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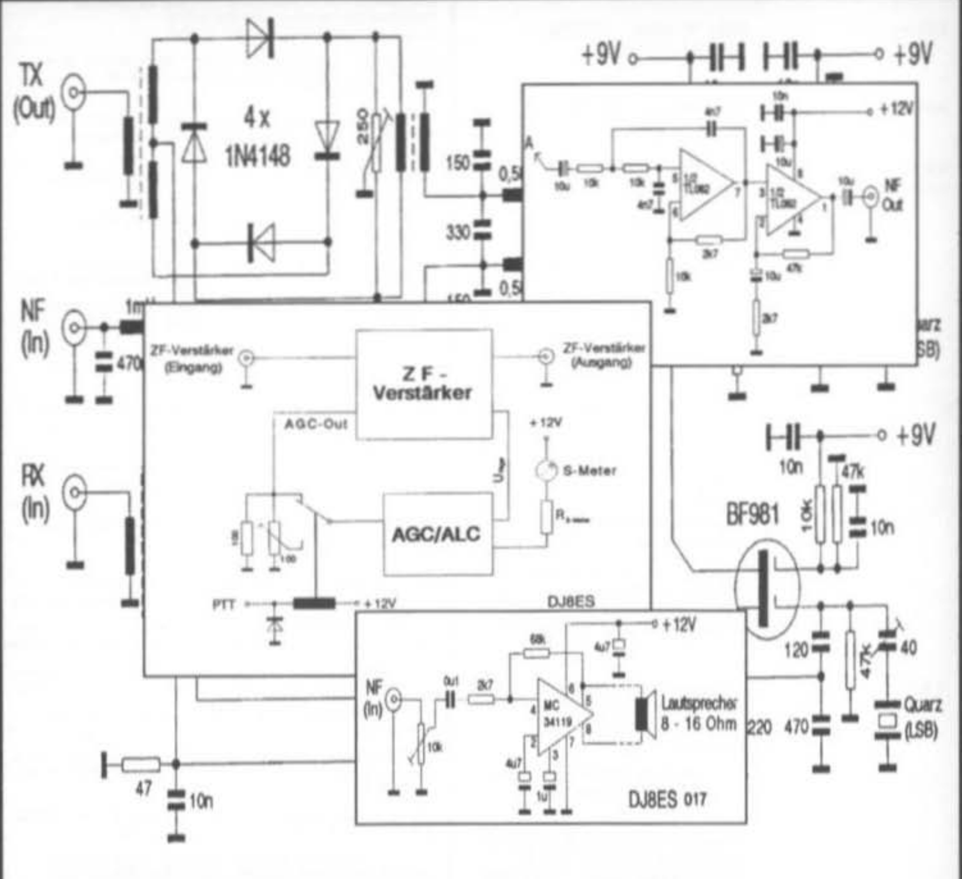
VHF COMMUNICATIONS

Volume No.25 . Summer . 2/1993 . £3.75



DJ8ES: SSB TRANSCEIVER FOR 50 MHz

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Volume No. 25 . Summer . Edition 2/1993

Published by: KM PUBLICATIONS, 5 Ware Orchard, Barby, Nr.Rugby, CV23 8UF, United Kingdom.

Publishers: KM PUBLICATIONS

Editors: Mike Wooding G6IQM, Krystyna Wooding

Advertising Manager: Mike Wooding G6IQM

VHF *COMMUNICATIONS* The international edition of the German publication UKW-Berichte is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is owned and published in the United Kingdom in Spring, Summer, Autumn and Winter by KM PUBLICATIONS.

The 1993 subscription price is £13.00. or national equivalent. Individual copies are available at £3.75, or national equivalent each. Subscriptions and orders of individual copies of the magazine should be addressed to the national representative, or - if not possible - directly to the publishers. All other enquiries should be addressed directly to the publishers.

Back issues, kits, as well as the blue plastic binders are obtainable from your national representative or direct from KM Publications.

Translated by: Inter-Ling Services Ltd., 16 Naseby Road, Rugby, CV22 5NH, U.K.

Printed in the United Kingdom by Apex Print, P.O. Box 111, 19 Church Street, Rugby, CV21 3PG.

Please address your orders or enquiries to your representative.

REPRESENTATIVES:

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FRANCE - Christianne Michèl F5SM, SM ELECTRONIC, 20 bis Avenue des Clairions, F-8900 AUXERRE, France. Telephone: (86) 46 96 59

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ELSEWHERE - KM PUBLICATIONS, address as for the U.K.

ISSN 0177-7505

Contents

KM Publications, 5 Ware Orchard, Barby, Rugby, CV23 8UF, U.K.

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Detlef Burchard, Box 14426, Nairobi, Kenya

MES-FETishism III

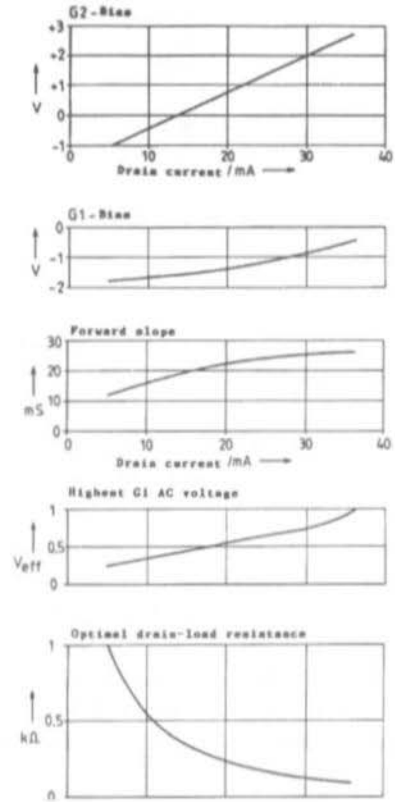
MES-FETs can be used for building power amplifiers too. Using simple parallel circuits there is no significant rise in the upper limiting frequency. An amplifier chain does allow us to press ever closer to the limiting frequency of slope. It also exhibits some "interesting" features with respect to cooling, feedback and matching.

1. POWER AMP'S HAVING NO MATCHING REQUIREMENTS

A MES-FET with the operating point I_D= 15mA, U_{G2S}= 0, S = 20ms and $R_{D\ opt}$= 330Ω can, when driven in the linear section of the characteristic curve, deliver about 20mW (+13dBm) on its drain resistor. This requires a control voltage of about $0.4V_{eff}$. At the same time one would reckon the drain AC voltage to be $2.5V_{eff}$, the voltage amplification being greater than six. Because of capacitance, the top frequency at the output is 800 MHz; at the input, where 330 ohms can be assumed, it amounts to 480 MHz. If we cascade stages of this kind, we soon fall short of the upper frequency of the VHF band (300 MHz). Circuit capacitance reduced the limiting frequency even further.

If we require a little more or less power, then a different operating point can be sought out by selecting a different G2 voltage. The settings for this linear operation can be taken from Fig.1, which shows an average CF300 device of group B.

Should significantly more power be required, two or more MES-FETs with roughly similar characteristic curves can be soldered on top of one another, as Fig.2 shows. Since they are not expensive, this can be more economic than looking for suitable components having higher power. Using this kind of circuitry the limiting frequencies can be maintained. Because of mutual heating, the technique reaches its limit at four devices.



a.

Fig.1: Determining the operating point (a) for Linear operation and AC values (b) for an aveage example of a CF300B with 6 to 8V Drain voltage

If a MES-FET works in a resonant circuit or in a selective network and if distortion of the drain current is not a problem, it can be driven up to a square wave. The amplitude of the fundamental wave current becomes virtually equal to the drain current; the operating resistance must be reduced to half. Under these conditions the MES-FET now produces 40mW (+16dBm) RF power, which may already be adequate for small transmitters. The efficiency of about 30% is similar to other devices of lower power.

All the same, the MES-FET requires only drive of wattless power up to high frequencies. Feedback is smaller too and we don't have to look for different devices, even with significant changes of frequency. The MES-FET will operate just as well at 900 MHz as at 27 MHz.

If the input of one of these low power amplifiers needs to be matched to the system resistance, a circuit as in Fig.3 can help. It has a 50Ω input, additional amplification and the noteworthy isolation between output and input of more than 60dB at 1 GHz.

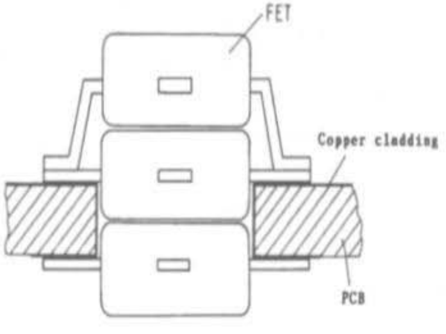


Fig.2: Parallel connection of several FETs

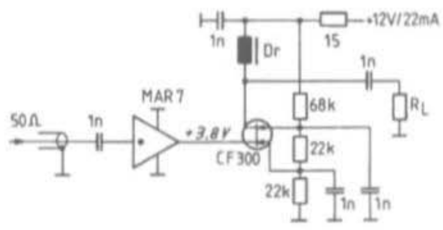


Fig.3: MES-FET PA with input matching

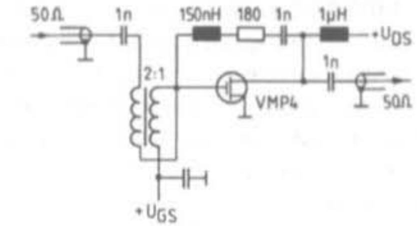


Fig.4: VMOS amplifier by Oxner (4)

2. POWER AMPLIFIERS REQUIRING MATCHING

It has already been mentioned earlier that a single MES-FET in a 50Ω system cannot give any amplification. Only when we use broadband or selective transformation on outputs and inputs does usable amplification arise. Broadband transformers can be produced in adequate quality up to around 500 MHz. They are not much help, however, when we want to advance MES-FETs into higher frequencies.

Of special significance in building receivers is an intermodulation-proof input amplifier. Modern receiver concepts with wide frequency ranges demand here broadband amplifiers with IM3 values of around +20dBm at the input. For the frequency range beneath 200 MHz the problem has been solved since Oxner (4) devised a concept with the VMOS transistor VMP4.

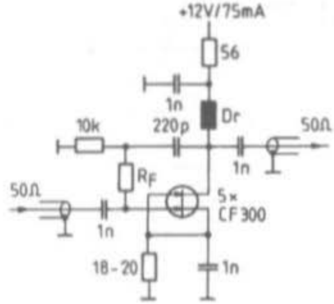


Fig.5: 12dB amplifier with 50Ω input and output

It turns up again in Meincke/Gundlach (3), albeit with different data for the same circuit (Fig.4).

A MES-FET circuit, which could do the same up to 1 GHz, would be desirable. Sadly, there are no power MOS-FETs for this application yet. We will try therefore to use five CF300 components in a circuit (Fig.5). The dimensioning follows the following formulae, which hold good for $R_{input} = R_{output} = Z_o$:

$$R_F = Z_o^2 \cdot S \qquad (1)$$

$$G = Z_o \cdot S - 1 = \frac{R_F}{Z_o} - 1 \qquad (2)$$

With S = 5 . 20ms = 100ms and Z_o = 50Ω R_F becomes 250Ω and G = 4(+12dB). The P_{-1dB} value on the input corresponds roughly to that of a single MES-FET

Data:	According to Oxner (4)	According to Meincke/Gundlach (3)
Amplification	12+/-0.5dB	10+/-0.5dB
Noise figure	2.4dB at 146 MHz	<7dB
Bandwidth	40-265 MHz	20-200 MHz
VSWR	No details	<2
P_{-1dB} (input)	+23dBm	+ 7dBm
IM3 (input)	+34dBm	+30dBm

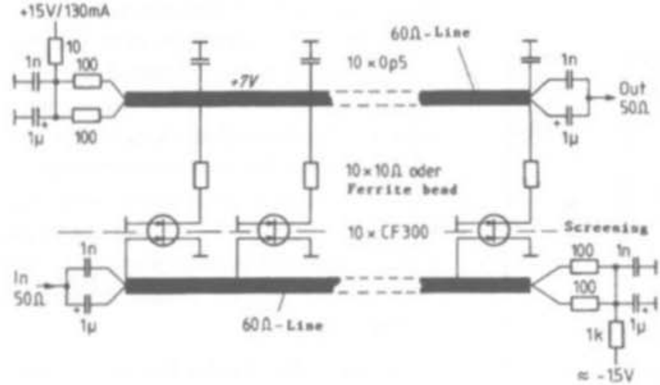


Fig.6: Amplifier chain; circuit principles

(+6dBm), as does the IM3 value (+16dBm). The limiting frequencies on the output and input are 1.5 GHz and 650 MHz respectively. A decrease of the first degree can be compensated with an inductance (approx. 15nH) in series with R_F and a frequency range of up to 1 GHz plus or minus 1dB can be achieved with construction that is otherwise favourable. The data, however, are still not so good as with a VMP4 below 200 MHz. All the same it is of advantage to avoid the input transformer, since the bandwidth limiting this causes is lost.

Theoretically it is possible to connect even more MES-FETs in parallel. This produces more slope or desired reserves for linearising the slope by means of source series resistors. This cannot be realised in practice because of problems with the heat produced. It is additionally possible to buy ready-made gain blocks with comparable specifications (e.g. Avantek MSA1104). For this reason another solution was sought.

3. AMPLIFIER CHAINS

This principle played a major role in the valve era because it represents the quasi-parallel connection of active elements, in which the S/C ratio increases proportionally with their number. We could, so to

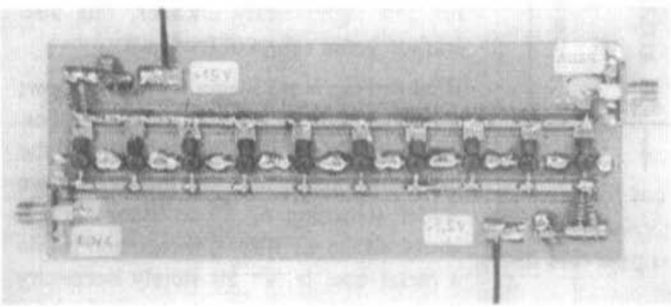


Fig.7: Amplifier chain, experimental construction using minimounts

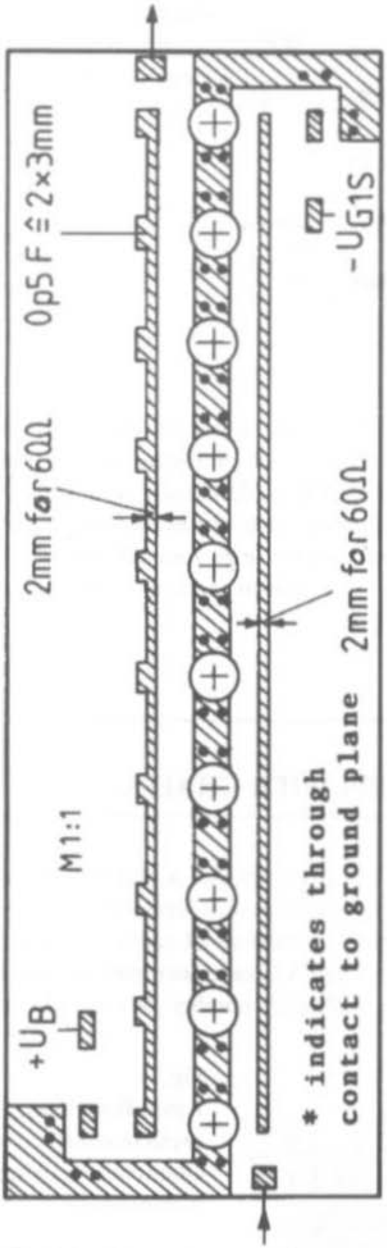


Fig.8: Circuit board layout for amplier chain. PCB material is 1.5mm Epoxy material, copperclad 35 microns on both sides

speak, advance to frequency bands for which no active elements exist. Anyone wanting to get better informed would do better to read older books (for example reference 2). In modern literature (3) you may find only a few sentences on amplifier chains. The principle can be seen from Fig.6. The MES-FETs are laid out in two rows so that their capacitances form a part of the track and no longer represent a problem.

There are defined rules for dimensioning amplifier chains, which are ignored here nevertheless. We will make use of the fact that a 60Ω line that is loaded every 15mm with 1pF forms a good approximation of a 50Ω line. So that the gate and drain conductors become exactly equal, additional 0.5pF capacitors are provided on the drains. An experimental construction on Minimount (Fig.7) produced on the first attempt a 3dB bandwidth of 1.4 GHz, giving vent to the suspicion that part of the attenuation measured arose from the not really suitable printed circuit board material used. Fig.8 shows the layout for epoxy PCB material. Construction on Teflon PCB material with SMD components is recommended if frequencies are to exceed 1 GHz significantly.

Siemens (1) give details of an amplifier chain with a frequency range above 2 GHz. It uses GaAs triodes. Using MES-FETs is not just significantly cheaper, this also leads to better values of feedback.

Heat sinking is no longer a problem now; we could pack them even closer together. There should not be coupling between the input and output line; for this reason we solder screening of 15 to 20mm height above all the CF300s. Construction inside a metal case is not absolutely necessary

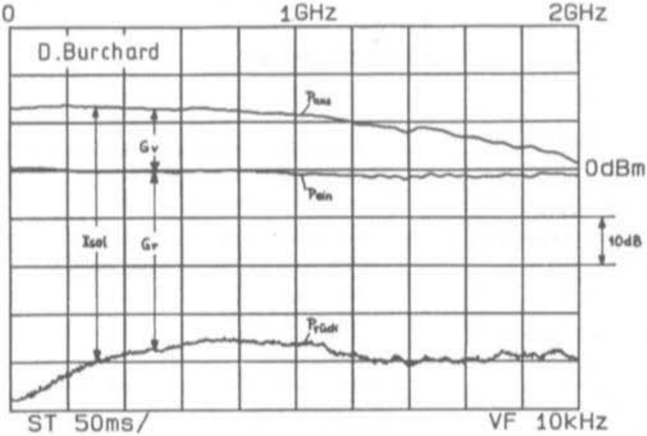


Fig.9: Frequency behaviour of the forward and reverse amplification of the amplifier in Fig.7

(the amplifier is stable and well-behaved in this open-top PCB construction) but modular cases would be recommended on account of the top frequencies produced.

Since my own measurement capabilities were inadequate, Mr N. Rohde of B&R Engineers was kind enough to undertake the tests of Fig.9, from which the forward and reverse amplification can be obtained.

Apart from the flat behaviour of the forward amplification (-1dB at 1 GHz), the very high reverse attenuation is also apparent. The isolation is 60dB at 100 MHz, 47dB at 1 GHz and still 41dB at 2 GHz. These are values directly related to the small G1 drain capacities of the MES-FET and definitely cannot be obtained with gain blocks.

