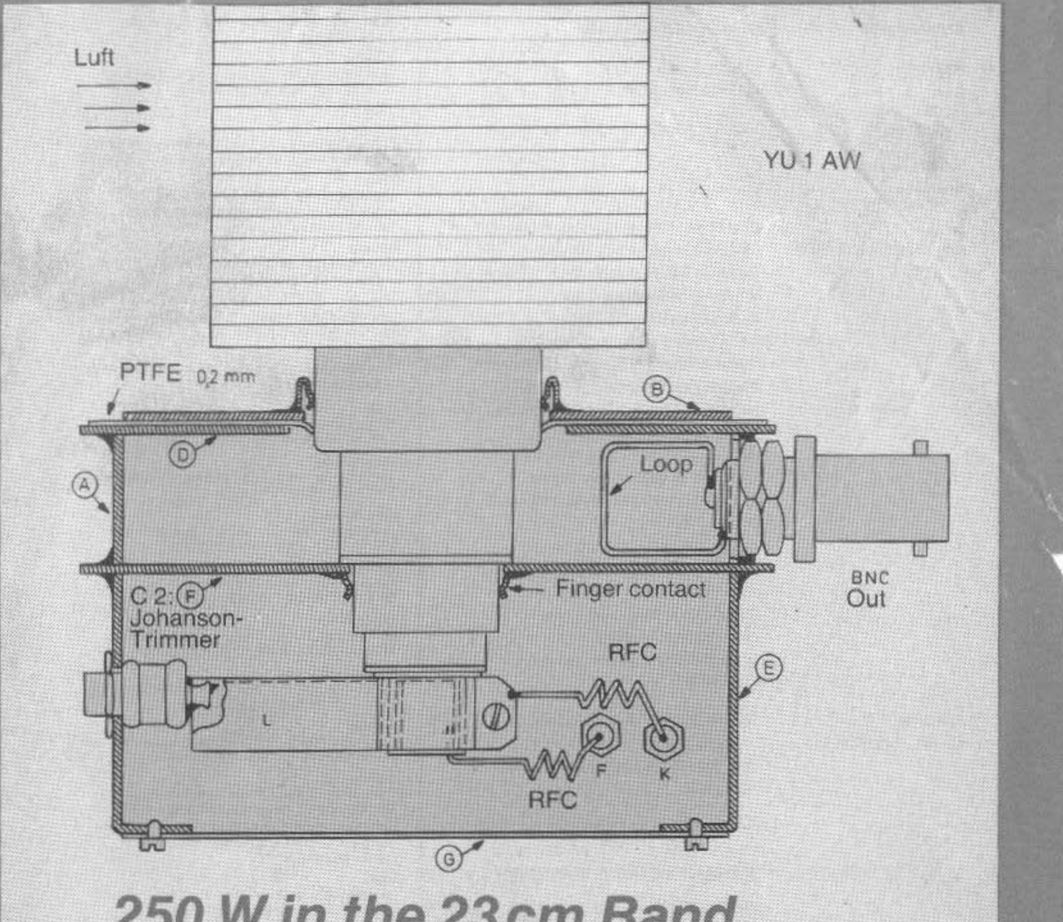
J 20 419 F

A Publication
for the Radio-Amateur
Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 19 · Summer · 2/1987 · DM 7.00



250 W in the 23 cm Band

VHF communications

A Publication for the Radio Amateur
Especially Covering VHF, UHF, and Microwaves

Volume No. 19 · Summer · Edition 2/1987

Published by: TERRY BITTAN oHG, P.O.Box 80, Jahnstraße 14, D-8523 BAIERSDORF Fed. Rep. of Germany Telephone (9133) 47-0 Telex 629 887 Telefax 0 91 33-47 18 Postgiro Nbg. 30455-858

Publishers: TERRY BITTAN oHG

Editors: Corrie Bittan
Colin J. Brock (Assistant)

Translator: Colin J. Brock, G 3 ISB / DJ Ø OK

Advertising manager: Corrie Bittan

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 published in Spring, Summer, Autumn and Winter. The 1987 subscription price is DM 24.00 or national equivalent per year. Individual copies are available at DM 7.00 or equivalent each. Subscriptions, orders of individual copies, purchase of PC-boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative, or – if not possible – directly to the publishers.

Printed in the Fed. Rep. of Germany by R. Reichenbach KG Krelingstr. 39 · 8500 Nuernberg.

We would be grateful if you would address your orders and queries to your representative.

Representatives

Austria
Verlag UKW-BERICHTE, Terry D. Bittan
POB 80, D-8523 Baiersdorf / W. Germany
Creditanstalt Bankverein, WIEN Kto. 17-90.599
PSchKto WIEN 1.169.146

Australia
W.I.A. P.O. Box 300, South Caulfield, 3162 VIC,
Phone 5285962

Belgium
HAM INTERNATIONAL, Brusselsesteenweg 428,
B-9218 GENT, PCR 000-1014257-25,
Tel. 00-32-91-312111

Denmark
Halskov Electronic, OZ 7 LX, Sigersted gamle Skole,
DK-4100 RINGSTED, Tel. 03-616162. Giro 7 29 68 00

France
Christiane Michèl, F 5 SM, SM Electronic,
20 bis, Avenue des Clairions, F-89000 AUXERRE
Tel. (86) 46 96 59

Finland
Erkki Hohenthal, SF-31 400 SOMERO
Joensuuntie 6, Tel. 924-46311

Holland
DOEVEN-ELEKTRONIKA, J. Doeven, Schutstraat 58,
NL-7901 EE HOOGEVEEN, Tel. 05280-69679

Israel
Z. Pomer, 4 X 4 KT, P.O. Box 222, K. MOZKIN 26114
Tel. 00972-4714078

Italy
Franco Armenghi, I 4 LCK, Via Sigonio 2,
I-40137 BOLOGNA, Tel. (051) 34 56 97

Luxembourg
TELECO, Jos. Faber, LX 1 DE, 5 - 9, Rue de la fontaine,
ESCH-SUR-ALZETTE, Tel. 53752

New Zealand
E. M. Zimmermann, ZL 1 AGQ, P. O. Box 31-261
Milford, AUCKLAND 9, Phone 492-744

Norway
Henning Theg, LA 4 YG, Postboks 70,
N-1324 LYSAKER, Postgirokonto 3 16 00 09

South Africa
HI-TECH BOOKS, P. O. Box 1142, RANDBURG,
Transvaal 2125, Tel. (011) 886-2020

Spain + Portugal
Julio A. Prieto Alonso, EA 4 CJ, MADRID-15,
Donoso Cortès 58 5°-B, Tel. 243.83.84

Sweden
Lars Pettersson, SM 4 IVE, Pl. 1254, Smégården Talby,
S-71500 ODENSBACKEN, Tel. 19-50223, Pg. 914379-3

Switzerland
Terry Bittan, Schweiz. Kreditanstalt ZÜRICH,
Kto. 469.253-41; PSchKto. ZÜRICH 80-54.849

Leo Kälin, HB 9 CKL, Funktechnik
Alte Landstr. 175, CH 8708 Männedorf
Tel. 01-9203535

USA
UV COMMS, K 3 BRS
P. O. Box 432, LANHAM, MD 20706
Phone 301-459-4924

ISSN 0177-7505

Contents

UKW**berichte** Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133 47-0. For Representatives see cover page 2

Dr. (Eng.) Ralph Oppelt, DB 2 NP

The Generation and Demodulation of SSB Signals using the Phasing Method
Part 1: Basic Theory

On account of certain advantages, proprietry amateur SSB equipment, without exception, utilize the filter method of generation and detection. It may be perhaps of interest for the active radio amateur to develop alternative circuit principles – for example an SSB transceiver using the phasing method. Such alternative solutions can bring very satisfactory results and in many aspects, more efficacious results than the slavish adherence to the normal commercial design principles.

In a two-part article, the possibility is presented of constructing an SSB/DSB/AM transceiver which uses no quartz filters and works with the phasing principle. This present article deals with the principle of phasing modulation and demodulation and also comparisons with the filter generation technique. A mathematical foundation for the design of the required Hilbert transformers (the AF phase processing transformers) is also presented to the amateur from the authors own work on the subject. The second part will be devoted to a practical realisation of the most important components necessary for the SSB phasing method – the Hilbert transformer encompassing two active fourth-order, all-pass filter chains.

1. SSB PHASING METHOD PRINCIPLES

The easily understood, and thereby the most familiar, filter method of producing SSB signals uses a narrow-band filter to select one sideband of a double sideband signal and reject both the other sideband and the carrier. The more subtle principle of the phasing method of generation will be explained with the aid of **fig. 1** in time sequence (mathematically simpler). In order to simplify the mathematics, the input AF signal is taken as comprising a steady amplitude single frequency tone ω. Also, all amplitudes are normalized at unity. **Fig. 1** shows that, due to the phase-shifting networks, $\Delta\varphi_{T1}$ and $\Delta\varphi_{T2}$ together with $\Delta\varphi_1$ and $\Delta\varphi_2$, the HF carrier ω_T and the AF modulating signal ω are present at the balanced mixer ports all at equal amplitude but with a 90° phase difference i.e. as sine and cosine signals. The output ports of the mixer thereby contain:

upper mixer:

$$\cos \omega_T t \cdot \sin \omega t = 1/2 [\sin (\omega_T + \omega) t - \sin (\omega_T - \omega) t] \qquad (1)$$

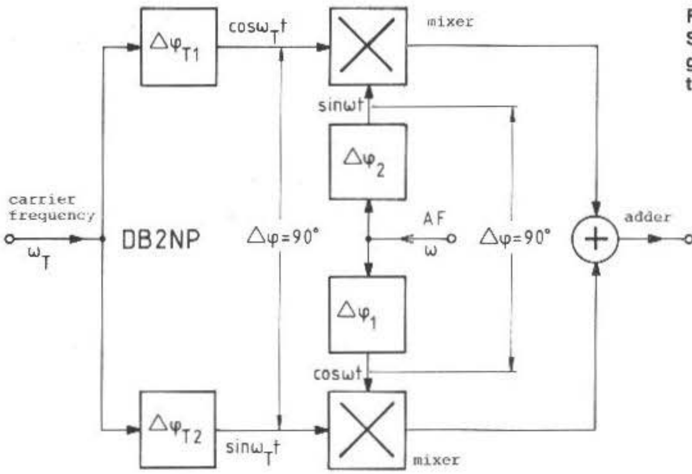


Fig. 1: SSB signal generation using the phasing method

lower mixer:

$$\sin \omega_T t \cdot \cos \omega t = 1/2 [\sin (\omega_T + \omega) t + \sin (\omega_T - \omega) t] \qquad (2)$$

Where the two mixer outputs are summed, only the term $\omega_T + \omega$ remains i.e. the upper side frequency. If equations (1) and (2) are substracted, on the other hand, only $\omega_T - \omega$ (the lower side frequency) remains. Should only an adder be used, as in **fig. 1**, i.e. no manipulation of the HF feeds, a sideband change, but with the advantage of it being not at HF, is possible by crossing the audio frequency lines cos ωt and sin ωt as shown by equations (3) and (4) below.

$$\cos \omega_T t \cdot \cos \omega t = 1/2 [\cos (\omega_T + \omega) t + \cos (\omega_T - \omega) t] \qquad (3)$$

$$\sin \omega_T t \cdot \sin \omega t = 1/2 [- \cos (\omega_T + \omega) t + \cos (\omega_T - \omega) t] \qquad (4)$$

The addition of (3) and (4) results in the production of the lower side frequency as opposed to the result obtained by adding the equations (1) and (2). For the sake of completeness, it must be mentioned that the arrangement of **fig. 1** can easily be applied to the production of both an AM and a DSB signal. It is only necessary to feed the AF ports of the two mixers with in-phase modulation signals e.g. cos ωt, instead of sin ωt and cos ωt, the HF port feed remaining the same, and the output product will be as follows:

upper mixer:

$$\cos \omega_T t \cdot \cos \omega t = 1/2 [\cos (\omega_T + \omega) t + \cos (\omega_T - \omega) t] \qquad (5)$$

lower mixer:

$$\sin \omega_T t \cdot \cos \omega t = 1/2 [\sin (\omega_T + \omega) t + \sin (\omega_T - \omega) t] \qquad (6)$$

It may be seen, that the addition of equations (5) and (6) leads to the production of two sinusoidal waves with sum and difference frequencies of ($\omega_T \pm \omega$) i.e. a double sideband signal. The

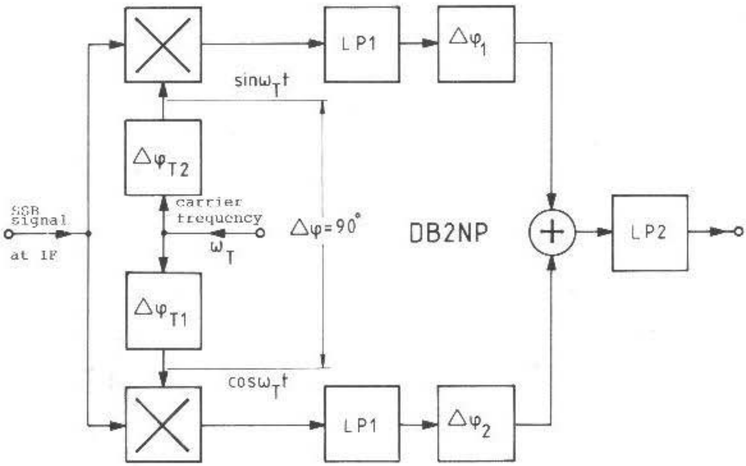


Fig. 2: Circuit arrangement for SSB demodulation using the phasing method

addition of a carrier completes the conversion to an AM signal. This could be carried out by the judicious unbalancing of the balanced mixers.

The re-configuration of the individual function blocks of **fig. 1** for the "true" demodulation of an SSB signal ("true" means that the RX hears one sideband only) is shown in **fig. 2**. The relevant mathematics are not explicitly considered in this case as they bear a strong resemblance to the form already worked out for the modulator.

Three low-pass filters have been introduced into the basic arrangement to form **fig. 2**. The identical low-pass filters (LP 1) after the mixer output ports are designed to eliminate the second carrier harmonic 2 ω_T and can be considered as simple RC or LC elements. The low-pass filter after the combiner, on the other hand, determines the received output band-pass and selectivity characteristics. It is therefore required to be a multi-pole band-pass filter with steep-sided flanks and may, in practice, be switched to provide optimal band-pass characteristics for either speech or CW. A sideband switch, similar to that of the modulator of **fig. 1**, may be carried out in the audio arms of the mixers' outputs. The inputs of the phase-shifters $\Delta\varphi_1$ and $\Delta\varphi_2$ are merely cross-connected to the outputs of the filters LP 1.

Since only one sideband is being received by the circuit depicted in **fig. 2**, a first-class reception of DSB signals is possible and without the changes in loudspeaker level which accompany the indefinate phase relationship between the conversion carrier ω_T and the suppressed carrier at the transmitter. It is recommended also, that a 300 Hz low limit should be made to the filters LP 1 (i.e. the lower limit of intelligible speech) in order that any carrier beating remains inaudible.

Whereas the carrier (i.e. oscillator ω_T) remains constant at, for example 9 MHz, the speech signals occupy, in reality, a spectrum from 300 to 3000 Hz – although for the purposes of analysis only a single side frequency was considered. This demands of the phase-shifters that they cause a 90° phase-shift at all frequencies within this spectrum and at the same relative amplitudes. Simple phase networks are liable to be frequency-dependent in either amplitude or phase. A correction to make one quantity constant results in the other being dependent upon frequency. The design of these phase-shift networks $\Delta\varphi_1$ and $\Delta\varphi_2$, to cover the entire speech band, is considerably more difficult than for the carrier phase-shift networks as the latter work at a single frequency.

2. FILTER AND PHASING METHODS – IMPORTANT FEATURES COMPARED

If the facility is required which permits the changing from upper to lower sideband at will, the phasing method is inherently simpler – only a suitable switch being required. The filter method involves, at the simplest, using another inserted carrier crystal or, much more expensively, keeping the carrier frequency fixed and switching in a filter for the other sideband.

The carrier suppression and the unwanted sideband suppression are undoubtably better using the filter method as a compensation technique must be used in the phasing method to ensure equal 90° phase shift across the whole of the speech band. This compensation is also affected by temperature and long-term stability considerations etc. The low conversion efficiency does not affect the transmit modulator to the same extend and for amateur purposes it is quite satisfactory. The phasing demodulator will be affected by compensation variations at low received signal levels which are in the presence (say 10 kHz removed) of a strong unwanted signal. The low-pass filter in the audio arms could easily separate the two signals but the mixers are located after the IF chain and will be overloaded. The only way to prevent this is to use a crystal band-pass filter in the IF chain. But that contradicts one of the tenants of this article i.e. not to use expensive crystal filters.

For the normal radio amateur that pursues his hobby for pleasure, even on contest-free days, a transceiver working entirely without crystal filters may be satisfactory. This is, perhaps, particularly the case when one considers the two-metre band. The higher complexity of the phasing method in terms of number of circuit elements (double the number of mixers, phase processors etc.) is offset by the fact that they are all cheaper than the components required for the filter method. The time taken to adjust and optimize the phasing rig (optimizing of carrier and sideband suppression) must also be taken into account as this aspect can almost be neglected with a ready-made proprietry crystal filter. In the second part of this article a method of lining-up a phasing modulator will be described which may be successfully carried out even by those who – quite by oversight – do not possess a spectrum analyzer.

As was made clear earlier in chapter 1, the phasing SSB method of processing is inherently suitable, not only for the simplicity of changing between upper and lower sidebands, but also for its ability to receive DSB and (carrier) AM with equal facility – all without using extra circuit components. Also, the sideband change facility results in a genuine change-over from one sideband to the other i.e. the reference carrier (suppressed) remains at exactly the same frequency. The most cost effective method to switch between LSB and USB for a filter modem is to shift the conversion (suppressed) carrier a few kHz and thereby utilizing only one crystal filter (the filter method, it will be recalled, would have to employ two balanced mixers and two crystal filters, USB and LSB, to perform the same feat of using a fixed frequency conversion carrier when a sideband switch facility is required – G3ISB).

3. CARRIER AND AUDIO PHASE NETWORK DESIGN

In chapter 1 it was stated that the design of the phase-shift networks $\Delta\varphi_{T1}$ and $\Delta\varphi_{T2}$ for the high frequency carrier signal is relatively easily carried out. The handiest form would be RC networks as shown in **fig. 3**. This shows a high-pass and a low-pass of the first order (1).

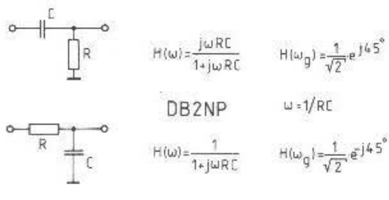


Fig. 3: High and low-pass first order and relevant transfer function H for the carrier phase networks $\Delta\varphi_{T1}$ and $\Delta\varphi_{T2}$.

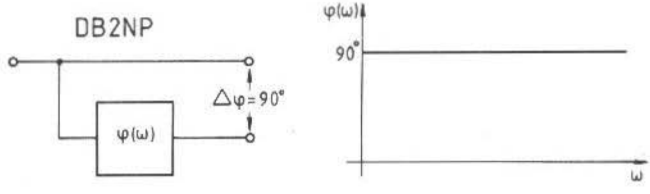


Fig. 4:
All-pass filter Hilbert transformer with constant phase versus frequency $\varphi(\omega) = 90°$

If these networks are driven at exactly their 3 dB limit frequency points, $\omega = \omega_g = 1/RC$, then the transfer function H (i.e. the relationship between complex input and output amplitudes) are both equal to $1 / \sqrt{2}$. Whilst the high-pass network shifts by $-45°$ at ω_g the low-pass network shifts by $+45°$ at ω_g. Both networks thus have the desired differential phase difference of 90° between each other. The 3 dB turn-over frequency of ω_g is, of course, arranged to be the modem carrier frequency ω_T. Thus a filter parameter is formed, in each case, by varying the quantity R until the 3 dB limit frequency $\omega_g = 1/RC$ – which is that of the carrier frequency ω_T.

The design of the audio frequency phase elements $\Delta\varphi_1$ and $\Delta\varphi_2$ cannot be satisfactorily accomplished using simple high and low-pass filter elements. This is because the speech band is required to have a constant amplitude response across the entire spectrum. The so-called all-pass filters are, however, suitable. The all-pass filters have a transfer function H which is constant for all frequencies but has a frequency dependent phase transfer characteristic. The described characteristic, however, is that of a constant 90° shift at all frequencies within the audio band (**fig. 4**).

Such a filter, known as the Hilbert filter or Hilbert transformer, does not, unfortunately, match the ideal characteristics presented in **fig. 4**. It can be approached by limiting the speech band in order that it is as narrow as possible consistent with intelligible speech. If necessary, the speech band may be limited by means of a band-pass filter inserted in the audio arm immediately before the phase-shift circuit. The second limitation concerns the 90° phase-shift in the output signal: a tolerance band, say $90° \pm 1°$ must be allowed in order that the complexity of the phase-shift network design is not greater than is necessary for the overall performance objectives of the modem.

The procedure for the signal processing is clear from **fig. 5**. The speech signal is split and presented to two all-pass filters $\Delta\varphi_1$ and $\Delta\varphi_2$ whose phase characteristics are also shown in **fig. 5**. Both phase networks may be of the all-pass type of the n th order i.e. consisting of a chain of first order all-pass elements. As the characteristic shows, a signal passed through such an all-pass filter chain has a phase shift at low frequencies approaching zero and at high frequencies the value of n 180°. Both all-pass chains can be so dimensioned that the phase characteristic of $\Delta\varphi_2$ approaches the limiting value n 180° 'earlier' than

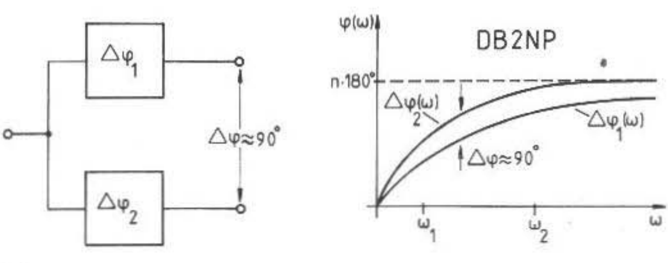
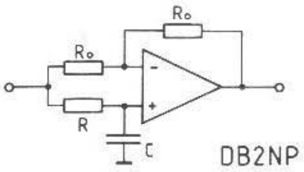


Fig. 5:
Two n th order all-pass filters together with phase characteristics for the realization of the Hilbert transformer for the range $\omega_1 \leq \omega \leq \omega_2$



$$H(\omega) = \frac{1 - j\omega\tau}{1 + j\omega\tau} = e^{-j2\arctan\omega\tau}$$

$$\Rightarrow |H(\omega)| = 1 \text{ und } \varphi(\omega) = -\sphericalangle H(\omega) = 2\arctan\omega\tau$$

$$\tau = R \cdot C$$

Fig. 6: Active first order all-pass filter and transfer function H (ω)

that of $\triangle\varphi_1$. In a certain frequency range ω_1 to ω_2, the condition $\triangle\varphi = \triangle\varphi_2 - \triangle\varphi_1 = 90°$ can be very well approximated.

