VCO with T9, which is tuneable between 380 and 420 MHz, the buffer amplifier, T8, the synthesiser module, I7, and the gain control amplifier, I8, and displays no peculiarities to distinguish it from a normal radio synthesiser. The synthesiser reference is an 8 MHz crystal. The buffered VCO signal

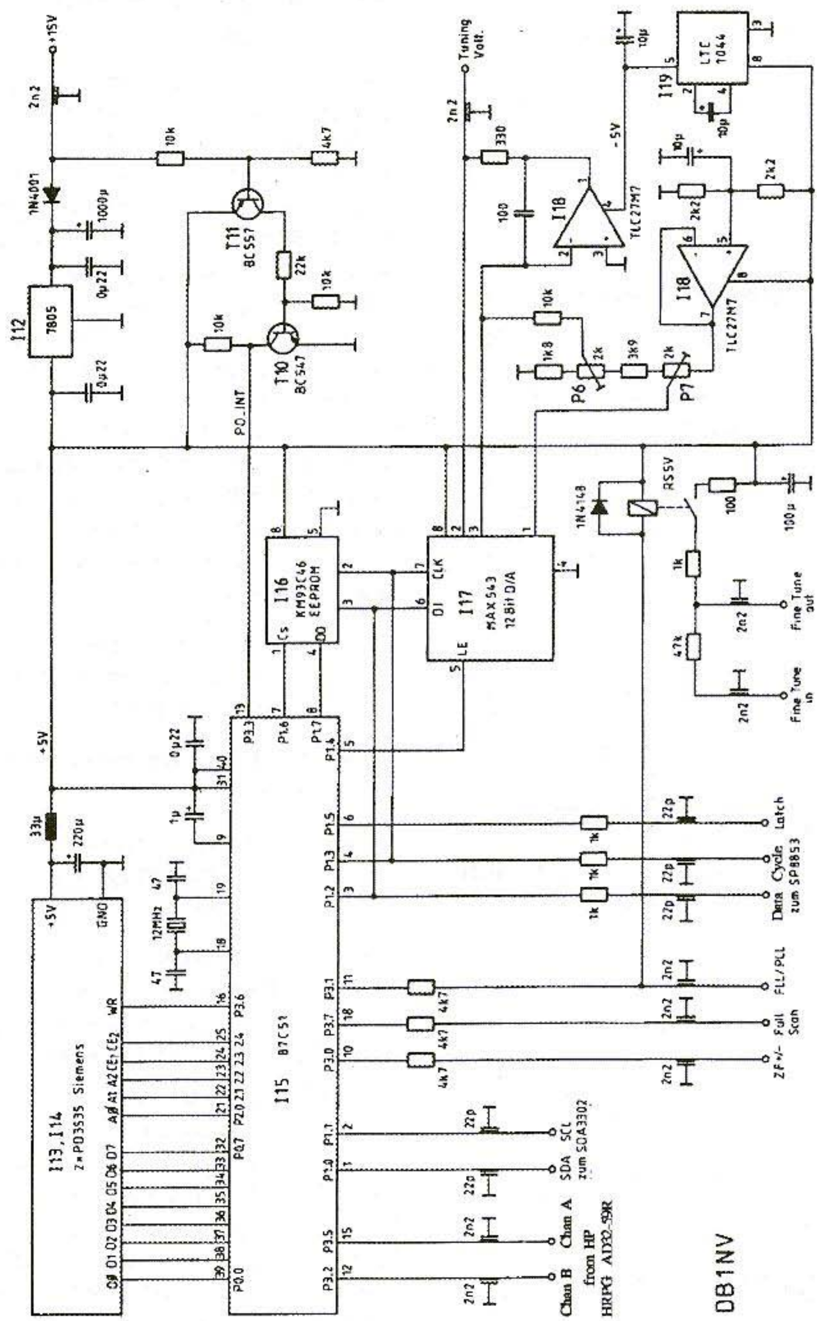


Fig.12: Synthesiser Cycle Control

is divided by 40 in I10 and I11. The transistor, T7, raises the ECL level to a maximum of 2 Vss. The unit dividing by 40 decides how low the noise level of the reference frequency generated will be. It has become apparent that ECL low power dividers, as they are used in normal synthesisers, create too much noise here, and so two decidedly venerable divider modules, the CA3199 and the SP8695, are brought into use. Since neither IC is now manufactured, and only left-over units can be obtained, the only long-term alternative is a combination of a divider:4, e.g. the SP8610, and a rapid TTL or CMOS divider, e.g. a 74F192.

To avoid disappointment due to the use of unsuitable divider modules, the influence of the divider chain on the noise characteristics of the synthesiser as a whole is documented in Figs. 9, 10 and 11. All the diagrams show the spectrum of a crystal oscillator at 60 MHz, with a representational width of 200 kHz/div (resolution 5 kHz, and video band width 1 kHz). In Fig. 9, the division from 400 to 10 MHz was undertaken using an SP8716, a low-power divider from Plessey. The noise base can be clearly observed. Fig. 10 shows a noise base which is about 7 dB lower. The divider chain, CA3199 and SP8695, was utilised here. Fig. 11 can be used as a comparison. Here the 10 MHz signal originates from an 8460A Hewlett-Packard standard signal generator, inside which it is generated, by division from a resonant cavity oscillator at 320 MHz.

These diagrams also illustrate the limitations of the circuit engineering used, which arise because of the multiplication of the 10 MHz reference signal with a signal-to-noise ratio of about -145 dB/Hz to the end frequency.

The precision synthesiser receives its control information through a serial three-wire interface. Pin 14 on I7 is the data signal. Pin 15 is the cycle and pin 16 the latch. Complete programming requires a 24-bit word and an 18-bit word, which are generated by the cycle control.

4.6. The Cycle Control

Fig. 12 shows the synthesiser cycle control. Even such a complicated control system can be created in the form of a micro-controller, using suitable software. The central module of the cycle control is I15, an 8-bit micro-controller of the 87C51 type, with 4K of EPROM and 128 bytes of RAM, with the current software occupying about 1.4 k of EPROM and 40 bytes of RAM. The micro-controller receives its input information from the following switches:

1. The incremental transmitter, connected to pin 12 and pin 15, plays the part of the main tuning knob. In the specimen unit, a relatively moderately priced optoelectronic incremental transmitter from Hewlett-Packard was used (type HRPG-AD32-59R), which supplies two square wave signals displaced through 90° with CMOS level (7). Other transmitters with "clean" signals (no mechanical transmitters!) should also be usable.

2. Other switch positions are covered by pins 10, 11 and 18. Pin 10 controls the addition or subtraction of the intermediate frequency of 463.5 MHz on the frequency display. If pin 10 is on 'High', the intermediate frequency is added. Pin 11 is used to switch between the synthesiser mode and the frequency control loop. A low level corresponds to the synthesiser mode. Pin 18 controls the average frequency. If it is set to 'Low', then the controller stores the previous frequency, switches to frequency control and selects 450 MHz. This is used in full sweep mode.

On the output side, the micro-controller controls an 8-character alpha-numeric LED frequency display, which consists of two PD3535 modules from Siemens. This is not the place for a precise description of these intelligent LED modules.

The interested reader is referred to the data sheets (8). LED modules of the same kind are also available from Hewlett-Packard. Since the modules strongly contaminate the power supply voltage, the LC filter (33 uH, 220 uF) should definitely be provided for in the 5-V circuit!

Pins 1 and 2 form the I^2C bus to control the main loop. It should be noted that these two circuits are wired up only with 22 pF feedthrough capacitors, since the usual blocking capacitors with 1 or 2.2 nF slur the data signals too much.