In the scheme of Fig.6 the amplifier has a frequency range of six decades (1.5 kHz to 1.4 GHz); larger coupling capacitors could easily bring this down lower. The amplification is, in agreement with previous considerations, 13dB. It can deliver below 1 GHz with 1dB compression +22dBm (160mW).

This corresponds to a P_{-1dB} value of +10dBm on the input, whilst the IM3 value there is measured as +18dBm. Input and output VSWR are below 1.5 and could be better if chip resistors were substituted for the line terminations visible in Fig.7.

The noise of an amplifier chain is clearly lower than that of a single MES-FET. The equivalent noise resistances of the individual transistors are also connected in quasi-parallel.

With ten MES-FETs the resulting noise resistance is in the order of magnitude of just a few tens of ohms. Certainly the line terminations contribute to the noise to the degree that a maximum always appears at $n \cdot \lambda$ (n = 0, 1, 2, etc; λ is the wavelength of the line) and a minimum exactly between these. With the measurements given here, the noise figure decreases to 800 MHz, then up again. Measured values are 4.2dB at 150 MHz and 3.1dB at 435 MHz.

4. SUMMARY

Although MES-FETs are first and foremost small-signal devices, they still have usable large-signal properties. Transmitter final stages of low power and also receiver input amplifiers with useful P_{-1dB} and IM3 values can be constructed with them. using the principle of amplifier chains, quantities of them can be used together to increase amplification, reduce noise or create reserves for linearisation measures (source-series antiphase coupling). Results are produced which approach those of VMOS solutions in the VHF range.

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(2) H. Meincke and F. W. Gundlach (1968): Taschenbuch der Hochfrequenztechnik, 3rd edition, chapter N, pp 955-963 Springer-Verlag, Berlin

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Dr Freek M. Schimmel, Ede, Holland

A simple Panorama add-on for Weather Satellite Receivers

Anyone who constantly monitors the frequency range of polar-orbiting satellites (137-138 MHz) will find the panorama add-on described now a handy accessory. The precondition is a receiver with manual tuning by means of a VCO, in which the VCO is sufficiently stable.

1. DISCUSSION OF PRINCIPLES

The panorama add-on generates a series of sawtooth voltages which are combined with the tuning voltage in such a way that the oscillator frequency is swept symmetrically about the tuning frequency.

The results are displayed on an oscilloscope with X and Y inputs. The sawtooth voltage is applied to the X input and provides for the horizontal deflection. The output signal of the demodulator is applied to the Y input and controls the vertical deflection (Fig.1).

If a signal is received now in the swept frequency range, a sharp vertical point appears on the oscilloscope screen, the height giving a measure of the signal level. Whereabouts on the horizontal axis this

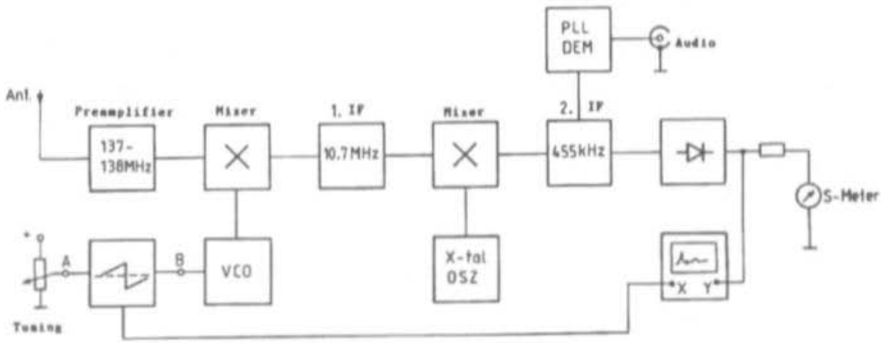
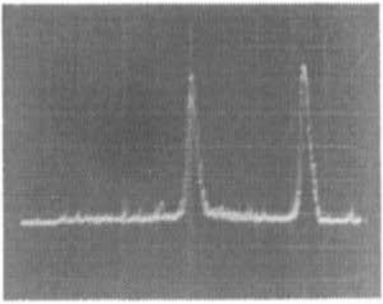


Fig.1:Block Diagram of the Receiver with Panorama add-on

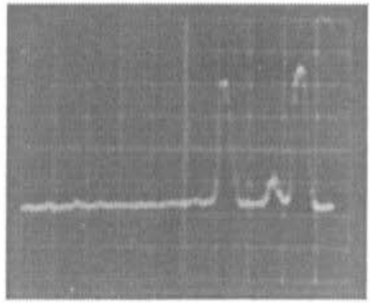
137.5 138.85



137.0 139.0

Fig.2a: Oscilloscope picture with two satellite signals

137.62 137.85



137.0 138.0

Fig.2b: Oscilloscope picture with three satellite signals

needle appears depends on the frequency of transmission, the tuning frequency, the frequency deviation and the linearity of the horizontal frequency scale.

If the frequency-to-voltage curve of the VCO is linear, the horizontal deflection and frequency deviation will be proportional. With a deviation of plus/minus 500 kHz and a maximum horizontal deflection on the 'scope screen of 10cm, one centimetre will always correspond to a range of 100 kHz. This gives per centimetre more than double the bandwidth needed for the polar-orbiting satellites in the range 137 to 138 MHz. If, for example, the receiver is tuned to 137.5 MHz, the satellites will be displayed side-by-side, conveniently separated (Fig.2).

METEOR3-4 (137.3 MHz) lies 2cm left of centre-screen, NOAA 10 and 12 (137.5 MHz) exactly in the middle, NOAA 9 and 11 (137.62 MHz) 1.2cm to the right of centre and METEOR2-19 (137.85 MHz) 3.5cm right from the centre.

Receivers having manual tuning and a large control knob generally use a potentiometer to vary the tuning voltage. In these cases it is quite easy to connect up the panorama add-on. The connecting wire between the wiper of the potentiometer and the VCO is simply severed and the add-on wired in-between.

Automatic frequency control cannot be used in this instance, so the AFC input is connected to deck. The tuning voltage required for a specific frequency in the range 137 to 138 MHz can be read off from Fig.3. The linearity for this application is quite sufficient. So, for example, we can take from the diagram that a frequency deviation of plus/minus 500 kHz will need a sawtooth voltage of around 0.8V peak to peak.

The diagram is obviously only valid for this special oscillator. That means that every other type of oscillator will need to have its frequency-voltage characteristics measured and drawn up as a diagram. After that you can determine the linearity and work out how large the sawtooth voltage needs to be for a defined deviation of plus/minus 500 kHz.

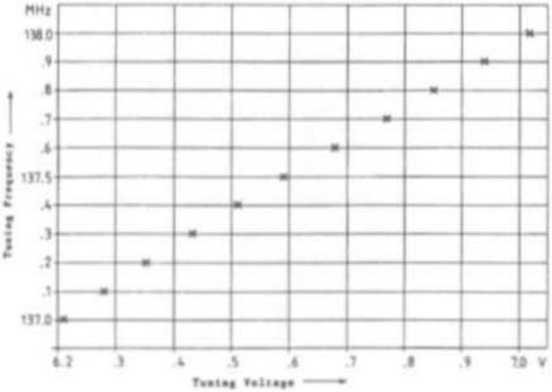


Fig.3: Frequency-to-Voltage characteristics of the Receiver

2. CIRCUIT DESCRIPTION

The forward/backward counters 4029 (IC2 and IC3), the resistor network and the NOR gate 4001 (IC4a) together make a sawtooth generator. The slope of the sawtooth flanks is very linear since they are formed of 256 steps of equal size.

The timer module 555 (IC1) produces the clock frequency, which passes via the analogue switch 4066 (IC9a) and the transistor stage TR1 to the clock inputs of IC2 and IC3. Switch SW1 (SWEEP ON) must be open so that analogue switches IC9a and IC9d are closed; the UP/DOWN inputs of the counter are then high. This has the result that a sawtooth voltage with positive flanks is produced.

Within certain limits the frequency of the sawtooth frequency is not fixed. A frequency of 16.6 Hz turned out to be useful, however, as this produced a relatively calm and flicker-free screen image on the oscilloscope even in the presence of 50 Hz and 100 Hz interference.

In terms of frequency stability, the choice of 4266 Hz as clock frequency places no special demands, since frequency deviation and horizontal deflection work in synchronism.

The clock frequency is set with trimpot P1. The sawtooth voltage passes through the op-amp IC5a working in non-inverting mode to reach inverter IC5b, which at the same time renders the signal symmetrical.

At the low-impedance X-output we now have a linear sawtooth voltage of plus/minus 2.5V for controlling the horizontal deflection. Setting symmetry is achieved with spindle potentiometer P2.

IC6a serves to attenuate the sawtooth signal, with step-switch S2 allowing three choices:

- +/- 500 Hz deviation for band monitoring
- +/- 150 Hz deviation for manual tuning
- +/- 30 Hz deviation for external frequency control.

The values of the trimmers P3, P4 and P5 are dependent on the slope of the measured voltage-to-frequency characteristic.

The tuning voltage of the receiver, removed from the wiper of the tuning control, passes the voltage follower of TR2 to the positive input of IC6b. The sym-

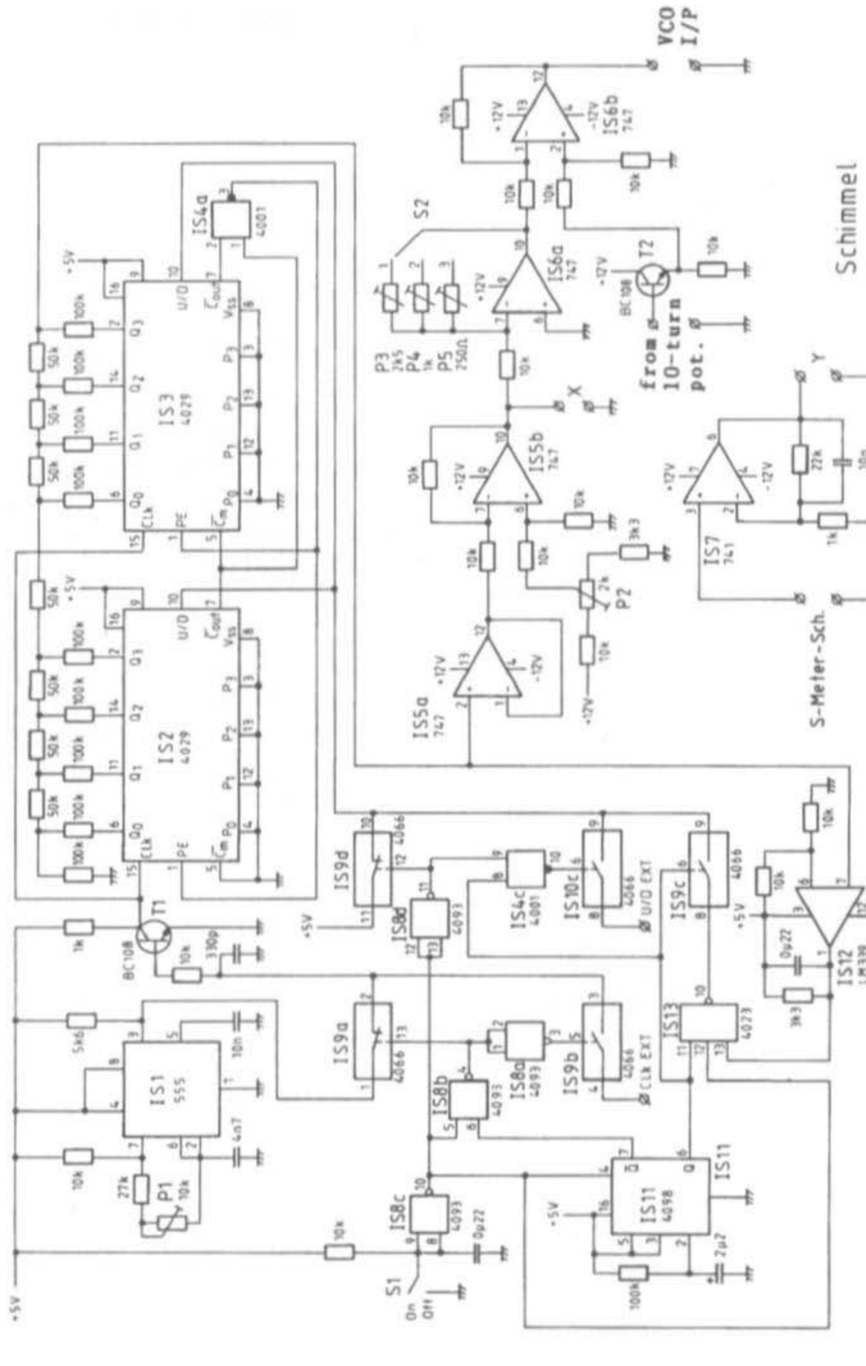


Fig.4: Circuit of the Panorama add-on

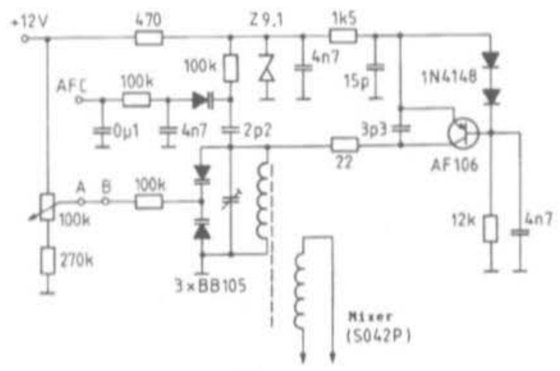


Fig.5: Circuit of theVCO

metrical sawtooth voltage connected to the other input goes to the differential amplifier, the output of which now has the tuning voltage with the sawtooth voltage overlaid. This signal is now led to the VCO where previously the wiper of the tuning control was previously connected direct.

The signal for vertical deflection is taken directly from the S- meter, amplified in IC7 and placed on the Y input (Fig.4).

If SW1 is closed (SWEEP OFF), IC9a and IC9d remain closed (for time t) as long as output Q of IC11 (4098) is low. At the same time IC9c is closed (Q is high) for the same interval t. Using comparator IC12 (LM339) the momentary value of the staircase voltage is compared with 2.5V. If a positive difference results (>0V) the UP/DOWN inputs of the counters are set low and the counters count backwards.

On the other hand, they will count forwards when the difference is negative (<0V). By means of this antiphase coupling the voltage on the X output becomes zero at the end of time period t. In the same way this voltage can assume a value of plus or minus (1/256) times 5 volts, that is plus or minus 20mV. This corresponds to a tuning error of +/- 1 kHz in switch setting 2 of SW2. This minor aberration can easily be compensated for by hand.

IC9a, 9c and 9d are opened at the end of time span t and the clock pulses from the 555 no longer reach the counter inputs. IC9b and IC10c are now closed. The counters can now be pulses forwards or backwards by pulses lying on the input CLK EXT. The direction of count depends on whether +5V or 0V is applied to the EXT U/D input.

If SW2 is in position 3, the frequency shift amounts to about 200 Hz per clock pulse. This simple digital frequency shift was the reason for using a digital, rather than analogue, sawtooth generator. The resistors around IC2, IC3, IC5, IC6 and IC12 should use metal film types on stability grounds.

The circuit can be built up on perf-board. The operating voltages should be found in the receiver as required.

3. ALIGNMENT

Offset alignment on the X output: put switch SW1 to position 1 (SWEEP OFF) and adjust P2 to 0V.

The display spot should now be exactly in the middle of the horizontal axis. Fine adjustment can be carried out easily with the horizontal shift of the 'scope.

The receiver is tuned to 137.5 MHz. The deviation is adjusted to +/-500 Hz with trimmer P3, retaining switch SW1 in the SWEEP ON position and SW2 on 1.

If no RF signal generator is available, a satellite signal can also be used direct. The X amplifier of the oscilloscope is adjusted so that the horizontal line of the picture matches the screen exactly.

Resistor values P3, P4 and P5 are proportional to the frequency deviation desired: so after adjustment of P3, the alignment of P4 and P5 is carried out with a resistance measurement set.

4. USER REPORT

Practical use has proved that the already eight years old weather satellite receiver of the author has sufficiently frequency stability. To avoid temperature drift following switch-on, the power supplies to the RF, mixer and oscillator stages of the receiver are never switched off. If the receiver is tuned to 137.5 MHz in the morning, then the signal of NOAA10/12, if in range, can be received all day without retuning. The possible Doppler shift is not taken into consideration.

The sole reason for checking the VCO for frequency instability is because a crystal oscillator is used as the second oscillator, converting 10.7 MHz to 455 kHz.

An improvement to the stability could be effected using the following measures:

- Stabilising the base voltage with two silicon diodes;
- Using metal film resistors and capacitors with small temperature coefficients;
- Adding a ten-turn potentiometer for

Dr.-Ing. Jochen Jirmann, DB 1 NV

Theory and Practice of the Frequency Synthesiser Part 1

Although nowadays frequency synthesisers form part of all radio equipment, it is clear that comparatively little is known about this technology in amateur circles. For older amateurs, in particular, there is even an aversion to synthesiser control transmitters. Here, the shock of the first "spurious wave synthesiser circuits" has probably gone too deep. It is the aim of this article to refresh basic knowledge and introduce some more recent developments, thus enabling the interested reader to analyse modern synthesiser concepts and to evaluate them by breaking them down into known functional units.

In the second part of the article, the part circuits of a synthesiser will be described in more detail and instructions are given for the dimensioning and the metrological investigation of synthesisers. The third part explains some interesting techniques, using two examples: a UHF synthesiser, which scans the range between 450 and 1,400 MHz in 50 kHz steps and can act as a local oscillator for the DB1NV 006...011 spectrum analyser, together with a fractional N-synthesiser, which can generate frequencies between 10 and 20 MHz with a resolution of 12.5 Hz.

1. REASONS FOR THE INTRODUCTION OF SYNTHESISER OSCILLATORS

The main reasons for the introduction of synthesiser oscillators in transmitters and receivers were the improved stability of the frequencies generated and the defined generation of frequency allocation schemes. In commercial radio, with established channels, this meant that operation was simplified if the complicated procedure for calibrating a VFO against a crystal calibrator (which also had to be repeated periodically) could be dispensed with. So the first synthesisers were already being developed in the 'fifties and were used in commercial short-wave radio, but there could be no thought of widespread use because of the valve technology.

With the progress in semi-conductor engineering and digital technology, new perspectives appeared in the use of synthesisers: in addition to the control of temperature drift and ageing, rapid changes to the generated frequency through hum or howl back could now be controlled, and the integrated digital circuits made it possible to replace the expensive mechanics of conventional VFO's, which meant equipment could become cheaper.