Fig. 6 shows an active first order all-pass filter and its accompanying transfer function H (ω) (1). Now it must be determined how many of such elements are required for the all-pass chain filters $\triangle\varphi_1$ and $\triangle\varphi_2$ and how the time-constant characteristic τ = RC are to be dimensioned. Only a final calculation is given here but a mathematical treatise on the subject may be found in (3).

The first step is to fix the frequency interval ω_1 to ω_2. The speech signal range $\omega_1 = 2\,\pi\,270$ Hz and $\omega_2 = 2\,\pi\,3600$ Hz offers some latitude, both above and below and gives a frequency ratio Ω of:

$$\Omega = \omega_1 / \omega_2 = 0.075 \qquad (7)$$

secondly, the number n of all-pass stages per phase-shift chain must be determined. In general, it may be said, that the approximation to the shift of 90° rises with the number of stages used. The lower limit to the number of phase-shift elements is set by the sideband suppression desired and in no account must less than two be used. In practice, a suitable number of stages would be four and that is used in this example. As each all-pass filter element requires one op-amp, and there are four op-amps cheaply obtainable in one IC package, the choice of four stages is well justified. Using Ω from equation (7) the first auxiliary quantity ε may be calculated:

$$\varepsilon = \frac{1}{2} \cdot \frac{1 - \sqrt{\Omega}}{1 + \sqrt{\Omega}} \qquad (8)$$

For $\Omega = 0.075$, $\varepsilon \approx 0.285$.

The next auxiliary quantity q is determined with the aid of ε in a rapidly convergent series:

$$q = \varepsilon + 2\,\varepsilon^5 + 15\,\varepsilon^9 + 150\,\varepsilon^{13} + 1707\,\varepsilon^{17} + \ldots (9)$$

Only the first two or three terms of this series are relevant and

$q \approx 0.289$ (as a comparison $150\,\varepsilon^{13} \approx 1.2 \times 10^{-5}$)

Finally, the auxiliary quantity k_v is required

$$k_v = \frac{4\nu + 1}{8\,n} \qquad (\nu = 0,1, \ldots.. n - 1) \qquad (10)$$

for n = 4 it follows that:

$k_0 = 1/32$, $k_1 = 5/32$, $k_2 = 9/32$ and $k_3 = 13/32$. With these values of k_v the **table** given below is formed.

	$\sin \pi k_v$	$\cos \pi k_v$	$\sin 3\pi k_v$	$\cos 3\pi k_v$	$\sin 5\pi k_v$	$\cos 5\pi k_v$
k_0	0.098	0.995	0.290	0.957	0.471	0.882
k_1	0.471	0.882	0.995	0.098	0.634	− 0.773
k_2	0.773	0.634	0.471	− 0.882	− 0.957	− 0.290
k_3	0.957	0.290	− 0.634	− 0.773	0.098	0.995

The time-constant of the four all-pass stages (ν = 0, 1, 2, 3) comprising the phase network $\triangle\varphi_2$ is given by:

$\tau_{\nu 2} = R_\nu C_\nu$

By means of the calculated auxiliary quantity q and the tabulated values given above, the time-constants are calculated as follows:

$$\tau_{\nu 2} = \frac{1}{\sqrt{\omega_1 \omega_2}} \cdot \frac{\cos \pi k_\nu + q^2 \cos 3 \pi k_\nu + q^6 \cos 5 \pi k_\nu + q^{12} \cos 7 \pi k_\nu + ...}{\sin \pi k_\nu - q^2 \sin 3 \pi k_\nu + q^6 \sin 5 \pi k_\nu - q^{12} \sin 7 \pi k_\nu \pm .....} \quad (11)$$

Here, also, only the first three or four terms of the series in the numerator and the denominator need be evaluated which is why **table 1** was not continued. Following the insertion of the values from the table all four all-pass stages of the network $\triangle\varphi_2$ are specified: τ_{02} = 2.3448 ms, τ_{12} = 369.52 μs, τ_{22} = 123.44 μs and τ_{32} = 36.173 μs. The time constant $\tau_{\nu 1}$ for the phase network $\triangle\varphi_1$ is simply found from the known $\tau_{\nu 2}$ according to equation (12)

$$\tau_{\nu 1} = (\tau_{\nu 2}\omega_1\omega_2)^{-1} \quad (12)$$

The following values were obtained for this example:
τ_{01} = 11.114 μs, τ_{11} = 70.524 μs, τ_{21} = 211.11 μs and τ_{31} = 720.43 μs.

In this manner, all eight RC elements are made known for the design of the fourth order all-pass chains. Since merely the product of R and C must have a definate value, the value of C, in practice, is selected first and measured, then the required resistor value is calculated. This procedure is advantageous because resistors are available in many more values and to a greater tolerance than capacitors.

The realization of all-pass stages with active filters has, above all, the advantage that every stage is decoupled from its fellows. A common characteristic impedance is thereby obviated as a by-product. Using the function φ (ω), given in **fig. 6** for a first order all-pass, the phase difference $\triangle\varphi = \triangle\varphi_2 - \triangle\varphi_1$ at the outputs of the chains calculated in the example are given as a function of frequency, eq. (13). It departs (purely calculated) in the frequency interval ω_1 to ω_2 by only 0.011° from the specified 90°.

$$\triangle\varphi(\omega) = 2 \sum_{\nu=0}^{3} (\arctan \omega \tau_{\nu 2} - \arctan \omega \tau_{\nu 1}); \quad (13)$$

4. LITERATURE

(1) Tietze U., Schenk Ch.:
„Halbleiter-Schaltungstechnik"
Springer-Verlag, Berlin-Heidelberg,
5. Aufl. 1980

(2) Schüßler H. W.:
„Netzwerke, Signale und Systeme", Band 2
Springer-Verlag, Berlin-Heidelberg, 1. Aufl.

(3) Wunsch G.:
„Zur praktischen Berechnung von
Zwei-Phasen-Netzwerken"
Nachrichtentechnik Heft 4, 1958,
Seite 154 - 158

Dieter Schwarzenau and Bernhard Kokot

Home-Constructed Frequency Counter Part 2: Conclusion

2.2.3. Construction of the 50 Ω Plug-ins.

Normally, the printed circuit boards are fixed to the extreme left of the front panel in order to conserve as much space as possible for panel-mounted components. The 50 Ω plug-ins, however, have the plastic-angle stock mounted the "wrong way round", thereby displacing the board about 9 mm in towards the middle of the front panel. The panel-mounted BNC socket pin (**see fig. 23**) then contacts the printed circuit board and can be soldered directly on to the appropriate track.

The component loading is done in accordance with **figs. 24 and 25**. The locating spot on the integrated resistor array should not be overlooked. Chip capacitors are used for the coupling and blocking capacitors at RF and are soldered directly to the conductor tracks underneath. The components marked with an asterik on the plug-in possessing the pre-scaler, should also be mounted underneath, on the conductor tracks, in order to reduce the effects of parasitic capacitance and inductance to as low a value as possible.

The transformer comprises two twin turns of 0.25 mm lacquered copper wire (CuL) on a double-holed ferrite core. The core size is 2.5 x 3.6 x 2.1 mm and bears the Siemens designation "U 17".

The overload indicator can either be a normal or a flashing LED, according to preference. When a normal LED is used, the dropper resistor R 817

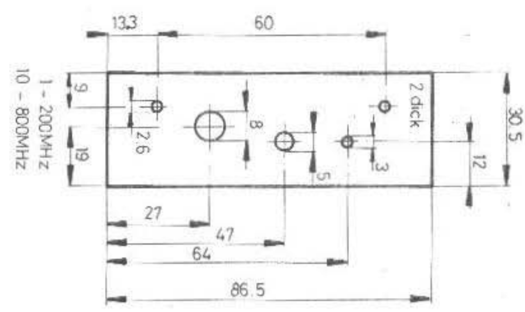
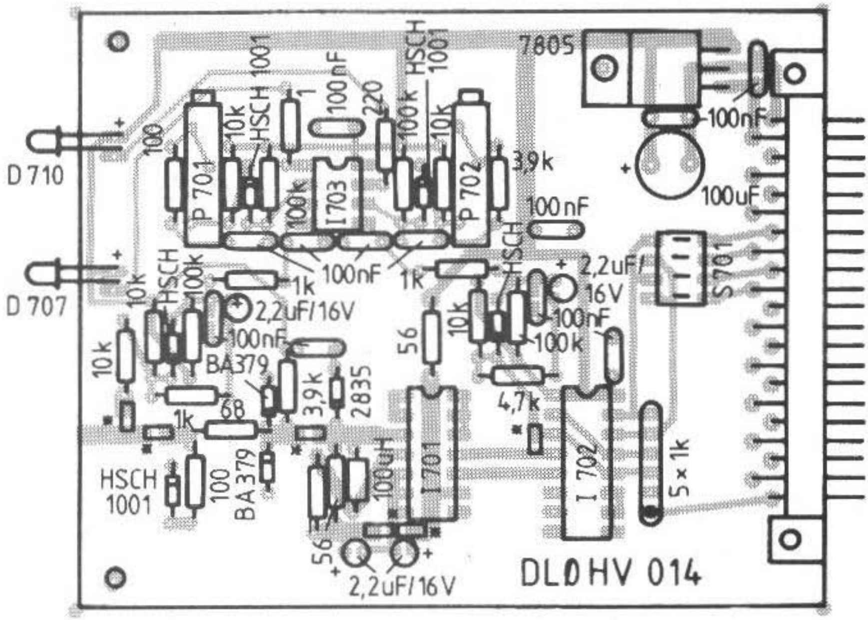


Fig. 23: 50 Ω Plug-in front panel dimensions



* 10 nF chip capacitor mounted on track side

Fig. 24: 50 Ω Pre-amplifier component location plan

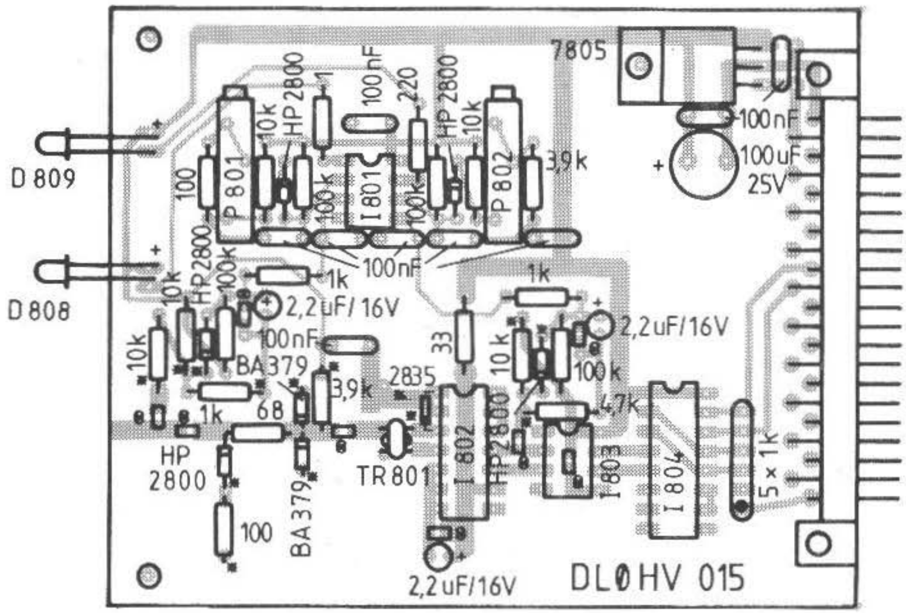
is raised to 220 Ω. The voltage regulator is screwed directly on to the PC board. The bonding of the front panel to the PC board is carried out by means of two solder tags, one fastened with two BNC socket mounting screws on the panel and the other soldered directly to the ground side of the PC board. The plastic-angle stock is filed as necessary, at the appropriate place, in order to complete the bonding.

2.2.4. Power Supply Construction

The power supply is accommodated in a separate box which forms the back wall of the enclosure. The PC board fits exactly into the box guide slots. The front of the box is drilled to allow the supply leads to be taken through. The back wall of the box is a detachable plastic plate with suitably shaped cut-outs for the mains switch, the IEC-norm mains plug, the 12 V connectors and the mains transformer which projects from the back of the cabinet (**fig. 26**). In the interest of greater mechanical stability, it may be better to replace this plate with one of aluminium.

The PC board is equipped in accordance with **fig. 27**. The connections to the integrated circuit voltage regulators are carefully bent in order that the regulator lies flat against the underside of the box when the board is in position. A 3.3 mm hole is drilled at exactly this spot in the box. An M 3 nut is glued to the regulator heat-sink in order that it can be screwed on to the housing wall after the PC board is in position.

At the indicated positions, six distance bolts, 27.5 mm long, are affixed. The four outer ones serve to secure the back wall and the other two hold the mains socket in place. Following the fitting of the back wall into position, the mains socket and mains switch are provided with con-



* mounted on track side, 10 nF chip capacitor on track side

Fig. 25 Pre-amplifier with divider component location plan

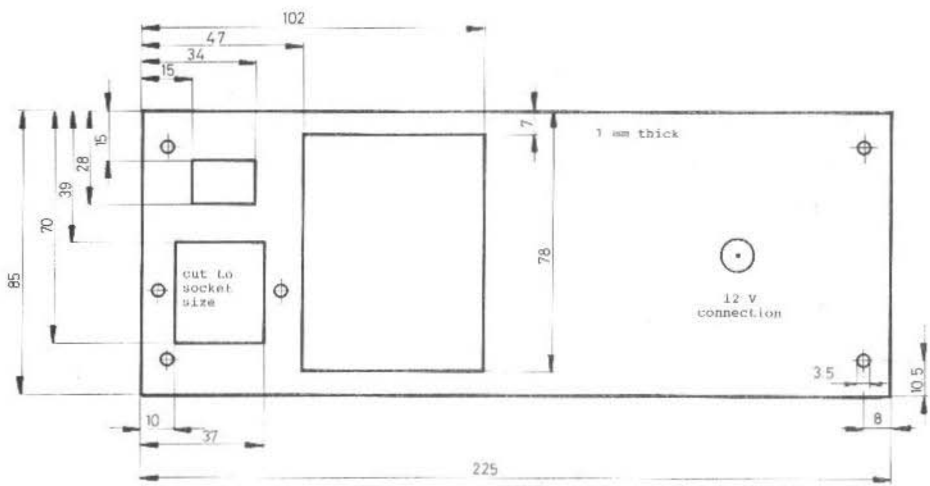


Fig. 26: Housing back wall dimensions

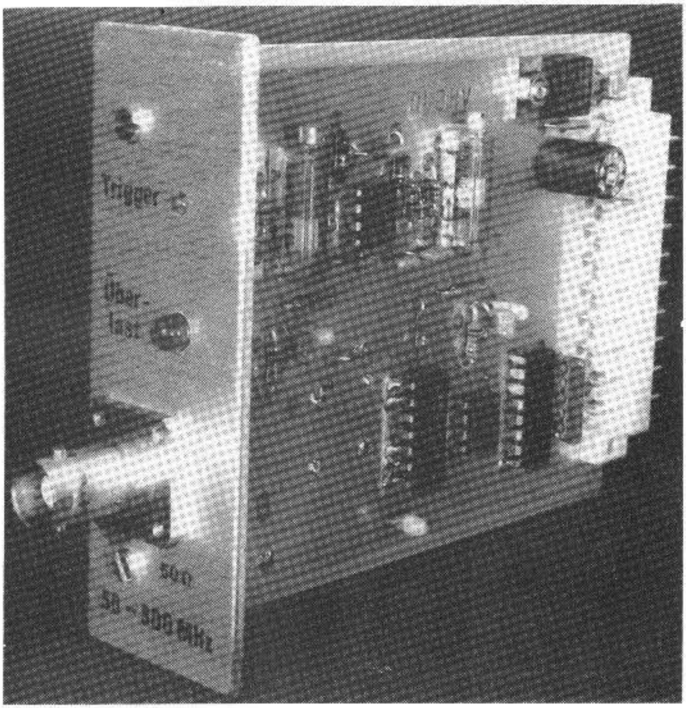


Photo E:
50 - 800 MHz Plug-in

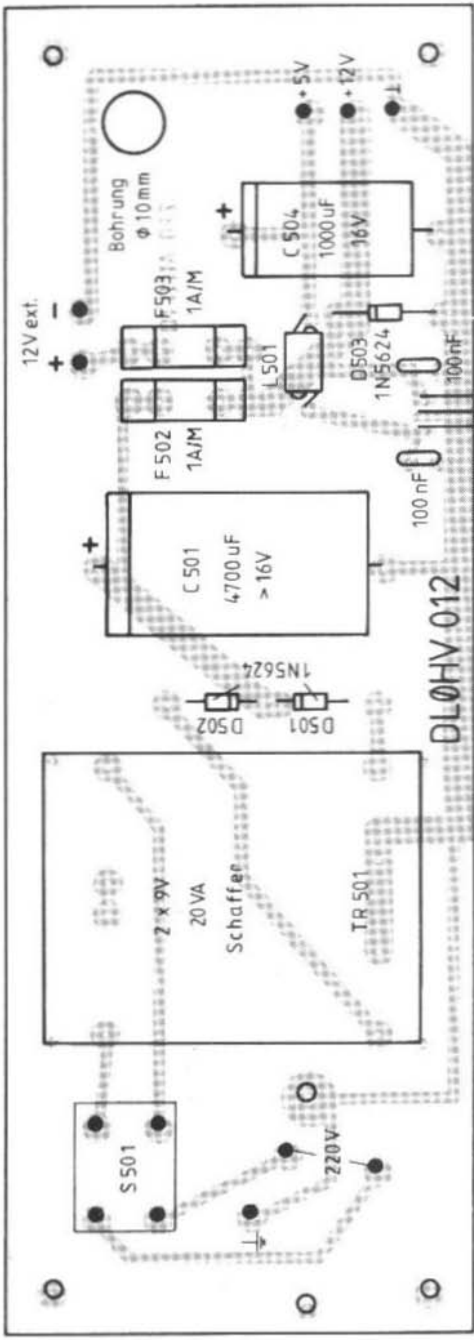
necting wires and both components mounted on to the panel. Eventually the ends of these connections are soldered to the PC board. The complete block is then slid into the guide channels of the box and the 3 mm set screw inserted to secure the voltage regulator to the box wall. It only remains now to lead the supply connections through the holes provided in the front wall.

2.2.5. Preparation of the Housing

As the sides of the cabinet are plastic, the aluminium top and bottom panels are not connected electrically to each other. This connection is provided by means of an aluminium plate 120 mm x 80 mm which is appropriately drilled and secured to the side of the housing where later the basic unit will be positioned, i.e. on the row of holes in the rear fixing rails. Lock washers should be used in order that the finished surface is pierced thus providing a good electrical contact.

The connection to the various plug-ins is enabled via 21 pole DIN 41 617 plug-in connectors which are also screwed to the rails and wired together. The socket is mounted from the front with two M 2.5 x 10 screws. The connections are made in accordance with **fig. 28**. As the ground pins lie on the outside of the pin field, solder tags, secured by the socket fixing screws, may be used to bond these connections directly to the housing. Lock washers should also be used here to ensure a solid ground connection.

The left-hand plug-in position uses two 21 pole sockets in juxtaposition, which, according to the PC board version, is either fastened to the ex-



Plug connections

1	ground
2	+12V unst.
3	+5V stab.
4	
5	
6	
7	input A1
8	input A2
9	input B1
10	input B2
11	
12	
13	
14	
15	
16	
17	reset
18	10 MHz
19	
20	
21	ground

Fig. 28:
Plug wiring plan

Fig. 27:
Power supply component location plan

treme left of the front panel or more towards the middle. Altogether then, five sockets are required.

The plug-ins should sit to the right when looking at the front of the cabinet. The PC board containing the time-base oscillator then lies in the middle and cannot be influenced by the surrounding environment.

The connections between sockets are made with either insulated connecting wire or with silvered wire over which an appropriate length of insulated sleeving has been slipped. The latter is the most elegant and tidiest solution. On one of the sockets, the power supply connections must be soldered. Finally, the plastic guide rails are mounted in the inside top and bottom. The left-hand plug-in aperature should be fitted with two parallel sets of guide rails.

3. COMMISSIONING AND FINAL ADJUSTMENT

Upon switching on the finished counter, the last four indicators should show "0" and the gate LED should start flashing at gate period rate. The last indicator, however, could show a "1". when using the high impedance plug-in, either a "0" or a "1" may be obtained in this last indicator simply by operating the trigger level control. Check that the gate LED blinks at a rate according to the gate time selected. A failure here could point to a time-base fault either in the digital control or in the oscillator/divider chain.

If all is well, the oscillator can be fine-tuned to its nominal frequency. This and other service operations are most conveniently carried out when the card is withdrawn from the unit. For this purpose, an "extensor" connector can be easily made from a 21 pole connector (plug and socket) and a length of multi-way cord.

The first thing to do is to adjust the crystal temperature. This is best measured using an electronic thermometer with a surface temperature probe. The probe tip is first smeared with heat conductive paste and then pressed firmly onto the crystal casing. The required temperature can then be adjusted by the multi-turn preset on the oscillator card. Owing to the thermal inertia of the system, it is advisable to wait a while after every adjustment before taking a temperature spot measurement. The nominal temperature to be attained is that given in the crystal manufacturer's specifications, or determined by means of the apparatus discussed in (1).

If another (stable) precision generator is available, the temperature adjustment may be carried out, using a little patience, in the following manner. The counter must be in a fully working condition. The reference frequency (from the calibrated source) is fed into one of the counter inputs and the longest gate time is measured. The temperature preset is then adjusted from minimum to maximum in small steps always allowing an appropriate pause between measurements. The correct temperature point is where the sign of the variation changes from positive to negative or vice versa.

To set the frequency of the oscillator, a known frequency source is necessary which is of as high a frequency as possible. This is fed to the input of the counter. The crystal trimmer capacitor is first of all soldered to the solder pins provided and then adjusted until the indicator display agrees with that of the reference. Of course, the whole set-up must be allowed to thermally stabilize first. When the oscillator has been accurately adjusted, a polystyrene cap should be glued over the oscillator card.