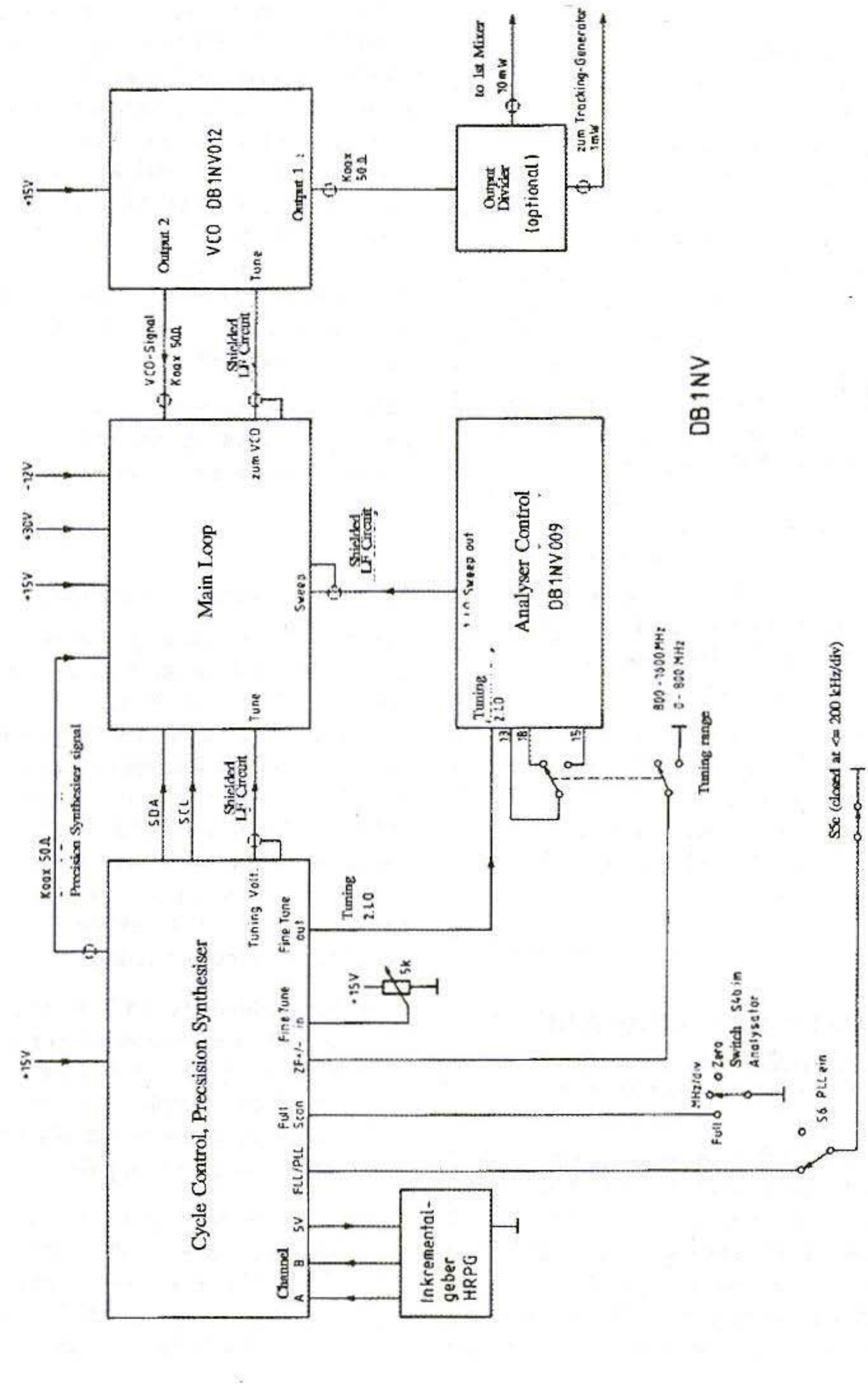
Pins 3, 4 and 6 represent the three-wire bus to the precision synthesiser. I17, a 12-bit digital/analogue converter, which generates the tuning voltage in frequency control loop mode, is connected to the same bus. I18 and I19 bring the output voltage of the D/A converter into the correct level range.

The precision tuning voltage for the second LO is looped through by the cycle control and the tuning range of the second oscillator is reduced by about 100 kHz in synthesiser mode by the relay. Thus a sufficiently precise tuning is possible between the 50-kHz steps of the synthesiser.

The EEPROM I16 has a special function. When it is switched off, the last average frequency entered is stored. A drop in the 15-V power supply is recognised by T10 and T11 and triggers an interrupt in the micro-controller. It initially covers the display with lines to save power and then saves some RAM contents into the EEPROM, which lasts for about 20 msec. The unusually large input capacitor of the I12 5-V controller (1,000 uF) thus keeps the 5-V power supply maintained for at least 50 msec..

After the writing procedure, the controller is also switched off, and can only be activated using a "Reset" switch. It then reads the EEPROM out first and resets the frequency which was set before the switch-off. This is not the place to describe the control software. Interested readers can ask for a program listing from the publishers.

Fig. 13 illustrates the interaction of the synthesiser assemblies and how they are integrated into the overall spectrum analyser circuit.



Interconnection of the Synthesiser Assemblies

5. ASSEMBLY

The author placed his assemblies on the known demonstration printed circuit boards with a continuous earth surface on the components side and solder islands on the solder side. The VCO is located in a Schuberth tinplate housing - 74 x 55 x 30 (mm.). The main loop comes into a housing with 148 x 74 x 30 (mm.) and the precision synthesiser and cycle control fit into a housing measuring 102 x 162 x 30 (mm.). The frequency display was similarly incorporated into a suitably sawn-out window of the housing, so that any trouble which may arise in relation to the perturbing radiation of the circuits between the display and the micro-controller disappears.

All high-frequency circuits are fed through SMB or SMC sleeves and all low-frequency circuits through feedthrough capacitors at 1 nF or 2.2 nF, unless otherwise noted in the circuit description.

6. COMMISSIONING AND CALIBRATION

First of all a warning - anyone who takes on the copying of such a complicated circuit as this synthesiser should have good high-frequency experience and at least some knowledge of measurement technique. Apart from absolute low-hum and low-noise laboratory network components (the author uses the E3610A and E3611A types from Hewlett-Packard, which are also interesting for the discriminating amateur), and a two-channel or, better, four-channel oscilloscope with a band width of 100 MHz, a meter going up to 1.5 GHz, with a 10-MHz time base output and a test oscillator going up to at least 500 MHz, is necessary. A good digital multimeter and a function generator should also be available.

Calibration requires much patience, and should be repeated, for the sake of security, after a run-in phase of some weeks.

6.1. Calibrating the Main Loop

The main loop assembly requires the most calibration work. First the voltages of + 15 V, + 30 V and - 12 V are fed in and the "noise test" is carried out. If the current consumption at 15 V is less than 200 mA and less than 30 mA at + 30 V and - 12 V, at least comparatively large short-circuits are not occurring. If the + 5 V are also present at pin 8 of the uPB582, a few DC checks can be undertaken:

- There should be 2.5 V at the gate of T2. The measuring point is very highly resistive! If the voltage falls sharply, one of the sampling diodes is possibly connected to the wrong pole.
- The voltage at pin 3 of I3A is measured. Due to the resistance tolerances, it will not be exactly 2.5 V. Then the voltage is measured at the drain of T3, and is set

precisely (!) to the value just measured, using P1.

- All other potentiometers are set to the central position.

- The standard signal generator is now connected to the VCO input at 400 MHz and 0 dBm and the oscilloscope or the frequency meter is used to check whether the uPB582 supplies 100 MHz at its outputs. 10 MHz is fed to the precision synthesiser input, and about 0.5 Vss out from the meter time base. If we now connect the cathode of D6 to + 5 V, then there will be a distorted 10 MHz oscillation value at 15 Vss at the collectors of T5 and T4. At the drain of T3, the oscilloscope indicates a beat frequency with an amplitude of 0.5 to 1 Vss, at a DC level of 2.5 V. The beat changes with the test oscillator tuning, and is repeated every 40 MHz. If nothing happens, check the pulse generator at D5. A normal oscilloscope can also be used to demonstrate the existence of a sampling pulse. Possibly only the connections of T2 are connected to the wrong poles, so that the pulse blocks the diode bridge again instead of sampling it.

- Potentiometers P2 and P3 serve to optimise the control loop noise, and are set to the lowest and flattest noise base only when the equipment in the spectrum analyser is ready.

- For further measurements, the cycle control and the VCO must be operative and connected up.

- The standard signal generator is set to 10 MHz and connected to the precision synthesiser input. The FLL/PLL input is earthed, so that the synthesiser element is activated. If we now introduce a frequency at the cycle control which gives us a synthesiser frequency which is divisible by 40 - e.g. 176.5 MHz (with the programmed intermediate frequency of 463.5 MHz, we then obtain 640 MHz), the corresponding frequency should then appear at the VCO output and should run with the standard signal generator tuning.

6.2. Main Loop Fault Location

If nothing happens and the VCO is at the upper or lower tuning limit, proceed as follows:

- Is the data communication to I6 working smoothly? It's frequently just a matter of SDA and SCL being mixed up. When the incremental transmitter is moving, the oscilloscope shows the unchanging bundles of pulses (the cycle) at SCL and the variable pulse specimens (the data) at SDA. Is pin 10 of I6 actually blank? It determines the bus address and must be open!

- Is there something wrong with the way the interception loop or the sampling discriminator is operating? To limit errors, earth pins 12 and 13 of the CMOS switch, I2, and give pin 5 a + 5 V charge. The interception loop is then permanently switched on and can be investigated at leisure. If the interception loop oscillates instead of engaging, then check the passive components at I3A to see if the correct elements have been fitted. If the sampling PLL is not operating correctly, then check the offset setting with P1 first.

6.3. Checking the Frequency Control Loop

If everything has gone well so far, the frequency control loop should be checked next. To do this, set the FLL/PLL input to "High" (leaving it open is sufficient) and feed a voltage variable between 0 and 3 V into the "Tune" input of the main loop. P4 is set in such a way that, with a tuning voltage amounting to -0.8 V, a VCO frequency is obtained of approximately 450 MHz, and with - 2.4 V we get 1,250 MHz. Precision calibration is carried out using the potentiometers in the cycle control. The VCO is now test-wobbled, using a 20 Hz delta signal from a function generator. The tuning voltage and the delta amplitude are set in such a way that the VCO is wobbled over the full range at a voltage of between 0.5 and 28 V. The delta at pin 1 of I3A should be approximately linear. This is a sign that the characteristic equalising using I3B and D10 - D13 suits the VCO being used. With other types of VCO, the resistance values should be matched to D10 - D13. No general rules can be laid down. The circuit reproduces a mirror image of the VCO characteristic from straight sections. The line gradient is determined with the longitudinal resistances at D10 to D13, while the voltage divider chain determines the application points of the straight sections.

Finally, the delta voltage is replaced by a 20 Hz low-amplitude square wave, and P5 is set in such a way that the VCO control follows the square wave rapidly and without spikes. This test should be carried out over the entire tuning range and the spikes should be minimised at the most critical point.

6.4. Putting the cycle control into operation

By comparison with the main loop, putting the cycle control loop and the precision synthesiser into operation is relatively easy.