The expenditure on synthesisers for continuously variable transmitters and receivers was for a long time too great; here a combination of a course-stage synthesiser with an accurate VFO for interpolation between the grid points was and is the state of the art.

The advantages of a synthesiser as control oscillator can nevertheless be summarised as follows:

- → Accurate frequency setting in given channel grids, irrespective of temperature drift and component ageing
- → Frequency stability specified for only one component, the mother crystal
- → Possibility of synchronisation through an externally applied standard frequency
- → Control of interference due to hum stray effect or howl back
- → Equipment becomes simpler and cheaper due to replacement of expensive VFO mechanics.

The core of most synthesiser circuits is one or more PLL's (Phase Locked Loops). Describing the mathematics of PLL circuits would take some time and not everyone can understand it. We shall therefore content ourselves with considering a concrete example. Anyone who wishes to go into it more deeply will find sufficient information in the literature, e.g. in (7).

2. BASIC STRUCTURE AND CHARACTERISTICS OF A PHASE LOCKED LOOP

A PLL has a structure as per Fig.1:

A voltage-controlled oscillator or VCO generates the output signal. One part of the

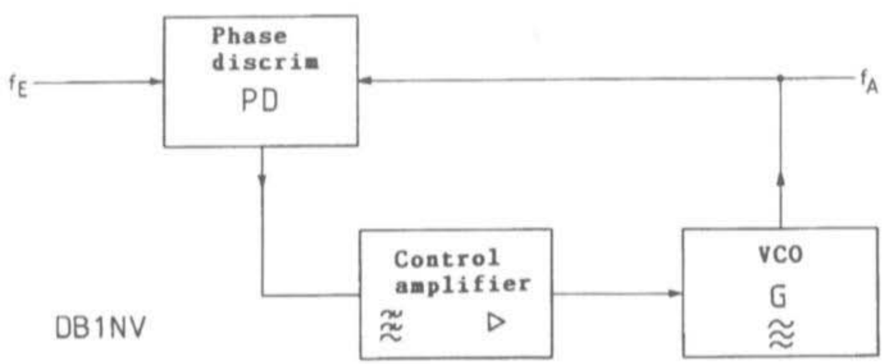


Fig.1: Basic structure of a Phase Locked Loop

output signal controls a phase monitor, to the second input of which is connected the input signal of the phase locked loop. The phase monitor or phase discriminator (often abbreviated to PD) supplies an output voltage, which is proportional to the phase difference of the two signals. An automatic volume control with low-pass characteristics controls the VCO in such a way that the phase displacement between the input signal and the VCO has a constant value (usually 0 degrees or 90 degrees).

At first sight, this circuit does not make much sense. The input signal comes out of the VCO again with the same frequency. However, a closer look reveals the following characteristics of the PLL control circuit:

- If the input signal has amplitude modulation, a suitable phase discriminator will ignore the AM and transmit only a frequency modulation (if present) to the VCO, i.e. the circuit works like a limiter amplifier.
- The frequency modulation of the input signal will be transmitted to the VCO only if the modulation frequency is lower than the cut-off of the control amplifier. This can be used to regain the carrier frequency from a frequency-modulated signal or, for example, to suppress unwanted high-frequency spurious modulations.
- The following behaviour with noisy input signals arises from what has just been said. A "slow" PLL circuit with a high-quality VCO with good noise characteristics will not follow the noisy input signal above its control cut-off. Only in the control range is the noise superposed on the VCO. This means that far away from the carrier the phase noise is determined by the high-quality VCO, and in the vicinity of the carrier by the input signal. This configuration is therefore suitable for removing wide-band noise and spurious modulations from the input signal. If, on the contrary, the VCO is poor, which generates a great deal of phase noise, for example, due to its wide tuning range, then its noise characteristics can be improved with a rapid PLL, provided a low-noise reference signal is available.
- With suitable phase discrimination it is also possible to synchronise the VCO to harmonics or subharmonics of the input signal. One application, for example, would be the synchronisation of an oscillator to the upper harmonic wave of a 1 MHz crystal oscillator. Here the phase control circuit forms a high-quality tuneable filter, the frequency of which is set by means of the pre-tuning of the VCO.

On the basis of these principles, the various PLL circuits can be divided into different application ranges in accordance with their control band widths:

❶ Extremely slow phase control circuits with control limit frequencies in the range of fractions of a Hertz are found in the preparation of standard frequencies. They can be used, for example, to link a VHF crystal oscillator, stable over brief periods, to a standard, stable over long periods, for example a rubidium standard, for drift compensation purposes.

A similar application is the synchronisation of a crystal oscillator with a normal frequency transmitter, e.g. DCF77. Here, however, the control must have a time constant of hours, so that the daytime propagation oscillations do not make themselves felt as a frequency modulation on the crystal frequency. The knowledge gained above can be well used here. The input signal of the phase control circuit has a spurious frequency modulation of which the lowest frequency is 1/day, since the propagation oscillations vary with the daily rhythm. In order to suppress this interference, an extremely low control limit frequency should be selected. For the crystal oscillator to attain a stability comparable with the DCF signal, it must itself have such a stable frequency that there is no noticeable drift in the course of a day. It can be seen that a link with a standard frequency transmitter is logical only if the local crystal oscillator is of correspondingly high quality. Anyone with a deeper interest in this topic will find further information, for example, in (1).

❷ Slow and medium-speed PLL control circuits with control band widths from a few Hertz up to app. 1 kHz are normally used to synchronise a free-wheeling LC oscillator with a crystal standard. In this way, temperature drift, ageing and, at higher control band widths, network hum and howl back too, are controlled. In this group, we find the standard synthesiser oscillators in radios. Above the control band width, the VCO can be frequency-modulated directly, which is normal for FM radios. Modulation fractions within the control band width are controlled by the PLL as disturbance variables.

The output spectrum of the synthesiser in the vicinity of the carrier (+/- about 100 Hz) is determined by the stability of the mother crystal, and further away (e.g. in the adjacent channel) by the short-time stability of the VCO, i.e. by the quality of the resonance circuit and the accuracy of the oscillator operating voltages. This discovery gives the lie to all those asserting that "it's all right if your VCO's crude, it will be kept on course by the mother crystal". If the VCO can be used without a PLL, but with just a tuning potentiometer and accurate tuning voltage for radio operation (apart from the temperature drift), then it's a successful construction.

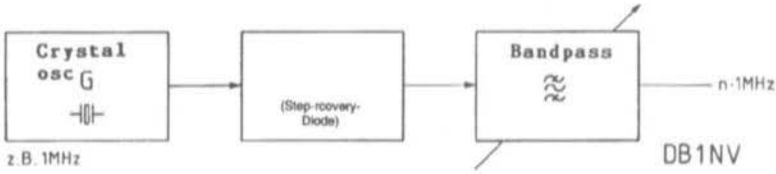


Fig.2: Frequency synthesis using the filtration process

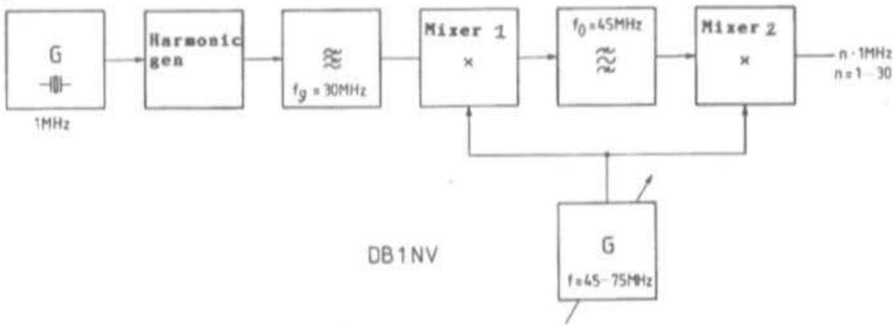


Fig.3: Expanded filtration process with inverse mixing

❸ Extremely rapid phase regulation circuits with regulation band widths of more than 100 kHz are used predominantly (apart from their use in digital transmission engineering) to connect oscillators with very broad tuning ranges to a low-noise reference. For one thing, VCO's tuneable over a broad band have low Q-values, and for another even the lowest noise and hum voltages act as spurious modulations. Synchronisation with a low-noise reference source through a broad-band PLL can improve the phase noise of the VCO by orders of magnitude. The main application of such regulation circuits lies in radio frequency measurement equipment, e.g. in synthesiser-tuned standard signal generators.

3. PROCESS FOR FREQUENCY SYNTHESIS

A frequency synthesiser circuit's job is to generate from an existing reference frequency (e.g. a crystal oscillator) an output frequency which can be adjusted in stages, the stability and reproducibility of which correspond to the reference. Phase locked loops are not necessarily compulsory here. The same end can also be achieved by means of frequency dividers, mixer stages and filters. Frequency synthesis processes involving the use of frequency division, mixing and reproduction are generally described as "direct synthesis", and processes using PLL's as "indirect synthesis".

So that the structure of complex synthesisers can be understood, the individual basic processes and the corresponding functional units are now presented:

3.1. The filtration process

For a long time, simple frequency synthesis processes have been used in short-wave receivers in which the oscillator signal is generated by mixing an accurate VFO with a reversible crystal ("band crystal"). This was done to avoid the VFO frequency range. This solution was acceptable for amateur equipment. For receivers with a continuous frequency range of 0 to 30 MHz, about 30 crystals were required if a VFO tuning range of 1 MHz was used. Since the crystals had to be individually tuned, the manufacturing and calibration costs were considerable.

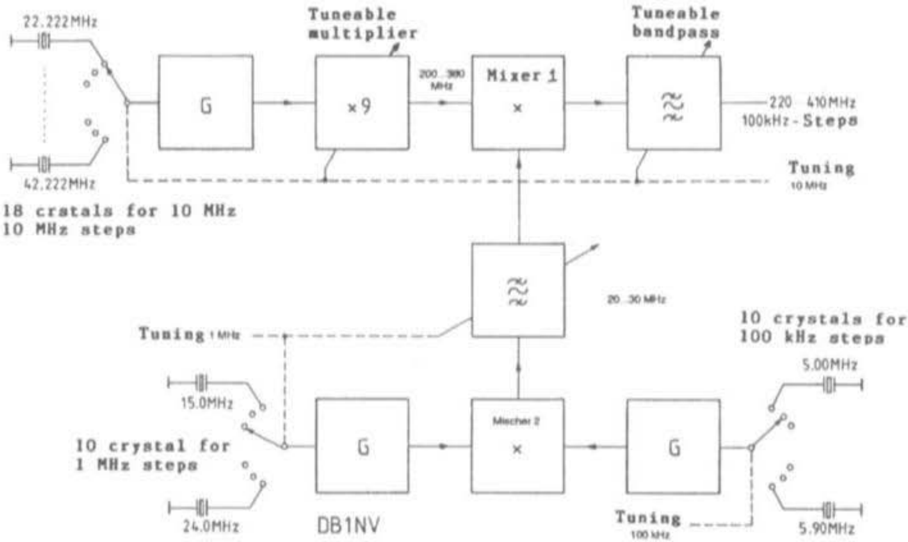


Fig.4: A VHF synthesiser with reversible crystal oscillators

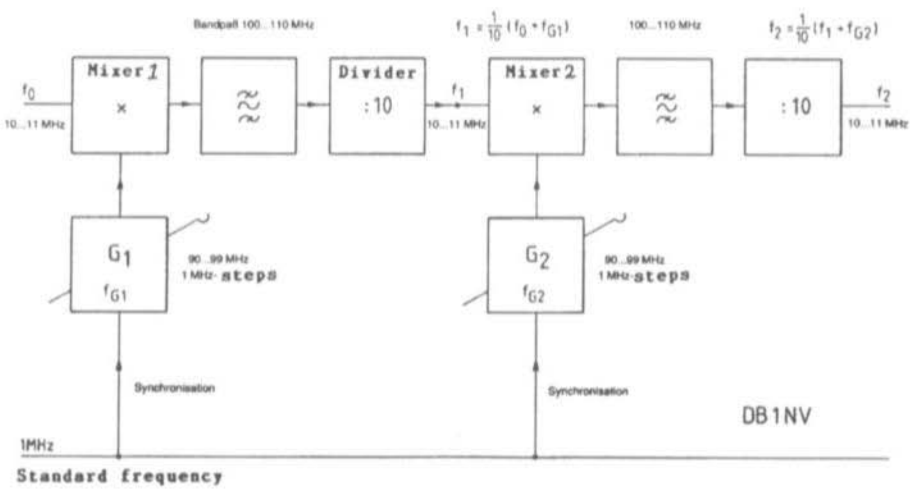


Fig.5: Structure of a Frequency Decade

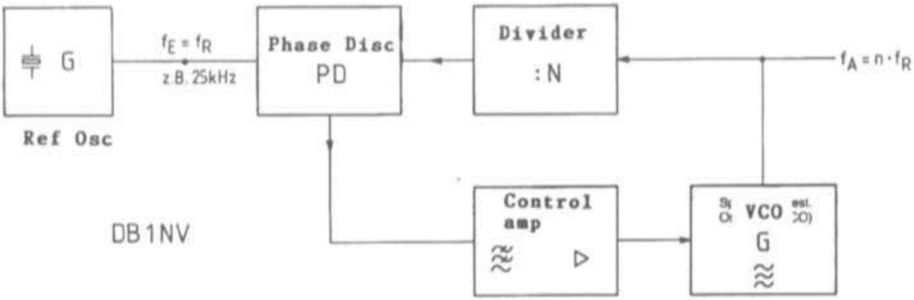


Fig.6: A Phase Locked Loop with frequency divider

An alternative was found in the shape of the following process (Fig. 2). A crystal oscillator (e.g. 1 MHz) powers a harmonic generator stage, at the output of which all the harmonics of the crystal frequency are included, up to a pre-set maximum frequency. The desired harmonic is selected by means of a tuneable filter and fed into the receiver circuit. Only one crystal is required, and apart from the VFO calibration the receiver requires only one crystal to be tuned. The cost of this technique lies in the filter, tuneable or reversible in stages, and increases with the requirement for the signal to be free of spurious emissions. The expense on the filter can be considerably reduced if the circuit is expanded, in accordance with Fig.3, by two mixers, a low-pass and a tuneable oscillator.

The harmonic spectrum generated by the crystal oscillator is spectrally limited through a deep pass and mixed with the signal from the tuneable oscillator. If the sum of a crystal harmonic and the oscillator is in the pass band of the subsequent band pass, it goes through to the second mixer and is mixed back into the original frequency position - hence the name, reverse mix process. The two mixers and the fixed-tuned filter are used to simulate a variable-frequency filter. But the best of it is that the tuneable oscillator used for mixing does not need to be specially stable, since its frequency drift has no influence on the output as long as the desired mixed product lies in the pass-band range of the band pass.

This process was developed in the 'fifties for commercial communications receivers by Dr. T.L.Wadley at the Racal company, and came into use in the first amateur all-band receivers, such as the Barlow-Wadley XCR30 or the Drake SSR1, about twenty years later.

3.2. The mixing process

If a large number of crystal-stabilised frequencies are to be generated in a pre-set channel grid, this can be achieved by mixing suitably staged crystal oscillators, the expense being even smaller than with individual crystals. Fig. 4 shows one example of how this system was used, e.g. in Collins military aircraft radios from the valve era; this synthesis concept generates frequencies in between 225 and 400 MHz in the 100 kHz grid. Two decimal-stepped reversible crystal oscillators directly generate the 100 kHz and 1 MHz stages. A ninefold crystal oscillator with a thermostat supplies the 10 MHz stages. Because of the many selective amplifier stages, to

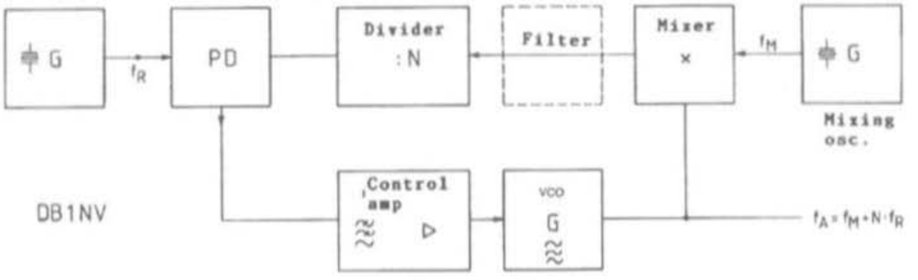


Fig.7: A Frequency Synthesiser with mixing

be synchronously mechanically tuned, the equipment was a miracle of light engineering. The cost of maintenance must have been considerable.

A similar process was perfected in the same period, which was referred to by the term "frequency decade", and which had the advantage that frequency resolutions of any fineness required could be obtained by means of cascading identical operational units. Frequency decades (also referred to as decadic synthesisers) have the great advantage that all oscillators run permanently at fixed frequencies and need to be reversed only on the output side. Since no PLL control circuits need to build up, frequencies can be changed more rapidly, which still secures a market share for the frequency decade even today.

Fig. 5 shows the basic structure of a chain consisting of two decades; an input frequency, Fo, of, for example, 10...11 MHz

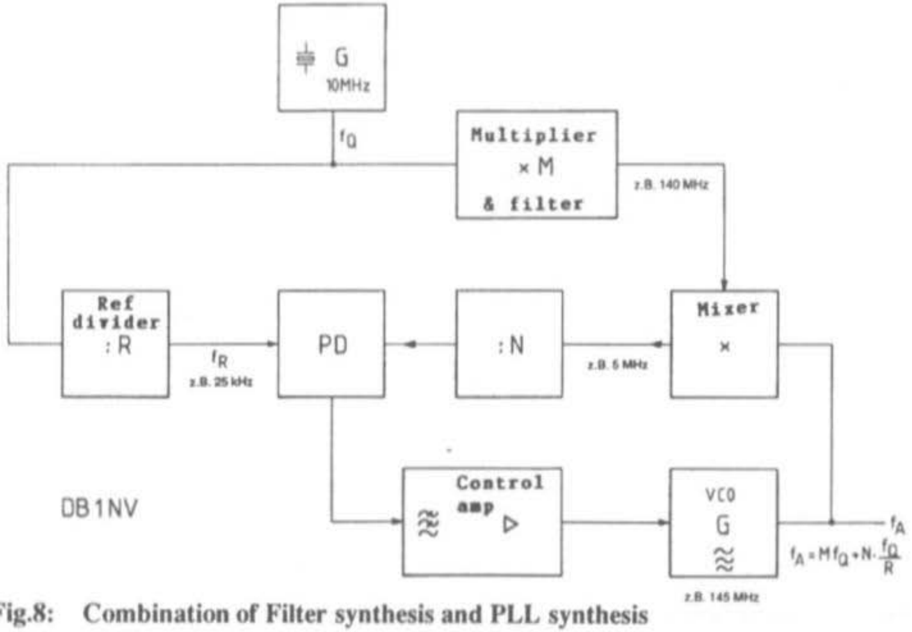


Fig.8: Combination of Filter synthesis and PLL synthesis

is mixed, through a mixer, with an oscillator tuneable in 1 MHz stages between 90 and 100 MHz, G1. Let the oscillator, G1, be synchronised with the decade's standard frequency of 1 MHz. In the simplest case, G1 can be a "locked oscillator", which is synchronised through direct coupling in of the 1 MHz reference. Another possibility is the use of the filter process mentioned in Section 1, with 10 reversible band passes. The total of Fo and G1 is filtered out using a travelling band pass filter and then divided by 10. The frequency, F1, at the output of the divider is then:

$$f_1 = (f_0 + f_{G1})/10 \qquad (1)$$

and acts as the input signal for a second identical decade stage with G2 as oscillator. The output frequency, F2, then equals:

$$f_2 = (f_1 + f_{G2})/10 = \frac{f_0}{100} + \frac{f_{G1}}{100} + \frac{f_{G2}}{10} \qquad (2)$$

As can be seen, by cascading the decades a signal can be generated with as fine a frequency resolution as desired.