Should individual display segments show variations in brilliance for any reason, they can be compensated by changing the value of the appropriate segment dropper resistor. If the brilliance of the whole numerical display must be adjusted to that of its fellows then it is better done by altering the 4.7 Ω resistor associated with it. The segment brightness control resistors are 15 Ω.

It can occur, that segments which should be unlit, in fact, exhibit a prominent glow. This is caused by the integrated circuit and may be diminished by placing a 10 kΩ resistor in the circuit marked with an asterik in the diagram. This resistor may be soldered under the PC board.

As the last two digits are driven directly, it is only meaningful to correct the brilliance of individual segments. If the brightness of the two digits is radically different from each other then it may be possible to effect some alleviation of this condition by changing around the digit IC modules in the IC socket holders until the total display looks satisfactory.

Finally, both the wire bridge links, corresponding to the counter control connections, are set to the correct position. This is accomplished according to the operational plug-in unit and shown in **fig. 29**. Changing the plug-in unit should result in a matching configuration of wire bridge links such that only input B can be fitted with a divider plug-in, as only in this position the matching gate time and positioning of the display decimal point can be carried out. Input A position is used only for the fitting of the high impedance or the 50 Ω plug-in units.

```
Connector link wiring plan


   Links      IInput B division factor
-----------+-----------------------
           I
           I
 o    o    I
 I    I    I
 o    o    I             1
           I
 o    o    I
           I
           I
-----------+-----------------------
           I
           I
 o    o    I
           I
 o    o    I             4
 I    I    I
 o    o    I
           I
           I
-----------+-----------------------
           I
           I
 o    o    I
 I         I
 o    o    I             10
      I    I
 o    o    I
```

Fig. 29: Connector link wiring plan

The adjustment to the plug-ins is limited to the setting of the comparator switch thresholds for the two 50 Ω units. For this purpose, a signal generator which can deliver a variable amplitude signal at between 100 and 150 MHz is required. A two-metre transceiver fitted with a suitable output attenuator would also do the job. First of all, the LED trigger threshold is adjusted using an input signal amplitude of 10 mV.

The counter should be allowed to operate until a stable count is indicated. The rear trimmer is then adjusted such that the upper LED is just illuminated. The switching threshold for the overload indicator is adjusted by the front preset resistor with the input amplitude set to 5 V. This adjustment uses the same procedure for the two 50 Ω plug-in units. The extensor cord would come in very handy for this procedure.

4. USING THE FREQUENCY COUNTER

The counter is very easy to operate but nevertheless a few words of advice about its use would not come amiss. The first consideration when using the counter is the choice of the plug-in unit. As, in general, the approximate frequency of the signal-under-test is known and also the approximate source impedance, the choice of unit presents no great difficulty. The choice of plug-in is made by the right-hand switch on the main-frame front panel. If, when using the 50 Ω plug-ins, the trigger LED illuminates, it indicates that the input signal has a sufficient working level. If the overload LED illuminates, or flashes, then the input amplitude to the counter must be attenuated, if necessary by a fixed attenuator. When using the high-impedance plug-ins, the gain should be advanced slowly from a low level. At each level the trigger is adjusted until the counting unit begins to operate – indicated by the flashing front-panel LED. If the trigger is not adbanced too far it remains on continuously. Do not attempt to connect a high-level signal when the sensitivity is set too high. The pre-amplifier can certainly handle the signal but because the transistors are driven to saturation a false count may be obtained. A BNC oscilloscope monitor point has been provided on the main-frame front panel in order that the nature of the test signal may be inspected before it arrives at the counter gate.

The desired gate period is now selected starting with the shortest, as then it will be quickly determined whether the display "freezes". Only when it does, so is it meaningful to try a longer gate time. If the display does not "freeze", it could be that the test signal is unstable, too weak or that it is subject to interference.

It is not necessary to wait for the full counting cycle when using the longer gate periods. Another count cycle can be manually commenced by pressing the reset button. The number of input pulses counted from the time the reset button will then be displayed.

If the accuracy of the internal crystal oscillator is insufficient for the measurement being undertaken, an external 10 MHz derived from a frequency standard may be fed into the right-hand BNC socket on the main-frame front panel (e.g. (4)). The amplitude of this signal must be greater than 100 mV. This input is DC-blocked and overload-protected. The switch-over from standard to internal time-base oscillators is achieved automatically.

Considering the limitations in the use of frequency counters, the commonly held view is, that when measuring a number of signals of different frequencies, the signal which has the greatest amplitude will be the one which the counter displays. If the signal has the form depected in **fig. 30**, i.e. a fundamental signal with a third harmonic, then indeed, the counter will display the larger fundamental providing the trigger threshold is set to point "1". Setting the trigger to point "2", however, will result in the counter reading twice the fundamental despite the fact that no signal of this frequency is present in the signal to be measured. Even in the case where the trigger level is set very close to the zero line, the strongest component will not be the one which the counter displays. This may be seen by considering **fig. 31** where the larger amplitude fundamental signal is mixed with a smaller amplitude 9th harmonic. If the counter is inherently capable of operation at this frequency then it will display a non-existent signal at five times the fundamental.

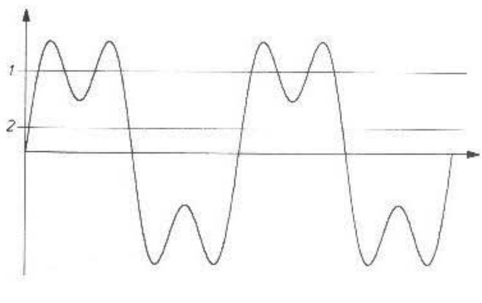


Fig. 30: Influence of the trigger threshold setting on the counter display in the presence of a third harmonic in the input signal

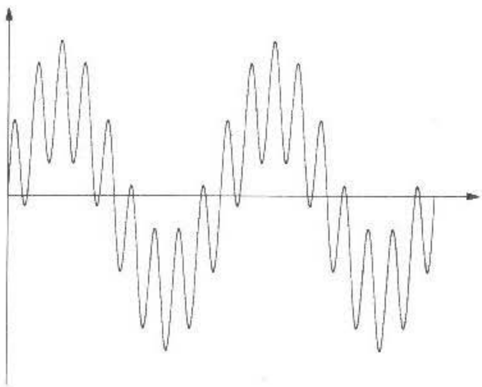


Fig. 31: Signal with ninth harmonic

5. FURTHER DEVELOPMENTS

At the time of writing, a 50 Ω plug-in is being evaluated for the frequency range 150 MHz to 1.5 GHz. The sensitivity is only 2 mV in the 23 cm band falling to 30 mV thereafter. The input is overload-protected in the same manner as the other plug-ins. This, however, is only operative for the case when the supply potentials are present. A solution to this problem is now being investigated. If there is sufficient interest, the details of this plug-in unit will be published at a later date. Unfortunately, the cost of the components is rather high and maybe beyond the reach of some people.

In the distant future, it is planned to build a 4 GHz plug-in. This may entail certain changes to the main-frame circuitry. The publication of this project will be conditional upon the component price being reasonable at that time.

6. REFERENCES

(1) Arnoldt, M.:
A Measuring System for Determining the Temperature Response of Crystals
VHF COMMUNICATIONS Vol. 12, Ed. 3/1980, Pages 159 - 168

(2) ELO-Universalzähler
ELO 8/1980, Page 48
ELO 9/1980, Page 62
ELO 10/1980, Page 70
ELO 4/1981, Page 59
ELO 5/1981, Page 56

(3) ELV-Serie 7000
1 GHz FRequenzzähler FZ 7000
ELV Nr. 17, Page 48

(4) Görl, R.:
A Standard Frequency Oscillator with an Accuracy of 10^{-8}
VHF COMMUNICATIONS Vol. 7, Ed. 2/1975, Pages 118 - 126

(5) Motorola:
MECL-Data Book 1982/83

(6) RCA:
Linear Integrated Circuits, Page 639
(Data sheet CA 3199)

(7) VALVO:
Integrierte Schaltungen in digitalen Rundfunk- und Fernsehempfängern 1980, Page 171
(Datenblatt SAB 1109)

7. PARTS LISTS

DL Ø HV 008: Counter Control PCB

R 101:	1 kΩ
102:	1 kΩ
103:	10 kΩ
104:	4 k 7 Ω
105:	4 k 7 Ω
106:	39 kΩ

1 Epoxy-glass board 100 x 80 x 1.5
1 Support bracket BICC-VERO 31774
2 M 2.5 x 6
2 M 2.5 x 8
4 M 2.5 nuts

C 101:	2.2 μF/16 V tantal RM 2.5
102:	100 n plate ceramic RM 5
103:	100 n plate ceramic RM 5
104:	100 n plate ceramic RM 5
105:	2.2 μF 16 tantal RM 2.5
T 101...103:	BC 237 C, BC 413 C or similar NPN-AF Trans.
I 101:	SN 74121 N
102:	SN 74 LS 42 N
103:	SN 74 LS 00 N
104:	SN 74 LS 490 N
105:	SN 74 LS 490 N
106:	SN 74 LS 151 N
107:	SN 74 LS 490 N
108:	SN 74 LS 490 N
109:	SN 74 LS 490 N

St 101:	multi-way connector DIN 41617 edge connector male SL 21 S
102:	edge connector female BL 1; 7 pole
103:	edge connector female BL 1; 6 pole
104:	edge connector male SL 2; 6 pole

DL Ø HV 009: Counter PCB

R 201:	330 Ω
202:	330 Ω
203:	470 Ω *
204:	1 K
205:	470 Ω *
206:	1 K
207:	470 Ω *
208:	470 Ω *
209:	470 Ω *
210:	1 K
211:	1 K
212:	100 Ω
213:	1 K
214:	470 Ω
215:	470 Ω * = 7 x 470 Ω integrated
216:	1 K L 08 - 1
217:	1 K
218:	1 K ° = 7 x 1 K integrated
219:	100 Ω L 08 - 1
220:	1 K °
221:	10 K
222:	1 K °
223:	10 K
224:	1 K °
225:	10 K
226:	1 K °
227:	10 K
228:	1 K + 5 x 1 K integrated
229:	1 K + L 06 – 1
230:	1 K +
231:	1 K +
232:	100 K
233:	100 K
234 - 247:	430 Ω
248 - 254:	4.7 Ω
255 - 261:	15 Ω
262 - 275:	470 Ω
I 201:	MC 10115 P
202:	MC 10104 P
203:	MC 10138 P
204:	SN 75140 P
205:	LM 339 N
206:	SN 74 LS 75 N
207:	SN 74 LS 247 N
208:	SN 74 LS 02 N
209:	SN 74 LS 290 N
210:	SN 74 LS 75 N
211:	SN 74 LS 247 N
212:	ICM 7208 IPI
C 201:	4.7 n FKC RM 5
202:	47 µF 16 V ELKO RM 5
203:	100 n plate ceramic RM 5
T 210...207:	BC 213 C, BC 415 C or equiv. PNP
208...214:	BC 237 C, BC 413 C or equiv. NPN
St 2:	edge connector male SL 1 6 contacts
202:	edge connector male SL 1 6 contacts

DK Ø HV 010: Display PCB

R 301:	150 Ω
302:	150 Ω
303:	330 Ω
304:	10 K
305:	10 K
306:	10 K
307:	15 Ω
308:	15 K
309:	15 Ω
310:	1 K
311:	1 K
312:	1 K
I 301:	SN 74 LS 138
302:	SN 74 LS 10
D 301:	LD 356 / 6
S 301:	rotary switch ITT SB 20 AD 4 x 3 way
S 302:	rotary switch ITT SB 20 AD 6 x 2 way

St301:	edge connector male SL 1 3 contacts
St302:	edge connector male SL 6 contacts

Misc.:
1 epoxy-glass PCB: 132 x 65 x 1.5 mm
1 ALU sheet 152.5 x 86.5 x 2 mm
1 Acrylic filter stock red
2 Wing knobs 13 mm Ø
2 caps for above
2 BNC UG 1094/U
1 Push-button (Rafi)

T 301...303:	BC 237 C, BC 413 C or equiv. NPN
DS 301...309:	HD 1105 R (Siemens)

DL Ø HV 011: Crystal Oscillator

R 401:	3.9 K 1 %
402:	3.9 K 1 %
403:	1.3 K 1 %
404:	10 K
405:	10 K
406:	1 M
407:	330 Ω
408:	10 Ω 1 %
409:	47 Ω
410:	1 K
411:	1 K
412:	1 K
413:	1 K
414:	100 K
415:	1 K
416:	1 K
417:	56 K
418:	10 K
419:	SAK 1000
P 401:	470 Ω multi-turn preset
D 401:	LD 356 / 6
402:	1 N 4148
403:	1 N 4148
I 401:	TLC 271 CP
402:	78 L 05 A
403:	MC 10116 P
404:	78 L 05 A
405:	SN 75 140 P
406:	SN 74 LS 122 N
407:	SN 74 LS 132 N
T 401:	BD 139-6
St401:	multiway connector DIN 41617 edge connector male SL 21 S
402:	edge connector female BL 1, 7 pole
403:	edge connector female BL 1, 3 pole
Q 401:	XS 6005 A (KVG) 10 MHz 30 pF 60° C

C 401:	– – – – – – – – – – –	
402:	2.2 µF 16 V tantal	RM 2.5
403:	100 n plate ceramic	RM 5
404:	100 n plate ceramic	RM 5
405:	4.7 p NPO ceramic	RM 2.5
406:	22 p NPO plate ceramic	RM 5 / 2.5
407:	5.6 p NPO ceramic	RM 2.5
408:	nc	
409:	100 n plate ceramic	RM 5
410:	100 n plate ceramic	RM 5
411:	2.2 µF 16 V tantal	RM 2.5
412:	100 n plate ceramic	RM 5
413:	100 n plate ceramic	RM 5
414:	100 n plate ceramic	RM 5
415:	10 µF 16 V tantal	RM 2.5
416:	1 µF 16 V tantal	RM 2.5
417:	2.5 p NPO trimmer	

Misc.:
1 epoxy-glass PCB; 100 x 80 x 1.5 mm
1 aluminium sheet; 25 x 11 x 2 mm
1 bracket BICC-VERO 31774
1 M 3 x 12
2 M 3 nuts
2 M 2.5 x 6
2 M 2.5 x 8
4 M 2.5 nuts

DL Ø HV 012: Supply Unit

C 501:	470 µF/40 V ELKO axial
502:	100 n plate ceramic

503:	100 n plate ceramic
504:	1000 μF/16 V ELKO axial
D 501:	1 N 5624
502:	1 N 5624
503:	1 N 5624
S 501:	toggle switch, 2 pol.
L 501:	wideband choke core (6 hole)
I 501:	MC 7805 CT
Tr501:	2 x 9 V/20 VA (Schaffer)
St501:	IEC mains socket (fused)
502:	external voltage supply socket Fa. Roka 520 0660
F 501:	125 mA
502:	1 A
503:	1 A

Misc.:

1 aluminium sheet 225 x 85 x 1 mm
6 spacing bolts M 3 x 28.5
11 M 3 x 8
2 M 3 c/s x 10
1 M 3 nut
4 self-tapping screws DIN 7971 2.9 x 6.5 mm
1 epoxy-glass PCB 225 x 80 x 2 mm

•

DL Ø HV 013 (a): High-Impedance Plug-in 10 Hz - 30 MHz

R 601:	910 k	1 %
602:	91 k	1 %
603:	10 k	1 %
604:	1 k	
605:	10 M	
606:	390 Ω	1 %
606:	180 Ω	1 %
608:	1 k	
609:	100 Ω	
610:	1.5 k	1 %
611:	1.5 k	1 %
612:	3.3 k	1 %
613:	3.3 k	1 %
614:	1.5 k	1 %
615:	1.5 k	1 %
616:	1 k	
617:	1 k	
618:	1 k	
619:	1 K	
620:	470 Ω	1 %
621:	470 Ω	1 %
622:	180 Ω	1 %
623:	330 Ω	1 %
624:	180 Ω	1 %
625:	1 k	1 %
626:	1 k	1 %
627:	1 k	1 %
628:	1 k	
629:	1 k	
630:	1 k	
631:	1 k	
632:	1 k	
633:	1 k	
634:	470 Ω	
635:	220 Ω	
D 601:	1 N 4148	
602:	1 N 4148	
603:	1 N 4148	
604:	LD 356/6	
T 601:	BF 256	
602:	2 N 5771	
603:	2 N 5771	
I 601:	MC 10116 P	
602:	MC 10116 P	
603:	MC 7805 CT	

C 601:	3.3 p	disc ceramic	400 V	RM 5
602:	33 p	disc ceramic	400 V	RM 5
603:	330 p	disc ceramic	400 V	RM 5
604:	4.7 n	disc ceramic	400 V	RM 5
605:				
606:	5.6 p	ceramic		RM 2.5
607:	47 μF	16 V Elko		RM 5
608:	100 n	plate ceramic		RM 5
609:	100 p	ceramic		RM 2.5
610:	100 n	ceramic		RM 5
611:	4.7 μF	16 V Elko		RM 2.5
612:	100 n	plate ceramic		RM 5
613:	22 p	ceramic		RM 2.5
614:	22 p	ceramic		RM 2.5
615:	100 n	plate ceramic		RM 5
616:	47 μ	6.3 V tantal		RM 2.5
617:	100 n	plate ceramic		RM 5
618:	100 n	plate ceramic		RM 5
619:	100 n	plate ceramic		RM 5

620:	100	n	plate ceramic	RM 5
621:	100	n	plate ceramic	RM 5
622:	100	μ	16 V Elko	RM 5
623:	100	n	plate ceramic	RM 5
624:	100	n	plate ceramic	RM 5

P 601: film track preset
Draloric 70 HDSC 47 k

St601: BNC UG – 1094/U
602: connector plug DIN 41617
edge connector male
SL 21 S

S 601: rotary switch ITT 2 x 6 way
Type SB 20 AD

misc:

1 epoxy-glass board 100 x 80 x 1.5 mm
1 aluminium sheet 86.5 x 30.5 x 2 mm
1 soldering tag for BNC socket
1 round 10 mm knob
1 10 mm cover
1 10 mm indicator disc
1 13 mm indicator knob
1 cover
2 M 2.5 x 6
2 M 2.5 x 8
4 M 2.5 nuts
1 support bracket BICC-VERO 31774
1 M 3 x 6
1 M 3 nuts
1 U-disc M 3

DL Ø HV 014: Plug-in 1 - 200 MHz

R 701:	100	Ω
702:		
703:	10	k
704:	100	k
705:	100	k
706:	10	k
707:	1	k
708:	10	k
709:	1	k
710:	1	Ω
711:	3.9	k
712:		
713:	10	k
714:	100	k
715:	100	k
716:	10	k
717:	4.7	k
718:	1	k
719:	220	Ω
720:	100	Ω 1 %
721:	68	Ω 1 %
722:	3.9	k 1 %
723:	56	Ω
724:	56	Ω
725:	56	Ω
726:	1	k
727:	1	k

5 x 1 k integrated
728: 1 k type I 06-1
729: 1 k

P 701: 1 k multi-turn preset
702: 1 k multi-turn preset

D 701:	HSCH 1001
702:	HSCH 1001
703:	HSCH 1001
704:	BA 379
705:	BA 379
706:	5082 - 2835
707:	CQX 21 red
708:	CQX 21 1001
709:	CQX 21 1001
710:	LD 356/6
I 701:	SAB 1009 BP
702:	MC 10116 P
703:	LM 393 N
704:	MC 7805 CT

DL Ø HV 014: Plug-in 1 - 200 MHz

C 701:	100	n	plate ceramic	RM 5
702:	100	n	plate ceramic	RM 5
703:	2.2	μF	16 V tantal	RM 2.5
704:	100	n	plate ceramic	RM 5
705:	10	n	plate ceramic	chip
706:	100	n	plate ceramic	RM 5
707:	100	n	plate ceramic	RM 5
708:	2.2	μF	16 V tantal	RM 2.5
709:	10	n	plate ceramic	chip
710:	100	n	plate ceramic	RM 5
711:	100	n	plate ceramic	RM 5
712:	10	n	plate ceramic	chip
713:	10	n	plate ceramic	chip

714:	10	n	plate ceramic	chip
715:	2.2	μF	16 V tantal	RM 2.5
716:	2.2	μF	16 V tantal	RM 2.5
717:	10	n	plate ceramic	chip
718:	100	n	plate ceramic	RM 5
719:	100	n	plate ceramic	RM 5
720:	100	μF	16 V Elko	RM 5
721:	100	n	plate ceramic	RM 5
722:	100	n	plate ceramic	RM 5

L 701: choke 100 μH

S 701: DIP switch 4 pole

1 epoxy-glass board 100 x 80 x 1.5 mm
1 aluminium sheet 86.5 x 30.5 x 2 mm
2 solder pins M 3
1 support bracket BICC-VERO 31774
1 M 3 x 6
1 M 3 nut
1 U-washer M 3
4 M 2.5 x 4
2 M 2.5 x 6
3 M 2.5 x 8
4 M 2.5 nuts

St 701: BNC UG 290 A/V
702: connector DIN 41617 edge connector male SL 21 S

DL Ø HV 015: Plug-in 50 - 800 MHz

R 801:	100 Ω	
802:	10 k	
803:	100 k	
804:	100 k	
805:	10 k	
806:	1 k	
807:	1 k	
808:	10 k	
809:	68 Ω	1 %
810:	100 Ω	1 %
811:	3.9 k	1 %
812:	4.7 k	
813:	10 k	
814:	100 k	
815:	100 k	
816:	10 k	
817:	1 Ω	
818:	3.9 k	
819:	33 Ω	
820:	1 k	
821:	220 Ω	
822:	1 k	integrated
823:	1 k	5 x 1 k
824:	1 k	Typ L 06-1
825:	1 k	

P 801: 1 k multi-turn preset
802: 1 k multi-turn preset

D 801:	5082-2800
802:	2800
803:	2800
804:	2800
805:	BA 379
806:	BA 379
807:	5082-2835
808:	CQX 21 red
809:	LD 356/6
810:	5082-2800

I 801:	LM 393 P
802:	SAB 1009 BP
803:	CA 3199 E
804:	MC 10116 P
805:	MC 7805 CT

C 801:	100	n	plate ceramic	RM 5
802:	100	n	plate ceramic	RM 5
803:	2.2	μF	16 V tantal	RM 2.5
804:	100	n	plate ceramic	RM 5
805:	10	n	plate ceramic	chip
806:	10	n	plate ceramic	chip
807:	100	n	plate ceramic	RM 5
808:	10	n	plate ceramic	chip
809:	2.2	μF	16 ØV tantal	RM 2.5
810:	10	n	plate ceramic	chip
811:	100	n	plate ceramic	RM 5
812:	100	n	plate ceramic	RM 5
813:	2.2	ΩF	16 V tantal	RM 2.5
814:	100	n	plate ceramic	RM 5
815:	10	n	plate ceramic	chip
816:	100	n	plate ceramic	RM 5
817:	100	n	plate ceramic	RM 5
818:	100	ΩF	16 V Elko	RM 5
819:	100	n	plate ceramic	RM 5
820:	100	n	plate ceramic	RM 5
821:	10	n	plate ceramic	chip

Tr 801: twin-hole core K 1 or U 17

St 801: BNC UG 290 A/U
802: multi-way connector DIN 41617
edge connector male SL 21 S

Misc.:

1 epoxy-glass 100 x 80 x 1.5 mm
1 aluminium sheet 86.5 x 30.5 x 2 mm
2 solder pins M 3
1 support bracket BICC-VERO 31774
1 M 3 x 6
1 M 3 nut
1 V washer M 3
4 M 2.5 x 4
2 M 2.5 x 6
2 M 2.5 x 8
4 M 2.5 nuts

Cabinet

1 cabinet BICC-VERO KMT sub rack system, order no. 127-31427
10 guide rails, order no. 127-32166 A
1 ALU plate 125 x 70 x 1.5 (mm)
18 M 2.5 x 10
18 M 2.5 nuts
18 lockwashers DIN 6798 A
M 3, Ad 6.0 St 0.4 mm
10 solder pins M 3
5 multi-way connectors DIN 41617
spring contact FL 21 S

Note: RM = lead spacing (mm)

Peter Gerber, HB 9 BNI

The Doppler Effect over Radio Links using Active or Passive Reflectors

This article describes the physical fundamentals necessary for an understanding of the Doppler Effect together with the necessary formulae. It also presents a method of calculating the influence of the Doppler Effect on contacts made via paths where both active and passive reflectors are involved. This calls for a sub-program which permits the calculation of the reflector co-ordinates to any desired point in time.