The cycle control is simply tested by feeding in the operational voltage. The version number - e.g. "LO V3.02" - then appears briefly on the LED display, followed by the last frequency set. When the equipment is first switched on, there are no meaningful values in the EEPROM, and meaningless values appear - e.g. "1B57,000". Now turn the tuning knob to set a meaningful frequency. If the tuning knob reacts incorrectly to this, swap the two channels of the incremental transmitter over. In the PLL mode, the tuning step is 50 kHz, and in the frequency control loop mode 250 kHz. Rapid turning automatically increases the step size.

If the collector of T10 is earthed, lines appear on the display, and some bundles of pulses can be observed on the lines leading to EEPROM I16. This means the switch-off interrupt is functioning. To switch on again, switch the operating voltage off and on.

Every movement of the rotary transmitter must produce measurable bundles of pulses at pins 3 and 4 of I15. If the PLL is switched on, then similar bundles of pulses can be measured at pins 1 and 2 of I15. In frequency control loop mode, a negative DC voltage, which varies with the tuning and lies somewhere between - 0.8 and - 2.4 V, can be measured at the "Tuning Voltage" output. The quiescent value can be set using P6, and the voltage dispersion using P7. In the final calibration, the two potentiometers are in turn set in such a way that the displayed and actual input frequencies at the band ends match.

6.5. Cycle Control Breakdown Service

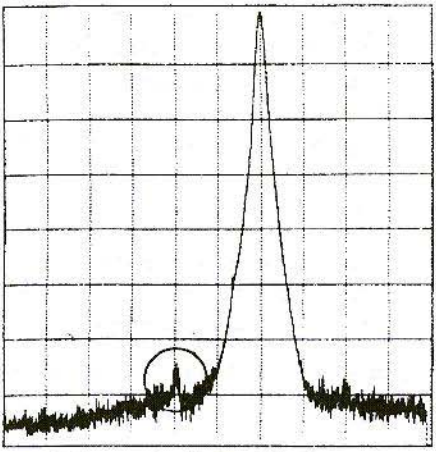
If nothing is happening at all:

- There is nothing on the display. Is the cycle oscillator of the micro-controller oscillating at pins 18 and 19? Is the reset working - i.e. is everything all right with the 1 uF capacitor at pin 9 of I15? Is the controller programmed, or has the programming been deleted by accident? Is pin 31 of I15 earthed or in the air? If so, the controller will try to switch on a non-existent external program memory.
- If there are lines on the screen, the switch-off interrupt has been triggered by an error at T10 or T11, or the current limitation is operating in the mains supply circuit.
- The rotary transmitter has no effect. Is there an interruption on the rotary transmitter's earth connection or on the 5 V supply?

It should be possible to isolate all other errors by following the direction of the line.

6.6. Putting the Precision Synthesiser into Operation

First calibrate the precision synthesiser's LC circuit at the collector of T9 in such a way that the VCO securely scans the range from 380 to 420 MHz with a 2 - 12 V tuning voltage. An external tuning voltage can be fed in for this test at the connection point of the two 1 kOhm resistances which lead from I8 (NE5534) to the tuning diodes. The output-side potentiometer, P8, is set in such a way that the level in the main loop is between 0.5 and 1 Vss. For the entire unit, the crystal trimmer of the precision synthesiser is then set in such a way that the VCO frequency measured is identical to the frequency which is displayed, + 463.5 MHz, if the spectrum analyser is operating in the 0 - 800 MHz range. In other ways, the precision synthesiser is like a conventional radio equipment synthesiser, so that fault-finding should not be all that difficult. Should the tuning not operate smoothly, first check the serial data lines for short circuits or transpositions.



Test Object: Synthesiser LO

Date: 10.09.93
Average Freq: 50 MHz
Resolution: 1 kHz
Ref Level: -20dBm
Storage Mode
Display width: 10 kHz/div
Video Filter: 100 Hz

Fig.14: Example of Further Measurement Options

Here too, if a blocking capacitor which is too large has been selected by mistake, the data or cycle signals can be slurred in such a way that the data exchange does not function.

7. TEST READINGS, FIRST OPERATIONAL EXPERIENCES AND FINAL COMMENTS

The phase jitter of the synthesiser alone was investigated using an HP spectrum analyser with phase jitter measurement facilities, and with a carrier interval of 20 kHz readings of - 107 dBc were obtained at the lower band end and - 102 dBc/Hz at the upper band end, which corresponded well with estimates. Once the equipment had been permanently incorporated into the analyser, the noise behaviour became somewhat worse, which leads to the conclusion that perturbing interference effects are present which have not yet been recognised. The first operational experiences have been positive in every way. Fig.14 provides an example for expanding the measurement options. Here, using two standard signal generators at 50 MHz, a spurious signal was simulated, which lies 20 kHz away from the wanted signal and 70 dB below it, and which represents the limit of recognition.

The synthesiser described is also universally applicable for use with simple standard signal generators or broadband test receivers. The precision synthesiser part of the circuit can also be used as a normal radio synthesiser up to the 23-cm. band if the VCO is modified accordingly. To assist the spectrum analyser to achieve the frequency precision of a simple test receiver, we require only a suitable synchronisation circuit for the second local oscillator at 450 MHz.

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Carl G. Lodström SM6MOM/W6

A Practical Loop Antenna for HF

By now at least ten years has past since my friend Christer, SM4DZR, visited me from Sweden. He saw my little bundle of 8 mm Aluminium tubes, 2 m long, that had been lying around for some three years earlier. He commented that "one day I ought to tune one like that to make a HF antenna."

1. ANTENNA DESIGN

I thought it was a great idea. Took one of them, bent it around a trash can. American trash cans have a diameter close to 2m/π. I flattened the ends and nailed it to a little piece of wooden board for a stand, tuned it with a variable capacitor, 16-385pF. The coupling out to a 50Ω load was by connecting a large capacitor in series with the grounded end of the loop:

After some tweaking around with the value of the fixed capacitor a very good match was achieved at 9.695 MHz, the frequency of interest, Radio Sweden International. The antenna worked very well, but the match got much worse the more it was tuned away from this frequency.

(So far has it not been determined if it will work on 9.59 MHz - Radio Norway - even after re-tuning.)

I connected it to a signal generator via a directional coupler. (Mini-Circuits PDC-10-1 about $20, get one! good from 0.5 to 500 MHz). On the oscilloscope I could see how the reflected wave approached zero only around

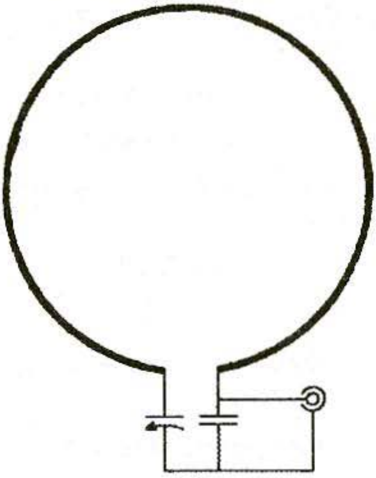

Fig.1: The first version with a Capacitive Divider

10 MHz., which was a pity since it worked well for the purpose, not much was done about it for a year or two. The nagging thought must have been there, though! Why did it only match in a small part of the tuneable band? One day it occurred to me that what I tuned (with the variable capacitor) was the loop, an inductance of course. The system becomes resistive at resonance. The loop should be divided down, as I had done, but not with a capacitor! I replaced it with a small inductor and there it was!

Testing it on various frequencies, modifying the small inductor, I found that with about 120nH it had a very good match (when tuned with the variable capacitor) from below 5 MHz to about 15 - 20 MHz. From 5 to 15 I claim the return loss to be about 30dB or better, a VSWR of less than 1.07:1 that is. It "decays" for some reason above 20 MHz. (Below 6 MHz, fixed capacitors are added to the variable).

I have found the loop itself to have an inductance of 1.85 µH. As mounted on the wooden board, there seems to be a capacitance between the ends of about 12pF, together with the 16pF minimum of the variable, that is a maximum operating frequency of 22 MHz, in very good agreement with observed performance. A smaller loop, or variable capacitor, would of course permit a higher operating frequency.

Also present in the model is a "resistance" related to the wavelength squared, about 14.5kΩ (in parallel or 2.25Ω in series) at 10 MHz. As the tuned frequency is varied, this resistance varies, and the antenna Q remains fairly constant at about 80. The high Q has a definite advantage in acting as a preselector for not so sophisticated solid state receivers. They usually need all the help they can get to cope with strong signals all over the band.

2. TRANSMITTING WITH THE ANTENNA

Actually going live on air with the antenna is something I have not done. In the model it is very simple to feed in an AC signal and see what happens. The model shows the following voltages across and currents through will be present for inputs of 1W / 10 W at 50Ω: 58 / 168 V and 460 / 1370 mA.

A fairly large (older type) of BC receiver tuning capacitor should be able to take 168 V (240 V peak) without arcing, but that is probably about it. A QRP antenna in other words, unless you put a big vacuum capacitor there.

3. ADVANTAGES AND DISADVANTAGES

The mathematics involved in designing an antenna of this construction was developed by Mr.T.L.Fu (1) and is essentially as follows:

Power that an antenna can deliver to a receiver:

$$P = A_e \times S \qquad (1)$$

where:

A_e = the effective aperture of the antenna, and

S = the Poynting vector (or power density) of the radio waves.

$$S = |E|^2 / R_o \qquad (2)$$

where:

R_o = $120\pi\ \Omega$ (the characteristic impedance of free space) and

E = the electric field strength.