However, in practise the simple structure is temperamental, since harmonics of the input signal and direct breakdown of the generators, G1 or G2, must also be taken into account in the mixing process, which increases the cost of the band pass filter or makes multiple frequency changing necessary. A relatively modern frequency decade with multiple shifting is, for example, the PTS 160 (2). The great advantage of a frequency decade is the rapid frequency changing; for this purpose, the mixing frequencies required for G1...Gn are centrally generated for all decades and only buffer stages and switches are used in the individual decades instead of the generators.

Frequency decades were already being built in the valve age and were the first synthesiser circuits which came into use in large numbers. Older amateurs will perhaps still remember the "Schomandl decades" with 50 to 100 valves.

3.3. The phase locked loop with adjustable divider

If the phase locked loop from Fig. 1 is expanded, in accordance with Fig. 6, by an adjustable frequency divider between the VCO and the phase detector, then we have the simplest form of a PLL frequency synthesiser. The reference frequency is selected as per the desired grid point interval (e.g. 25 kHz), and by selecting a suitable divider factor we can use any whole-number multiple of the reference frequency. The VCO is dimensioned in

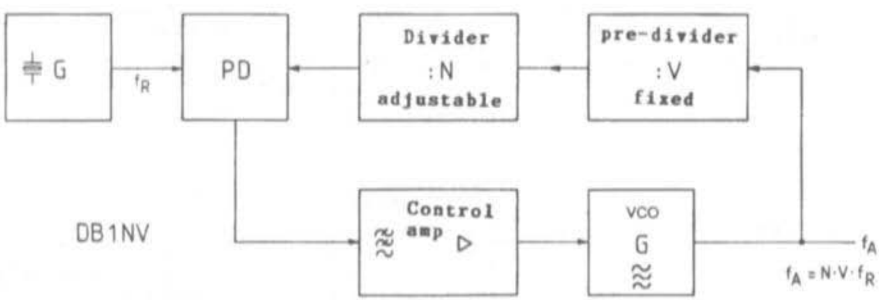


Fig.9: PLL frequency range expansion with fixed pre-divider

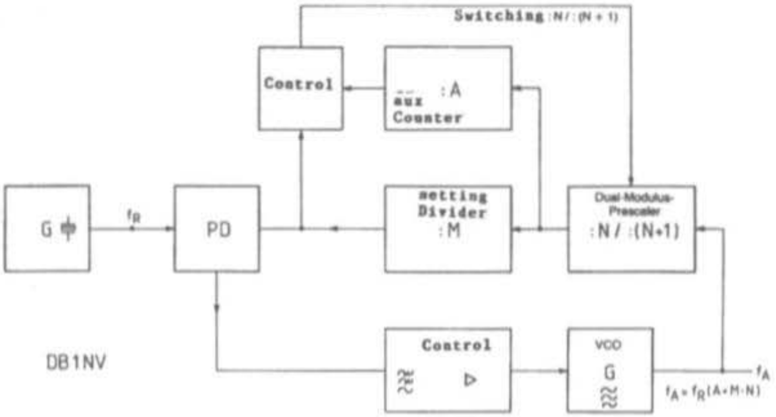


Fig.10: Frequency range expansion with dual-modulus pre-divider

accordance with the required tuning range and the low-pass filter in the control amplifier is to be selected in such a way that, firstly, the control loop builds up rapidly and stabily and, secondly, the residues of the reference frequency which is still present at the phase detector output do not reach the VCO and generate a spurious modulation there in the form of secondary lines with the interval of the reference frequency. Many sins were committed by the first radio sets with PLL synthesisers in this context, because the steep spike pulses from the frequency divider and phase monitor reached the VCO in every possible way and laid the foundation of the ''lattice fence image'' of the synthesiser technique in amateur circles.

The basic circuit described is suitable for the generation of frequencies below app. 50 MHz in wide channel grids above approximately 10 kHz. At higher frequencies, creating adjustable frequency dividers becomes more and more difficult and expensive, so that it is better to use one of the three following circuit variations. For narrower channel grids, the build-up time for the control loop becomes too long, since then the limit frequency of the control amplifier must be selected low, in the interests of adequate suppression of spurious emissions. Controlling the hum and howl back of the VCO is thus no longer possible. In this case, it is better to turn to multi-loop PLL's, fractional-N techniques, direct digital synthesis or a combination thereof.

3.4. Phase locked loop with mixing

If the VCO is to work at frequencies which can no longer be economically processed using an adjustable frequency divider, then a process in accordance with Fig.7 can be used. The output frequency of the VCO is mixed down using a crystal-stabilised reference frequency into the operating range of the frequency divider (usually below 10...20 MHz). Should the tuning range of the VCO exceed these frequencies, then several reversible crystal oscillators are used and switched appropriately. This circuit is still used, even today, in many synthesisers for VHF equipment, in particular for FM portable sets. It can be structured so as to save more energy than the pre-divider processes described in the next two sections. The disadvantage of the mixing process is that the output frequency of at least two crystal oscillators is determined, which impairs the long-time stability. This can be got round if the mixing oscillator frequency is generated, as per Fig. 8, by multiplying and filtering the reference crystal. By switching over the filter, a large VCO tuning range can also be covered. This circuit can be set up as a combination of a PLL synthesiser with a filter synthesiser as per Section 3.1.. Many radio equipment manufacturers go still further and also use the reference crystal as a mixed oscillator for conversion to a second intermediate frequency or as a BFO. The goal is thus achieved of taking the reception or transmission frequency back to an individual reference frequency.

3.5. The phase locked loop with pre-divider

In cost-critical mass production applications, e.g. in audio systems, the cost of an additional crystal oscillator or a multiplier with a filter is higher than that of a rapid fixed frequency divider. A PLL with a fixed pre-divider in accordance with Fig. 9 is the standard circuit in tuning synthesisers for VHF radio equipment, television receivers and video recorders. The fixed pre-division operates as though the adjustable frequency divider could be adjusted, not in single steps, but only in multiples of the pre-division. This is equivalent to saying that the reference frequency can not be selected at the phase discriminator to be equal to the channel interval, but must be selected to be smaller by the pre-division. This has the following consequences. Owing to the lower phase monitoring frequency, the filter is to be dimensioned at a lower frequency to remove the monitoring frequency residues, which brings about a slower build-up of the loop after a frequency change. Moreover, the slower loop can control bad interference on the VCO frequency, e.g. hum or howl back. But these disadvantages cause little interference in the audio systems sector: there is no continuous frequency switching, as with a radio set operating as a transceiver, the build-up time of the synthesiser for a channel change is not critical, whereas, for example, on a TV set the picture and line

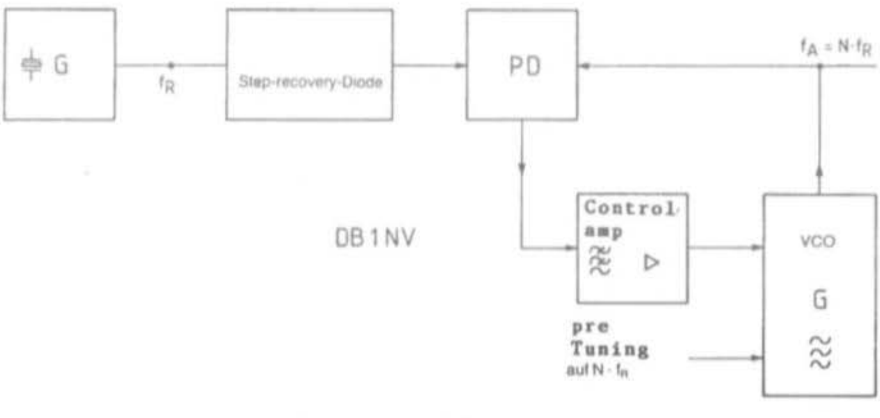


Fig.11: A PLL circuit with frequency multiplication

synchronisation must also build up, and the requirements in relation to the accuracy of the VCO in the vicinity of the carrier are considerably more relaxed than in SSB or telegraph mode.

PLL's with fixed pre-division are hardly ever used in radios. The options described in Section 3.6. are more often made use of, with a reversible pre-divider.

3.6. The phase locked loop with reversible pre-divider

The phase locked loop with reversible pre-divider combines the advantage of the circuit options above, lower expense as against a mix, with the higher phase monitoring frequency of the mixing PLL. Phase locked loops with reversible pre-dividers are also described in the literature as PLL's with dual-mode pre-scalers or as swallow counter PLL's. The key component here is a rapid pre-divider, which can be switched using a control signal to the divider factors N and (N + 1); possible divider factors are 10/11, 40/41, 80/81 or 128/129. The structure of a PLL with a dual-mode pre-divider is shown in Fig. 10, and largely corresponds to a PLL with a fixed pre-divider, except that, parallel to the adjustable frequency divider (here usually known as an M-divider) a further adjustable counter is mounted, which uses logic to switch the pre-divider between the divider factors N and (N + 1). The procedure is as follows. When the M-divider overflows, both A-dividers and M-dividers are loaded with their pre-set values and begin to count upwards in "common mode", with the pre-divider switched to (N + 1). If the A-divider has counted down to zero, the pre-divider is switched to N and the A-divider is stopped until the next overflow from the M-divider.

The cycle then begins again. The overflow pulse from the M-counter controls the phase discriminator. M pulses per cycle are needed at the output of the pre-divider, so the pulses from the pre-dividers A are switched to (N + 1) and (M - A) pulses are switched to N. The VCO also has to supply oscillation cycles in a cycle of:

$$A(N + 1) + (M - A)N$$

By conversion, the total divider factor thus becomes A + MN, which corresponds to the desired result. In spite of the pre-division, the phase monitoring frequency can be set in stages using the auxiliary divider A. The A-divider makes a kind of interpolation possible between the frequency stages pre-set by the pre-division of N times phase monitoring frequency. Since the switching cycle between N and N + 1 is synchronous with the reference frequency, these processes are not noticeable as spurious modulations to the phase detector signal. Modern integrated PLL circuits contain the logic and the A-divider to drive a dual-mode pre-divider, and the structure of such a synthesiser thus becomes relatively simple. This circuit has therefore become a standard module of many modern radios.

3.7. The phase locked loop with multiplication

The disadvantage of the PLL circuit presented above with a frequency divider in the control loop is that a spurious modulation of the VCO in its phase variation is divided down before it arrives at the phase discriminator. The obvious idea of simply raising the amplification of the control amplifier in order thus to improve the control of VCO instabilities is not practicable for two reasons. Firstly, both the frequency divider and the phase discriminator generate extremely small phase instabilities, which can not be distinguished from the phase noise of the VCO. This can even go so far that the noise contributions of the divider and the phase discriminator predominate at the control input of the VCO, and the PLL can even make the noise characteristics worse. Secondly, a phase discriminator can supply only an endless quantity of "readings" per period of its input signal for phase deviation, a digital discriminator supplies, for example, one or two readings per period. This means that faster phase oscillations in the VCO signal can not be controlled in principle. So the control band width of a divider PLL is also restricted to a fraction of the phase monitoring frequency.

The solution to these problems is a phase discriminator which acts directly on the final frequency of the VCO and is driven by means of the reference frequency; Fig. 11 gives an outline diagram of this. At higher frequencies, PLL's of this type can no longer be created using the standard circuits for phase discriminators consisting of logic gates, so we have to turn to the following solution. The reference frequency is converted into a sequence of extremely short pulses using a step-recovery diode and fed into an analogue phase discriminator, for example a ring mixer, at the other end of which is the VCO signal. The equal fraction at the mixer output follows the VCO up in familiar fashion. The functioning of the phase monitoring can also be understood as mixing a harmonic with an intermediate frequency of zero. In principle, the VCO can be locked to each harmonic of the reference frequency, so a pre-tuning of the VCO in the vicinity of a locking point, e.g. using a digital-analogue converter, is still necessary for the defined setting. The advantage of this circuit engineering is that, with reference frequencies in the megahertz range, a control band width of several hundred Hz can be obtained, so that even wide-band control of the VCO phase noise is possible. This type of PLL

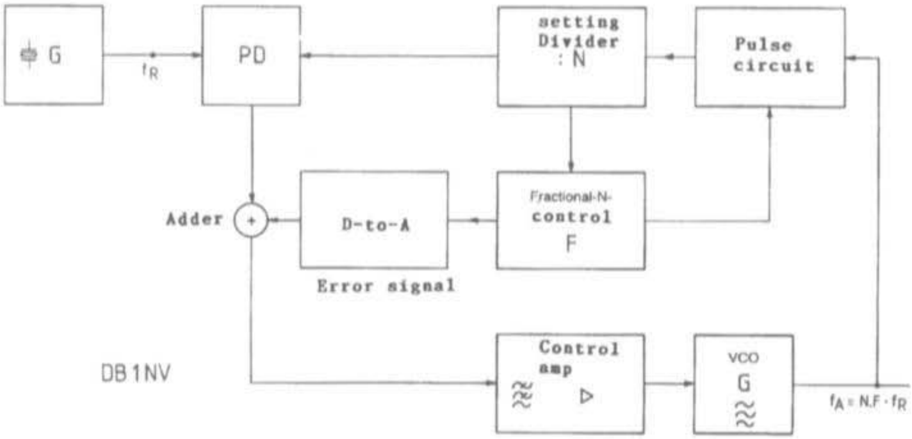


Fig.12: Basic principle of Fractional-N Synthesis

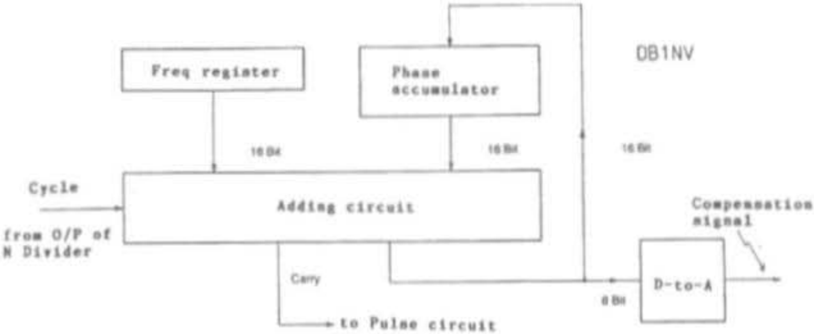


Fig.13: Mode of operation of Phase accumulator

can be regarded as a kind of variable-frequency filter, which precisely filters out one frequency from the "lattice fence" of the reference spectrum.

The step-recovery diode and mixer sub-assemblies are in practise manufactured as so-called sampling mixers or sampling phase discriminators. We shall come back to this interesting circuit.

3.8. Digital interpolation

The disadvantage of the synthesiser concept described above is that the phase monitoring frequency pre-determines the frequency grid. A narrow frequency grid means a low phase monitoring frequency and thus a long build-up time after a frequency change and insufficient control of interference on the VCO. We therefore looked for ways to develop synthesisers with fine frequency resolution and satisfactory control behaviour at the same time.

The first digital interpolation process has only modest claims to freedom from harmonics, but it forms the basis for an understanding of fractional-N engineering. Digital interpolation was introduced in the amateur literature quite some years ago through the "Suedwind" radio (3).

The basic concept is simple. If we periodically switch over the divider ratio between N and N + 1 in a divider-synthesiser, using any circuit engineering, the phase discriminator in the middle is presented with a frequency which lies between the frequencies represented by N and N + 1. By varying the keying ratio, we can set the "interpolated" frequency. For example, should the divider factor N represent the frequency of 10,000 MHz and should N + 1 represent 10,010 MHz, then, if we switch over in the keying ratio 1:1, we get an average output frequency of 10,005 MHz. For a setting of 9 time units N and one time unit N + 1, we get an output frequency of 10,001 MHz. But since the phase discriminator never receives the reference frequency presented, but a frequency which is too high or too low, a control voltage will form which tries to tighten up the VCO. If the control loop is fast enough, then the VCO will jump backwards and forwards between the two frequencies. If we make the control loop slow, then the control amplifier averages out the oscillations, but the build-up behaviour becomes sluggish. The final effect is that no significant advantage is obtained as against a PLL with a lower reference frequency and thus a finer resolution. With additional expense, it is certainly possible to set the frequency lying nearest in the channel grid immediately when there is a frequency change, with a small control time constant, and then to activate the interpolation logic with a large control time constant. However, the result bears no relation to the expense.

3.9. Fractional-N technology

Fractional-N synthesisers use the idea of cyclical switching between N and N + 1 explained above to interpolate between the frequencies pre-determined by N. The decisive additional concept is to compensate for the periodic interference voltage arising at the output of the phase discriminator instead of filtering it out. Thus the characteristics of the basic synthesiser with a high monitoring frequency can be retained, but with the fractional-N technology supplying the fine interpolation resolution.

Reference Period	Phase Accumulator Contents	
	F=0.2	F=0.6
1	0	0
2	2	6
3	4	2*
4	6	8
5	8	4*
6	0*	0*
7	2	6
8	4	2*
9	6	8
10	8	4*
11	0*	0*
12	2	6
13	4	2*
14	6	8
15	8	4*

Fig. 12 shows an initial approach to solving the problem. A normal synthesiser forms the basic equipment. Between the VCO and the adjustable frequency divider a circuit is introduced which I might loosely describe as a "pulse claw circuit". It cuts out precisely one period from the signal coming from the VCO if it is activated. If the circuit is activated once in a counting period of the setting divider, the VCO must then generate N + 1 pulses in that time instead of N. The "pulse claw circuit" thus brings about the switching of the divider between N and N + 1 without it being necessary to re-programme the divider. The switching is controlled by a circuit unit which is also new, the scheduler. So far the structure corresponds to a synthesiser using digital interpolation.