1. PRINCIPLES

In the case of a communication path which is continually varying in distance between transmitter and receiver, the received frequency is not the same frequency as that which was transmitted. This frequency alteration is called the Doppler Effect after its discovery by the physicist Doppler. A similar effect is well known at audio frequencies when, for example, a railway locomotive with a continuous horn is passing an observer. The pitch of the horn always sounds higher to the observer when the locomotive is approaching than when it is departing. For this acoustical phenomenon the physical theory is extremely complicated as it must always be decided what is moving relative to whom taking into consideration the sender, the receiver and the communication medium.

As electromagnetic waves exhibit no measurable relative movement to the medium through which it passes (experiments by Michelson and Einstein's theory of relativity) at least one variable may be eliminated, and that leaves only the relative movements of transmitter and receiver. When considering the case where relative speeds approaching light is concerned, the theory is also very complicated such as the velocity components in the line-of-sight as also the transverse components give rise to a Doppler Effect. With small relative velocities, and that includes all amateur contact paths, only the line-of-sight component V_s plays a part in the Doppler

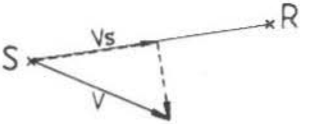


Fig. 1

S = Sender
R = Receiver

Fig. 2

Effect (**fig. 1 and fig. 2**). In practice, the case of transmitter and receiver moving away from each other is so small, and occurs so seldom for the amateur anyway (see example below). The Doppler Effect, however, plays a very prominent role in paths, using either active or passive reflectors, made in space via satellites and EME (4).

2. PHYSICAL FUNDAMENTALS

The transmitter, of frequency f, moving with a velocity v directly to a receiver (**fig. 2**) will be measured at the receiver as: –

$$f_r = f\,(1 - v / c) \qquad (1)$$

where v is considered a negative quantity when the sender S and the receiver R approach each other. The received frequency f_r is then higher than the transmitted frequency f.

Example:

The sender, with a frequency of 145 MHz, moves towards the receiver at a speed of 120 km/h. What will be the received frequency?

$v = -120$ km/h $= -33.33$ m/sec.
$f = 145$ MHz $= 145 \times 10^6$ 1/sec.
$c = 300 \times 10^6$ m/sec.

$$f_r = 145'000'000 \times (1 + \frac{33.33}{300'000'000})$$

$$= 145 \times 10^6 \times 1.000000111 = 145'000'016 \text{ Hz}$$

The received frequency f_r, is therefore 16 Hz higher than the send frequency which, normally, has no practical consequence. Transposing the eq. 1 the following quantities may be obtained:

Change in frequency $\quad df = f\,(-v / c)$
Change in wavelength $\quad dl = l\,(v / c)$

The point to watch is that all quantities are presented to the equation in the same units i.e. metres, metres per second, hertz. Also, that the velocity sign is negative when the transmitter approaches the receiver. The main problem, when calculating the Doppler Effect on signals via reflectors (active or passive), is that of determining the velocity of sender and receiver in the direct path. The sender, receiver and the reflector are mostly moving in complicated relative orbits in space. The sender on earth moves with the earth, the satellite in a canted ellipse, with variable speed and direction and the receiver, again, with the earth. Active satellites have the additional consideration of the alteration in frequency due to frequency changing processes within the satellites equipment. These "conversion functions" must also be taken into account when calculating the received frequency.

3. CALCULATIONS

The calculated received frequency of a signal, transmitted via active or passive reflectors, is arrived at by using the following steps:

- Transmitted frequency f_{u1} (uplink 1)
- Uplink Doppler Effect
- Received frequency f_{u2} (uplink 2)
- Transit function
- Transmitted frequency f_{d1} (downlink 1)
- Downlink Doppler Effect
- Received frequency f_{d2} (downlink 2)

If the transmitter and the receiver are both located at the same place then the same relative velocity is used for the Doppler Effect calculation. If a passive reflector is involved, the transit function is simply $f_{u2} = f_{d1}$.

It has already been mentioned that the chief problem lies in the calculation of the relative velocities of the participating stations in the direct point to point path. The radio amateur OK 1 DAT gives, for example, in (1) the following formula for the Doppler Effect over an EME link (but without considering the influence of the elliptical nature of the moon's orbit):

$$df = -\frac{2\,\pi\,6370 \cos BE \sin AH \cos DE}{24 \times 3.6\,\lambda\,(1 + 0.034 \cos AH)}$$

When calculating the effect over satellites, the formula is even more complicated because the elliptical orbit must be taken into account.

Using a small computer for calculating the orbit of the transponder, a simple procedure may be employed.

The principle lies in the fact that the distance from the sender/transponder and the receiver/transponder can be arrived at relatively easily. Should the distance to two different, but not too far apart in the time frame, points be calculated, the average line-of-sight velocity in this time frame may be taken, namely the velocity component in the most direct path.

The distance calculation is carried out simply by converting the main polar co-ordinates to the Cartesian form and then using three-dimensional Pythagoras. The necessary formulae for this are given as follows:

Conversion from polar to Cartesian co-ordinates given that a longtitude information is L
a latitude information is B
a radius information is R

The Cartesian co-ordinates are X, Y and Z:

$$X = R \cos B \cos L$$
$$Y = R \cos B \sin L$$
$$Z = R \sin B$$

When these co-ordinates have been calculated for both transmitter and receiver and the Cartesian co-ordinates subtracted from each other viz.

$$DX = X1 - X2$$
$$DY = Y1 - Y2$$
$$DZ = Z1 - Z2$$

then the instantaneous distance is as follows: –

$$\text{Distance } E = \sqrt{(DX \times DX + DY \times DY + DZ \times DZ)}$$

A point to watch is that the transmitter and the receiver (= transponder) should use polar co-ordinates which are compatible with each other, as for example, those given in **table 1**.

A general procedure for the calculating process

EARTH			TRANSPONDER		
Longtitude	**Latitude**	**Radius**	**Longtitude**	**Latitude**	**Radius**
longtitude	latitude	earth radius	longtitude subsatellite point	latitude subsatellite point	Distance from earth's centre
local startime	latitude	earth radius	right ascension	declension	geocentric distance
longtitude	latitude	earth radius	hour angle (GHA)	declension	geocentric distance
0	latitude	earth radius	local hour angle	declension	geocentric distance

Table 1: Mutually matching polar co-ordinates

may be given as follows:

- Calculate the polar co-ordinates of the QTH and the transponder.
- Calculate distance E 1
- Convert QTH into Cartesian co-ordinates
- Convert transponder into Cartesian co-ordinates
- Subtract the Cartesian co-ordinates
- Use Pythagoras to find the distance
- Calculate the polar co-ordinates of the QTH and the transponder for a later point in time (e.g. after 60 s)
- Calculate the difference in the distances and divide by the time difference to obtain the velocity
- Calculate the Doppler Effect of the link

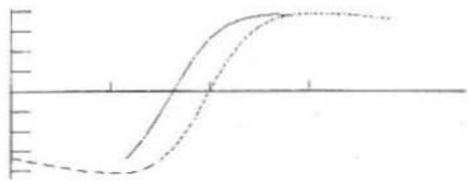
When a passive transponder is used, the Doppler shift is multiplied by a factor of 2. Active transponders require separate calculations for uplink, transit through the transponder, and downlink.

4. APPLICATION EXAMPLES

The Doppler Effect for an EME contact via the AMSAT-OSCAR 10 was calculated using the above mentioned formulae as a base.

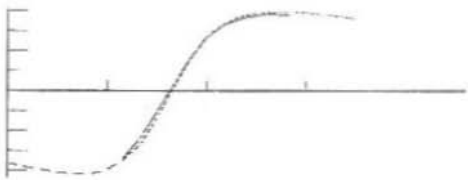
The calculation for the EME path was made using both the given formulae and that given by OK 1 DAT. The results on the whole, were compatible, the maximum difference amounted to 10 Hz for a transmitter frequency of 144 MHz (Doppler shift + 350 Hz to − 350 Hz). This discrepancy arose from the influence of the elliptical nature of the moon's orbit which was approximated in the OK 1 DAT calculation.

There was no alternative method available of calculating the effect when assessing the OSCAR 10 link. The Doppler Effect on this link, on the other hand, amounted to over 3000 Hz. This shift can be easily measured with amateur equipment. Suitable measurements were carried out and evaluated (**fig. 3 and fig. 4**). The calculations are so exact that, even in modest circumstances, the orbit data of OSCAR 10 may be evaluated.



DATA: TIME SCALE: 9.30 UT
ARG.PERIG. 40 RAAN 121
TIME 0 MIN FREQ.SHIFT 1400 HZ
EPOCHE G: 2776.00899

Fig. 3: Comparison between measured (solid line) and calculated (dotted line) Doppler shift using OSCAR 10 before adjustment. The satellite, according to the calculation, is a little premature.



DATA: TIME SCALE: 9.30 UT
ARG.PERIG. 40 RAAN 121
TIME -10 MIN FREQ.SHIFT 1400 HZ
EPOCHE G: 2776.00205

Fig. 4: The same data following the matching of the satellite's data (time shift about 10 min.)

5. REFERENCES

(1) OK 1 DAT, ohne Titel, handelt vom Doppler-Effekt bei EME, DUBUS Buch # 2, p 28 ff

(2) Autorenteam: Meyers Handbuch über das Weltall, Mannheim, 1974

(3) O. Montenbruck: Grundlagen der Ephemeridenrechnung, Verlag Sterne und Weltraum, München, 1984

(4) Gerber, P., HB 9 BNI: EME für jedermann? "old man" 9/1986, Seite 23 - 26

Dragoslav Dobričić, YU 1 AW

A 250 W 23 cm-Band Power Amplifier

A power amplifier is introduced which has been designed and built for communications using the Earth-Moon-Earth (EME) path. Its main characteristics are reliability and stability together with a dependable achievement of its design specifications. No special tools will be necessary for its construction.

1. CONCEPT

When the author first began to consider a power increase, there were no solutions available for the 23 cm band which did not involve complicated work with machine tools on the anode resonators. Also, all published constructions in the projected power range utilized more than one PA tube thus complicating both the construction and the tuning procedure. Following the evaluation of all the available literature and the formulation of a suitable mathematical model, a program was written for a personal computer. This program enabled an exact calculation of the resonator dimensions for all the frequencies of interest upon presentation of the particular PA valve's parameters.

Perhaps the commonest 23 cm power amplifiers use the 2 C 39, or its equivalents and derivatives, either singly or up to six valves in a ring formation feeding a common resonator. On account of the constructional advantages involved, and also its compact dimensions, the author felt that the Siemens valve type YD 1270 would be more suitable. This tube is used, above all, in UHF television transmitters. Compared with the 2 C 39 it has twice the anode dissipation and output power. Also, its output capacity is some 50 % greater and its maximum usable frequency is about 3 GHz – its use here at 1.3 GHz lies well within its capabilities.

The author selected the well-tried and tested construction: a grounded grid amplifier with an anode resonator in the TM_{010} mode and a half-wave coaxial cathode resonator (**fig. 1**).

2. CONSTRUCTION

2.1. The Cathode Circuit

In order to make the construction as easy as possible, the cathode circuit is fashioned as a loaded $\lambda/2$ coaxial line which mechanically supports the tube, thus obviating the requirement for the

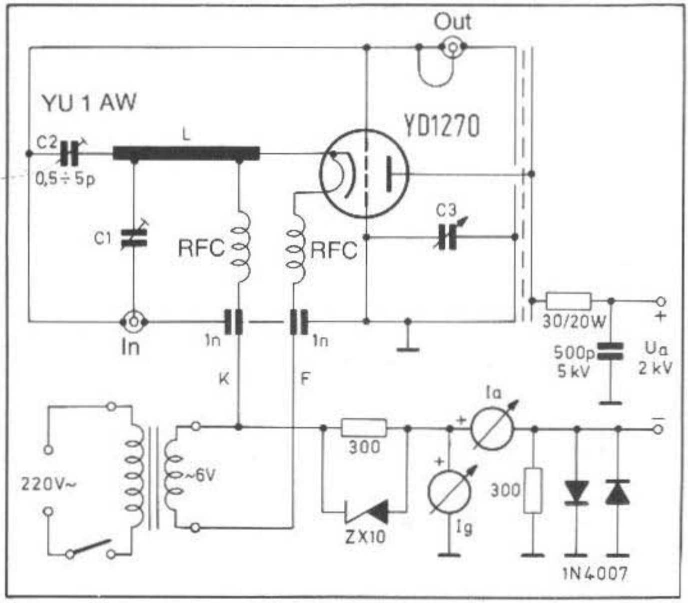


Fig. 1:
Circuit diagram of the 23 cm power amplifier

original tube socket – the latter being both expensive and not easily obtainable. The cathode cavity (**fig. 2**) was made fairly large in order that a series of other necessary components could be accommodated within it. The resonator itself, has a particular form (**fig. 3 and 4**) in order to achieve a very low circuit impedance. This is necessary when, in spite of the high tube input capacitance and the fact that a portion of the λ/2 line is inside the tube, a usable length of line is available outside the tube. This idea has already been published in the VHF Communications (see ref. 1).

The trimmer C 2 is tuned for resonance at 1296 MHz at a capacitance of 1 pF. The satellite band of 1269 MHz can be tuned with a higher value of capacitance. In any event, a high-quality version of this component is required, the author used a Johanson 5200 (0.5 - 5 pF).

The resonator L has a typical impedance of 75 Ω and is brought to resonance at 1330 MHz with a minimal C 2 capacitance of 0.5 pF. Capacitor C 1 loads the circuit and is fabricated from a single

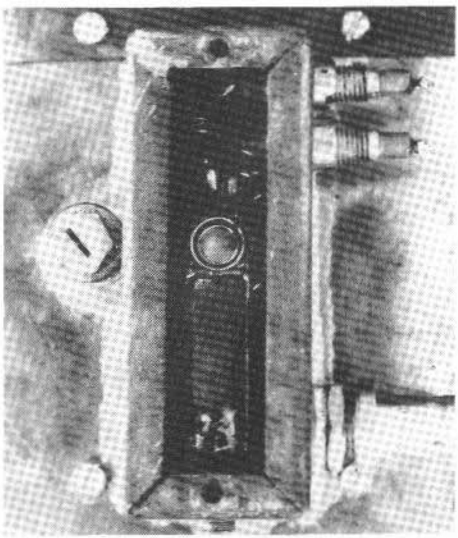


Fig. 2: The cathode cavity and heater chokes

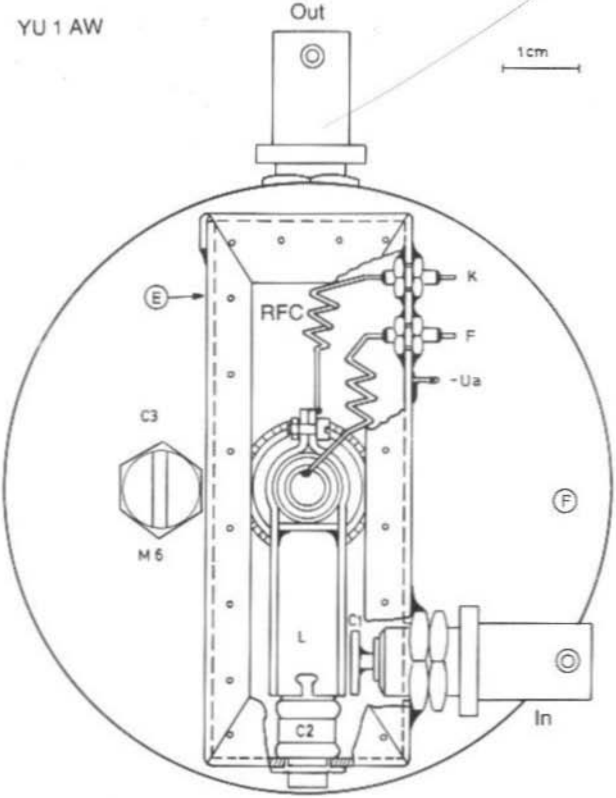


Fig. 3:
The amplifier viewed from below together with the cathode cavity

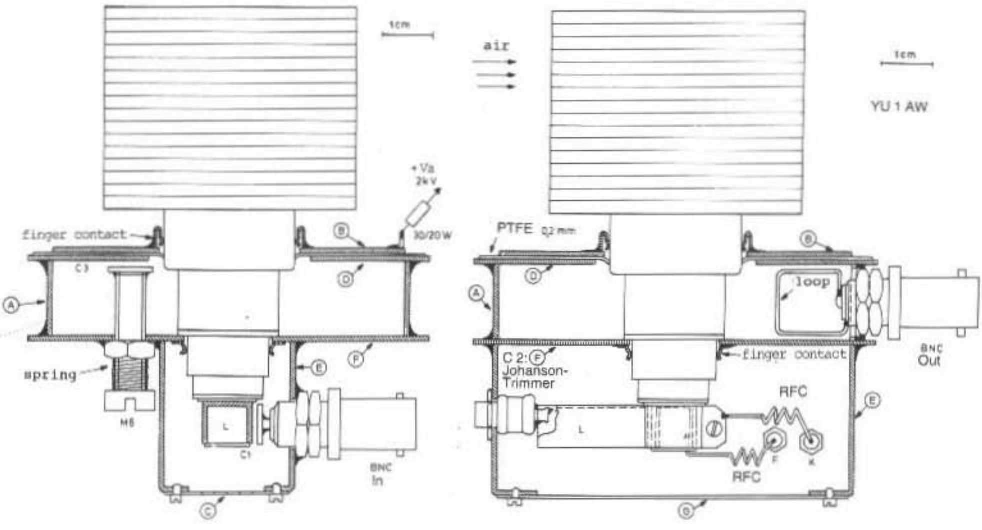


Fig. 4: Two cross sectional side aspects separated from each other by 90°

hole mounting BNC connector. The centre conductor of the BNC socket carries an end-plate of 8 mm diameter. The capacitance to the resonator is varied as the body of the BNC socket is revolved. Optimal coupling occurs when C 1 lies between 0.1 and 0.2 pF.

2.2. The Anode Resonator

The anode resonator takes the form of a round cavity with the PA valve in the centre (**fig. 4**). This indicates the TM_{010}-mode operation with its optimum power output and simple fabrication.

The anode resonator is tuned by C 3, a 6 mm brass screw with a 9 or 10 mm diameter brass disc soldered to its end. This must be positioned right against the PA valve where the electrical field is at its strongest. This simple and effective tuning arrangement matches the resonator field distribution very well indeed. Higher mode parasitic oscillations are well-suppressed and the desired dominant mode is strongly maintained. The resonator is so dimensioned that, when the capacitance of C 3 is at a minimum, the resonant frequency is 1360 MHz. It must have a capacitance of about 1 pF to tune the anode to 1296 MHz. The tolerance of C 3 is sufficient to allow for any spread in the valve's output capacity. The calculated loaded output circuit Q is 120.

The output coupling to the antenna is inductive which is favourable for this high output capacity,

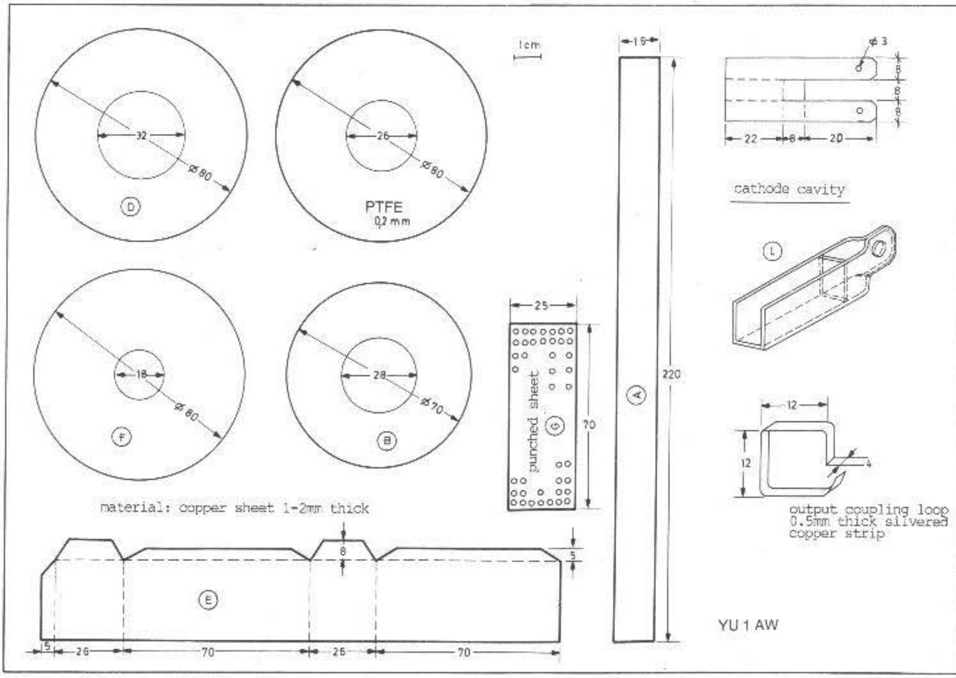


Fig. 5: 23 cm Power amplifier piece-parts

compact-resonator design. The output coupling loop was given special attention. Both due to its form and its deployment, it was possible to obtain an optimum load coupling to the resonator. The degree of coupling is determined by the angle by which the coupling loop cuts the resonator magnetic field. As the loop is soldered to a rotatable BNC single-hole socket, a precise control over the output coupling is possible.