For antennas the effective aperture and directivity relates like:

$$D = (4\pi/\lambda^2)A_e \qquad (3)$$

where:

D = antenna directivity

λ = wavelength

Combining Equations 1 to 3 yields the expression:

$$P = (\lambda^2/4\pi)D(|E|^2/R_o)$$

$$= (\lambda^2|E^2|/480\pi^2)D \qquad (4)$$

For loop antennas the directivity has been shown as:

$$D = D_o\ n\ k \qquad (5)$$

where:

D_o = directivity of a loss-less loop antenna,

n = the loop efficiency, and

k = the matching loss between the antenna and the receiver.

By reciprocity, loop efficiency can be derived by treating the antenna as a radiator. Assuming that current *i* flows through the loop, the loop efficiency is given by:

n = radiated power/power from source

$$= i^2R_r/i^2(R_r+R_l) = R_r(R_r+R_l) \qquad (6)$$

where:

R_r = radiation resistance, and

R_l = loss resistance.

The radiation resistance is directly related to the loop area by:

$$R_r = 20[(2\pi/\lambda)^2A]^2 \qquad (7)$$

Applying (7) to this antenna, R_r is 150E-6, 2.41E-3 and 56.3E-3Ω at 5, 10 and 22 MHz respective.

As calculated earlier, the loop current at 1 W input is 460 mA. For 10 MHz, for example, with 2.41mΩ the radiated power is 0.51 mW! -33 dB! A disastrous efficiency. Good enough reason not to use it for transmission.

For receiving it works great. There is apparently a margin of much more than 33dB in the received signals, as compared to a "perfect antenna!"

The efficiency can of course not be compared to a full size beam, but neither can the size. This can be especially detrimental when transmitting. Compared to large antennas, the received signals can be expected to be weaker, but so is the noise. A lot of times low signal strength is not the problem in reception. According to experience and (2) the background noise is high anyway on HF, making signal to noise ratio more important than signal strength.

Small antennas (relative to he wavelength) like wires and whips are "probes in the electric field." This antenna is a "probe in the magnetic field." A radio wave is comprised by both an electric and a magnetic component. Near the ground or walls with re-enforcement iron bars, the electric field is reflected to some degree. The incident wave and the reflex are out of phase cancelling each other. The magnetic components add to each other near a reflective surface. An obvious advantage for a probe in the magnetic field.

One evening, driving back from Sunnyvale, CA, I even received Radio Sweden very well with this antenna and a Yaesu FRG-7 receiver in the front seat of my solid 1964 Oldsmobile car! Power line noise resulting from the electric supply companies inability to maintain their insulators seem to be more of an electric than magnetic nature near the lines. Inside the near field another advantage -.unless you want to listen to it.

Since most electric distribution companies still believe in hanging their lines in the air, as opposed to putting a cable in the ground, this advantage is valid almost everywhere..

Noise from power lines or neon signs can sometimes be quite a problem. With this antenna they can often be cancelled by simply turning it around and tilting it. Now you can hear really weak signals! In the evening both Radio Sweden International (during good conditions) and WWV on 10 MHz puts about 100µV on the receiver input with this antenna. I have been able to see WWV direct on the oscilloscope using just the antenna and fine tune my frequency standard with a Lissajou figure! Tuning an inductor with the oscilloscope input capacitance give an even better reading!

Later I rebuilt the loop, bending it into a square shape. I used a HP 606A signal generator instead of the trash can. Less of a heresy I thought. The performance is the same, but now it fits in a suitcase. Even later, I equipped a briefcase with a loop (one turn) of Litz wire in the lid and a variable capacitor mounted onto the side wall. A piece of RG-174 with a 3.5mm plug is coiled up and plugs into the SONY 7600 receiver antenna receptacle. Any conductor will do, a copper foil tape may be a good choice. Even the metal trim, if any, on the suitcase lid can be used. Make sure there is no closed loop of metal somewhere in the lid.

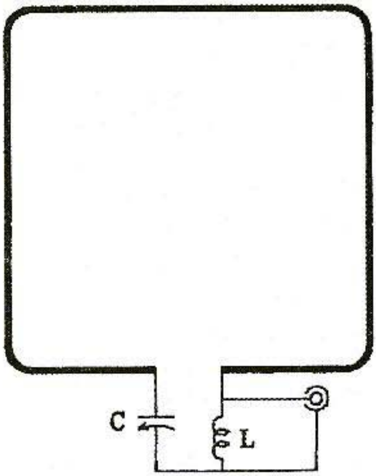

Fig.2: The improved version with an Inductive Divider

3. CONSTRUCTION

The dimensions of the tube are not all that important, but a 2 m long conductor, grounded at one end, will become resonant at about 37 MHz, so you cannot get any higher than that, in a simple mode.

Take what you have, wrap it around what you want, nail it to the bedpost or the antique oak dinner table in the living room! Anything for an audience! Go to a fleamarket and pick up one of these beautiful old capacitors. Stick it in the washer with the dinner dishes (I am not kidding) and oil the bearings as soon as you get it out of there. Contact oil on the wipers grounding the rotor. If the insulator is phenolic, not ceramic, glass or sapphire, it may take a few days for it to dry up, and you will not get the full Q of the antenna until then, but it will get there.

For the small coil, 4 - 5 turns on a 10 - 8mm ballpoint pen will do. This coil depends on the size of the loop, though. it is not particularly critical for reception, your receiver is probably not a very good 50Ω load anyway. For transmitting, tweak the coil until you can tune the loop with the variable capacitor to a perfect match. It is truly possible to get a perfect match.

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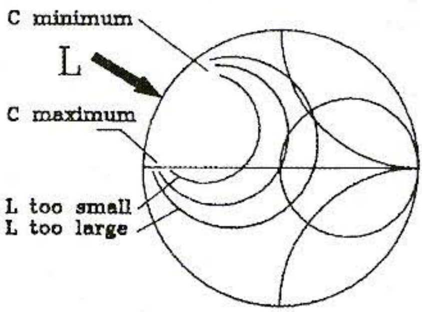

Fig.3: Effects on the Reflection Coefficient from Tuning and from varying the value of the Small Inductor

Wolfgang Schneider, DJ 8 ES

A Hybrid Antenna Switch for the 23cm Band

Even in amateur radio technology, the trend is towards compact, self-contained assemblies. The Mitsubishi MD004H hybrid module offers an interesting form of electronic hybrid antenna switch. When it is combined with a power amplifier and/or a pre-amplifier, there are many possible applications in the frequency range around 1.3 GHz.

1. PRINCIPLE OF ELECTRONIC HYBRID ANTENNA SWITCH OPERATION

Fig.1 shows the outline structure of an electronic hybrid antenna switch. Its operation can be explained in few words:

When the turn-on voltage, Us, is applied, both diodes (D1, D2) become conductive. The TX connection is switched through directly to the antenna. The short circuit which the diode, D2, generates at the RX connection is transformed by means of a $\lambda/4$ circuit. The antenna thus sees a high-ohmic RX connection.

Without the turn-on voltage, both diodes are high-ohmic. Thus the TX connection is cut, and the receiver is connected through the $\lambda/4$ circuit.

2. CHARACTERISTICS OF THE MD004H

The MD004H hybrid antenna switch module operates in the 1,200 to 1,300 MHz frequency range. The maximum power level that can be switched through is 50 W. A switching current amounting to 50 mA (Imax = 100 mA) is required for this purpose for the diodes.

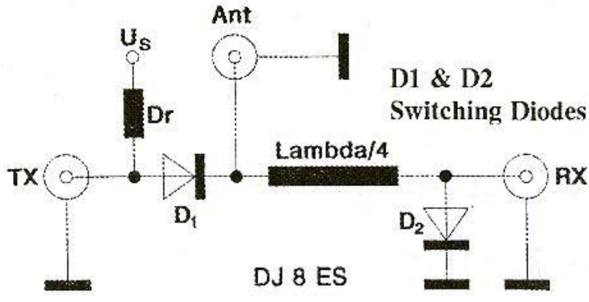

Fig.1: Basic Structure of an electronic Antenna Duplexer

Practical experiments show insertion damping levels of between 0.5 and 1 dB. This applies to both the transmission and the reception branches. This is thoroughly acceptable for many applications, such as, for example, ATV, PR, or even portable SSB stations.

The decisive advantage lies in the rapid conversion from a transmitter to a receiver. This is essential for certain applications.

For example, packet radio, a popular type of application, requires conversion times of less than 1 ms. Normal coax relays of this power class are all to be set to 50ms here.

3. PRACTICAL APPLICATION

A multitude of possible applications are conceivable. The author combines the hybrid antenna switch module with a type M57762 power amplifier (Mitsubishi) and a GaAs pre-amplifier (2 x MGF 1302).

The operating voltage for the power module is used for the transmission/reception conversion. It is fed through a protective resistor to set the switching current of the diodes (Is = 50 mA). The equations below can be used to calculate the resistance, Rv.

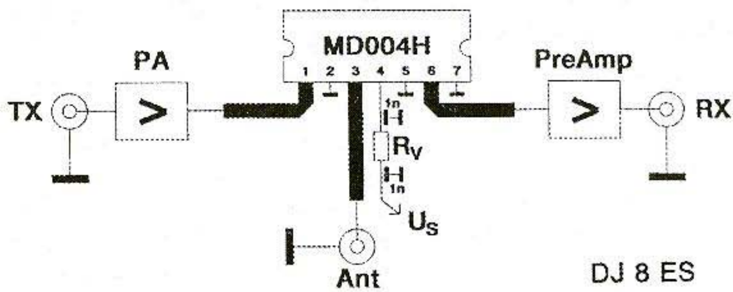

Fig.2: Switching the MD004H Hybrid Antenna Switch Module

RV = (Us - 1.4 V) / 50 mA

PV = (Us - 1.4 V) x 50 mA

The entire circuit, with a PA, a preamp and a switching module, is assembled as an open-air structure. The baseplate of a standard tinplate housing measuring 74 x 55.5 x 111 (mm.) is used as a carrier here.