As a result of the periodic switching between N and N +1, the output voltage of the phase monitor now contains an AC voltage fraction which would modulate the VCO and cause interference. In order to compensate for this interference fraction, an additional stage is introduced between the phase discriminator and the control amplifier which receives its signal from a digital / analogue converter. The D/A converter is provided with the digital compensation information by the scheduler.

In the literature, for example in (4) or (5),with regard to the obtaining of the compensation signal, it says only that it can be calculated from the interpolation frequency. In order to be able to understand the procedure and thus construct the scheduler, let us start with the following imaginary experiment. Suppose the synthesiser is operating at an output frequency of 10,001 MHz and a phase monitoring frequency of 100 kHz. Let the setting

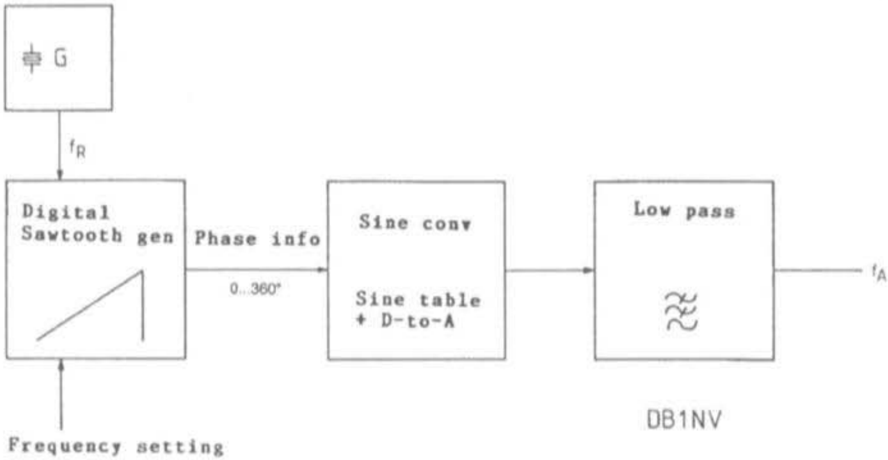


Fig.14: Analogue version of a DDS Synthesiser

divider be programmed to N = 100. Normally this setting is not stable and the phase monitor will pull the VCO down to 10,000 MHz. If we keep the VCO at its frequency, then the phase difference between its divided down signal and the reference frequency will increase with each reference period. Since the actual VCO oscillation is higher than corresponds to the channel grid, a staircase-type voltage sets in, in a negative direction, at the phase discriminator output, which wants to pull the VCO downwards. Without further measures, the staircase-type voltage works against the lower selection limit of the phase discriminator. But if we cut out one pulse from the VCO signal, with the help of the pulse claw circuit, then the phase of the divided VCO signal is pushed back a certain amount, the output voltage of the phase monitor jumps upward by a certain amount and remains within its operating range. The scheduler must thus activate the pulse claw circuit periodically and thus keep the phase monitor in the operating range. The interference signal created by the switching between N and N + 1 is consequently a sawtooth, the amplitude of which depends on the height of the phase jump in the N/(N + 1) switching and thus on the divider factor N.

The interference is successfully compensated for if the D/A converter generates a sawtooth of equivalent size but displaced through 180 degrees. The control amplifier records the total of the two signals, which in the ideal case is an equal fraction, and keeps the VCO at the frequency of 10,001 MHz, i.e. between the grid points.

The scheduler is constructed in accordance with a process which is known by the name of phase accumulation. The outline diagram in Fig. 13 shows that a so-called phase accumulator is linked to an adder in such a way that its contents are periodically incremented by the contents of the frequency register (it contains the fractional part of the divider ratio). This process is repeated every time the frequency divider overflows. The overflow of the adder drives the "pulse claw circuit" and the upper bits from the phase accumulator form the input signal of the D/A

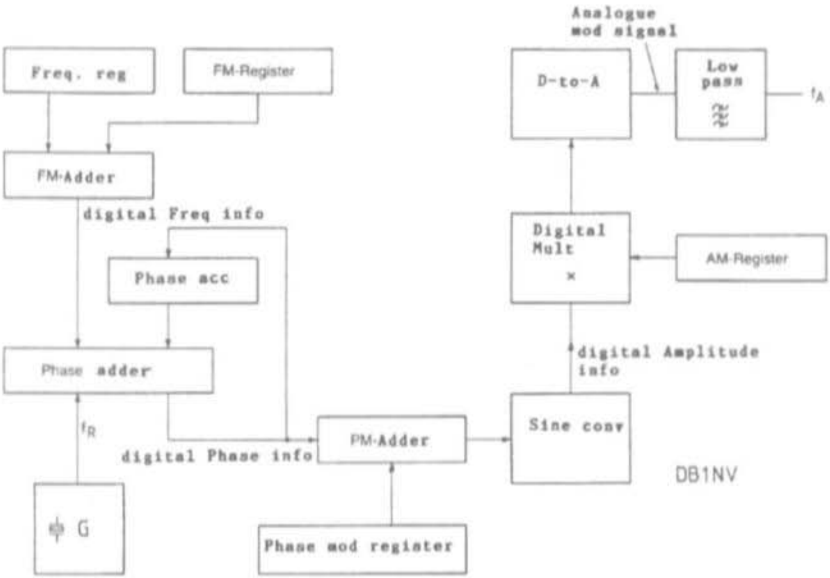


Fig.15: DDS Synthesiser with digital modulation

converter for the interference signal compensation. As already mentioned, the compensation signal has still to be multiplied by the N-divider ratio set. The simplest way to do this is by means of a superimposed pulse width modulation. We can, for example, control the switching-on time of the compensation signal proportionally to the periodic duration of the VCO. With a high N and thus a high VCO frequency, we then automatically get a shorter duty cycle for the compensation signal.

The word width of the phase accumulator corresponds to the number of "fractional" frequency steps between two N-stages. If, for example, we wish to insert 100 fractional frequency stages between the grid stages, then the phase accumulator has a word width of two decimal places. The way the phase accumulation functions can be seen more clearly in the following table. Here a word width of one decimal place is assumed and the phase accumulator contents are shown for two frequency settings, namely for a broken fraction of 0.2 and 0.6.

For the additions marked with a star, an overflow takes place, which triggers a switch to the divider ratio N + 1 for one reference cycle by means of the "pulse claw circuit". In the left-hand example, the system is switched to N + 1 for one cycle in five, which corresponds to an average divider factor of N + 0.2. In the right-hand example, the switching takes place in three cycles in five, which corresponds to a division of N + 0.6. In the left-hand example, the sawtooth pattern characteristic of the phase accumulator contents and thus the correction signal can be recognised. In the right-hand case, the pattern of the correction signal is less easily appreciable.

As can be seen, the additional electronics for a fractional-N synthesiser can be created using standard circuits, but the expenditure on digital technology is considerable. The digital operational units are therefore combined into industrial circuits to form application specific integrated circuits (ASICs). Examples of this can be found, for example, in (4) and (5). Perhaps one or two of our readers can look at these equipment handbooks in the QRL. Since the ASIC path is not open to the normal amateur, little attention has been paid to this circuit technology up to now.

The author tried a way of creating digital logic using the little man's ASIC, a single-chip micro-controller. The main problem here was that the fractional-N synthesis was not good enough. Since many switching processes are taking place so slowly that they are within the control band width of the PLL, extreme demands have to be made on the de-coupling of the individual sub-assemblies, and the cost of this is an order of magnitude higher than for a "normal synthesiser". A detailed description of the synthesiser circuit developed by the author will follow in Part 3 of this article.

3.10. Direct Digital Synthesis (DDS)

Synthesisers using the circuit technologies described so far can generate high-quality signals of almost any frequency. But there are applications which require signals in the upper kilohertz range and can live with compromises as to quality, and therefore need a more favourably priced synthesiser. Conventional technology is not very appropriate for an integrated and thus cheap solution. It is worth striving for a synthesis process, which can largely be put into practise using digital circuits and requires a digital/analogue converter only as a final stage. In an extreme case, such a synthesiser can be created as an integrated circuit. A modulation fraction allowing, for example, digital frequency modulation, phase modulation or amplitude modulation can be integrated into a DDS circuit without significant additional expense. Even multiphase signals, e.g. frequency-adjustable three-phase current in the drive engineering, or 90-degree signals to generate an SSB using the phase method, can be easily generated without tuning.

The basis of direct digital synthesis is the so-called Shannon's theorem, which is of sweeping significance in digital signal processing.

If applied to digital frequency synthesis, Shannon's theorem says the following. For the accurate generation of any signal with an extremely high spectral component at a frequency, f, at least 2f scanning readings per second are required. For the precise representation of a sine at 1 MHz, however, it is not necessary actually to generate a sine. It is sufficient to produce a sequence of "scanning readings" at a minimum of 2 MHz in order to give a full description of the sine. The conversion of the scanning readings into the signal is taken care of by a steep-flanked low pass filter.

The sinus amplitudes can also be calculated at a finite number of sequential points in time, converted into analogue voltages through a D/A converter, and finally filtered through a low pass to generate a perfect sine. The only restriction is that at least two readings have to be calculated per period.

Since many readers will not be familiar with digital signal processing, the mode of operation of a direct digital synthesiser should be clarified using the following analogue example as per Fig. 14.

The first operational unit is a digital sawtooth generator which generates a sawtooth voltage, more precisely a step function. The sawtooth is interpreted as an angle specification of 0...360 degrees and drives a sine-wave converter which transmits the sine-wave corresponding to the angle. A low-pass filter removes unwanted components from the signal. The mode of operation is thus immediately clear. Each time the sawtooth or phase angle moves from 0 to 360 degrees, the sine-wave converter supplies a complete sine oscillation. The steeper the sawtooth becomes, the more sine oscillations are generated per time unit and the higher the output frequency of the synthesiser.

Since all circuit sections are digitally created, the question of the appropriate word length has to be clarified, both in the sawtooth generator and in the sine-wave converter. The longer the word length is, the more finely can the amplitude and phase be resolved, but the expenditure increases immediately. A common compromise is a phase resolution of 32 bits and a resolution of the sine-wave converter with 8 to 12 bits.

We have already come to know the digital sawtooth generator as a combination of phase accumulator, adder and frequency register in the fractional-N synthesiser. If we expand the frequency register somewhat, we then have the possibility of a digital FM, as shown in Fig. 15. Instead of the frequency register, two registers are used, the contents of which reach the adding device of the phase accumulator through an adder. For example, the main tuning frequency of a transceiver can be established in the first register, with the file for RIT in register 2, or else register 2 is

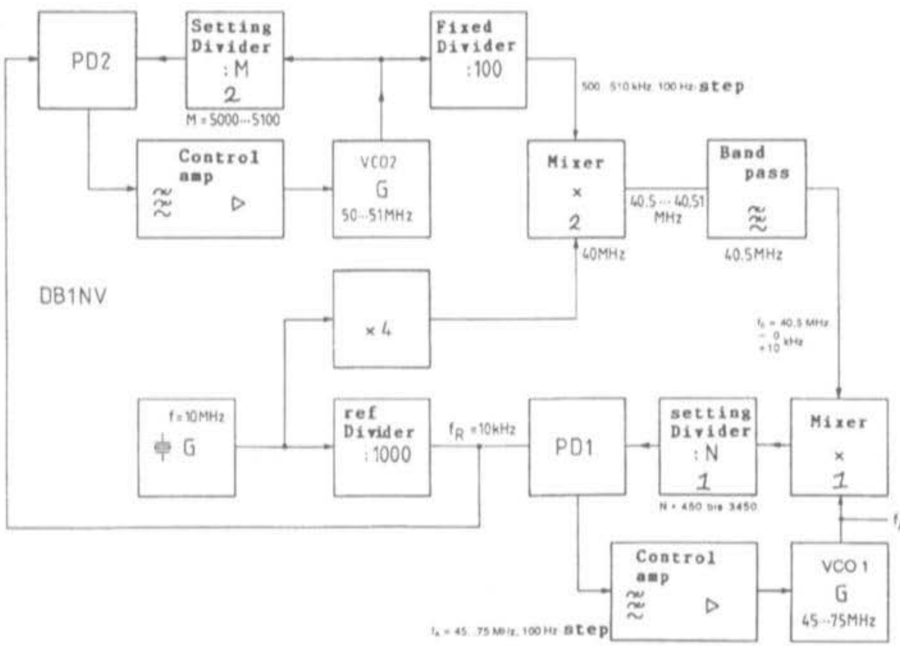


Fig.16: Multi-loop Synthesis

"fed" with digitised speech and thus receives a digital frequency modulation. The phase accumulator and the frequency register usually have a word length of 32 bits. The mode of operation of the circuit is the same as above. If the phase accumulator is topped up with an increment established by the frequency register, the digital sawtooth is generated, but here the overflow of the phase accumulator is not used for anything else. The upper 8 to 12 bits of the phase accumulator usually control the sine-wave converter directly, and an adder can also be introduced between the two, which is connected to a "phase register". An additive phase displacement can thus be generated which can be used for digital phase modulation. The sine-wave converter receives the phase information and converts it into digital sine amplitudes. This is most simply done using a sine table, which is housed in a ROM. The memory requirement can be reduced by a factor of 4 by using the symmetry of a sine function, in that only the first quadrant is stored and all the other values are generated through appropriate manipulation of the plus or minus signs. The digital sine values control a D/A converter which generates the output signal. Should an amplitude modulation be required, a digital multiplier can be introduced between the sine ROM and the converter which is controlled by an amplitude register. Finally, a low-pass filter smoothes out the sine function generated and removes unwanted spectral fractions.

The functions of a DSS synthesiser can even be carried out by means of a pure software solution using a micro-processor. Some time ago, the author succeeded in controlling a frequency converter using DDS software on an 87C51 single-chip micro-controller. Three sine signals were generated, each with a phase displacement of 120 degrees, from 0 to 600 Hz, using a 16-bit phase accumulator at a clock frequency of app. 13 kHz. The frequency resolution was 0.2 Hz. Some experiments in the low-frequency range are thus promising.

Anyone interested in learning more about the innermost workings of DDS circuits should refer to a document from a private company (6). Perhaps there is even someone reading this who has already experimented with DDS circuits and can report on his or her experiences.

3.11. Multi-loop synthesiser

To obtain fine frequency resolutions without the disadvantage of low phase monitoring, there is another way, apart from the DDS process just mentioned, which makes use of cascaded synthesisers with a moderate step width (10 kHz each), but is not so expensive as a frequency decade. The two-loop resolution is the way shown for a standard radio synthesiser with a frequency resolution of 100 Hz. Even smaller frequency steps can be generated using a third loop. But the simpler solution uses a stretched crystal oscillator for fractional detuning.

A suitable two-loop synthesiser for a short-wave receiver of 0...30 MHz and a first intermediate frequency of 45 MHz might look like Fig. 16. A VCO from 45 to 75 MHz (VCO 1) generates the oscillator signal of the receiver. The VCO is mixed down with a frequency of 40.5 MHz (initially taken as fixed) to a frequency of 4.5 to 34.5 MHz. A setting divider (N = 450...3450) divides the signal to 10 kHz and a phase discriminator (PD1) compares it with the correspondingly divided signal of a 10 MHz mother crystal. The phase discriminator PD1 follows VCO 1 up. So far, the circuit corresponds to a normal synthesiser with a mixed-down VCO signal and a step width of 10 kHz. A 40.5 MHz mixing frequency is now also synthesised for interpolation. The second PLL circuit is used to do this with VCO 2, the second setting divider (M = 5000...5100) and the phase discriminator, PD2. This circuit section generates a frequency of from 50 to 51 MHz, also with a step width of 10 kHz. If this frequency is divided by means of a fixed divider (: 100), it can then be mixed with the signal from the fourfold mother crystal (40 MHz) and the hum frequency of 40.5 to 40.51 MHz can be filtered out. This signal forms the mixing frequency of the first loop. It can be seen that, in spite of the high phase monitoring frequency of 10 kHz, this circuit has a tuning resolution of 100 Hz. Should this still not be enough, a further mixer can be introduced into the second loop, which is again controlled by the total signal from the mother crystal and by a divided down third loop. This makes it possible to attain step widths of 1 Hz.

The risk that harmonics will form due to unwanted mixture products grows with the number of loops, so that careful screening and de-coupling are required to generate a low-harmonic signal. So we would not use more than two or three cascaded loops. An up-to-date approach to generating extremely fine resolutions, for example, is to replace the fine loop at 500 to 510 kHz by a DDS synthesiser, which can be manufactured at this kind of low frequencies as a customer-specific CMOS circuit.

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Wolfgang Schneider, DJ 8 ES

SSB Transceiver for 50 MHz using 50Ω Modules Part-3 Conclusion

(Revised version of the presentation given at the 1991 Weinheim VHF Convention)

5.2.3. Commissioning

When the operating voltage (+ 15 V) is connected, the current consumption is 16 mA. The external S-meter, together with the PIN diode controller in the intermediate-frequency amplifier, is connected later. Both outputs of the sub-assembly (U-Control and S-meter) are initially set so that + 12 V is available at the corresponding connection. The 10 kOhm trimmers are provided at the OP's for this purpose. The control voltage for the PIN diodes is tuned again afterwards, say in the ready-made transceiver, to the control application desired.

The longitudinal resistance is to be calculated for the S-meter (RS-meter). The entire reception branch must be taken into consideration for this purpose. In the specimen layout, this consists of:

- → External pre-amplifier
- → Reception mixer
- → Intermediate-frequency amplifier
- → Field strength indicator.

With the configuration given above, an S9 signal fed in appears on the aerial (-73 dBm at 50Ω), with - 29 dBm at the AGC/ALC output of the intermediate-frequency amplifier. At the output of the NE614 (pin 5), the voltage is app. 3.5 V, depending on the characteristic. This corresponds to a value of + 8 V at the S-meter output. For the S-meter used (50μA, 3kΩ), the value thus calculated for the compensating resistance, RS-meter, is 120kΩ

Theoretical considerations had to be confirmed in the fully assembled transceiver. Only now was it possible to carry out fine tuning using the 10kΩ trim potentiometer at the O/P.