3. MECHANICAL CONSTRUCTION

The metal parts of the amplifier were cut from 1 to 2 mm thick copper sheet. It may be seen that no special tools are necessary in order to make this high-performance 23 cm power amplifier: tin snips, large soldering iron, hand drill and some files are sufficient.

All metal parts are soldered together on the outside, as solder in the resonator cavity causes losses. All screws protruding into the cavity must be of either brass or copper. The strip A (**fig. 5**) is bent into a ring and soldered to parts D and F thus forming the output cavity resonator. The 0.2 to 0.3 mm thick teflon (PTFE) foil must be smooth, without holes or cracks.

The output coupling loop is formed from a 4 mm wide, and only 0.5 to 0.6 mm thick, strip of copper or brass stock. In order to bring it into the cavity, a slot is cut in the resonator wall A. The hack-saw cut is made through the BNC mounting flange nut (which has already been soldered to the cavity wall) from top to bottom of the cavity so that the slot lies in the same plane as the tube's axis and also, incidently, to the HF wall currents. The positioning of this slot is such that the Q of the cavity is in no way compromised.

Both BNC sockets have two nuts, one being used as the mounting flange and the other as a locknut to keep the loop in position after the coupling has been adjusted. The centre conductor of the output coupling BNC socket is cut as short as possible in order that the loop has the maximum coupling i.e. in the closest proximity to the wall of the cavity where the field is the most intense.

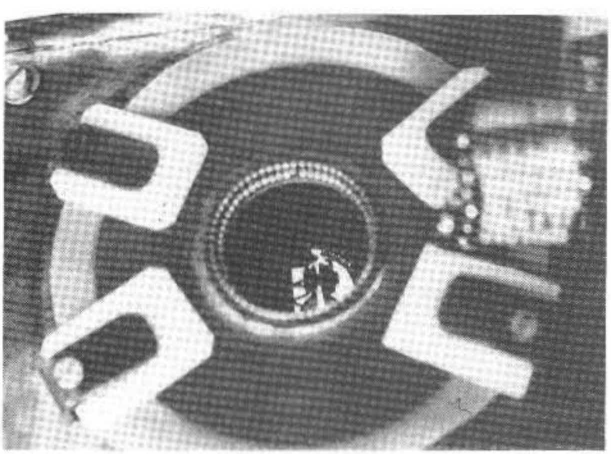


**Fig. 6:
View into the anode cavity**

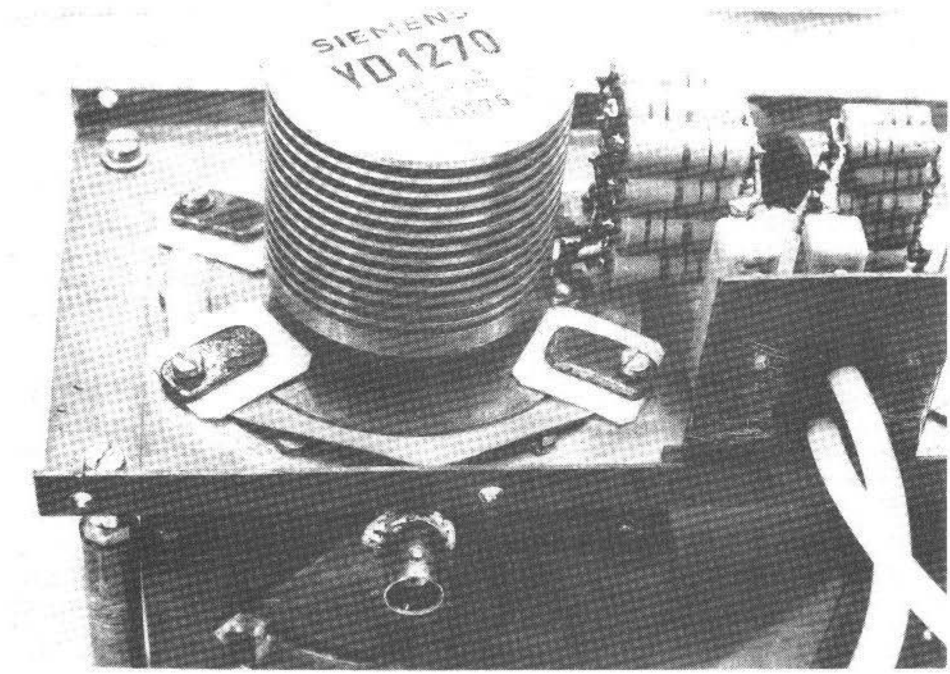


Fig. 7: The completed 250 W amplifier without cooling duct

The valve's anode and grid make contact via finger contact stock (**fig. 6**). The strip is cut to the required length and carefully soldered to the appropriate inner holes of plates B and F. Any poor contact here with the tube's electrodes will lead either to a power loss due to arcing or to instability.

The tube is pushed through the plate B until the anode is firmly held by the finger contacts. The teflon plate lies between plate B and the top of the cavity and is held in place by the grip exerted by the finger strip on the cavity bottom plate F and that of the finger strip of plate B.

After the tube has been inserted, the cathode cavity L is clamped to the cathode by means of a screw through the extended ends of the cavity. Both heater chokes are self-supporting and have an internal diameter of 4 mm and wound from 0.3 mm copper wire (CuL) to form a 40 mm long coil.

4. POWER SUPPLY

The amplifier requires only a heater and an anode supply, the grid bias is developed automatically across the power Zener diode Z x 10 (10 V, 12 W).

The heater potential is 6 VAC. This potential should be reduced slightly at UHF and SHF owing to the additional back-heating effect at these frequencies.

The heater secondary windings must be isolated from earth because of the automatically derived bias.

An anode supply voltage of 2000 volts is required and is protected by a number of small resistors (**fig. 7**), connected in order that the total array can dissipate some 20 W. The anode disc is then held in position with four small insulated retaining plates.

Of course, the tube must be cooled by a powerful fan blower. A chimney made from perspex, or some similar material, conducts the air coolant to the tubes cooling fins and surrounding anode area. This was omitted from fig. 7 on the grounds of clarity but its provision should be effected with little difficulty.

In order to better stabilize the tube heat dissipation, it may be better to switch the blower off during breaks in transmission or perhaps arrange it to work at a reduced performance. Also during sender pauses, the heater potential could, with advantage, be reduced.

5. TUNING

The setting-up commences with fixing the output coupling loop so that it is as near as possible to the cavity wall and is in a plane parallel to the main axis of the tube, as in fig. 4. Looser coupling is effected by turning the loop on its axis towards the 90° position, where it becomes zero. The optimal coupling will occur when the angle of the loop is some 30° to the start position.

The tuning consists of supplying the amplifier, applying a little RF drive power and adjusting C 1 and C 2 alternatively until a maximum anode current has been achieved. It is assumed, of course, that the amplifier has been either coupled to a low VSWR antenna or preferably a dummy load. The latter could comprise several lengths of poor quality (for this frequency) 50 Ω cable.

Capacitor C 3 is now turned in order to obtain the maximal output. The drive is then increased in steps, each time re-adjusting the three trimmers.

Maximum power is then achieved by trimming the angle of the output coupling loop, each adjustment followed (always) by a retuning of C 3. The final adjustment is made when the anode current has been driven to 250 mA coincident with a maximum output power.

A word of caution: All tuning operations should be undertaken in the space of a few seconds, as the input power is in the region of 500 W! The amplifier must never be tuned off-load or with partial load otherwise the resulting voltage build-up in the anode cavity could spell the end for the YD 1270.

Test Parameters

When the amplifier has been correctly tuned, the following test readings may be expected: –

V_a	=	2000 V	P_{in}	=	500 W
V_{g1}	=	– 10 V	P_{out}	=	235 W
I_{ao}	=	35 mA	Gain	=	13 dB
$I_{a\,(max)}$	=	250 mA	V_f	=	5.9 VAC
I_{g1}	=	30 mA			

Using increased cooling power, an output of some 300 W may be expected but the risk of loosing the tube is also higher.

The amplifier described has been in operation for some time now and without any problems worth mentioning. It has been used alone for many EME contacts and is now driving a 1 kW power amplifier.

6. REFERENCES

(1) Editors: A Power Amplifier for the 23 cm Band Equipped with the 2 C 39 Tube. VHF COMMUNICATIONS Vol. 8, Ed. 4/1976, Pages 222 - 231

Joachim Kestler, DK 1 OF

A 10 kHz - 30 MHz Receiver Front End Part 2 (Concluding)

4. THE FREQUENCY DIVIDER UNIT

4.1. The Circuit

Referring to **fig. 13**, the oscillator signal is fed to a MOSFET buffer stage T 1 via Pt 1. This stage isolates the oscillator from the effects of the HF stages. An emitter follower T 2 effects an impedance change in order that the wideband amplifier I 1 can be driven by a low impedance. A 15 volt supply potential is required for these stages which is fed in via Pt 2 at a current consumption of about 40 mA.

Transistor T 3 functions as a switching transistor, supplying TTL-level signals to the divide-by-four scaler I 2. Providing 15 V is applied at Pt 5, a signal fo/4 is supplied to the output Y via I 4 pins 5, 6, 9 and 8. When, on the other hand, Pt 5 is low

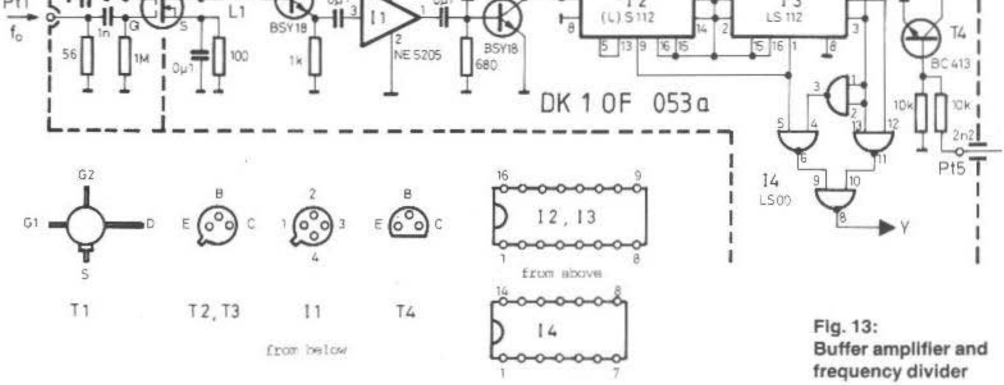


Fig. 13: Buffer amplifier and frequency divider

(zero volts) I 3 is activated which divides the signal yet again, this time by 2. The output signal fo/8 is then supplied via I 4 pin 12, 11, 10 and 8 to point Y. Transistor T 4 is an inverter/level-changer for the switching potential applied to Pt 5. The integrated circuits I 3 and 4 can be changed, if desired, with the low consumption LS (low-power Schottky) TTL version. The chip I 2 may, under certain circumstances, require an S (Schottky) type as the guaranteed clock frequency of the LS 112 is only 30 MHz whereas for this application 40.7 MHz is required. The tested limit frequency of about a dozen examples of LS 112 of various manufacturers, however, lay over 60 MHz.

The digital section, described above, is supplied via Pt 4 by + 5 volts at a current consumption of approx. 25 mA when I 2 is an fo signal output to the receiver input-frequency counter to which it supplies about 70 mV at 50 Ω. The input sensitivity of the circuit versus frequency is shown in **fig. 14**. When the PLL module DK 1 OF 046 was described in ref. 2, it was mentioned that it required at its input a distortion-free, sinusoidal input signal of 20 mV RMS amplitude. **Fig. 15** shows how this demand has been elegantly fulfilled. It comprises a three-stage, low-pass filter (L2, L3 and L4) which is supplied at its input with the divided oscillator signal from point Y at TTL level. The low-pass filter's limit frequency is adjustable by means of varactor diodes D1, D2 and D3. The control chip I 5 is so adjusted that the HF voltage on rectifiers D4/D5 remains constant. Frequency f_y can then be taken from Pt 6 at the required level of 20 mV in 50 Ω. The level regulator stage is supplied via Pt 7 with + 15 V at approx. 3 mA.

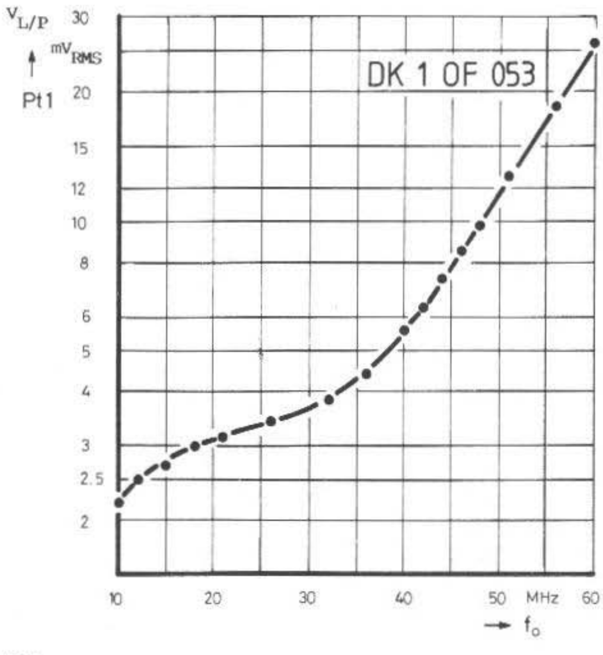


Fig. 14:
Input voltage versus frequency

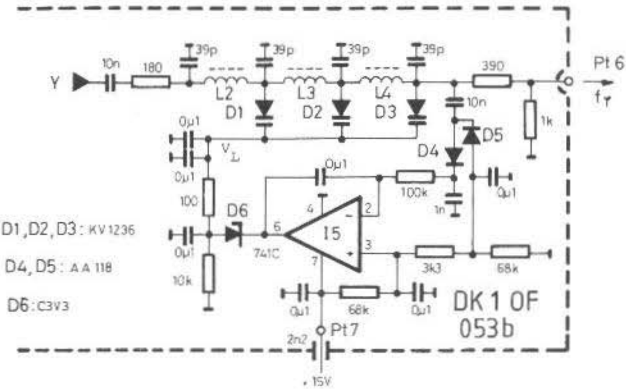
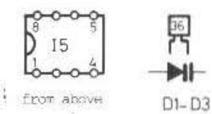


Fig. 15: **Output low-pass and level control**



4.2. Construction

For these units, a PCB was also designed using a single-sided 109 mm x 34 mm glass-epoxy material. It is housed in a tin-plate box and carries the designation DK 1 OF 053. The layout plan can be seen in **fig. 16** and a completed example is shown in the photo of **fig. 17**.

4.3. Components

I 1:	Integrated wideband amplifier NE 5205 EC (Valvo Signetics)
I 2:	SN 74 S 112 N or SN 74 LS 112 N (see text)
I 3:	SN 74 LS 112 N
I 4:	SN 74 LS 00 N
I 5:	Operational amplifier 741 C DIP-8 package
T 1:	BF 900, BF 963 or similar
T 2, T 3:	BSY 18, 2 N 708, 2 N 914 or similar
T 4:	BC 413, BC 108 or any Si-NPN-transistor
D1, D2, D3:	Varicap diode KV 1236

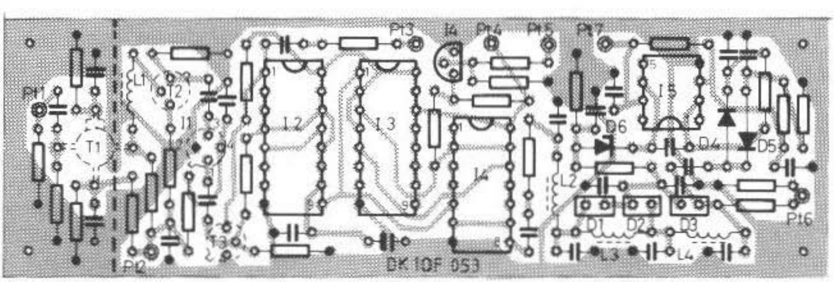


Fig. 16: Component layout of buffer amplifier and divider card

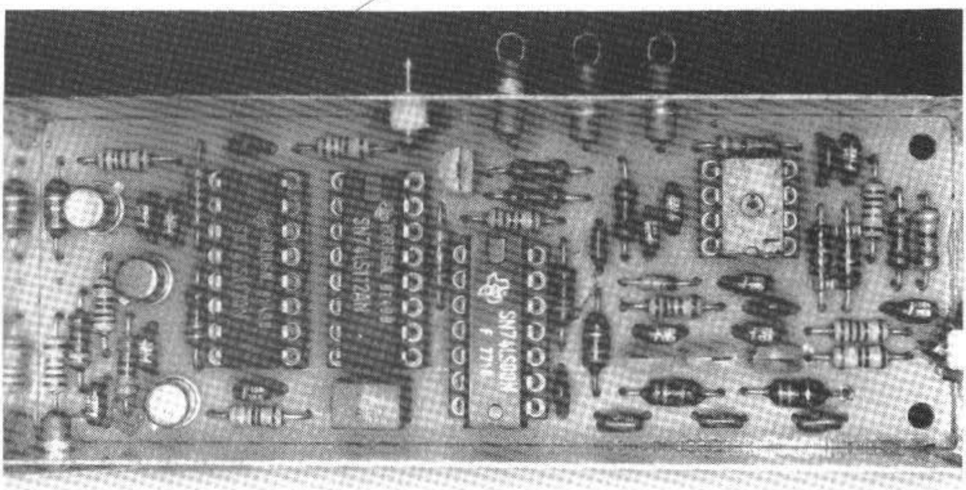


Fig. 17: Prototype of buffer amplifier and divider unit

D4, D5:	AA 118 or similar Ge diode
D6:	Z diode 3.3 V
L 1:	Ferrite choke 3.3 μH (eg. Siemens B 78108-T 1332-K)
L 2, L 3, L 4:	Ferrite choke 33 μH (eg. Siemens B 78108-T 1333-K)

All capacitors smaller than 1 μF are ceramics or plate types.

4.4. Setting up

Following the application of + 15 V to Pt 2, the working point of tranistor T 1 is checked. A drain current of 6 - 10 mA should be evident, this is best checked by measuring the potential difference across the source resistor of T 1. Should there be a departure from the above specified current, then another BF 900 should be tried. The DC voltage at I 1 pin 4 should lie between 5.5 V and 7 V. Afterwards, Pt 7 is also supplied with + 15 V and Pt 4 with + 5 V. A signal-generator, set to a frequency of between 10 MHz and 40 MHz, is fed to Pt 1 and a frequency counter to Pt 6. This latter instrument should indicate an input signal which has been divided by either 4 or 8 depending upon the presence or absence resp. of the + 15 V supplied to Pt 5.

Finally, the automatic level circuit can be checked by terminating Pt 6 with 50 ohms and plotting the control voltage V_v versus output frequency characteristic. The specified characteristic is shown in **fig. 18**. Departures from this curve are not critical provided I 5 is not working under saturation conditions – its output voltage at pin 6 should lie between 4 V and 12 V as f_v is shifted from 2.6 MHz to 5.2 MHz.

5. Deployment of the Modules

In order to obtain a better perspective and also to clear up any possible confusions, the wiring plan of the various modules of the project is shown in **fig. 19**. The following table describes the function of the control elements.

Table 1:

P 1:	Main tuning potentiometer (Sine-Cosine pot) may be replaced by module DK 1 OF 050 (5) if digital tuning is preferred.
P 2:	Frequency fine-tune (see ref. 2)
P 3:	Frequency course-tune
S 1:	Tune-rate switch: course/fine-tune
S 2:	Frequency range switch A: 0 - 2 MHz B: 2 - 4 MHz C: 4 - 10 MHz D: 10 - 20 MHz E: 20 - 30 MHz
S 3:	Input pre-amplifier on/off

The 15 V supply voltage is most conveniently provided by a three-terminal regulator, such as a 7815, which will carry some 350 mA. The DC supply to this regulator should be constant and not subject to the sort of variations which would occur if an AF output stage, for example, were supplied by the same line. The 5 VDC supply for Pt 4/053 can be taken from the counter or from a separate 50 mA stabilized supply. The VCO of module 046, working at 5 - 6 MHz is not necessary and can therefore remain, for the moment, without active elements.

