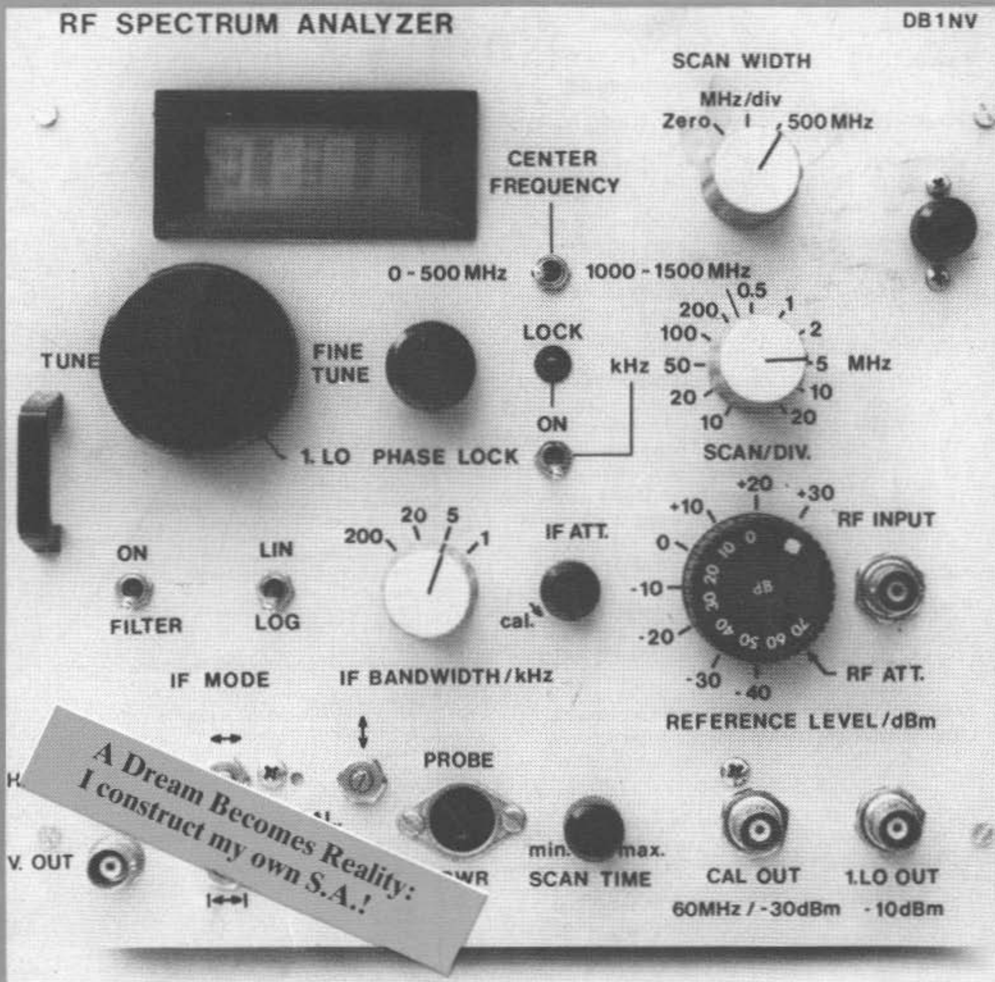
J 20 419 F

A Publication
for the Radio-Amateur

Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 19 · Autumn · 3/1987 · DM 7.00



VHF communications

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

Volume No. 19 · Autumn · Edition 3/1987

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 Michel, 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

Peter Lytz, OH 2 AVP, Gesterbystingen 14 E 49
SF- 02410 Kyrkslätt, Tel.358/0/2 98 17 61

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

Israel
Doron Jacobi 4Z4RG, P.O. Box 6382
HAIFA, Israel 31063

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 Cortes 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
Timekit
P. O. Box 22277,
Cleveland, Ohio 44122
Phone: (216) 464-3820

ISSN 0177-7505

Contents

UKWberichte 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 2: Signal Processing for a SSB/DSB/AM-Transceiver without using Crystal Filters

A practical realization of phasing SSB processing will now be described, the design principles of which were expounded in VHF COMMUNICATIONS 2/1987. The experimental prototype presented here contains the phase-shifting networks and mixer stages for both sender and receiver. The single-side-band signal is capable of being switched between upper and lower sideband and the term double-side-band refers to two sidebands carrying identical modulation but having a suppressed carrier. The AM signal is, of course, the DSB signal complete with full carrier.

Switching between send and receive as well as between the three modes is accomplished at audio frequency by means of CMOS switches. Switching at HF is not required in any form.

The complete circuit has been divided into modules each of which is capable of being withdrawn from the complete system, modified and tested in isolation. Should certain facilities (e.g. USB or Rx) not be required, the appropriate circuit portion can be left out and bridged in the wiring plan.