6. THE MODULATOR/DE-MODULATOR UNIT

The modulator/de-modulator unit consists of:

- → SSB modulator
- → SSB de-modulator
- → USB/LSB side-band oscillator
- → Microphone amplifier
- → Low-frequency amplifier

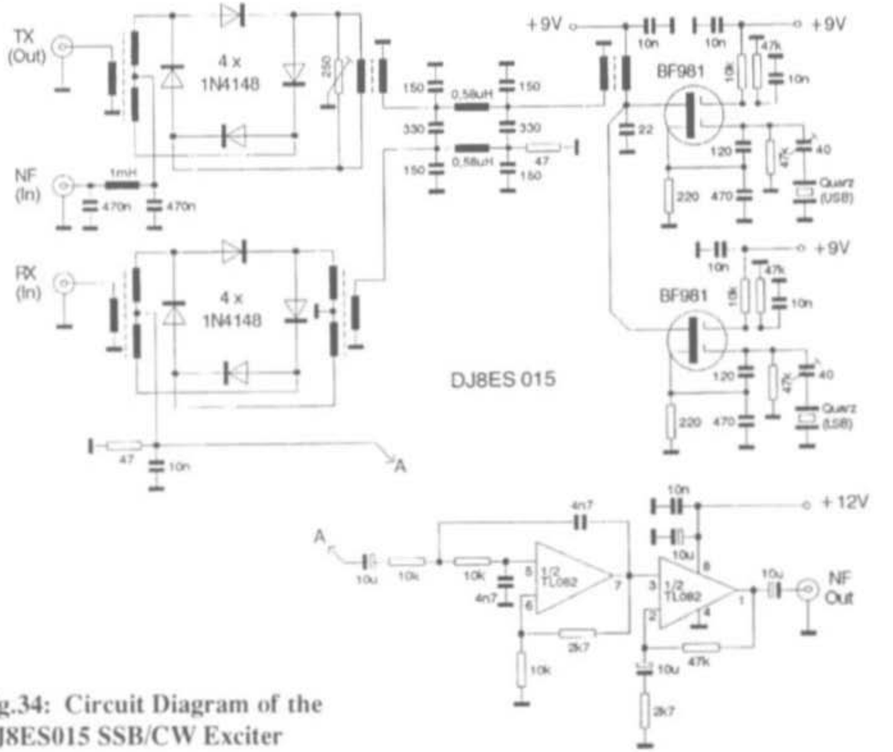


Fig.34: Circuit Diagram of the DJ8ES015 SSB/CW Exciter

6.1. The SSB Exciter

The side-band oscillator has a dual-gate MOSFET (BF 981). The crystal frequency can be tapped through the 40pF trimmer. There is room on the printed circuit board for a second side-band oscillator. Both operate on the same drain oscillation circuit. Switching between USB and LSB is done by means of the G2 voltage associated with the oscillator in question. A capacitively coupled hybrid divides the oscillator power up between the modulator and the de-modulator.

The amplified microphone signal is fed to the modulator, a ring mixer with 4 x 1N4148, through an LC low pass (fg = 4 kHz). A trim potentiometer (250Ω) is used for circuit balancing.

From the output of the SSB de-modulator, the low-frequency signal goes through a two-pole low pass to reach the low-frequency pre-amplifier. This two-stage amplifier is fitted with an operational amplifier (TL082) (Fig. 34).

6.1.1. Assembly instructions and commissioning

The dimensions of the SSB exciter printed circuit board are 72 mm x 72 mm, suitable for a standard tinplate housing.

The two side-band oscillators have a mirror image structure. The drain circuit is common to both. This filter, originally laid out for 10.7 MHz, is brought into resonance with 9 MHz through the 22pF capacitor, which is connected in parallel.

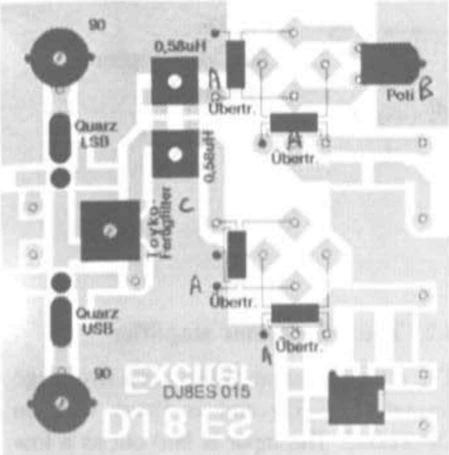


Fig.35a: SSB Exciter; only a few components on the earth plane

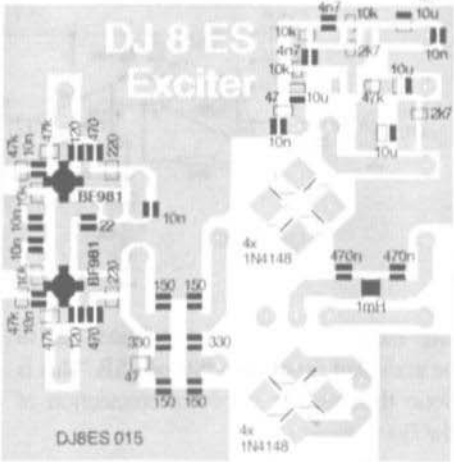


Fig.35b: DJ8ES015 SSB Exciter from the foil side

6.1.2. Parts list

2 x BF981 (Valvo)

1 x TL082, 8-pin DIL housing

8 x 1N4148, SMD format (Minimelf)

1 x 1 mH choke, SMD format

2 x Neosid 0.58 μH filter, BV5036 (or/bl)

1 x Toko 10.7 MHz filter, green (10 mm)

1 x 250Ω potentiometer (10/5 mm)

2 x 40pF trimmer (10 mm, red)

Capacitors and resistors SMD format

6.1.3. Commissioning

After the connection of the operational voltages of + 12 V and + 9 V for the

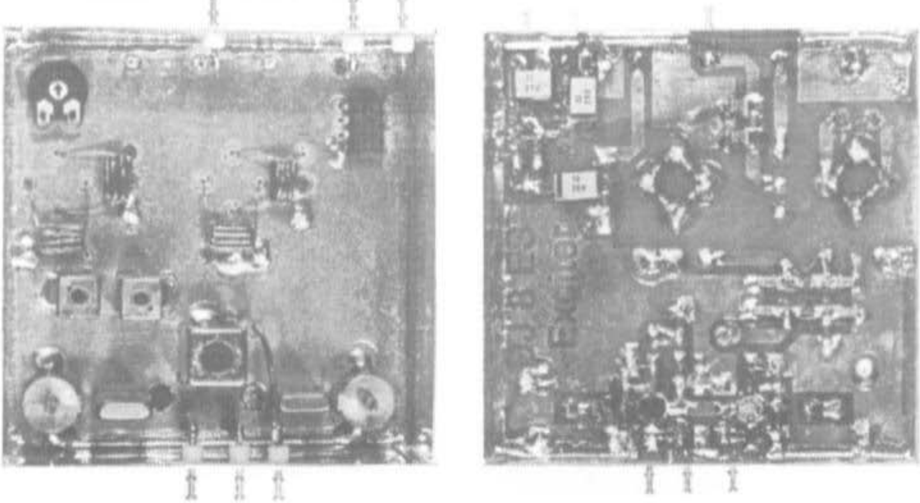


Fig.36a, b: Specimen of the SSB Exciter, assembly and foil side

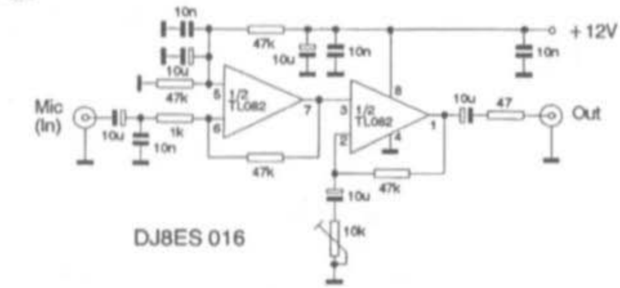


Fig.37: The Michrophone Amplifier

side-band oscillators, the oscillator has to be activated for either USB or LSB. This is done through the selective connection of the G2 voltage.

The precise frequency is set by means of the individual 90pF trimmer. Then tune the filter in the drain circuit to the maximum signal (10 mW).

The potentiometer for circuit balancing the SSB modulator is in the central position. Fine tuning at the minimum oscillator level at the mixer output (TX-out) terminates the tuning required. 48dB carrier suppression should be obtained.

The SSB exciter element is now ready for operation.

6.2. The microphone amplifier

The microphone amplifier has a two-stage format (Fig. 37) and is fitted with an OP TL082. The input is laid out as a low pass. A normal dynamic microphone is used, with an impedance of 600Ω. The output voltage required for the SSB modulator (app. 1 Vss) is set by means of the 10 kΩ potentiometer.

6.2.1. Assembly instructions and commissioning

The amplifier takes the form of a double-sided copper-coated epoxy printed circuit board with the dimensions 34 x 34 mm. The element is placed directly onto a 4-pin microphone socket.

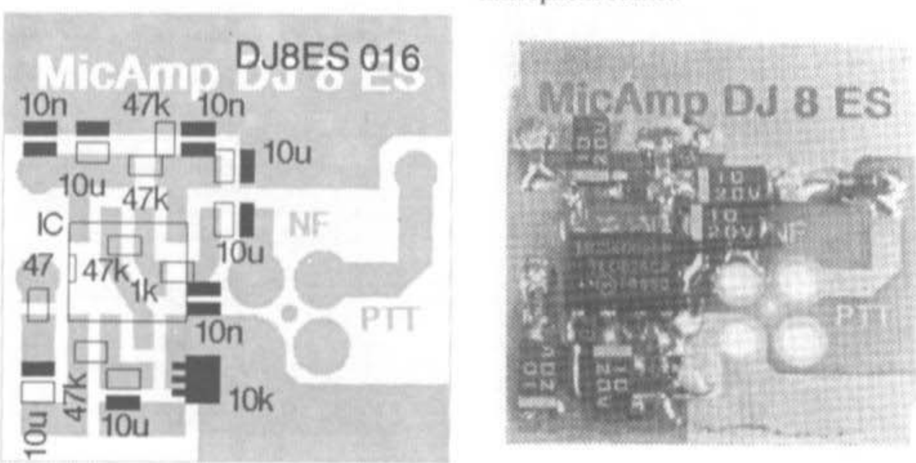


Fig.38: DJ8ES016 Components side and a completed unit prior to housing

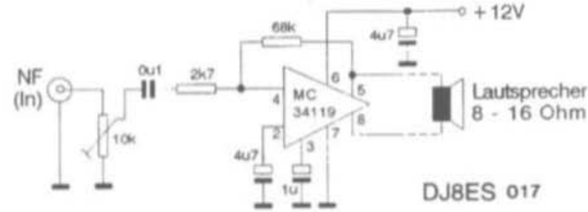
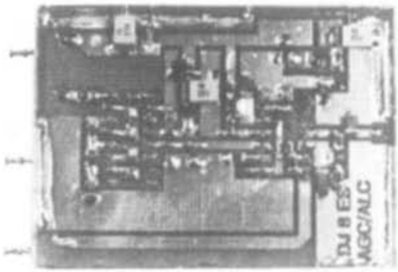


Fig.40: The Low-frequency Amplifier

In contrast to what is otherwise the standard procedure, the printed circuit board is not put into the tinplate housing until after assembly (Fig. 38). So the microphone socket has to be precisely fitted into the housing cover. Only the OP (TL081) is in the 8-pin DIL housing. All the other components are SMD formats. Teflon bushings are used for the connections.



6.3. The Low-frequency amplifier

The heart of the low-frequency amplifier is the MC 34119 integrated circuit. This IC is available as an SMD format in an SO-8 housing. The component supplies an output power of 200mW into 8Ω. The minimal external wiring here produces a simpler, but more efficient, low-frequency amplifier (Fig. 40).

6.3.1. Assembly instructions

Like the microphone amplifier before it, the printed circuit board for the low-frequency amplifier has the dimensions 34 mm x 34 mm.

As Fig. 41 shows, the component is directly soldered onto the connections of the noise level controller. Alternatively, of course, a housing can be provided. The necessary number of bushings then have to be provided for the connections (loudspeaker, low-frequency input, power supply).

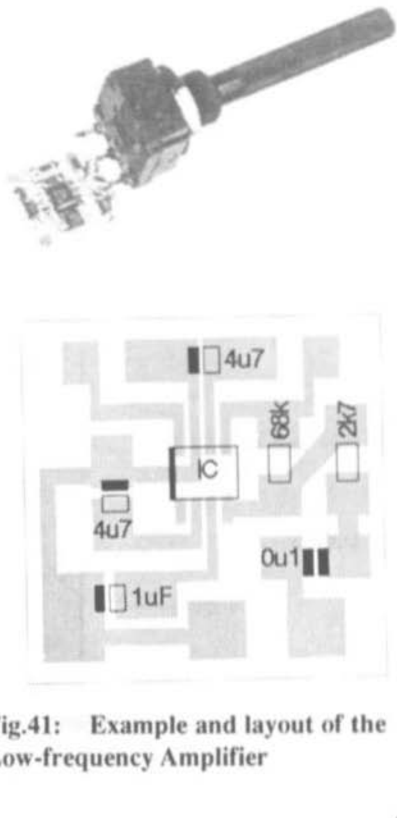


Fig.41: Example and layout of the Low-frequency Amplifier

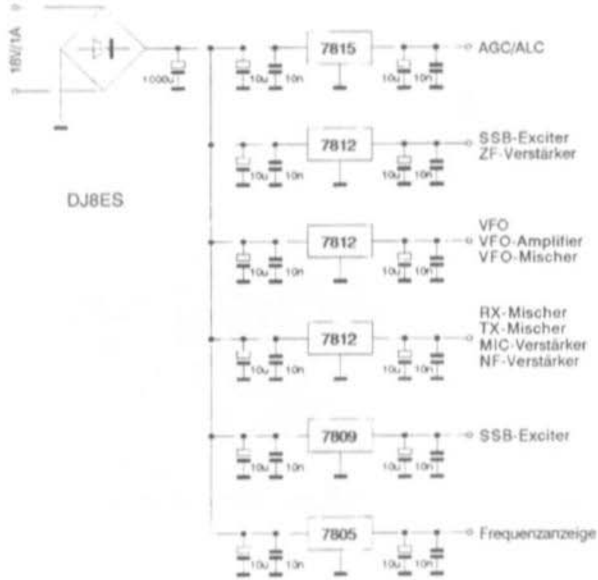


Fig.42: Separately stabilised voltages for the individual sub-assemblies

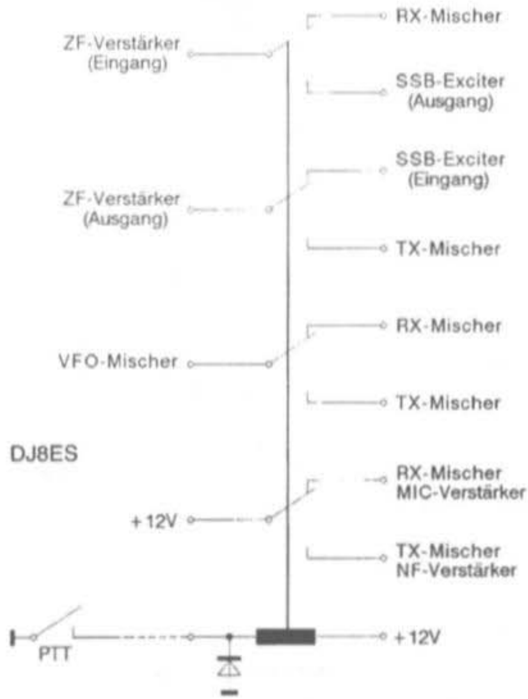


Fig.43: Transmission and Reception switching in the Transceiver

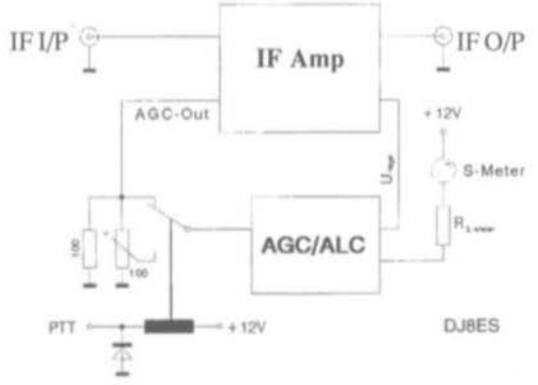


Fig.44: Interconnection of the IF Amplifier and the AGC/ALC unit

7. INTER-CONNECTING COMPONENTS

Once the individual printed circuit boards have been assembled and pre-tuned, as described in the section in question, all the functional units can be inter-connected. The incorporation into an equipment housing will not be considered here. This is very much dependent on the wishes and ideas of the individual constructor. The electrical interaction of the components is more important.

Various supply voltages are required in the transceiver:

- → + 15V AGC/ALC component
- → + 12V SSB exciter, microphone amplifier, low-frequency amplifier, reception mixer, transmission mixer, oscillator, VFO amplifier, VFO mixer, S-meter
- → + 9 V SSB Exciter
- → + 5 V Frequency display

Not only the various voltages but also operational considerations play a part in the circuit drafting for the mains section. Thus, for example, the VFO, the Audio amplifier and the mixer each need to have a separate voltage stabiliser provided. This prevents the oscillator from being affected by feedback from other operational assemblies (Fig. 42).

The intermediate-frequency amplifier is used not only for reception but also in the case of transmission. Suitable reversing facilities are provided for here. This component is wired to be controlled by the PTT contact. The same is also true for the oscillator (41 to 42 MHz) and the supply voltages for the transmission-reception branch (Fig. 43).

The intermediate-frequency amplifier and the AGC/ALC components are in direct relation to one another (Fig. 44).

The intermediate-frequency signal at 9 MHz is de-coupled through the 20dB coupler, terminated at 50Ω The input of the AGC/ALC component is assembled with the NE614 field strength indicator. The input impedance is 1 kΩ The feedback circuit is terminated through the U-Control output.

The S-meter can now be calibrated in receive. A wired-up pre-amplifier or a transverter should also be considered here.

The precise setting point of the control for the intermediate-frequency amplifier is similarly determined next.

Only a small amount of the de-coupled SSB signal is fed to the NE614 in transmission mode. The potentiometer should be set in such a way that approximately 100µW is present at the output of the intermediate-frequency amplifier.

8. SUMMARY

As can be seen, standard components used in 50Ω engineering can be used to produce even circuits as lavish as the transceiver presented here. Special attention was paid to good technical data and reliable operating characteristics. Moreover, only components which form part of the stock of any reasonably well-equipped DIY specialist are used.

As was said right at the start, this article is not intended to give a detailed description of the construction of a 50 MHz transceiver. The idea is rather to provide some ideas and show the possibilities for putting such a concept into practice.