6. TEST RESULTS

The following data represents that taken from the complete front-end under the following de-

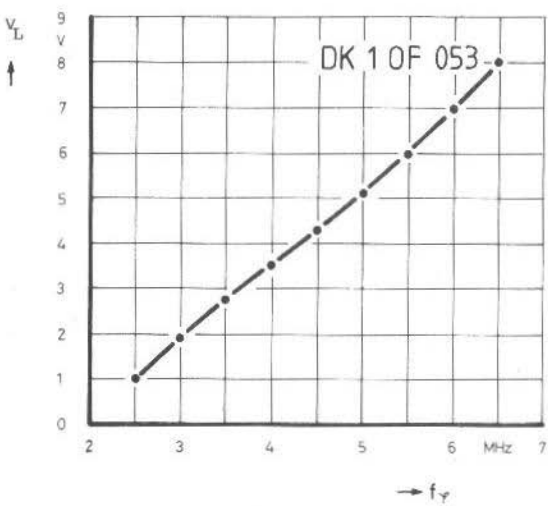


Fig. 18:
Level control voltage V_L versus f_y

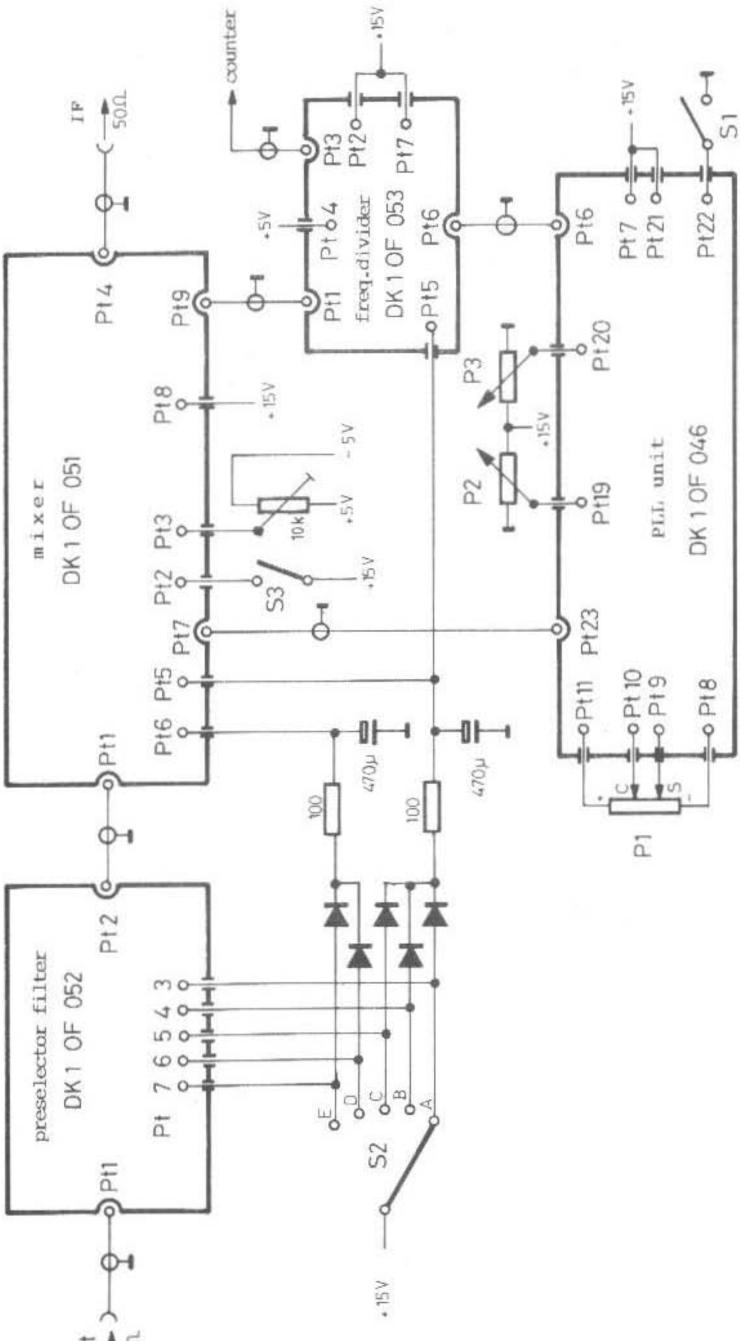


Fig. 19: Module wiring plan of complete front-end

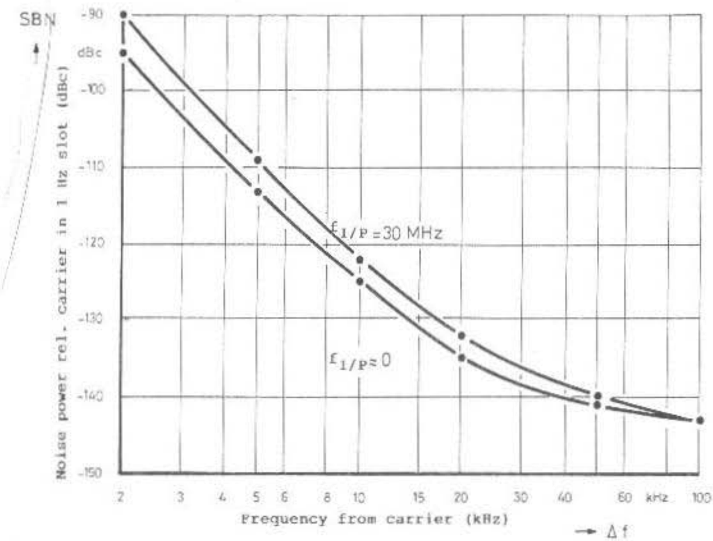


Fig. 20: Oscillator noise spectrum

signated working conditions: –

„A"	Pre-amplifier switched off
„B"	Pre-amplifier GPD 462 with 3 dB attenuator
„C"	Pre-amplifier NE 5205 without attenuator

6.1. Through Gain, Noise Figure and Intercept Point

Note: IF Amplifier noise figure F = 3 dB

„A"	G = − 7 dB	F = 11 dB	IP_3 = +23 dBm
„B"	G = + 3 dB	F = 8 dB	IP_3 = + 5 dBm
„C"	G = +13 dB	F = 6.5 dB	IP_3 = − 2 dBm

6.2. Suppression:

Image suppression

f_{in} < 2 MHz	a_s > 100 dB
f_{in} = 10 MHz	a_s = 85 dB
f_{in} = 20 MHz	a_s = 92 dB
f_{in} = 30 MHz	a_s > 100 dB

IF break-through:

without trap:	57 dB
with trap:	86 dB

These values refer to receive ranges C and D. In ranges A, B and E the IF break-through is greater than 100 dB owing to the additional influence of the pre-selector filters.

6.3. Oscillator Data

Noise spectrum:

at 10 kHz distance:	< −120 dB in 1 Hz bandwidth
at 100 kHz distance:	< −140 dB in 1 Hz bandwidth

see **fig. 20**

Frequency drift:

Ranges A, B, C:	< 40 Hz per hour
Ranges D, E:	< 80 Hz per hour

Output to frequency counter:
70 mV ≙ −10 dBm/50 Ω

Oscillator (radiation) signal at antenna input:
< -80 dBm when fo $\leq$ 30 MHz
< -100 dBm when fo $>$ 30 MHz

6.4. Other data

HF input impedance:	50 Ω
Output impedance at IF:	50 Ω
Supply requirements:	+15 V (approx.) at 350 mA
	+ 5 V (max) at 50 mA
	− 5 V (approx.) at 1 mA

7. LITERATURE

(1) Kestler, J.:
PLL Oscillators with Delay Lines,
Part 1: Fundamentals.
VHF COMMUNICATIONS Vol. 16,
Ed. 4/1984, Pages 211 - 220

(2) Kestler, J.:
PLL Oscillators with Delay Lines,
Part 2: A Shortwave VFO from 5 to 6 MHz
VHF COMMUNICATIONS Vol. 17,
Ed. 1/1985, Pages 46 - 54

(3) Kestler, J.:
PLL Oscillators with Delay Lines,
Part 3: Oscillator Module for the 2 Metre Band
VHF COMMUNICATIONS Vol. 17,
Ed. 2/1985, Pages 112 -120

(4) Kestler, J.:
PLL Oscillators with Delay Lines,
Part 4: Carrier Noise Sidebands.
VHF COMMUNICATIONS Vol. 17,
Ed. 3/1985, Pages 138 - 140

(5) Kestler, J.:
PLL Oscillators with Delay Lines,
Part 5: Digital Frequency Tuning.
VHF COMMUNICATIONS Vol. 19,
Ed. 1/1987, Pages 2 - 12

(6) Kestler, J.:
A Universal Converter for HF and VHF.
VHF COMMUNICATIONS Vol. 8,
Ed. 3/1976, Pages 159 - 174

(7) Krug, F.:
A Versatile IF Module Suitable for 2 m Receivers or as a IF Module for the SHF Bands, Part 1.
VHF COMMUNICATIONS Vol. 13,
Ed. 4/1981, Pages 244 - 250

(8) Krug, F.:
A Versatile IF Module Suitable for 2m Receivers or as a IF Module for the SHF Bands, Part 2.
VHF COMMUNICATIONS Vol. 14,
Ed. 2/1982, Pages 112 - 124

(9) Krug, F.:
A Versatile IF Module Suitable for 2 m Receivers or as a IF Module for the SHF Bands, Part 3.
VHF COMMUNICATIONS Vol. 14,
Ed. 3/1982, Pages 172 - 189

(10) Krug, F.:
A Versatile IF Module Suitable for 2 m Receivers or as a IF Module for the SHF Bands, Part 4.
VHF COMMUNICATIONS Vol. 14,
Ed. 4/1982, Pages 239 - 252

(11) Krug, F.:
A Versatile IF Module Suitable for 2 m Receivers or as a IF Module for the SHF Bands, Part 5.
VHF COMMUNICATIONS Vol.15,
Ed. 1/1983, Pages 49 - 60

(12) Krug, F.:
A Versatile IF Module Suitable for 2 m Receivers or as a IF Module for the SHF Bands, Part 6.
VHF COMMUNICATIONS Vol. 15,
Ed. 2/1983, Pages 103 - 111

Jochen Jirmann, DB 1 NV

Integrated Single-Stage Broadband 0 to 2 GHz Amplifier

Not long ago, "Mini-Circuits" brought out an integrated HF amplifier which covered the frequency range 0 to 2 GHz with varying degrees of amplification. As may be seen from fig. 1, the amplifier needs merely two DC blocking capacitors and a resistor to establish the working point, in order to put it into operation. The price also is interesting, which, according to version, lies between 5 and 10 DM.

As far as its innards are concerned, it could only be established that a mixed current-voltage feedback, bipolar microwave transistor is employed. The feedback network fixes the working point of the transistor and the circuit diagram is given in **fig. 2**.

Experiments have shown that the amplifier is easy to work with and is conservatively specified, inasmuch, that the amplification minimal value was found to be higher than that given in the accompanying data-sheet. The gain falls very gradually above the high limit frequency so that even at 3 - 4 GHz there is 6 to 8 dB available. The input and output return-loss lies around 15 to 20 dB. The frequency phase characteristics displayed a few kinks above 2.5 GHz which will have to be taken into account when disigning amplification for extremely wideband signals. The amplifier is capable of delivering a power from 50 to 100 mW when the working point is set somewhat higher than that recommended in the data-sheet. Increasing the working current, however, has no

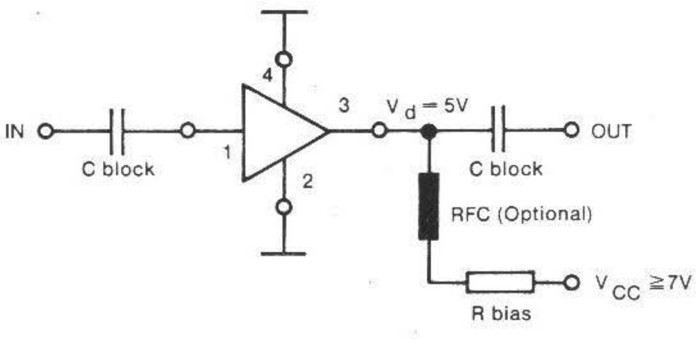
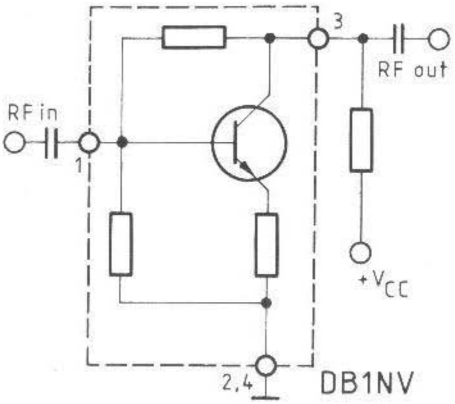


Fig. 1: Integrated wideband amplifier circuit diagram



positive effects, as the amplification and the maximal output power both fall and the limiting frequency is reduced.

Besides its application as a wideband instrument amplifier, it may also be considered for use as a pre-amplifier for a GHz frequency counter, a wideband buffer for an oscillator, a ringmixer local oscillator driver amplifier or as an HF pre-amplifier in a simple VHF-UHF receiver.

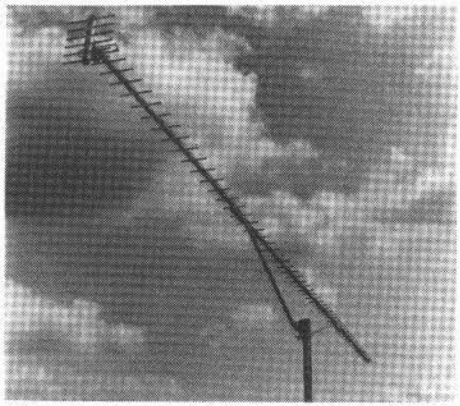
Fig. 2: The current voltage feed-back is also integrated

New High-Gain Yagi Antennas

The SHF 6964 is a special antenna for the **space communication allocation** of the 24 cm band. The maximum gain of this long Yagi is 19.9 dB_d at 1269 MHz and falls off quite quickly, as with all high-gain Yagis, with increasing frequency. We do not, therefore, recommend this type of antenna for operation at 1296 MHz but for **ATV applications** at 1152 MHz it is eminently suitable. There is no 24 cm ATV antenna on the world market which possesses more gain.

The mechanics are precise, the gain frequency-swept and optimised. Measurements carried out during heavy rain show that the antenna is not detuned by moisture.

Length:	5 m
Gain: 22 dB_i, i. e.	19.9 dB_d
Beam-width:	13.6°
Front / Back ratio:	26 dB
Side-lobes:	− 17 dB
VSWR ref. 50 Ω:	1.2 : 1
Mast mounting: clip (max).	52 mm
Stock-No. 0103	**Price: DM 298.—**



The SHF 1693 is a special version for the **reception of METEOSAT 2**. This unobtrusive alternative to a 90 cm diameter parabolic antenna enables, with the aid of a modern pre-amplifier or down-converter, noise-free weather picture reception.

Length:	3 m
Gain: 20.1 dB_i, i. e.	18 dB_d
Beam-width:	16.8°
Front / Back ratio:	25 dB
Side-lobes:	− 17 dB
Stock-No 0102	**Price: DM 440.−**

Jochen Jirmann, DB 1 NV

The SDA 4211 - An Interesting UHF Prescaler

Since television receivers were equipped with synthesized tuners, favourably priced ECL pre-scalers with integrated pre-amplifiers, for use up to one Gigahertz, have become available.. Well-known types are the U 264/U 664 (÷ 64) or the U 666 (÷ 256) from Telefunken. Circuits for the decimalisation of the division ratio have been described in VHF Communications and nothing should stand in the way of their use in amateur built frequency counters.

Unfortunately, the frequency of operation of these ICs only extends as far as 1100 to 1200 MHz and only a very few selected examples can work at frequencies in the 23 cm band. Recently, however, a divider chip with the designation SDA 4211 and a guaranteed frequency range from 70 to 1300 MHz, has become available from Siemens. The division factors are 64 and 256, switchable, and the consumption is only 25 mA at 5 VDC.

The pin-out is exactly that of the U 264/U 664 except that pin 1 is unconnected and pin 5 switches the division factor. When pin 5 is earthed, the chip divides by 256, when it is at + 5 V the division factor changes to 64. The sensitivity, according to the data sheet, is 11 mV at 70 MHz, 5 mV at 250 MHz and 40 mV at 1300 MHz. The author's experiments – carried out on a few examples – show that the SDA 4211 will work perfectly when driven by a ainusoidal input from 5 MHz (200 mV) to 1600 MHz (220 mV 50 ohm or 0 dBm).

The chip costs about 8 DM each in quantities of 10, which puts it well within most amateur budgets.

Ralph Berres, DF 6 WU

Television Receiver Field-Strength Indicator

A received level indicator is for the amateur television operator very advantageous. I have designed a field-strength indicator in the form of a moving column of light along the top of the TV screen. This is more suitable as there is no room to mount a moving coil instrument (portable TV) and in any case the TV housing should not be drilled into.

1. FUNCTIONAL DESCRIPTION

A_s may be seen from the circuit diagram of **fig. 1**, there is nothing very difficult about the practical realization of this circuit. The transistor T1 to-

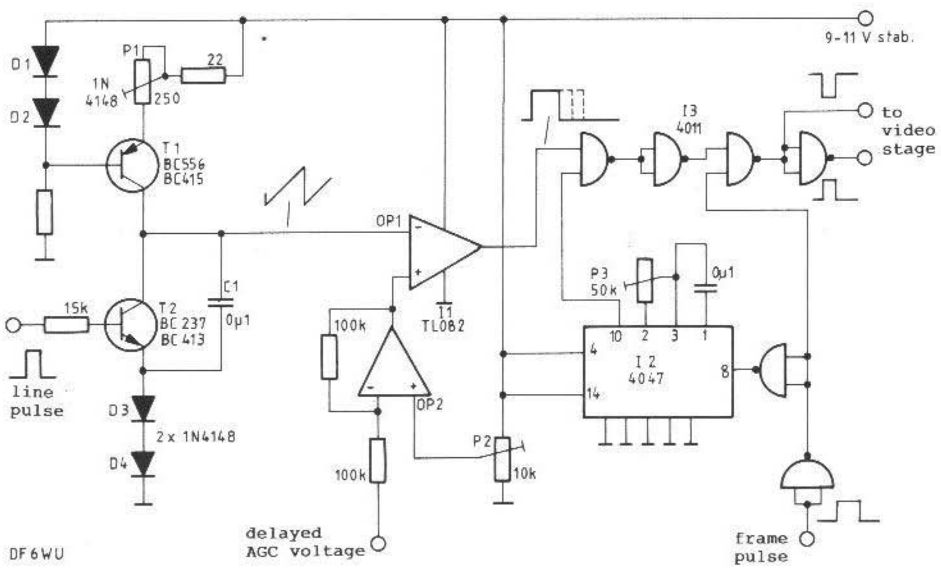


Fig. 1: TV receiver field-strength indicater

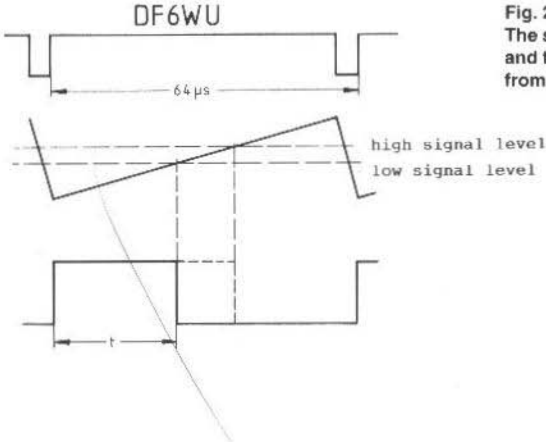


Fig. 2:
The signal control voltage, saw-tooth wave and the variable duration pulse derived from them

gether with potentiometer P1 form a constant current source which charges capacitor C1. The charging time is controlled by P1 and lies in the region of 60 μs. The capacitor C1 is being discharged by transistor T2 during the period of every line impulse. This produces a saw-tooth wave which is synchronized to line frequency and the amplitude of which is dependent upon the charging current – the latter being controlled by P1.

The diodes D3 and D4 prevent the saw-tooth wave from reaching zero volts because otherwise the following operational amplifier OP 1 will receive a voltage overload at its input terminals.

The AGC voltage of a television tuner is, in general, 6 to 9 V in the absence of signal and reduces with increasing signal-strength. For our purposes, this behaviour must be reversed and that is accomplished by OP 2 connected as an adder. The potentiometer P2 at its non-inverted input controls the scaling value of the AGC control voltage.

The inverted AGC voltage is then taken to OP 1 which is used as a comparator. When the AGC voltage at any instant is more positive than the saw-tooth wave, the output of the comparator goes directly to the potential of the postive rail $+ V_b$. The length of time it stays at this potential is dependent upon the AGC voltage which, in turn, is dependent upon the incoming signal-strength. This sequence is repeated for the duration of line time (**fig. 2**) at line frequency.

In order that half of the screen, from top to bottom, does not "white out", the incoming signal must, in some way, be connected with the frame-pulse. For this purpose a C-MOS type 4047 mono stable trigger is used. Its output pulse duration is controlled by P3, this being the control for the width of the moving column of light indicator. This signal is now gated with the column-length signal in the 4011 NAND gate.

The following gating with the frame pulse is intended to prevent TV sets, having automatic brilliance control, and which, in the vertical blanking time, produces a white line for every colour channel, from throwing the colour symmetry out of balance. It also ensures that the moving column starts only at the end of the frame pulse.

2. INSTALLATION AND ADJUSTMENT

The few components employed are loaded quite simply on to a piece of vero-board and connected up. The main problem is the determination of suitable circuit interface points which will accommodate the board without causing any deterioration to the rest of the picture. Unfortunately, no specific instructions can be given here as there are simply too many basically different circuit concepts.

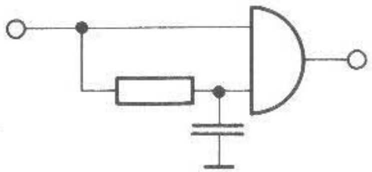


Fig. 3: 5μs delay circuit. R = 10 KΩ, C = 470 pF

In the author's set, the "sandcastle" pulse was used to derive the line pulse. Difficulties can occur if the picture blanking pulse is used since it is not possible to blank brilliance and black at the same time. In this case, it is recommended to delay the line pulse by a further 5 μs using the simple gate delay circuit of **fig. 3**.

The frame pulse can be obtained from the limiter which, as the same time, is used for the synchronization of the vertical deflection. The accompanying spurious pulses are rendered harmless by the second gate. After the module has been completed and successfully connected into the TV circuitry, P1 and P2 are turned to midposition and P3 to maximum. The set is then switched on and a very strong signal is tuned-in. The indicator column of light should be visible. With an oscilloscope, connected to the output of the unit, a pulse should be visible whose width is a function of the field-strength.

The length of the column is now adjusted with P1 such that it nearly reaches the right-hand edge of the screen. The signal is removed and P2 adjusted until the column is now nearly at the left-hand extremity of the screen. These two pot'meters should be iterated until the column moves satisfactory from extreme left to extreme right with no, and full, signal respectively. Finally the column width is set by P3.

Werner Rahe, DC 8 NR

Switched-Capacitor Audio Filter

Variable Bandwidth, Tunable Centre-Frequency, Steep-Sided Flanks

For some years now, an integrated circuit filter has been available which functions using the "switched-capacitor" principle. It enables the relatively easy construction of tunable filters with steep-sided flanks and possessing low-drift characteristics (1). The requisite resistance required is simulated in this type of filter by a switched capacitor. The switching signal supplies an external clock generator through which, at the same time, shifts the limit frequency of the filter.

An astronomical price was being asked for such ICs when they first made an appearance in the early 80s, but now they are reasonably affordable, a 300 Hz to 3000 Hz tunable filter design can be presented in these pages. Both high and low-break frequencies may be independently adjustable in order to achieve the desired passband and centre frequency. The flank slope remains constant at 30 dB per octave for the lower flank and 100 dB per octave for the more important upper flank.

1. INTRODUCTION

The building of a tunable audio filter is no longer a secret since the development of the active filter circuit (2). Within limits the bandwidth may be varied (3). Variable tuning whilst maintaining the same flank slope combined with variable bandwidth has, until now, necessitated the same sort of design artistry as that required for the Indian rope trick.

Radio receivers could, in part, have these desirable characteristics realized in the high-frequency portion with switched filters, passband tuning and IF shift. A continuously variable, independent of bandwidth, tuning, however, can only be achieved with much complication and is encountered only in high-priced professional shortwave equipment. In the higher bands it is not used at all.