5. CIRCUIT CONCEPT

The circuit concept of the equipment to be described is embodied in the block diagram of **fig. 7**. It comprises the following features: –

a) The audio signal processing: pre-amplifier, dynamic limiter and audio frequency bounding.
b) The Hilbert transformer which performs the 90° phase-shift for the audio signal. It consists of two all-pass chains which are capable of being upgraded to a 4th-order filter. These units are used for both send and receive functions.
c) The 90° phase-shift network for the HF carrier
d) Two push-pull mixers

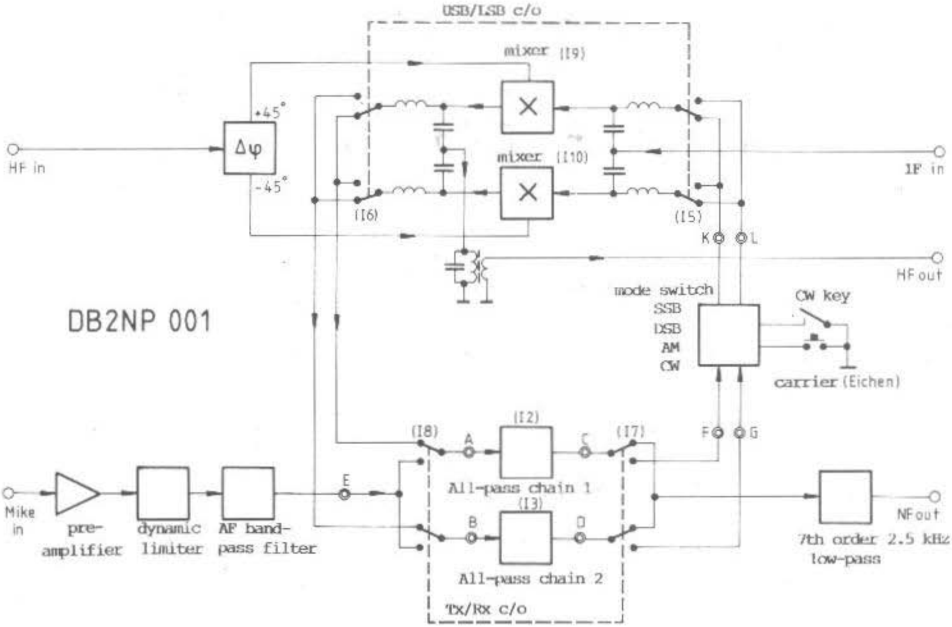


Fig. 7: Block diagram of phase-method of processing SSB/DSB/AM signals showing signal paths. Points A...L correspond to the same points in fig. 2. Switched to receive (LSB).

e) A 2.5 kHz low-pass 7th-order filter to determine the adjacent channel selectivity in the receive mode.

f) All Tx/Rx and mode switching together with the necessary AF switches and frequency diplexers.

Only the circuit necessary for the processing of the SSB signal for the transceiver has been considered, all other components of the transceiver necessary for a complete installation such as up/down converters for the HF band, AF amplification and RF power amplification, have been omitted. For one thing there is an abundance of circuit literature on these topics and also the unit described here has universal applications.

The signal paths are easily followed on the block diagram. The similarity between the **figs. 1 and 2** of the theoretical part of this article and the transmit and receive conditions respectively in **fig. 7** can be clearly seen. A general functional description will therefore not be given again here.

6. CIRCUIT DETAILS

The circuit technical realization of the block diagram of **fig. 7** is detailed in the schematic of **fig. 8**. The semi-conductors were carefully chosen so that they could be easily obtained and are also quite inexpensive. This had to be, as otherwise the whole point of the inexpensive design resulting from the omission of the crystal filter would be nullified by the cost of the other components. The circuit high-lights, outlined in 5 a to 5 f, will now be explained in more detail:

a)

The processing of the audio signal is effected by three of four operational amplifiers encapsulated in I 1. The first op.-amp. forms a pre-amplifier, the gain of which is controlled between 1 and 22 times, by means of P5. The CR combination at the