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Dr. Volker Grassmann, DF 5 AI

Long-Delayed Radio Echoes, Observations and Interpretations

The final propagation speed of electromagnetic waves generates a time delay between the transmission and the reception of a radio or light signal. In terrestrial radio traffic, this time delay can be ignored for most practical applications. Only with particularly long propagation paths does the effect become perceptible. In the short-wave range, radio waves can go all the way round the world, under special conditions, which leads to a time delay of about 138 ms. (round-the-worldechoes). Even more impressive are the echoes reflected from the Moon, where the propagation path corresponds to twice the distance between the Earth and the Moon (time delay of about 2.5 s.). The phenomenon described below can be explained by neither of these processes. LDE's (long delayed echoes) can involve a time difference measured in tens of seconds between transmission and reception.

1. First Observations

The first systematic observations were reported by Hals, Stoermer and van der Pol (1).A previous accidental observation had been logged by Hals, who had received long delayed echoes from transmitter PCJJ in Hilversum in the vicinity of Oslo in 1927 (time delay app. 3 seconds). However, experiments carried out over many months on 9.5 MHz were initially fruitless. In October, 1928, long delayed echoes could finally be demonstrated to exist without any doubt, with delay times of between 3 and 30 seconds. Further observations were also reported in subsequent months by other experimenters. (8) contains a comprehensive historical retrospective, in which a further LDE observation from Hals is mentioned, with time delays of 260 s. and 195 s. However, after 1935 interest in this unusual phenomenon appeared to diminish.

2. Scientific Observations

In the years 1947 - 1949, renewed experiments took place involving observations of long delayed echoes. The research was carried out in Cambridge by Budden and Yates (2). They transmitted at 13.4 MHz with a power of 30kW and at 20.6 MHz with a power of 1kW. The aerial used was a half-wave dipole, the structural height of which was intended to make vertical radiation possible.

Unhindered by the ionosphere, the radio waves were to travel on into space, for there were "well-founded indications" that "ionised clouds" were being emitted from the Sun towards the Earth, which were taken into consideration by the authors as possible reflectors for radio waves (note: in 1951, Biermann drew the conclusion from observations of a comet's tail that a particle stream must exist, moving away from the Sun, which we now refer to as the solar wind.) Budden and Yates published their experimental results in 1952.

Altogether, about 27,000 test signals were transmitted. Round the world echoes were certainly frequently detected, but no long delayed echoes. (Note: the round the world echoes indicate that the aerial radiation system selected could not suppress the ionospheric wave propagation, as had been intended.) The authors conjectured that the transmitter frequencies selected, high in comparison with previous experiments, were as inappropriate as was the vertical radiation. They further conjectured that there was an "earthbound" origin for the long delayed echoes, for the rapid solar particles should have triggered a Doppler effect, which had, however, never been reported by previous experimenters, and had been, to some extent, explicitly denied (see, for example, (1)).

Laboratory experiments on echo effects in plasmas aroused interest in long delayed echoes at Stanford University. The mechanisms discovered in the laboratory also appeared to be conceivable under the conditions of ionospheric plasma, but this conjecture had to be verified by more precise research. Nevertheless, experiments were carried out using a 20kW

Year	Call Sign	Band	Delay Time	Mode	Echo Signal
1932	W6ADP	28 MHz	18 s	CW	Own
1950/51	W5LUU	7 MHz	5 s	CW	Own
1965	K6EV	14 MHz	3 - 4 s	SSB	Own
1967	W5VY	28 MHz	3 s	SSB	Own
1968	W5LFM	10 MHz	0.5 s	Time Marking	Station RID
1968	W6KPC	28 MHz	1 s	SSB	Other
1969	W6OL	14 MHz	6 - 10 s	SSB	Other
1969	K6CAZ	2 MHz	2 s	SSB	Other/Own

Table: LDE reports, as per (3)

transmitter (5 to 25 MHz) and a log-periodic aerial, although the equipment was modified several times during the experiments, between 1967 and 1970 (4).

The first long delayed echoes were registered in October, 1968, but had to be rejected as internal interference signals. A pulse diagram was finally to improve the unambiguous identification of the actual signal transmitted. The authors mention the difficulty of automatic plotting and the particular capability of human hearing to recognise signals of this kind among background interference. In January and February, 1970, three long delayed echoes were received at 11.02 MHz and 10.62 MHz, with time delays of 15 s. and about 20 s.. By December, 1971, the number had risen to 31, according to (12).

An automatic observation apparatus was described by Duffet-Smith of Cambridge in 1975 (6). Experiments using a 250W transmitter produced no results. The automatic recording system proved not very effective, due to short-wave interference. The author explains that all earlier research had confirmed human hearing as the most sensitive detector. On the basis of the negative results of his research, evaluations of the phenomenon of long delayed radio echoes tended to be cautious.

A critical evaluation of previous research was published by scientists at Stanford University in 1985 (12). Some experiments were undertaken, using a vertically radiating frame aerial, on the WWV time signal transmitter (5.865 MHz).However, the seven echoes registered (time delay between 1.5 s. and 18.3 s.) could not finally be regarded as authentic. The earlier measurements from Stanford University (see above) were also put in doubt and were linked with technical side effects. The observation data made available by Goodacre (VE2AEJ/3 - see below) were also treated with caution. Vidmar and Crawford demonstrated, on the basis of their own observations, that if stricter criteria were applied the supposed radio echoes generally proved questionable. But the comprehensive reports stretching back over the past fifty years ruled out any doubt as to the existence of long delayed echoes.

3. Observations From Radio Amateurs

Radio amateurs have contributed the most comprehensive observational material concerning long delayed echoes. There is no other field of radio wave propagation where radio amateurs' observations have obtained a comparable degree of respect in the scientific literature. The LDE observations listed in the table have been obtained from the QST.

Goodacre (VE2AEJ/3) reported, in a scientific essay, on eight possible LDE effects, which he had been able to pick up between November, 1978 and January, 1979 in the 28 MHz band(9). A five-element Yagi and a 400 W-powered transmitter were used in the experiments near Ottawa. Groups of three to nine pulses were transmitted at a rate of 130 - 150 Hz. The pulses were generated using a continuously switching relay, which was switched by means of a Morse key in series. The radiation was being transmitted towards the Western horizon when ground-to-ground communication in the 10 m. band began to fail, and 10 m. connections were already impossible on other compass bearings. The questionable band points were noted while tape recordings were being made, so that the only observation findings subjected to evaluation were those which had already been considered as suspected LDE's while the experiments were going on (avoidance of false interpretations arising from the known "copying effect" of magnetic tapes). The tape recordings were finally studied using an oscilloscope.

Hardly any attention had been paid to long delayed echoes in the German-speaking world up until then. But it may be assumed that some observations made had not been publicised because the unusual effect was interpreted as an attempt at deception. For example, Schwarzbeck (DL1BU) very probably reported long delayed echoes, in addition to the normal round the world echoes, as a "Morse dot delayed by several seconds" which remained present when the frequency was changed (10).

During an RTTY contest in October, 1986, DJ4ZF and DL6QH received the last 40 to 50 teleprinter characters (45.45 Bauds) of their own transmission (14). The frequency of the echo was app. 300 Hz lower than the transmission frequency. The 20 m. band displayed the typical effect of the "close" before the breakdown of transmission, with the echoes exhibiting a "West Coast or Alaskan character". An output power of 750W was transmitted on a monoband ground plane. The authenticity of the echoes can not be guaranteed. The observers interpreted the phenomenon as an attempted deception by another radio amateur and did not hear of the LDE phenomenon till later.

4. Observations in the UHF Range

The most spectacular amateur observations report long delayed radio echoes in EME connections. Rasmussen (OZ9CR) reported "ghost echoes", in a scientific journal, which he had picked up at 1,296 MHz (7). In a Moon echo experiment on 7. 7. '74 (8 m. dish aerial operating at 500W), it proved possible to receive echoes which arrived about 2 seconds after the genuine Moon echoes. These echoes remained noticeable, even when the aerial was not aligned precisely with the Moon and the usual Moon echoes were absent. The phenomenon lasted for 20 minutes.

Note: doubts were cast on the observation data reported by Rasmussen on the basis of the Moon position specified, and Rasmussen therefore corrected the data on 28. 5. '74 (see (8)).

(8) refers to further LDE observations in connection with EME experiments. The Dubus Magazine also has a short report on long delayed radio echoes which were picked up by YU1AW on 432 MHz (11). The additional echoes followed the Moon echoes with a time delay of about 2 s. and were slightly displaced in frequency.

When the various LDE reports were scanned, a possible common factor in the data from the YU1AW observations in the 70 cm. band and the 10 m. observations of DL1BU (see above) became clear. Unfortunately, the actual observation times could no longer be reconstructed. A survey of long delayed echoes published by the author (see (13)) unfortunately yielded no evaluable results.

5. Interpretations

The attempts at interpretation which accompanied the observations of previous decades can not be summarised here. We should, of course, point out that the boldest attempts at explanations assumed that an extra-territorial space probe was reflecting the terrestrial radio signals (see, for example, (5) and (8)).

Since the abortive scatter experiments on the solar wind, at the latest (see above), there is final agreement that reflections from astronomical objects can not be held responsible for the radio echoes. In near-Earth space, there is only one object with a sufficient back scatter cross-section to cause radio echoes which are observable, even with low-powered transmitters. But radio amateurs know from their own Moon experiments that the echoes are relatively weak, with a delay time of (2.5 ±0.5) s.. The next possible object is Venus, which can approach as close as 0.27 astronomical units to the Earth (1 astronomical unit corresponds to the average distance between the Sun and the Earth). For this special astronomic configuration, a time delay of app. 270 s. can be calculated for the journey there and back.

Finally, the screen effect of the ionosphere on short electro-magnetic waves must be considered. It is generally accepted today that long delayed radio echoes are an effect, the origin of which must be located in the terrestrial ionosphere or magnetosphere.

The observations reported in (10) verify that, in principle, short electro-magnetic waves can go several times round the world. Seven times round the world yields a propagation path with a length of app. 0.97 light seconds, which is 290,000 km.. The estimate made by Schwarzbeck for this example also points out that the round the world echoes may be subjected to path attenuation effects, which are significantly reduced as against free space propagation. However, "normal" round the world effects do not seem able to explain the radio echoes under consideration here. At least, there are no indications that the delay times observable for LDE's are multiples of 138 ms.. Nor do the LDE reports mention any multiple echoes analogous to Schwarzbeck's observations. The delay time of 8 s. which is typical for the LDE's would mean they had to go round the world another 58 times. For a time delay of 40 s., it would appear necessary to go round the world 289 times.

However, the radio echoes picked up must be accepted as experimental facts. If we assume the delay times observed to be identical to the "lifetime" of the transmission signal radiated, and the signal propagation speed to be 300,000 km./s., then we necessarily obtain long propagation paths, irrespective of the geometry of the propagation route (curved propagation path around the world, straight back and forth lines in space, etc.). The resultant difficulties of interpretation can be eliminated if we dispense with one or both of these assumptions.

In a gas, variations in pressure or density can take the form of wave or oscillation conditions (see, for example, sound waves in air). The ionosphere represents a special mixture of gases, for something like a thousandth of the gas particles are ionised, i.e. are present as positively charged ions and negatively charged electrons. The motion of the charged particle, incorporated in the neutral gas, is thus subjected to the effect of additional electrical and magnetic forces. In the partial gas formed from ions and electrons, special wave phenomena can arise ("plasma waves"). Taking ionospheric plasma waves into consideration appears to be an important step forward in the physical explanation of long delayed radio echoes. For example, there can be an interaction between the materially connected plasma waves and electro-magnetic waves, under suitable conditions.

A mechanism is proposed in (4), in the context of a qualitative model, in which part of the electro-magnetic energy is bound in a longitudinal plasma wave. In this context, "longitudinal" refers to propagation along a field line of the Earth's magnetic field. Here, energy-rich electrons, which move around this field line in a spiral, reinforce a plasma wave of this type. The propagation speed of the plasma waves is usually several orders of magnitude below the speed of light. There are thus relatively long propagation times even for comparatively short propagation paths of a few hundred kilometres.

The radio signal is thus, in a way, intermediately stored in a plasma wave until a process analogous to the initial condition creates a new electro-magnetic wave, which can then be observed by the receiver station.

Of course, the authors refer to the circumstance that long delayed radio echoes at high short-wave frequencies can not be satisfactorily explained by this model. This is especially true of amateur observations in the 10 m. band. To wit, with the mechanism described we must assume from the start that the reciprocal effects take place at the height at which the frequency of the radio wave exactly corresponds to the local plasma frequency. This frequency is generally considerably lower than 28 MHz. But the attractiveness of this model lies in the identification of a "natural" time-lag device. A group velocity of 1 km./s. for the plasma waves and a reciprocal effect region of app. 10 km. diameter (4) give delay times of 10 seconds.

The theory is skilfully advanced further in (8) and is used to explain the observations of OZ9CR at 1,296 MHz. Two radio transmitters, at frequencies f1 and f2, are scanning the same region in the ionosphere, unbeknown to one another. The two electro-magnetic waves cause a non-linear reciprocal effect and generate a plasma wave at the differential frequency δf, i.e. f2 - f1, which should simultaneously correspond to the local plasma frequency. If f1 = 1,296 MHz and f2 = 1,303 MHz, for example, the frequency obtained is 7 MHz. After a certain propagation time, the plasma wave again interacts with the electro-magnetic wave, f2, and generates an electro-magnetic wave with a frequency of f1 = f2 - δf, which appears to the puzzled observer to be an echo of his or her own transmission.

6. CONTRIBUTIONS FROM RADIO AMATEURS

With long delayed echoes, we are clearly dealing with an unpredictable and comparatively rare phenomenon. The results of scientific observations indicate that LDE's are not very often traced by deliberate observation. The opportunities for amateur radio enthusiasts to make observations are far less subject to geographical restrictions, or those dependent on local time. In particular, short wave amateurs form a world-wide network of observers always ready for action. The favourable "observation statistics" for amateur radio enthusiasts have conceivably been a decisive factor in the numerous LDE observations in which they have been involved.

The author also endorses the opinion expressed by Budden and Yates, that radiation directed towards the horizon yields better observation results than vertical radiation. True, according to (4) radiation along the local line of the Earth's magnetic field could be more advantageous. The horizontal radiation might possibly increase the probability of observation, since a larger segment of the ionosphere can be scanned by the aerial, by comparison with vertical sounding. The typical radiation characteristic of amateur aerials could thus turn out to be advantageous.

The previous reports confirm that long delayed radio echoes attract attention mainly in non-automated radio traffic, which is connected with the particular capability of human hearing to recognise the acoustic pattern. This state of affairs also corresponds, to a large extent, with the typical conditions for amateur radio traffic.

The experimental conditions for amateur radio enthusiasts thus seem particularly suitable for LDE observations. Radio operators professionally active in the short-wave range also come into consideration as competent observers to the same extent (e.g. marine radio, here especially wireless telegraphy operators). However, there are no corresponding observation reports. It must thus be assumed that some observations are interpreted as attempts at deception and thus do not receive general attention. Conceivably, numerous amateur observations are also not taken into account for the same reason.

It has already been mentioned that the reliable identification of a "natural" LDE presents a difficult task, even if the observer takes great care. Apart from unwanted technical effects, in particular, attempts at deception can in no case be ruled out, and it is to be feared that less serious radio amateurs contribute to deliberate falsification, in view of the fact that the LDE phenomenon is better known in such circles. The continuous self-sabotage should be recalled here which amateur radio circles inflict on themselves in research into VHF radio wave propagation (e.g. by interference in the 2-m. beacon band). Moreover, it can not be ruled out that the exposure of genuine attempts at deception also puts the credibility of subsequent amateur observations in doubt. In that event, scientific institutions could be forced to leave amateur reports out of consideration in future, on grounds of principle. An apparently harmless joke involving a fake LDE can thus do serious damage to the image of amateur radio.

Above all, observers should describe the measures which they have taken in attempting to exclude cheating. Short transmissions using changing frequencies are a relatively simple procedure for excluding potential trouble-makers. Unfortunately, this procedure can be used in practise only when you are re-receiving your own transmissions. If long delayed echoes are received from other transmitters, enquiries should be made as to whether other transmitters on other frequencies are also displaying LDE phenomena. Reports submitted on deliberate or accidental LDE observations should contain the following additional data:

- → Date, start time and end time of observation
- → Location of observation station
- → Observation frequency
- → Echo characteristics (re-receiving of own transmission or observation of another transmitter, duration of echo, time delay, field strength)
- → Transmission power and radiation characteristic, direction of aerial.

Please send appropriate observation reports to the author (address: 68, Hausfeld, 5600 Wuppertal 23). Plottable data will be passed on to the Max Planck Institute for Aeronomy, Lindau/Harz.

Acknowledgement: The author is grateful to Dr. K.Schlegel, of the Max Planck Institute for Aeronomy, for useful information and ideas.

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(13) Long delayed echoes on EME-circuits, Grassmann, V. Dubus Magazine, no. 3, 1987, p. 244

(14) Personal report Grobauer, H.

Erich Zimmermann, HB 9 MIN

A Simple dB-Linear S-Meter for Microwave Applications

The runaway development of integrated circuits for the mobile radio sector also means more and more new and useful modules are being showered on radio amateurs. Thus, for instance, integrated intermediate frequency circuits for cellular applications have characteristics which make it possible to display reception levels on a dB-linear scale.

1. SUMMARY

Available for selection are the MC 3356P from Motorola and the NE 604 AN and NE 605 AN from Signetics-Philips. The wiring required by the Motorola module is extremely simple and problem-free.

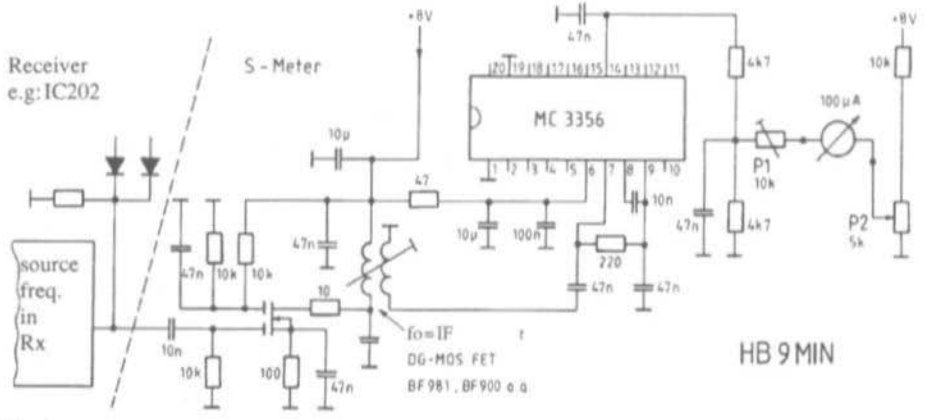


Fig.1

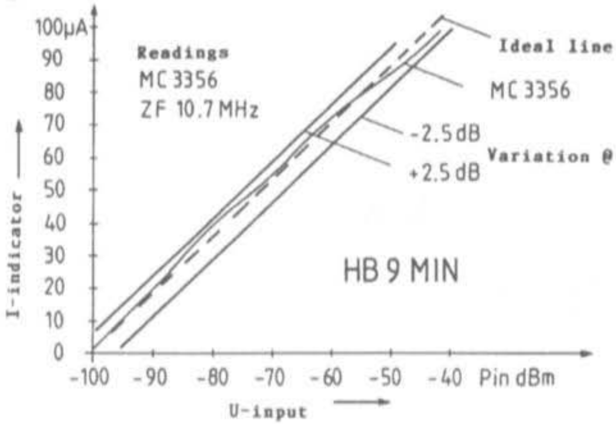


Fig.2:

The wide dynamic range provides for 70dB at 10.7 MHz, and still gives 58dB at 21.4 MHz, with an error of ±1 dB.