The audio stage of a receiver requires some sort of selectivity or filtering in order to suppress some of the wideband noise coming in from the higher stages. Every effort should be made in a high-performance multi-mode receiver to tailor the passband to that of the incoming signal in order to optimize the overall signal-to-noise ratio. The filter requirements for a wideband – only receiver, such as FM ATV, are not too high, – a simple low pass being sufficient to produce an acceptable flank slope and no great demands on selectivity. The situation is quite a bit different, when, in a multi-mode receiver RTTY or CW is being used and the IF does not possess a suitable narrow-band IF filter. If the signal-to-noise cannot be optimized at IF, it must be done at audio. This, however, is much more difficult to accomplish at AF as the demands for an audio filter are difficult to fulfil; 60 dB per octave flanks together with an extremely narrow bandwidth. Such a filter should be the main design aim. In addition comes the requirement for continuous bandwidth control, an ability to control the midband frequency and also the insertion gain of the filter. Since flank steepness, bandwidth, and the ability to handle impulses are mutually dependent properties, some compromise must be tolerated.

2. FILTER CIRCUITS

2.1. True Bandpass Circuits

Simple filter circuits which would satisfy some of the criteria considered so far, do not exist. An example is the universally employed "state variable filter" (4), (also known as the "universal filter" because other filter functions are also possible from the basic design), which will serve to illustrate the point. This filter circuit is represented in advertising literature as one of the new wonder weapons against interference and noise. Closer inspection reveals, however, that only very modest demands can be met with it (3). This filter, it is true, has a good midband Q and amplification – both variable quantities using potentiometers. This does not mean, however, that the two quantities are independent of one another.

They are, of course, related by the formula $Q = fo/B$, where B is the bandwidth in Hertz. It may be easily appreciated, that altering the resonant frequency Q and keeping the relative bandwidth $b = B/fo$ constant, the absolute bandwidth climbs or falls with the resonant frequency. That means, that as the filter is tuned to a higher frequency, the bandwidth becomes larger.

A worth-while flank slope in the vicinity of both turn-over frequencies, can only be achieved by raising the resonant frequency – that is, by increasing the Q or making the bandwidth smaller. The passband of such filters only looks good between the 20 dB points and problems with CW ringing and stability are no problem since the Q is seldom greater than 10. After the 20 dB points, the filter selectivity is as wide open as a barn door deteriorating to only 6 dB per octave – that of a basic second-order filter. Larger bandwidths, such as that used for telephony transmission, with the requisite steep flanks, are not at all possible. Using the same sort of complexity as the universal filter, 3 or 4 operational amplifiers, bi-quadratic filters are the better choice. This is because fo and Q are in a linear relationship with one another; increasing the fo also increases the Q thus preserving the absolute value of the bandwidth.

The same sort of thing can be achieved by using a bandpass filter with several feedback stages – something which can only be done with operational amplifiers. The reason for the existence of simple narrow-band circuits (2) is that they can be dimensioned so that the centre frequency and the amplification can be independently adjustable, i.e. when the tunable range is not too great. Unfortunately, the possibility does not exist to make the bandwidth adjustable and independent of the afore-mentioned variables. It must be considered as a fixed quantity. In principle, these types of filters employ similar bandpassed, cascaded stages, some with offset midband frequencies, in order to achieve the desired overall bandpass. The frequency adjustment becomes very difficult as the number of stages increases.

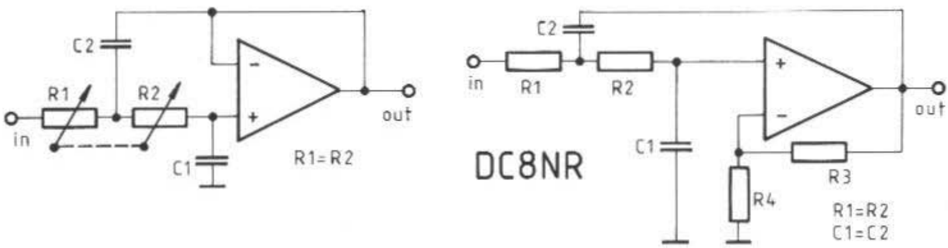


Fig. 1: 2nd order low-pass (simple coupling, controlled source)
Left: LP with unity gain, right: LP with identical components

2.2. Compounded Bandpass Circuits

In view of the relative shortcomings of employing genuine bandpass circuits, the tendency has been to construct the filter from cascaded high and low-pass elements. The flank slope is then dependent upon the order of filter and the number of stages. If the limit frequencies are successfully made variable then a variable bandwidth with constant flank slopes will have been achieved.

Of the equally suitable circuits available, the filter with single coupling (key filter) seems to be the most suitable because it has the smallest number of components. **Fig. 1** shows a 2nd-order low-pass filter. The requisite high-pass is obtained merely by changing the values of resistors and capacitors. Owing mainly to practical reasons, the tuning of such a filter is only carried out by a variation of the resistors only.

In order to obtain a higher-order filter, several second-order basic elements are connected together in cascade. Fortunately, there are designs available which allow the use of equal-value resistances in each stage. If the emphasis is given to amplification, for which filters of a higher order are favourable owing to the high degree of op-amp. feedback (**fig. 1 left**), the filter type is

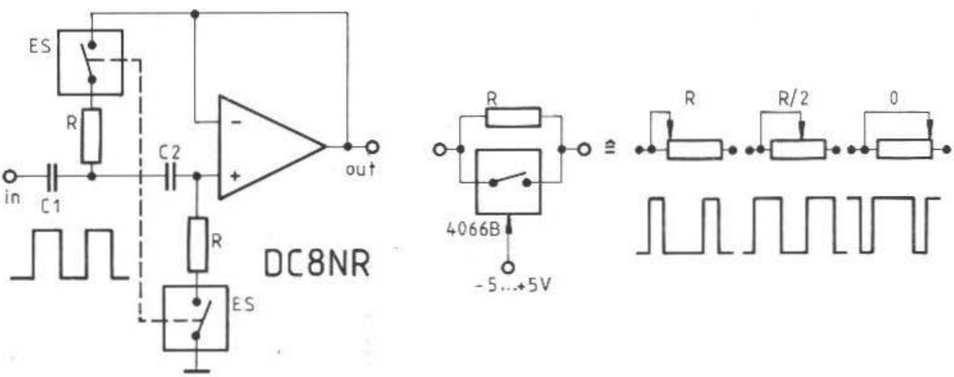


Fig.2 : Simulation of a variable resistor by an electronic switch with controlled duty factor.
Left: high-pass structure

determined by the relationship of C 1 to C 2. If the same value of resistors and capacitors are employed, the amplification of the operational amplifier is fixed to a pre-determined value by R 3 and R 4 (**fig. 1 right**). With the exception of the Butterworth filter, it is not sufficient merely to connect identical elements together in cascade in order to achieve a certain bandpass. Every element has its own limit frequency and Q. The complex design of such filters is simplified greatly by the use of tables (4) and (5).

Using a double potentiometer, the limit frequency of a second-order filter may be varied over quite a large range. A flank slope of 60 dB per octave is only achievable with a Butterworth filter design of the tenth order, or with a Tschebyscheff filter of the sixth order (has passband ripple of 3 dB). Now the difficulty may be clearly seen: The problem of obtaining a six or ten-gang potentiometer with better than a 1 % tracking (tighter tolerance when the higher flank slopes are required). It can be done, but it would stretch the resources of the component industry somewhat. This requirement may be obviated by using switched, fixed-value high-tolerance resistors instead of the ganged potentiometers but then the continuous adjustment is also dispensed with.

2.3. Switched Resistors

Instead of switching the resistors mechanically, they can be switched by electronic means. **Fig. 2** shows an electronic switch, either in shunt or series, with a fixed frequency-determining resistor. The electrically operated switch is constantly being opened and closed but if the keying duty factor is varied, the effective resistance of the combination is also varied in sympathy. **Fig. 2 (right)** shows the arrangement for a parallel connection of electronic switch and fixed resistor. A pre-requisite for this type of circuit is that the switching frequency always lies well above that of the highest signal frequency and the time constants of the filters.

There are many economical electronic switches using C-MOS chips to choose from, above all, perhaps, the 4066. The "on" resistance of these switches is, however, finite and what is worse, it varies not only by manufacturer but from one specimen to another. This may be countered by placing a higher value of resistor in series with all the switches thus making the differential of the switched resistances much smaller as a percentage to the total resistance.

Impulse generators having a variable duty-factor to operate the switches may be easily made with the 7555. The switching frequency must be chosen to be well above the audio frequency so it, and its harmonics, may be easily removed using simple low-pass filters.

Using only a single control voltage to all the switched resistors combinations, a variable ganged potentiometer has been simulated which is capable of tuning quite complex filter arrangements.

2.4. Switched Capacitors

The integration of complex filter circuits into a chip, using the techniques outlined above, has now run into a serious fundamental problem. This is, that the lower values of resistances and capacitances cannot be realized on a chip with the sort of precision required for filter work. This has lead to the development of the switched capacity or SC-filter. The technological manufacturing problems of switching resistances are obviated by switching a capacitor instead. This is done in the following manner. The basic circuit of an active low-pass filter is shown in **fig. 3 (left)** and represents the circuit from which all switched capacitor filters are derived.

Taking the well-known fact that an RC combination has a time constant $\tau = R.\ C$, it may be said that a resistance in the presence of a capacitance and a time constant may be written: –

$$R = \tau / C, \text{ but } \tau = 1 / f, \text{ therefore } R = 1 / fC \qquad (1)$$

Now it must be shown that a capacitor, such as that in **fig. 3 (right)**, connected to a switch which is being operated at fs, represents a resistor of value R. Putting a Voltage V_1 at position 1 in the circuit input, the capacitor receives a charge Q 1 $= C_1\ V_1$. When the switch is now changed to position 2, the capacitor is discharged to a voltage V_2. This occurs in time t during which a current i flows. The transferred charge is then

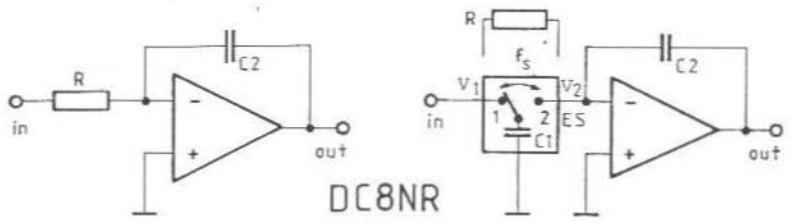


Fig. 3: Basic integrator circuit and its simulation by an SC-arrangement

$$Q = C_1 (V_1 - V_2) = i\,t \qquad (2)$$

If this switching sequence is periodic with a frequency $f_s = 1 / T$, quantities of charge will be transferred every unit of time. If the switch stays in position 1 for the same time as position 2, the period of the switching frequency is

$$T = 2\,t \qquad (3)$$

Substituting in equation (2)

$$i / 2\,fs = C_1 (V_1 - V_2) \qquad (4)$$

As this current i flows for half the time, its average value is i / 2. One can think of this as flowing for the whole of the time. The equivalent ohmic resistance through which the same current would flow is found by Ohm's law.

$$R = \frac{V_1 - V_2}{i / 2} \qquad (5)$$

Substituting in equation (4) gives the analogy

$$R = 1 / fs\, C_1 \qquad (6)$$

Using the SC arrangement, a resistor has thus been simulated whose effective value depends upon a capacitance and the applied switching frequency. The frequency f 1, at which the amplification of the integrator is unity, is:

$$f\,1 = 1 / 2\,\pi\, R\, C_2$$

Using eq. (6)

$$f\,1 = \frac{fs}{2\,\pi}\ \frac{C_1}{C_2} \qquad (7)$$

To remain at any set filter frequency depends upon the stability of the switching frequency and that of the capacity ratio between the two capacitors. The latter is assured by the inherent stability of the MOS integrated circuit. The capacitance ranges from about 0.1 to 100 pF yielding ratios of between 10^{-3} and 10^{+3}.

In **fig. 4** is a practical form of the circuit with an anti-phase controlled MOS−FET switch. In the integrated filter, however, there are several hidden refinements, one of which, is a device to eliminate parasitic capacitances. In general, they are superior to discretely built filters.

2.5. Characteristics of Switched Filters

As opposed to the advantâges of this new filter techniques, there is one disadvantage – the so-called aliasing effect. If a signal is fed through an

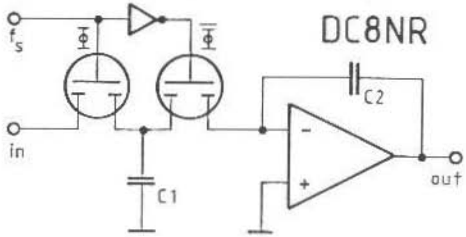


Fig. 4: Practical realization using a MOS-FET switch

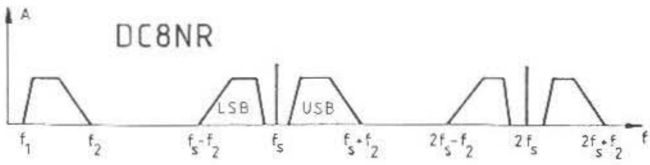


Fig. 5: Output spectrum of a SC-filter (amplitudes if IM products and harmonics not to scale)

SC-filter, the output spectrum will contain a spectrum limited version of the input plus a component of the switching frequency fs together with its harmonics and intermodulation products. Apparently, nothing happens which does not occur in a modulation or a mixing process. As **fig. 5** shows, the filter bandpass is disposed about the switching frequency fs at intervals of nfs, representing a normal amplitude modulation process. The upper sideband (USB) is a mirror image of the lower sideband (LSB) and it can be thought of as the two sidebands being folded about the carrier fs.

If the switching frequency is lower than twice that of the highest signal frequency to be passed, intermodulation will be the result. There will be an overlapping of signals from the main bandpass and those carried from the switching frequency's lower sideband. In like manner there will be interference between the switching frequency's USB and its harmonic's lower sideband. **Fig. 6** should make this clear. In order to counter this effect, a normal low-pass filter, known as an anti-aliasing filter, is always employed before an SC-filter in order to limit the extent of the signal's upper frequency components.

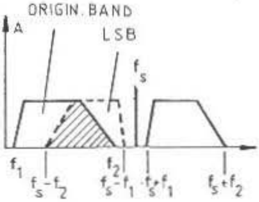


Fig. 6: Production of alias signals at fs < 2f. Interference to signal by LSB components from signal modulated fs carrier.

But in general, these considerations don't cause too many problems except perhaps when SC high-pass filters are concerned. This is because the chip manufacturer has solved the problem of the switching frequency choice and also the maximum permissable signal frequency component. In fact, the SC-filter switching frequency is usually 30 to 1000 times higher than the highest signal frequency. The switching frequency stability is therefore still not compromized as the highest signal frequency is of very low proportions. In general the duty-cycle of the switching frequency is chosen to be 1 : 1 (50 %).

For practical reasons, an analogue low-pass filter is also added to the output of an SC-filter in order to reduce the level of any residual aliasing and switching components. This is particularly important when a second SC-filter is to be connected in tandem. The first should posses the analogue input and output filters.

Another important requirement for the SC-filter is the provision of a well-filtered supply potential. Most voltage regulators possess a rather high-level noise floor complete with regulation spikes, both of which will compromise the efficacy for the filter. Where it is at all possible, a battery supply should be used.

The afore-mentioned theoretical considerations represent an attempt to aquaint the reader with the development and tendencies in the somewhat difficult field of filter technology. He then has a starting point from which suitable filter circuits may be chosen for the solution of specific filtering requirements. The long march of progress can be recalled, from bulky, loss-prone LC-filters of bygone days, to discretely constructed active RC-filters and on to the highly integrated complex filter chips of today. The SC-filter represents the latest state-of-the-art in analogue filter technology. From their manner of function, they actually occupy a place half way between analogue and digital filter technology.

The following part of the article will deal with the dimensioning, the realization and the test-data of tunable SC-bandpass filters.

3. THE PRACTICAL DESIGN

A tunable bandpass filter, using the SC-filter technique as described above, will now be considered in some detail.

3.1. Choice of Filter Modules

Similar to the case when ICs first came on to the market, the number and variety of filter integrated circuits is beginning to be confusing. They range from very simple ICs to multi-pole arrangements of either fixed or user-controlled characteristics. Many chips have an integrated, switching-frequency generator or an additional operational amplifier which serves as a low-component-count, anti-aliasing or clean-up filter. Some ICs contain several independent types of filter or also programable types. A review can be obtained from ref. (1) and ref. (6).

The price of such ICs is relatively high at present, making their use a matter of some consideration – perhaps the requirement may be fulfilled with a simpler filter. Also, because they are still regarded as being "high technology" products, their sale has been restricted to manufacturers and distributors.

The choice, finally, fell on two chips manufactured by EG & G Reticon which seem, at present, to be ahead in their field. They are monolithic SC-filters in a mini DIP package. The short-form data is: –

R 5609: Elliptical (Cauer) low pass, 7th order and 6 nodal points,
working frequency range: 0.1 to 25 kHz
switching frequency: 100 x limit frequency (typ)
flank slope: 100 dB per octave
passband ripple: smaller than 0.2 dB

R 5611: Tschebyscheff high pass, 5th order,
working frequency: up to 8 kHz
switching frequency: 500 x limit frequency (typ)
flank slope: 30 dB per octave
passband ripple: smaller than 0.6 dB

Both these types have a dynamic range of over 75 dB and an insertion loss of zerc dB. The distortion is less than 0.3 %. The supply potentials are +/− 4 V to +/− 11 V. Input voltages of up to 12 V pp may be handled at the higher supply voltage (7).

The inputs are integrated diode-protected but in view of their price it is just as well to treat them as carefully as any unprotected MOS-FET device as far as anti-static measures are concerned (8).

3.2. Circuit Description

Using the above ICs, a bandpass filter will now be built following the circuit shown in **fig.7**. As such a filter is normally connected immediately following the receiver demodulator, the input signal is first of all amplified in the first half of a times-two, low-noise operational amplifier, I 1, by an adjustable (with P1) level of between 6 and 20 dB. The low-pass network of Ch 1 and C 1 keeps high-frequency components from entering the filter.

I 1b forms a Butterworth anti-aliasing filter with a limit frequency of 3.7 kHz. The SC high-pass I 2 follows. When two filters are connected in tandem, it is better to connect the high-pass at the beginning because the noise and intermodulation from the following low-pass will be reduced.

As not all receivers can supply the dual voltages required for this unit, I 2 pin 7, normally grounded, is fed with half the supply voltage in order that only one + 13.8 VDC power source is required. This IC must then have half the supply voltage at its input.

The switching frequency generator, I 3, comprises a 4046 C-MOS PLL oscillator with a 1 : 1 (50 %) duty cycle. The usual 555 IC is incapable

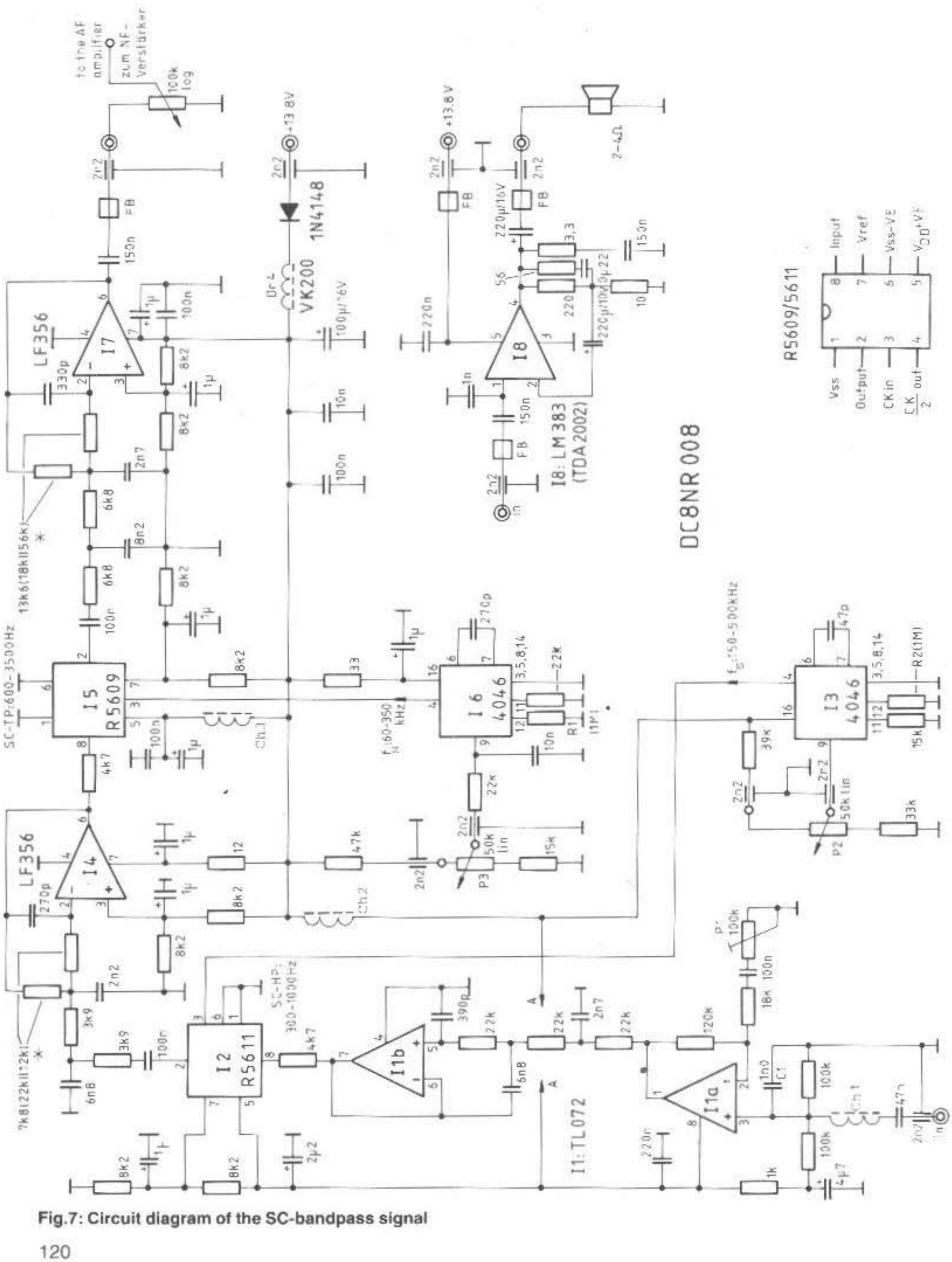
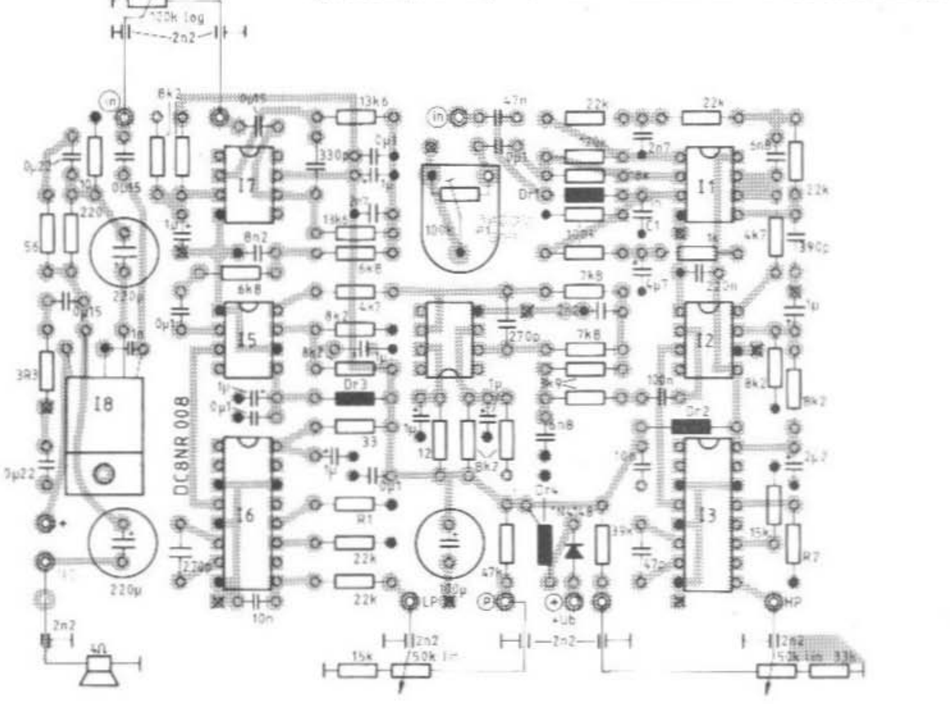


Fig.7: Circuit diagram of the SC-bandpass signal

Fig. 8: Component layout plan of 111 x 72 mm SC-filter board DC 8 NR 008



of supplying such a constant duty cycle independent of frequency.