Both modules are screwed on (don't forget the heat conducting paste!). The connections are made directly to N-cable terminal sleeves, using UT-141 semi-rigid cable. An advantageously priced alternative is the terminal sleeve for RG58 6 mm. cable.

It is advantageous if the operating voltages for the pre-amplifier and the high-level stage module are supplied using feedthrough capacitors. The control voltage can be measured off from the amplifier module through the pre-amplifier.

4. LITERATURE

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Angel Vilaseca, HB 9 SLV and Jean-Pierre Morel, HB 9 RKR

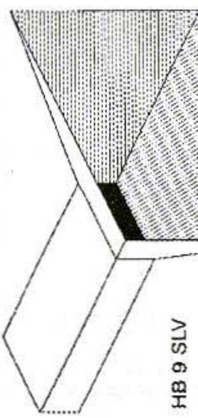
A Strip Line Antenna for 10 GHz

The manufacture of antenna always involves mechanical work, and antenna for 10 GHz are no exception. In order to construct, for example, a horn antenna (Fig.1), a parabolic type reflector or a slot waveguide radiator, you need a well-equipped mechanical workshop. By contrast, the antenna described here consists of nothing more than a printed circuit, and thus requires almost no mechanical machining.

1. DESCRIPTION AND TYPICAL APPLICATIONS

A strip line antenna can be used either alone or in conjunction with a parabolic type reflector. Indeed, the strip line antenna has a multitude of applications in conjunction with a parabolic type reflector. Depending on the number and

Fig.1: A Horn Antenna



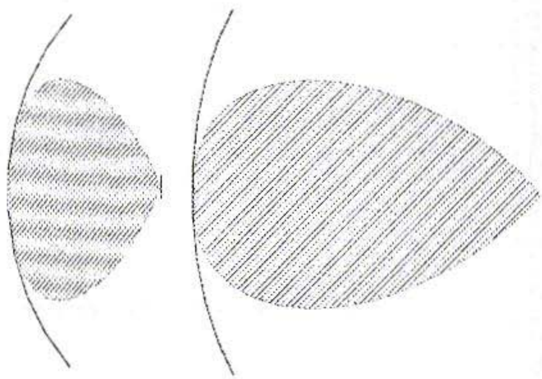


Fig2a: Deep Reflector: the Focal Distance is short, the F/D is small. The radiated field pattern of the exciter must be wide.
Fig.2b: Flat Reflector: longer focal distance, the F?D is large. The exciter must this irradiate at a narrow angle.

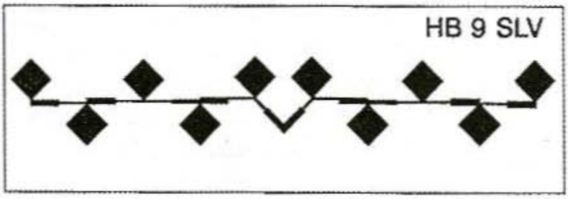


Fig.3:
The Antenna's Printed Circuit Board

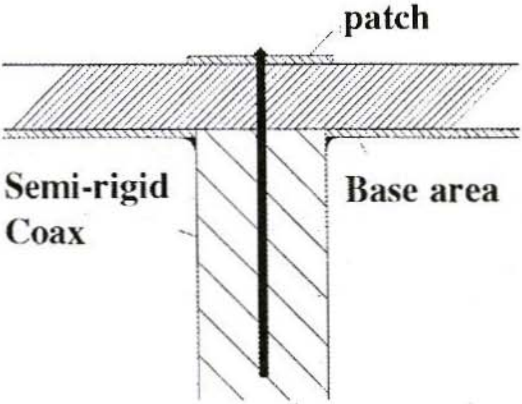


Fig.4:
Antenna powered through Coax Cables

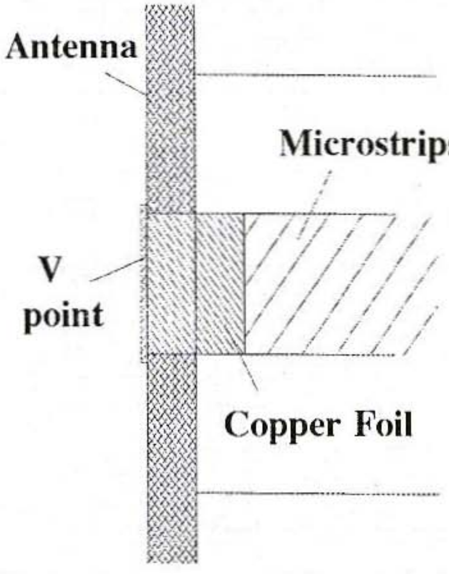


Fig.5a: Antenna powered through Strip Line

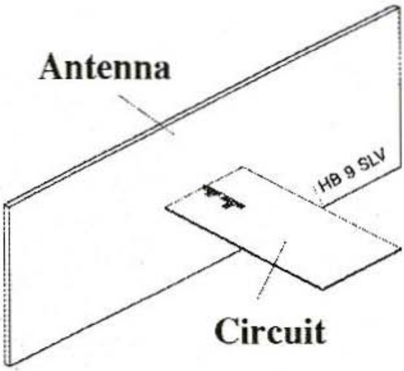


Fig.5b: Direct connection between a Circuit and the Antenna

layout of the elements, you can aim for a narrow or wide lobe, in order to be able to illuminate different parabolic type reflectors, flat or deep (Fig.2).

A further advantage of this type of aerial is that, from the point of view of weight and dimensions, it can be connected directly to the input of a receiver or the output of a transmitter. In this way, losses through wave guides or transmit-receive relays can be avoided.

Fig.3 shows the antenna. It consists of 10 diamond-shaped patches in form. The individual patches are connected by strip lines, the length and width of which bring about the desired impedance conversion and phase angle rotation.

The printed circuit material is naturally glass Teflon (Di-Clad Keene). Fig.3 shows the size and position of the individual patches and the connections between them, and lays down the total volume of the printed circuit board. The reverse side is entirely copper-coated and acts as an earth surface - as indeed is the case for all other strip line circuits.

The strip line antenna is excited in the centre of the printed circuit board at the point of the V, the impedance being 50Ω. A semi-rigid cable can thus be connected (Fig.4) or, as shown in Fig.5, the desired circuit can be soldered on directly.

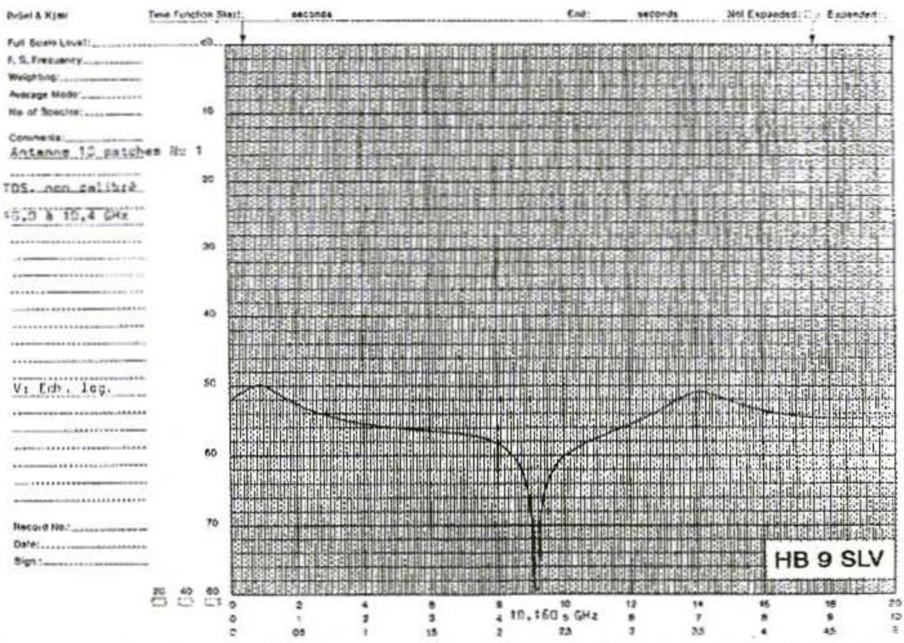


Fig.6: SWR of Antenna plotted against Frequency

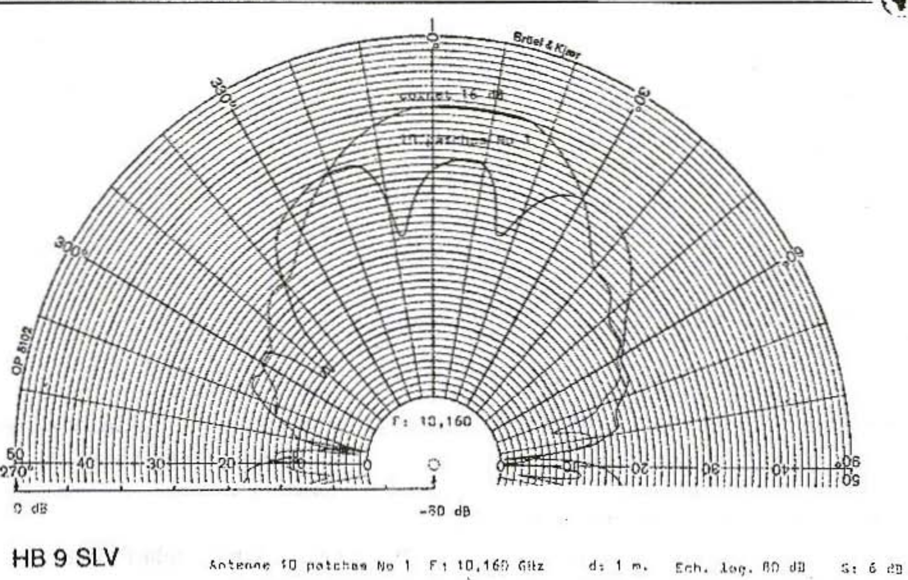


Fig.7: Stripline Antenna Radiated Field Pattern (cornet = cone)