The two integrated circuits from Signetics-Philips, NE 604 AN and NE 605 AN, are even designed for 85dB dynamics with an error of ±2 dB. Unfortunately, the wiring required is more expensive, since an additional intermediate-frequency filter is needed.

The signal is passed on to pin-7 of the MC 3356P. For correct operation, the input voltage range lies between 30µV (-77dBm) and 80mV (-1dBm), measured at pin-20. From pin-14, the output signal goes through a voltage divider and a pre-resistance and arrives at the indicator. With the given circuit, a range of 50dB can be displayed linearly on an indicator with a full-scale reading of 100µA. The gradient is thus 2µA/dB (Fig.2).

2. DESCRIPTION OF CIRCUIT

The circuit shown here with the MC 3356P operates in the broad band, from approximately 5 MHz to 22 MHz. Only the input circuit of the pre-amplifier has to be in resonance with the intermediate frequency.

The intermediate-frequency signal is transmitted from the crystal filter of the adjuster to the MOS-FET pre-amplifier (Fig.1). The latter has to amplify the signal and simultaneously minimise the load of the crystal filter. The drain oscillation circuit is tuned to the intermediate frequency in use.

3. PRE-TUNING

1. An intermediate-frequency signal is fed in to pin-7 at -70dBm. The indicator is set precisely to 0µA using potentiometer P2.

2. Raise the signal level to -20dBm and tune the indicator deflection to 100µA (full-scale reading). The linearity is controlled by reducing the signal in 10dB stages.

Should it not be possible to calibrate, P1 can be replaced by an 8.2kΩ resistor.

4. INSTRUCTIONS FOR ASSEMBLY

❶ The completed circuit should be pre-tuned and tested for correct operation before assembly, together with the indicator, using a standard signal generator and a reducer (corresponding to 3).

❷ The signal for the S-meter is drawn off after the SSB crystal filter. To keep the additional capacity load low, a UHF-FET with a small Cg1S should be used.

❸ The MOS-FET amplifier should be mounted in close proximity to the crystal filter. Using the SMD technique for assembly is advantageous. The remaining part of the circuit is placed in a free corner of the equipment and connected by means of thin coaxial cable.

❹ No AGC stages should be present between the RX input and the S-meter. Should there be an AGC, disconnect it and replace it by fixed voltages.

❺ The display range of the S-meter incorporated into the RX is dependent on the amplification of the equipment between the aerial socket and the crystal filter output.

Examples of MOS-FET stages:

AGC prior to G1; now connect G1 at 0 V DC.

AGC prior to G2; now connect G2 (DC), at about 4 V higher than the voltage connected to the source (measured against the earth).

5. FINAL TUNING

❶ The built-in S-meter is now adjusted to the equipment. The output circuit of the S-meter pre-amplifier is tuned to the maximum reading by means of a test signal at the receiver input.

❷ Now increase the input signal in 10 dB stages until the indicator registers 100 micro-A as precisely as possible. Set the pointer to full-scale reading (100 micro-A) using P2. The linearity is checked by reducing the input signal. Finally the equipment is marked with an appropriate scale.

❸ Example: a signal of -80dBm gives, e.g., 96µA. So this is corrected to 100µA using P2. Thus we can obtain: 100µA = -80dBm, 80µA = -90dBm, 60µA = -100dBm.

6. LITERATURE

(1) Specification sheet MC 3356 from Motorola

(2) Specification sheet NE 604 / NE 605 from Signetics / Philips

Eugen Berberich, DL 8 ZX

Output Wiring of GaAsFET Amplifiers

The wiring of GaAsFET pre-stages in the output circuit is still giving grounds for discussion.

1. POSSIBLE SOLUTION METHODS

The solution using the 4:1 transformer (1) requires skill in the construction of the repeater and is always linked with unwanted resonances at higher frequencies. The RX input impedance acting as terminator is also strongly dependent on frequency. So it is not surprising if instability effects occur with the relatively large gate drain.

Other authors (2) provide the drain by means of a high resistance from a high voltage (12 V), with the residual voltage decaying. The author has also found this solution to be effective. The HF decoupling is usually provided through a coupling capacitor or a coil for matching, or a 4:1 transformer (3), which in this case is dampened by a resistance and the Ri of the amplifier. Such a pre-amplifier often tends to oscillate at a frequency higher than the operating frequency, because the reactive components have a stronger effect at higher frequencies.

In the drain wirings described above, the transistor in the first place "sees" an inductance, which represents a frequency-dependent operating resistance. It is thus already pre-determined that at higher frequencies the amplification is also greater, and so the influence of the Cg-d increases. The 4:1 repeater connected also works less and less effectively as the frequency rises, since the leakage inductance becomes noticeable. The connected RX input represents a frequency-dependent load, which scarcely imposes any burden on the 4:1 repeater outside the desired frequency range. The transistor output "sees" no real load in a wide frequency range. Figs. 1a to 1d show the usual drain wirings.

2. PI TRANSFORMATION

My proposed circuit uses a pi circuit (low-pass) as a transformer, going from app. 250 to 50Ω at the output. Starting from the internal resistance of the transis-

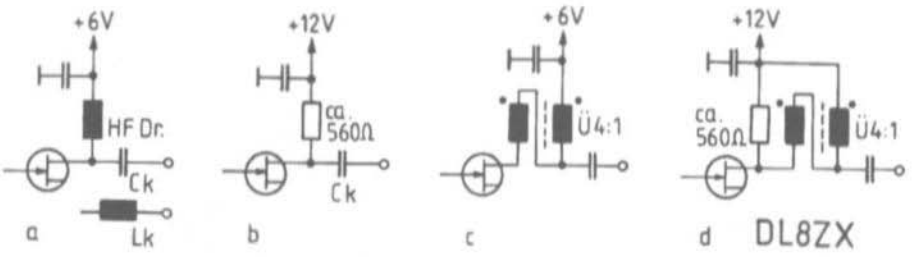


Fig.1a to d: Standard Drain wirings

tor output (app. 500Ω) and the parallel circuit of the mains resistance (also app. 500Ω), we obtain a source resistance of approximately 250Ω.

As is often demonstrated at the aerial output in transmission amplifiers, various impedance values can be matched using a pi filter network. The advantages of this circuit are, firstly, the correct matching, but also, and of particular advantage, that the drain connection is capacity-loaded for frequencies other than the operating frequency.

This means a high-frequency short-circuit for high frequencies outside the desired range. The capacitor, C1, on the drain must therefore be selected to have low induction (SMD type) and must be mounted directly on the drain. The components can now be calculated using the standard pi filter formulae. The transistor output and switching capacity is added to C1 and must be taken into consideration.

Providing power through a resistance of app. 500Ω has the disadvantage that part of the high-frequency output is lost. With the high amplification resources of modern GaAsFETs, we usually do not need all the amplification possible, since with amplification levels of 20dB there is already a danger of saturation for the receiver.

Nor are the active modules in commercial amplifier engineering designed completely for maximum amplification. Instead, they bring about linear and stable behaviour by means of negative feedback. Nor is the once standard neutralisation used. The solution in (3) takes a fresh look at standard practise from valve engineering.

Unfortunately, very little is written about negative feedback in amateur publications. With regard to HF amplifiers, you could ferret out something on linearity to help reduce the inter-modulation. Unfortunately, negative feedback in preamps usually makes the noise factor worse, except as shown in (4).

3. PROPOSED CONSTRUCTION

Now to the circuit options I propose for pre-amplifiers (Fig.2). One example is calculated for the 2-m. band. Table 1 shows the values for the 70cm. and 23cm. bands. Q = 12 is taken as the operational quality.

$$C_1 = \frac{Q}{\omega \cdot R_1} = \frac{12}{6{,}28 \cdot 145 \cdot 10^6 \cdot 250} \approx 52\,\text{pF}$$

$$X_{C1} = \frac{R_1}{Q} = \frac{250}{12} = 20{,}8\,\Omega$$

$$X_{C2} \approx 9{,}5$$

$$C_2 = \frac{1}{\omega \cdot X_{C2}} = \frac{1}{6{,}28 \cdot 145 \cdot 10^6 \cdot 9{,}5} = 115\ \text{pF}$$

$$L_1 \approx \frac{X_{C1} + X_{C2}}{\omega} = \frac{20{,}8 + 9{,}5}{6{,}28 \cdot 145 \cdot 10^6} = 33{,}3\ \text{nH}$$

$$X_{C2} = \frac{R_2}{\sqrt{\frac{Q^2 \cdot R_2}{R_1} - 1}} =$$

$$= \frac{50}{\sqrt{\frac{144 \cdot 50}{250} - 1}} = 9{,}48$$

The closest standard values can be used for the capacitors. Problems arise in bringing about very low inductance levels. For the 2 m. coil, 3 windings were used with a diameter of 4 mm. (0.4 CuL wire). Matching can be done using a suitable ferrite core. For the higher bands, we just need a shorter wire strap! C2, laid out as a trimmer, allows the matching to be tuned. C1 should not be laid out as a trimmer because of the high internal inductance, although this could optimise the matching.

Amateurs who have access to a network analyser can use the test rig shown in Fig.3 for tuning. C2 and L1 act as tuning elements (change length of wire strap). In addition, this output wiring can also be successfully used for mixed stages; e.g. for UHF/SHF mixers which convert, for example, to 2 m. Here the mixer operates as a drain circuit for the input in the UHF/SHF range, since the drain connection through C1 for the input frequency is earthed.

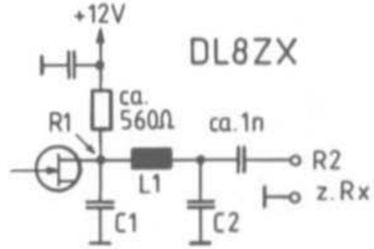


Fig.2: π-transformation at the output of a GaAsFET amplifier

Band	L_1	C_1	C_2
2 m	33 nH	52 pF	115 pF
70 cm	11 nH	18 pF	39 pF
23 cm	3,7 nH	6 pF	13 pF

Table 1

4. LITERATURE

(1) W 6 PO: Some GaAsFET Preamplifiers
Eimac Applikation Note As-4931

(2) YT 3 MV: A very low-noise pre-amplifier for the L-band
VHF Communications 2/92, p. 90

(3) Detlev Burchard: MES-FETishism 2
VHF Communications 4/92, p. 223

(4) DJ 7 VY: New-type pre-amplifiers for 145 MHz and 435 MHz receivers

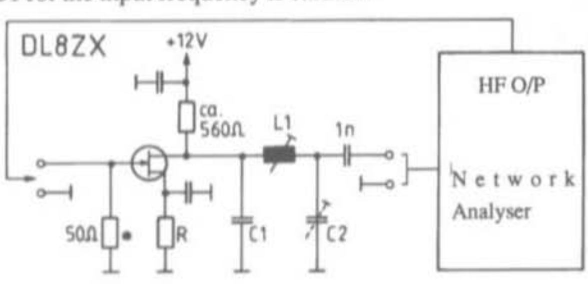


Fig.3: Test rig with Network Analyser

*** soldered in for tuning only**

Mike Wooding, G6IQM

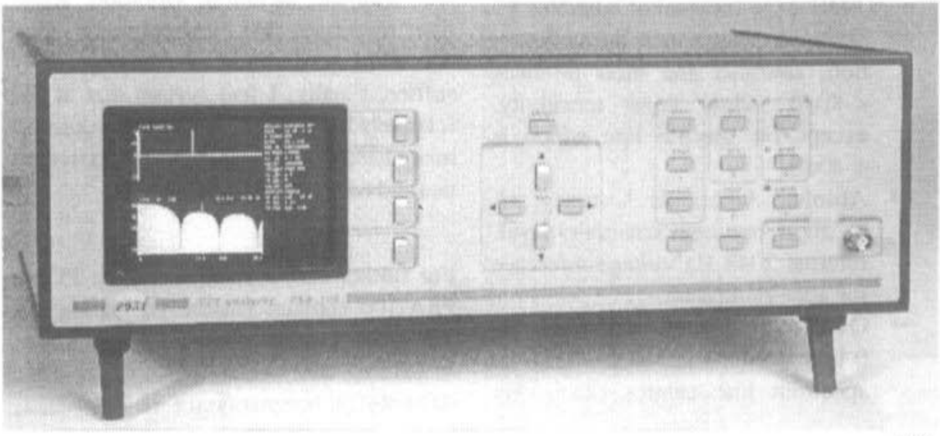
The Pont PSA-100 FFT Spectrum Analyser

Review

The Pont PSA-100 is an Audio Spectrum Analyser operating in the frequency range DC to 25.6 kHz. Although reviews of such equipment do not usually appear in VHF Communications, I thought it worth a brief mention as many of our readers are professional engineers, and such instruments are used widely in broadcast and audio engineering.

The unit has an impressive range of features for what is a relatively low-cost instrument. Using powerful Digital Signal Processing (DSP) techniques a range of measurements is made available, including basic audio spectrum analysis, auto-correlation, cepstrum analysis and a true RMS voltmeter mode.

The instrument is very neat and compact, measuring 440 x 320 x 145mm and weighing only 6kg. The spectra or voltage measurements are displayed on a 5'' CRT, with an auxiliary composite video output to drive an external monitor and a parallel interface for a graphics printer.



There are few user controls, the various measurement modes and parameters being set up using on-screen multi-level menus.

As standard the instrument is fitted with 16k of battery backed-up RAM for recording measured spectra and measurement set-ups. Optional extra memory boards can be installed if required. Also available as optional extras are a GPIB/IEEE and RS-232C interface cards, to facilitate automatic measurements under external computer control.

As previously mentioned, the measured spectra are displayed on a 5'' CRT as a 400 line image, with either a single full-screen image or two separate spectra, one in the top half and the other in the bottom half of the display area. The span of the image can be over the whole 25 kHz band or over any sub span (in powers of 2) down to a minimum of 12.5 Hz and the spectrum can be zoomed by up to 32768 times around any line. The instrument also has 'smart' cursors for spectral measurements. In the RMS voltmeter mode the display consists of large alphanumeric characters with the reading user-selectable in Volt or dB units.

The basic parameters of the instrument are:

- Input Sensitivity: 1mV to 31.6V in 10dB steps (manual or autorange)
- Dynamic Range: non-linear distortion, spurious and alias products <-80dB below input sensitivity, except 1st spectral line which is <-40dB.
- Absolute Amplitude Accuracy: +/- 0.2dB at any input sensitivity level. Internal 2048 Hz voltage reference for auto-calibration.
- Overall Frequency Response: +/- 0.1dB from 10 Hz to 25.6 kHz at spectrum line centres (25.6 kHz span). Maximum +/- 0.18dB with zoom filter (span less than 25.6 kHz or in zoom mode).
- Measurement Modes: Baseband Spectrum, Zoom. Autocorrelation, Cepstrum, 1/3 Octave, Full Octave and digital Voltmeter.

In conclusion, this instrument represents real 'value-for-money' in these difficult times, with a price range that begins at £2200 for the basic model. Many of the instrument's features are normally only found on much more expensive units and the basic accuracy of the measurement modes is more than enough for most modern measurement needs. The control functions are very 'user friendly' and the on-screen menus easy to navigate.

My only real criticism, and this is only because I regularly use a Dynamic Signal Analyser with a much higher specification and price!, is that the bandwidth of the instrument is only 25.6 kHz,. This could tend to make audio harmonic measurements less meaningful, as the highest 2nd harmonic measurement than can be made is on a fundamental of 12.8 kHz, and the third harmonic on a fundamental of 8.5 kHz. That aside, for many other types of measurement, including vibration, audio filter responses, and electroacoustic, then the instrument could more than fully suffice. Finally, I feel certain that at the relatively low cost the instrument could fill many needs in schools, colleges, universities and educational institutes.

For further information about the PONT PSA-100 contact: Manor Technology, 8 Manor Road, East Tytherley, Salisbury, Wiltshire, SP5 1LN, U.K. Telephone: 0794 40923; International + 44 794 40923.

MATERIAL PRICE LIST OF EQUIPMENT

described in VHF COMMUNICATIONS

DB1NV	**Digital Image-Store for the Spectrum Analyser**		**Art. No.**	**Ed. 4/1991**
PCB	DB1NV 010		6477	DM 44.00
Components	Processor P80C31; 12 ICs; 1 Reg. IC, Transistor; Zener Diodes; Silicon Diodes; Chokes; RAM; EPROM DB1NV 010; 4 x 2k Trimpots; Crystal		6478	DM 276.00
DB1NV	**Tracking Generator for the Spectrum Analyser**		**Art. No.**	**Ed. 1/1992**
PCB	DB1NV 011		6479	DM 31.00
F6ILR F6BXC	**A Digital Slow-Scan Television Transmit Coder**		**Art. No.**	**ED. 3/1992**
PCB	SSTVCODE 1 (KM Publications)		SSTV1	£ 28.50
DB1NV	**Broadband VCO's using Microstrip Techiniques**		**Art. No.**	**Ed. 4/1992**
PCB	DB1NV 012		6480	DM 33.00
PCB	DB1NV 013		6481	DM 33.00
Components	400 - 1250 MHz 3 x BB619; 1 x BB811; 1 x BFG96; 2 x AT42085; 1 x BFQ69; 2 x 2.2nF & 1 x 27pF Feed-through Cap.; SMC Connectors; 2 x 0.47 H SMD Choke; 1 x housing 74x55x30mm; 1 PCB DB1NV 012		6482	DM 81.00
Components	450 - 1450 MHz as above but: 1 x BFG65 instead of BFG96		6483	DM 81.00
Components	800 - 1900 MHz as above but: 4 x BB811 instead of BB619 and PCB DB1NV 013 instead of 012		6484	DM 85.00
Tuning Diode	BB619	10 off	10450	DM 20.00
Tuning Diode	BB811	10 off	10451	DM 31.50

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