There is no point in making the high-pass variable over the total range of 300 to 3000 Hz as the midband frequency for telegraph filters lies, in general, between 600 and 1000 Hz. The tunable range was therefore confined to between 300 and 1000 Hz which requires a switching frequency, at an fs/fg ratio of 500, of between 150 and 500 kHz. The tunable limits of P2 are set by fixed resistors at hot and cold ends of the potentiometer. Resistors R1/R2, which serve to preset the lower tunable limit, can be dispensed with.

The VCO of the 4046 is capable of generating a clock frequency of up to 1.5 MHz. The working frequency is determined by the voltage at pin 9 and the capacitor connected between pins 6 and 7. Under no circumstances should the input (pin 3) of the unused phase comparators be left open-circuit, otherwise a 100 kHz spurious frequency will be generated.

After the SC high-pass follows a low-pass stage with a multiple feedback coupling, I 4. This is a restoration filter with the arbitrarily set limit frequency of 14 kHz (approx.). This frequency being the standard frequency used for determining components specifications.

The chip I 5 is the SC-low-pass which is associated with the switching frequency generator I 6.

The latter, also, has a restoration filter following it with a limit frequency of 6.6 kHz.

The filters of I 4 and I 7 work at unity gain and have a Bessel characteristic. Owing to the flat attenuation flanks of Bessel filters, a 3rd-order working was chosen. It takes a little more mathematical effort but the result is an economical stage using only one operational amplifier. Even at the lower tunable limit, over 50 dB of insertion loss is obtainable.

No multi-purpose operational amplifiers were used as a deliberate design policy in order to avoid cross-coupling between the various filter elements . A small AF power amplifier, I 8, serves to make the unit's employment universal.

3.3. Construction

With previous painful experience of constructing filters using unfavourable physical parameters, this one was built on double-sided, printed circuit board (**fig. 8**). The PCB, designated DC 8 NR 008 is drilled and fits into a tin-plate box of dimensions 111 mm x 72 mm x 30 mm. The depth of the box is determined by the installed height of the largest component, this being the 220 μF electrolytic capacitor. The component side of the board must have all drillings which carry non-grounded component leads, countersunk with a 3 mm drill in order to prevent inadvertent earthing to the ground plane. A suggested order for component mounting is as follows: –

1) Insert the IC holders
2) Through-contact the earth connections with wire (or from component leads) at the drillings marked with an x
3) Solder in the components (with certain exceptions)
4) Four filter/parallel resistors and an HF blocking capacitor are soldered onto the PCB track side. They are all marked with an * in fig. 7.

All inputs and outputs are effected via feedthrough capacitors and taken to board mounted chokes. This measure is necessary to prevent any mixing between stray RF fields and harmonics of the switching frequency leading to interference in the receiver.

3.4. Components

R 5609, R 5611:
EG & G Instruments GmbH Reticon, Muenchen, Tel. 089-92 69 20 (Price early 1986 about DM 60.- ea.)
All frequency determining capacitors:
Multi-layer ceramic or Styroflex/Polypropylene, tolerance 5 % or 10 %, RM 5 or 7.5 mm
All coupling and filter capacitors: Tolerance 20 %, RM 5 mm
All feedthrough capacitors:
1 - 2.2 nF soldered-in
All resistors:
1/8 W, 5 %, RM 10 mm (form 0207, E 24 series)
Ch 1...Ch 3:
Single CuL wire wound on 3 mm ferrite bead.

3.5. Commissioning

Before connecting the supplies, it is advisable to check the board for solder-bridges or component mislocations as well as dry or unsoldered joints. Any mistakes could prove costly. The ICs should be inserted in one-at-a-time but first checking the potentials at the sockets of the empty holders. The ICs should be inserted and checked in the following order: 1, 3, 2, 4, 6, 5, 7.

3.6. Test Results

The point-by-point characteristic of the filter is given in **fig. 9**. The narrow curve represents the smallest bandwidth which may be obtained by shifting the low and the high flanks together to their furthest extremity. The results speak for themselves and fully endorse the manufacturers specification. The measured dynamic range was limited, in the author's case, by the low supply potentials employed, which, in turn, necessitated a reduction in the signal-generator input amplitude. It was, however, 65 dB, the lower limit being determined by circuit noise.

Signal input voltage:	Min. 0.9 V, max. 3.8 V (peak – peak)
Output voltage:	Max. 5.5 V (peak – peak)
Gain of unit:	Min. 3.2 dB, max. 15.7 dB

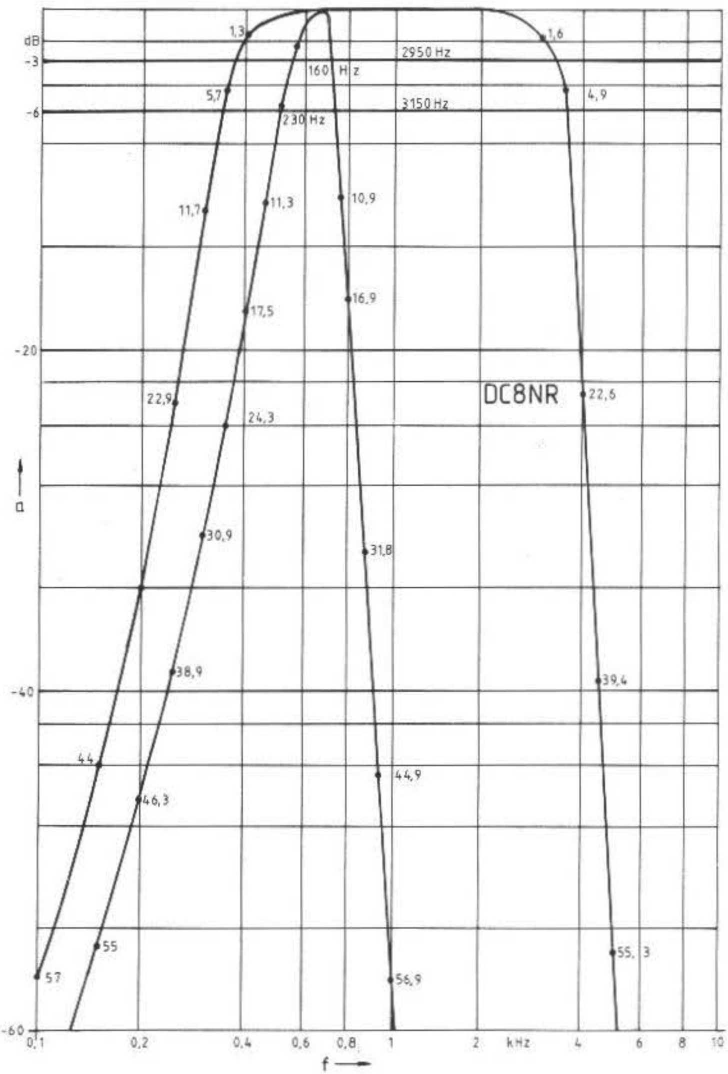


Fig. 9: Attenuation versus frequency of SC-bandpass filter

Supply consumption: 30 mA approx. at 13.8 VDC (without power amp.)

Power amplifier output: 3 W in 4 Ω or 5.5 W in 2 Ω (owing to inadequate heat-sink, do not work at full output continuously)

Unfortunately, there was no suitable equipment available with which the impulse characteristics could be tested but high-speed telegraphy at 100 Hz bandwidth produced no discernable deterioration to received quality. It is recommended that this filter unit be employed as far back towards the receiver second-mixer as possible. The unit

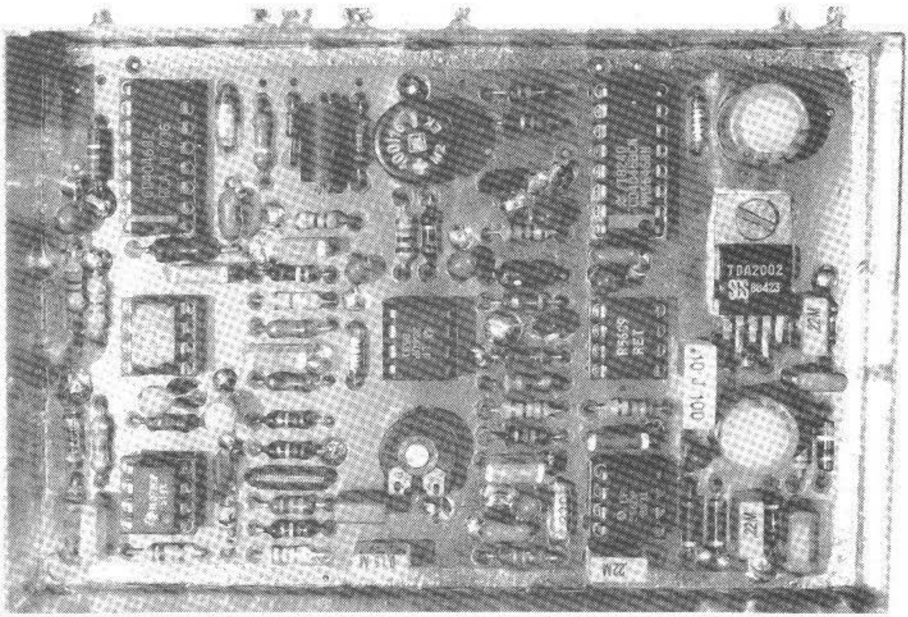


Fig. 10: Prototype housed in its tin-plate box

should never be employed at the receiver loud-speaker terminals owing to noise deterioration.

3.7. Observations and Suggestions

When testing, it was accidentally discovered that the TDA 2002 distorted the signal with approx. − 30 dB of rectified mains frequency caused by a form of modulation hum. Other power amplifier IC's exhibited the same deterioration, e.g. TDA 1037 and TDA 2030, whilst others, TBA 820 and TBA 810 S did not. This raises the possibility that the affected ICs have an internal circuit quirk which accentuates this type of problem and that it cannot be eliminated by simple filtering measures. Therefore, if a common supply is used for the power amplifier, additional filters should be used in its supply leg.

It was a little surprising to learn that the SC high-pass flank slope only had a manufacturer's specification of 30 dB per octave. Theoretically, a Tschebyscheff 5th-order filter (w = 0.5) of this type should have a slope of 42 dB per octave. Is that an indication that no pure high-pass filter ICs can be expected on the market?

Reticon offers the R 5609 in Butterworth form (type R 5613) for the requirements that a constant group-delay over the band-pass is more important than a steep-sided flank. Also, of particular interest, is a 4th-order notch filter (R 5612) with a notch depth of over 50 dB, and which also has the same data as other ICs in the series. It could be inserted between the band-pass filter and the output power amplifier. The audio frequency output stages of a receiver could hardly be improved with such a system. It will not, however, make up for a receiver's deficiencies in the IF and RF departments.

References (9) and (10) are brand new and serve to indicate the extent to which these switched-capacitor filters are being used nowadays in industry.

4. REFERENCES

(1) Arnoldt, M: Digitale Schalterfilter
Franzis-Verlag München, 1984

(2) Hempel, M.: Durchstimmbares NF-Filter für CW-Empfang
cq-DL Heft 8/1974, S. 478 - 480

(3) Moliere, Th.: Aktive und passive NF-Filter – Erfahrungsbericht
cq-DL Heft 12/1977, S. 467 - 470

(4) Tietze/Schenk: Halbleiter-Schaltungstechnik
Springer-Verlag, Berlin

(5) Lancaster, D.: Das Aktiv-Filter-Kochbuch
IWT-Verlag, Vaterstetten, 1982

(6) Ohne Verfasser: Aktivfilter-IC's mit geschalteten Kapazitäten.
Elrad 8/1983, S. 57 - 83

(7) Reticon: Data sheet R 5609, R 5611, R 5612
Sunnyvale, CA/USA

(8) Reticon: General handling and operating considerations for MOS-integrated circuit devices and sub-asasemblies.
Applic. note 177
8000 München 80, Hohenlindener Str. 12

(9) Steer, C.: Narrow-minded Filtering
73 Amateur Radio, Dec. 1986, P. 48 - 49

(10) Reberga, Taruttis: Empfangsfilter für Telegrafiesignale
Elektronik Informationen 12/1986, S. 54 - 56

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 2 / 1987 of VHF COMMUNICATIONS

Frequency Counter DL0HV 008 - 015			Art.Nr.	**Ed. 4/1986 and 2/1987**
PCB	DL0HV 008	double-sided, drilled, silvered	6186	DM 34.-
PCB	DL0HV 009	double-sided, drilled, silvered	6187	DM 35.-
PCB	DL0HV 010	double-sided, drilled, silvered	6188	DM 34.-
PCB	DL0HV 011	double-sided, drilled, silvered	6189	DM 34.-
PCB	DL0HV 012	single-sided with layout plan undrilled	6190	DM 36.-
PCB	DL0HV 013	double-sided, drilled, silvered	6191	DM 34.-
PCB	DL0HV 013a	double-sided, drilled, silvered	6192	DM 32.-
PCB	DL0HV 014	double-sided, drilled, silvered	6193	DM 34.-
PCB	DL0HV 015	double-sided, drilled, silvered	6194	DM 34.-
DK1OF	**HF Receiver Front-End (10 kHz - 30 MHz)**		Art.Nr.	**Ed. 1 + 2/ 1987**
PCB	DK1OF 051	(Mixer) through-contacted	6296	DM 38.-
Components	DK1OF 051	2 ICs, 1 HL ringmixer, 5 transistors, 13 diodes, 2 relays, 1 heat-sink, 1 foil trimmer, 8 chokes, 3 ring cores, 2 m silv. wire, 47 resistors, 45 ceram., 4 tant. and 5 feed-thro. caps., 4 teflon feed-thro. insulators	6297	DM 375.-
Kit	**DK1OF 051**	**complete with all above parts**	**6298**	**DM 395.-**
PCB	DK1OF 052	(Prescaler filter) through-contacted	6299	DM 47.-
Components	DK1OF 052	10 relays, 24 Vogt coil-kits, 5 m CuLS, 5 m silv. wire, 15 resistors, 47 cer., 5 feed-thro. caps., 2 teflon feed-thro.	6300	DM 330.-
Kit	**DK1OF 052**	**complete with all above parts**	**6301**	**DM 360.-**
PCB	DK1OF 053	(Freq. divider) drilled with comp. plan	6302	DM 27.-
Components	DK1OF 053	5 ICs, 4 transistors, 6 diodes, 4 RCFs, 26 resistors, 21 cer., 4 feed-thro. caps. and 1 elko capacitor, 3 teflon feed-thro.	6303	DM 98.-
Kit	**DK1OF 053**	**complete with all above parts**	**6304**	**DM 120.-**
DC8NR	**Steep-Flanks, Switched-Capacity Filter**		Art.Nr.	**Ed. 2/1987**
PCB	DC8NR 008	double-sided, drilled	6219	DM 44.-
Components	DC8NR 008	upon request		

VHF COMMUNICATIONS – Selected Articles on a Common Topic

1. Antennas: Fundamentals
2. Antennas for 2 m and 70 cm
3. Antennas for 23 cm and 13 cm
4. Microwave Antennas
5. Amateur Television (ATV)
6. Crystal Oscillators: XOs and VXOs
7. VFOs
8. Synthesizers
9. RF and AF filters
10. Frequency Counters and Dividers
11. Noise-Figure and Noise-Spectrum Measurements
12. Simple Test Equipment
13. HF Power Measurements
14. Shortwave and IF Circuits
15. Mini Radio Direction Finder for 2 m and 70 cm
16. Converters and Pre-amps for 2 m and 70 cm
17. Converters and Pre-amps, for 23 cm and 13 cm
18. Transverters and PAs for 2 m
19. Transverters and PAs for 70 cm
20. Transverters and PAs for 23 cm and 13 cm
21. Circuits for 9 cm and 6 cm
22. 10 GHz Technology Part 1
23. 10 GHz Technology Part 2
24. FM Equipment for 3 cm and 1.5 cm

Single-theme collection, including binder and postage, **only DM 29,50**

THEME: Antenna Technology – Fundamentals

On the subject of antenna technology and propagation, there are 11 selected articles from VHF Communications at the favourable price of

only DM 29,50 (including postage)

There are approx. 100 pages of basic articles covering circular polarisation, mobile antennas, optimal dimensioning of Yagi antennas, rotors and their control, as well as the optimal stacking of directional arrays.

Also, this collection contains a further 500 pages of interesting publications carefully selected from VHF Communications.

THEME: Antennas for 2 m and 70 cm

On the subject of antenna technology for the 2 m and 70 cm bands, there are 10 selected articles from VHF Communications at the favourable price of

only DM 29,50 (including postage)

There are approx. 75 pages of articles on the dimensioning and optimization of Yagi antennas for both linear and circular polarization. On the simple omni antenna for 2 m and 70 cm, the extra long Yagi antennas and helical antennas, to the high-performance array of YUØB – the whole of the subject matter for these bands is covered.

UKW**berichte** Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133 47-0. For Representatives see cover page 2

THEME: Amateur Television (ATV)

On the theme of amateur television, there are 9 selected articles from VHF Communications in a blue binder at the very favourable price of

DM 29,50 (including postage)

There are approx. 90 pages of detailed constructional descriptions of all the modules necessary for the construction of a 70 cm band, AM-ATV transmitter and a colour test-image generator together with worth-while information on the subject matter.

These are only three examples from a total of 24 theme collections listed in the above table.
Every collection comprises 9 to 11 VHF Communications articles in a blue binder. As well as the subject articles, every collection contains almost 500 pages of interesting publications carefully selected from VHF Communications.

Reduced Prices for VHF COMMUNICATIONS !!!

(from 1.3.1987)	Volume	Individual copy
Subscription 1987	**DM 24.00**	**each DM 7.00**
VHF COMMUNICATIONS 1986	DM 24.00	each DM 7.00
VHF COMMUNICATIONS 1985	DM 20.00	each DM 6.00
VHF COMMUNICATIONS 1980 to 1984	DM 16.00	each DM 4.50
VHF COMMUNICATIONS 1975 to 1979	DM 12.00	each DM 3.50
Individual copies out of elder, incomplete volumes, as long as stock lasts:		
1/1970, 2/1970, 1/1972, 2/1972, 4/1972,		each DM 2.00
2/1973, 4/1973, 1/1974, 3/1974		each DM 2.00
Plastic binder for 3 volumes		DM 9.00

All prices including surface mail.

When ordering 3 complete volumes, a free binder is included!

As we cannot guarantee the reprint of back issues which are older than 6 years, this would be a good opportunaty to complete your collection.

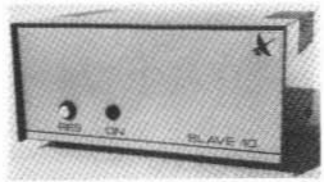
UKW**berichte** Terry D. Bittan · Jahnstr. 14 · Postfach 80 · D-8523 Baiersdorf
Tel. West Germany 9133 47-0. For Representatives see cover page 2

New · New · New · Now available ex stock

Interface „slave 10“

for the satellite rotator systems
KR 5400 and KR 5600

Stock-Nr. 1001 DM 590.—
(incl. connection cable)



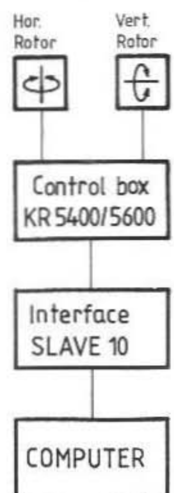
- **fully-automatic antenna tracking system for satellite communications**
- **connection to any computer possible via RS 232**
- **resolution of the dual-channel A / D converter amounts to 10 bits**
- **OSCAR 10 software for the C 64 available**
- **connection to existing rotator systems possible**

Table of commands:

command		response		function
R	CR	R	CR	rotation clockwise
L	CR	L	CR	rotation counter clockwise
U	CR	U	CR	rotation up
D	CR	D	CR	rotation down
S	CR	S	CR	all rotators stop
V	CR	V	CR	rotator stop vert.
H	CR	H	CR	rotator stop horiz.
G xxxxyyyy	CR	G	CR	preset position
F	CR	F xxxxyyyy	CR	interrogation position

xxxx: Vertical position (4 digits)
yyyy: Horizontal position (4 digits)
CR: CARRIAGE RETURN

System's block diagram



Technical data:

Data exchange:	3-wire asynchron. full duplex input and output negative or positive
Data format:	1 start bit 8 data bits 2 stop bits
Baud rate:	1200 B / s
Power supply:	14 V unstab. via control box KR 5400 or KR 5600
Dimensions:	w x h x d = 160 x 80 x 130 mm

Special accessories:

Software on diskette for C 64	Art. nr. 1100	DM	48.—

Satellite rotator systems:

KR 5400	Art. nr. 1013	DM	809.—
KR 5600	Art. nr. 1014	DM	1070.—