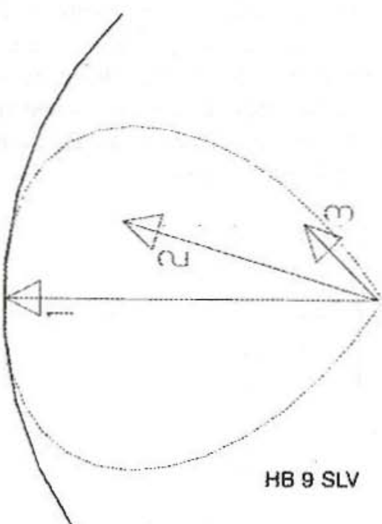
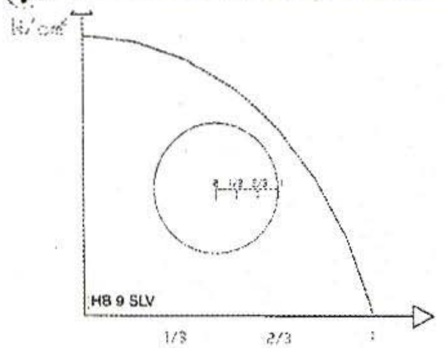


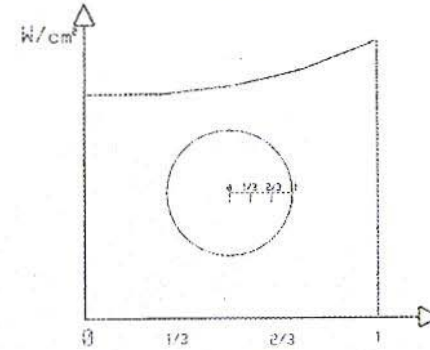
Fig.8: Example of Excitation of a Parabolic-type Reflector at various Angles:
1. 50% of HF; 2. 40% of HF; 3. 10% of HF

Fig.9: Area 1 corresponds to only 10% of the Exciter radiation. By contrast, Area 3 corresponds to half the total Area, but receives only 10% of the HF



Reflector radius

Fig.10: Distribution of radiation of a Horn plotted over the reflecter radius



Reflector radius

Fig.11: Distribution of radiation of an ideal Horn plotted over the reflecter radius

2. AERIAL MEASUREMENTS

2.1. The Standing Wave

This is a very narrow-band antenna. The best voltage standing wave ratio is 1:1 with an average frequency of 10.160 GHz (Fig.6).

2.2. The Radiated Field Pattern

Fig.7 shows the radiated field pattern for the strip line aerial with 10 patches. By contrast with a horn antenna with a 16dB gain, the strip line antenna is wider at the front. This leads to the conclusion that this type of antenna should be very efficient as an exciter for parabolic antennae.

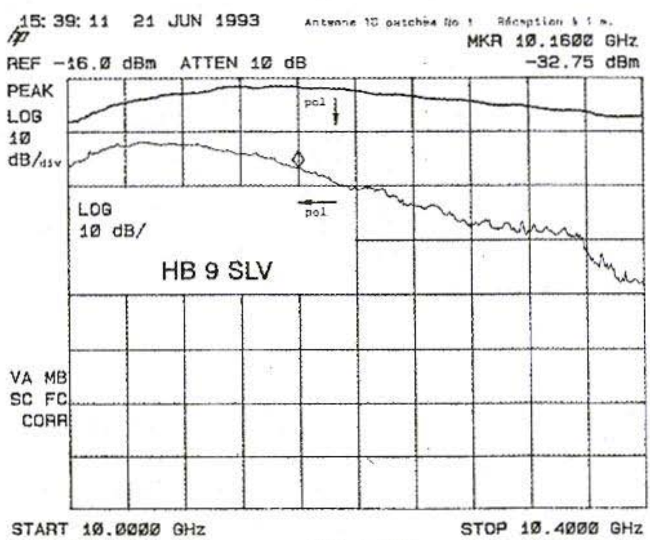


Fig.12: Measurement Curves
- Radiation plotted against Frequency
- Attenuation of Divergent Polarisation

Fig.8 shows that a parabolic type antenna normally receives more radiation from the exciter in the centre than on the edges - even though the surface away from the centre is considerably larger. For example, the central area of the parabolic type antenna marked "1" in Fig.8 can receive about 50% of the exciter radiation. By contrast, the area marked "3", which is several times larger (Fig.9), receives only about 10%.

To obtain greater efficiency, the excitation should be carried out as per Fig.10; i.e. the energy per square centimetre should be constant over the entire surface of the dish, and should very quickly become zero as it approaches the edge.

A strip line antenna with a curve in accordance with Fig.6 comes considerably closer to the ideal than a horn antenna. This would mean that a parabolic type reflector with a strip line aerial as an exciter has greater efficiency and achieves about 1dB more gain than a horn antenna (at about 16dB). Moreover, this layout produces fewer and less pronounced side lobes.

The number of patches plays an important role in the radiated field pattern of the aerial. As with a Yagi antenna, we can expect more patches to mean a narrower radiated field pattern (Fig.11).

2.3. Polarisation

Fig.12 shows that the maximum radiation for the antenna is obtained at 10.160 GHz, i.e. this is the frequency with the lowest SWR. It can also be observed that the attenuation of the cross-polarisation of the antenna ranges from a minimum value of app. 10dB at 10 GHz to considerably higher values - about 30dB at 10.4 GHz.

At the rated frequency of 10.16 GHz, the attenuation is approximately 15dB, which is very tight for a duplex circuit with varying levels of polarisation (Fig.13).

3. SUMMARY

This aerial demonstrates a concept which is still used by relatively few radio amateurs. Since the dimensions of the individual exciters at 10 GHz are very small, this technology is thoroughly suitable for this frequency range. A strip line antenna is also simple and cheap to construct. It's no accident that this type of antenna is often used for TV satellite installations.

(Fig.13 overleaf)

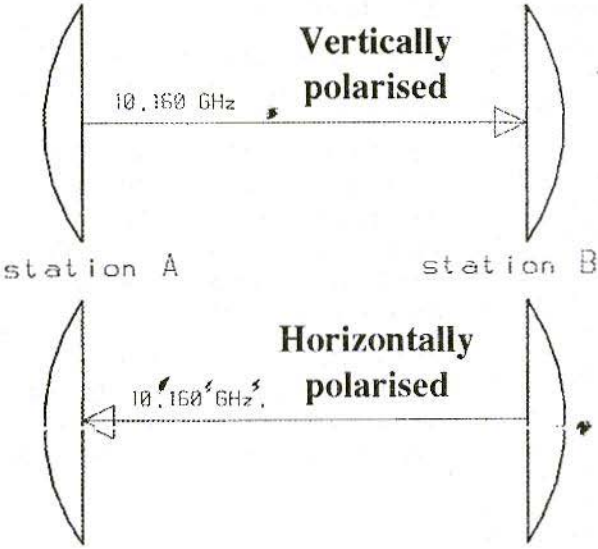


Fig.13:
A Duplex circuit with varying Polarisation

Michael Kuhne, DB 6 NT

A UHF-SHF Marker Generator

The marker generator described below is intended to simplify and facilitate the monitoring of frequencies and the calibration of SHF transverters in ranges from the 23cm band right up to the 3cm band.

1. INTRODUCTION

The active radio amateur is always faced by the same problem - the precise frequency of his or her SHF station has to be checked. However, all that is usually available is a frequency meter with a good 10 MHz oscillator - which can be used only on measurements up to 2 GHz.

Now, if a 10 GHz transverter has to be measured, things are already getting difficult. So a frequency which the frequency meter can still process has to be included in the multiplier chain. Very few units have provided for corresponding outputs at which measurements could be taken, so the unit has

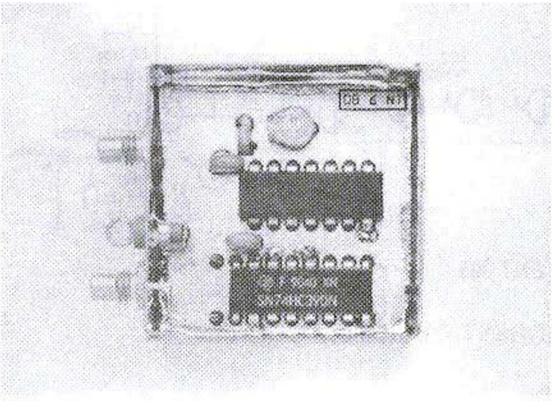


Fig.1: The Marker Generator

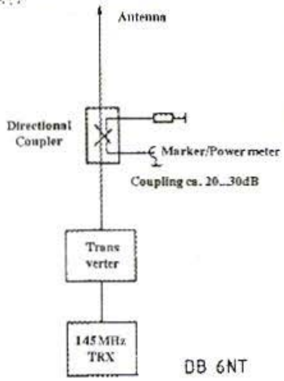


Fig.2: Measurement Rig with Directional Coupler

to be opened up for this purpose and possible taken apart, to provide better access to the individual assemblies. Receiving an SHF radio beacon is a considerably easier option. Of course, if there are no beacons within a suitable distance, it makes sense to use a marker generator (Fig.1).

2. USING THE GENERATOR

The marker uses an external 10 MHz timebase as a control source. Since, as already mentioned, commercial meters are often available with high-precision TCXO's or OCXO's, which almost always have a 10 MHz output, it is not necessary to take on the additional work of constructing a precise time-base.

At intervals of 1 MHz, the marker generates signals which can be received with a good signal-to-noise ratio up to 10 GHz. The precision of the marker stands or falls with the precision of the external 10 MHz timebase.

A directional coupler looped into the antenna feed (Fig.2) makes it possible to monitor the frequency even during the station's normal operation. To this end, the marker's signal is fed into the coupler. The output power of the station can thus also be monitored using the coupler.

Since the marker generates a wide frequency spectrum, you must also make sure that only the desired frequencies are received, without any of the intermediate frequency break-through. It would also make sense to use a suitable band filter between the marker and the test piece.

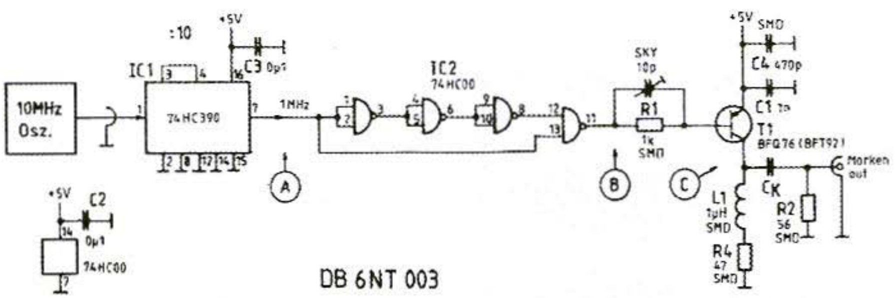


Fig.3: Circuit Diagram of the DB6NT 003 Marker Generator

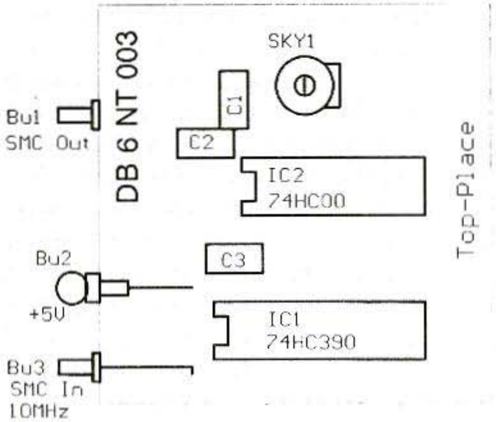


Fig.4: Component Overlay - top side

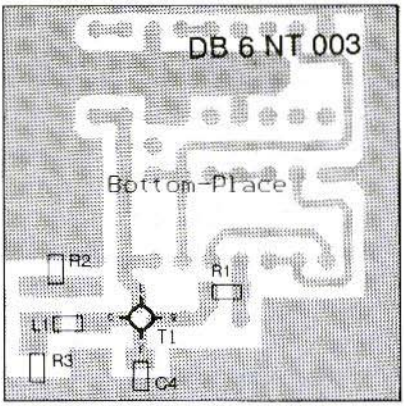


Fig.5: SMD Components - under side

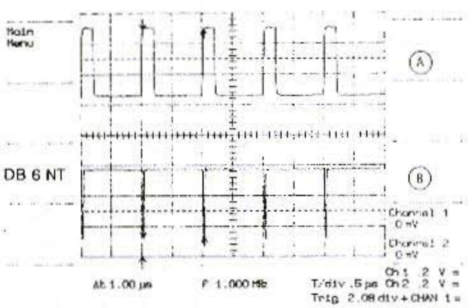


Fig6a: Signal at Test Points A and B

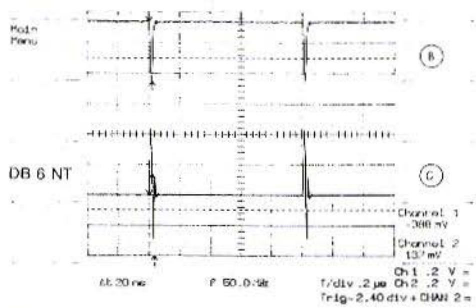
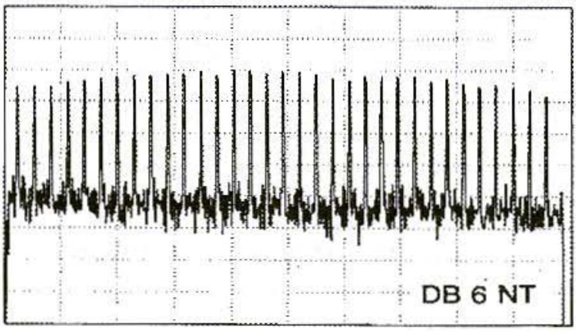


Fig6b: Signal at Test Points C and D

Spektralanalysator DB1NV, Version 2.04 vom 24.10.93, HPGL - Plot



Vertical:	5dB / Div
Display Width:	30 MHz
Centre Frequency:	2320 MHz
Resolution:	3 MHz / Div
Video Filter:	100 kHz

Fig.7: A 30 MHz span plot of the Marker Generator around 2320 MHz

3. CIRCUIT DESCRIPTION

The 10 MHz signal supplied from outside arrives at the divider, IC1, a 74HC390, which divides the signal down to 1 MHz (Fig.3). The subsequent gates of the IC2 (74HC00) generate spike pulses from it, which are fed through the RC combination, consisting of a 10pF Sky Trimmer and a 1kΩ SMD resistor, on the base of T1 (BFQ 76). Here, the transistor generates the harmonic waves desired in this case. These are decoupled through a printed capacitor, which acts as a high pass filter. The 10pF Sky-Trimmer serves merely to set a constant marker amplitude. Apart from that, the capacitor is fully screwed in (max. capacity 10pF).

If the marker is to be used only for frequencies below 3 GHz, the BFT 92 type, in a plastic housing, can also be used as the transistor, T1.

4. ASSEMBLY INSTRUCTIONS

The entire circuit is mounted on a small epoxy printed circuit board measuring 35 x 35 mm. The printed circuit board is soldered into a suitable tinplate housing (37 x 37 x 30 mm). The conventional components, the feedthrough capacitor and the two SMC connectors, are mounted as per Fig.4. Fig.5 shows how the SMD components are mounted on the foil side. When all the components have been mounted and when the operating voltage of + 5 V has been fed in, the equipment should be ready to operate. Fig's.6a and 6b show the measurement curves which are obtained at the points marked A, B and C on the wiring diagram. Fig.7 shown the marker generator output signal on a Spectrum Analyser.

5. COMPONENT LIST

IC1	74HC390N
IC2	74HC00N
T1	BFQ76 (BFT92) SMD
SKY1	10pF SKY trimming capacitor, black
C1	1nF RM2.5 ceramic capacitor
C2/C3	0.1µF RM2.5 ceramic capacitor
C4	470pF RM 2.5 ceramic capacitor
R1	1kΩ SMD
R2	56Ω SMD
R3	47Ω SMD
L1	1µH SMD choke

1 off Tinplate housing, 37 x 37 x 30mm
2 off SMC flanged connectors
1off 1nF feedthrough capacitor

Carl G. Lodstrom, SM6MOM/W6

Measurements on Resonance in Capacitors

I have found it very useful to take advantage of self resonance in regular capacitors for low impedance coupling of one point to another in VHF and UHF circuits.

In various articles I have seen claims that component lead inductances are of the order of 1nH/mm. I have had better luck with an estimate of 0.5 nH/mm.

1. INTRODUCTION

The idea is not new. The author of (1) showed this in a Wireless World article September 29th 1933! A graph in (1) shows the optimum decoupling capacitance, assuming a total length of 1 inch, for various frequencies. It boils down to 0.71 nH/mm, which is close to my findings.

Lately I read in a VHF Communications article that small surface mount capacitors are not good enough for decoupling, having too much inductance. Of course, they must have more inductance than disc capacitors mounted direct between ground and what is to be decoupled, but just how bad are they? My curiosity got peaked, so I decided to check this out, as well as to more accurately measure some regular capacitors with leads.

Whilst working for E.F.Johnson during the summer of 1976 I had access to, among many other goodies, a HP 4815A vector impedance meter. I was well aware of the fact that a physical capacitor will behave as an inductor, given a sufficiently high frequency, but had never thought much of the transition region. One day I was taking measurements with the 4815 and a few regular ceramic disc capacitors. I was struck by the pronounced resonance, and by how low the resistance became at resonance, often less than 1Ω.

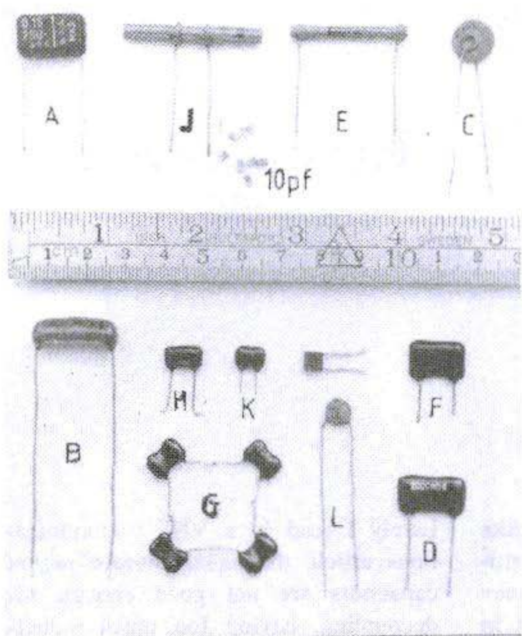


Fig.1: The Capacitors as detailed in Table.1

2. The Test 'Method'

This finding inspired a method to measure the self resonant frequencies of the ceramic trimmers manufactured at E.F.J. (At moderate frequencies, the circuit in (2) and (3) can also be used for this test).

At first I measured the resonant frequencies (from a selection of capacitors with leads, Fig.1) with a Grid Dip Meter. Both leads soldered in a parallel fashion to a ground plane, in one case I soldered two and four identical capacitors together and measured the resonant frequency of the resulting loop.

Later I used a piece of micro stripline, the 'Method' as mentioned above, shunted by the capacitor to be measured. The microstrip connected between a sweeper and a home-built μW RF Detector, or a 275 MHz Oscilloscope (HP 1725A) when the frequency allowed.

The results for the leaded capacotors are shown in Table.1 and give values in the range of 0.4-0.7 nH/mm. See Table 1.

After I made these measurements I got the idea to try a Siemens "MK" (if I remember the type name right) capacitor, 1.5nF/ 25OV. I t can be mounted on the pins and then upside down, eliminating the inductance of the pins. See Figs.2a and 2b.

I soldered it across a microstrip line as shown in Fig.2a, standing on the ends of the pins. It resonated at 44 MHz (suppressing the signal at the other end of the microstrip by 40 dB!) As I turned it upside down the resonance moved to 96 MHz, also with a 40 dB suppression. The body inductance must then be 1.83nH and the total (with pins) 8.72nH. The difference, 6.89nH divided by the length of two 5.5 mm legs yields 0.63nH/mm.

The body is 7.6 mm long; 0.24nH/mm. This constant may be applicable for the capacitors in Table 1 as well.

I then connected 9 chip capacitors of 10pF each, so that they formed a half circle standing on a ground plane. The most superb GD-Meter in the authors

	Res. MHz	Cap. (pF)	leg (mm)	body (mm)	L (nH)	nH/mm
A	1.91	154000	29	25	45.08715	.5432187
B	7.7	4800	60	20	89.00565	.6357546
C	17	2160	35	7	40.5778-1	.5269846
D	27	1000	19	11	34.74663	.7091150
E	30	480	30	25	58.63494	.6898229
F	74	300	14..5	8.5	15.41898	.4111727
G	98	220	10.5	6	11.98852	.4440191
H	91	180	13.5	6.3	16.99358	.5103178
I	75	120	31	9	37.52636	.5285403
K	138	100	12.5	3.5	13.30093	.4666994
L	69	98	42	8	54.28952	.5901035
					Average:	.5505226

Table 1: Self Resonance in Small Capacitors

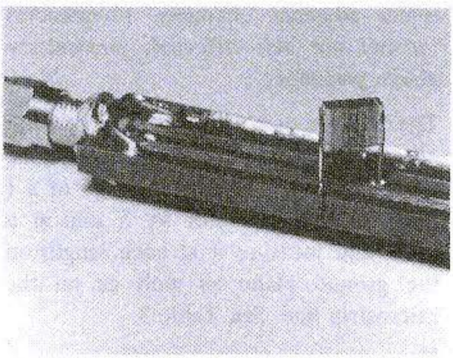


Fig.2a: The Siemens 1.5nF Capacitor with legs down

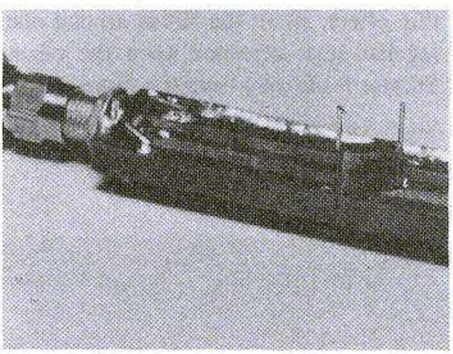


Fig.2b: The Siemens 1.5nF Capacitor with legs up

collection ends at 940 MHz, and indicated a possible resonance just above. With a signal source and an RF detector I placed a loop from each on either side of the arc (Fig.3). The resonance in the capacitors gave a very nice peak at 1010 MHz. The 3dB BW was about 10 MHz, indicating a Q of 100! Assuming no inductance in the ground plane, each capacitor should then have a parasitic inductance of 2.48nH. If the ground plane contributes 1nH, each capacitor would have 1.48nH. The distance between the grounded ends of the arc is 20 mm.

2.1 Introduce the microstrip line and the "Method", Fig. 4.

The 10pF chip capacitor resonated at 1210 MHz , consistent with an inductance of 1.73nH. Could 20mm of groundplane have added 0.75nH? resulting in 37.5 pH/mm?

The measurement confirms the statement in (1) as valid, but one may also

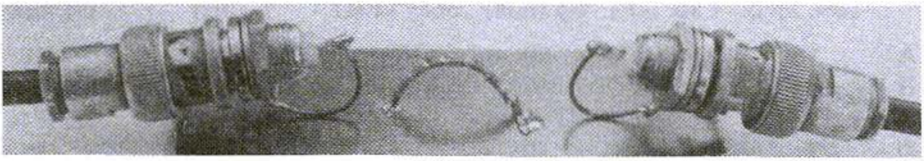


Fig.3: The 9 x 10pF Chip Capacitors and Coupling Loops

take advantage of the series resonance. Select the capacitance value that resonates with about 1.73nH! For a wider band decoupling, you can use several capacitors of varying value or added lead lengths. Just make sure they do not couple inductively to each other.

I tried five chip capacitors across the microstrip, and the results in are in Table 2.

	Res. MHz	Cap. (pF)	Ind. (nH)
M	1210	10	1.73
N	660	33	1.76
O	405	100	1.54
P	210	330	1.71
Q	140	1000	1.29

Table 2

The capacitor dimensions are 3.25 x 1.6 x 0.6 mm. The gap they bridged from the microstrip line to the ground plane was 2.7 mm, so about 0.6nH/mm is applicable here.

If they all really had 1.73nH, which is likely, the 100pF should have resonated at 383 MHz and the 1nF at 121 MHz. I have not investigated the deviations. The capacitors have 5% tolerance.

The dip at resonance, when introduced in the microstrip line, was between 30 and 40dB. This is a pretty good rejection, although not sharp enough to reject adjacent channels. Frequencies far out are also affected, particularly above resonance.

The measurements had so far not been all conclusive, so I decided to try one more approach: To cut the legs of a (long legged) capacitor by 5 mm at a time, and measure it at each length on the ground plane as well as on the microstrip line. See Table 3.

The specific inductance (nH/mm) seems to increase as the leads get shorter. Wider separation of the legs will have this effect, when the fields around one leg has less influence upon the other. Shorter leads may have the same effect.

Fig.5 shows an RF sweep with 5mm long legs on the 100pF capacitor.

Fig.4: The Micro Stripline with a Chip Capacitor mounted for measurement

Each leg Length (mm)	Res.freq. gnd.pl (MHz)	Res.freq. µ-strip (MHz)	Ind gnd.pl. (nH)	Ind µ-strip (nH)	Ind. gnd.pl. (nH/mm)	Ind. µ-strip (nH/mm)
43.67	67	66	55.321	57.010	.680	.701
40	69	68	52.160	53.705	.704	.725
35	74	74	45.349	45.349	.708	.708
30	81	81.5	37.850	37.387	.700	.692
25	88	88.5	32.068	31.706	.728	.720
20	101	100	24.344	24.833	.716	.730
15	109	115	20.901	18.777	.870	.782
10	137	128	13.231	15.157	.945	1.08
5	191	170	6.807	8.593	1.70	2.14

Note: 6mm of body path length is subtracted before the last two columns are calculated.

Table 3: Cutting down the legs of a 102pF disc capacitor

3. CONCLUSION

It is well known not to use excessively large de-coupling capacitors in RF work. They can be as harmful as not large enough. For applications operating at one frequency, or a band, the advantages of resonant decoupling, or coupling, can improve performance such as gain, stability or output power.

4. LITERATURE

(1) M. G. Scroggie, Radio Laboratory Handbook, 7th edition, pp 415-416

(2) D. Burchard, Crystal Testing, VHF Communications 3/93 p 163.

(3) D. Burchard, Don't be Afraid of High-Frequency Transformers, VHF Communications 2/94 p 99.

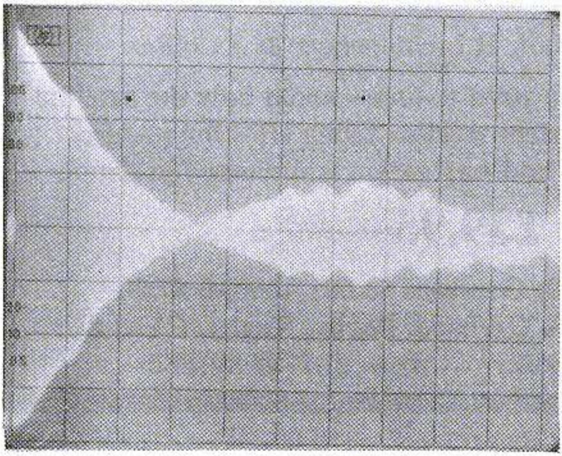


Fig.5: 0-500 MHz Sweep (RF into the Oscilloscope) with the cut-down (5mm legs left) 100pF Capacitor on the Micro Stripline, showing resonance at app. 170 MHz

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