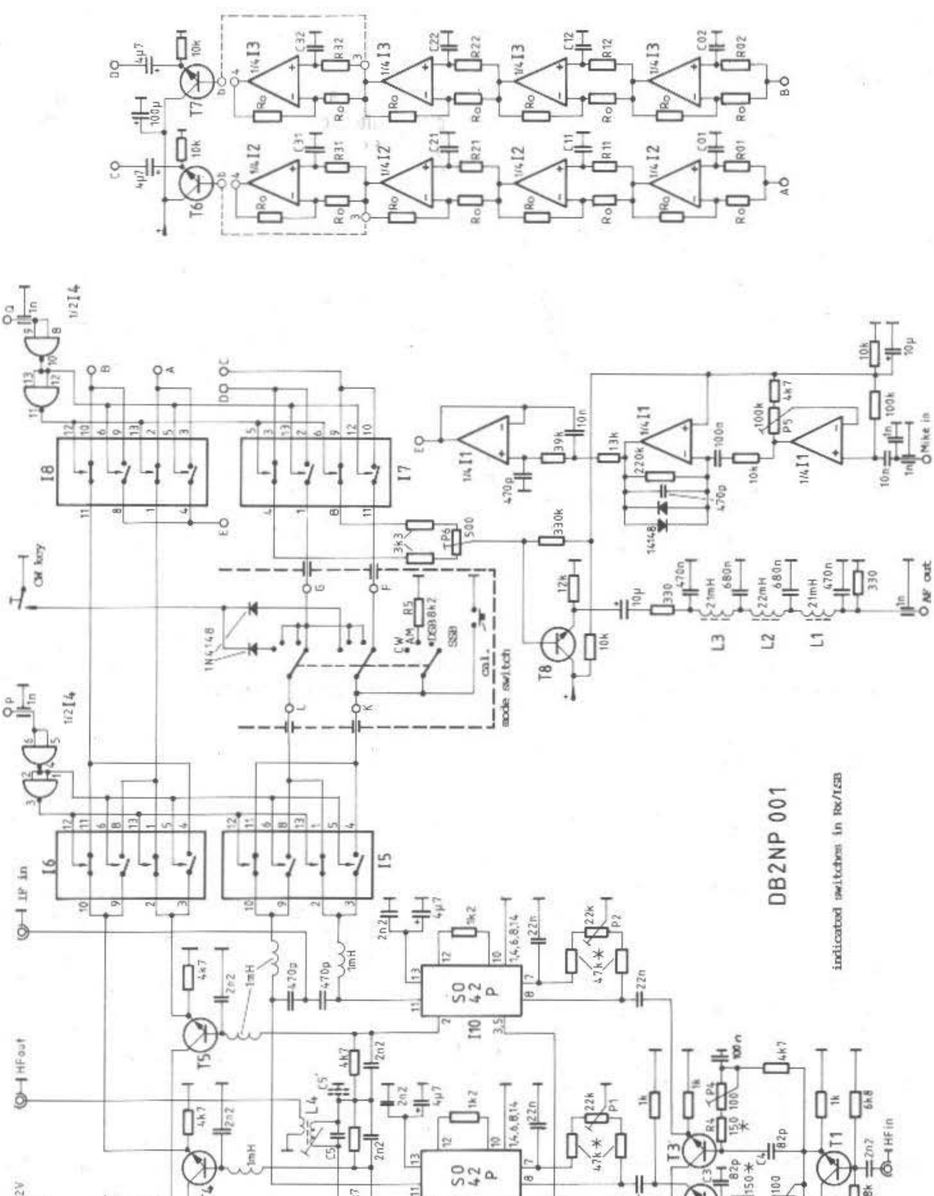


Fig. 8: Circuit schematic of the SSB/DSB/AM processor. Points A...E are taken from the PCB for further connections. Points k...n are connected on the PCB but are shown unconnected to conserve clarity. Circuit shown switched to receive (LSB). Supply: 9 V (14 mA) to 12 V (19 mA).

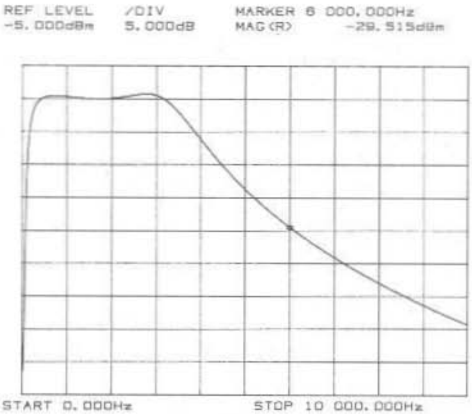


Fig. 9: Measured small-signal frequency response of the speech processor pre-amplifier, dynamic range limiter and band-pass filter.

input attenuates the sub-audio frequencies. The second op-amp. amplifies by a factor of 22 and also acts as a dynamic compressor using the two anti-connected feed-back diodes. The third op-amp. is connected in a third-order, low-pass filter circuit which suppresses signal components above about 3 kHz. The small signal frequency response of this circuit is shown in **fig. 9**. It is important that the low-pass filtering takes place after dynamic compressing because, in spite of the relatively soft compression afforded by the diodes, harmonics would be generated.

b)

The all-pass filter chains of the Hilbert transformer are built around two quadruple operational amplifiers TBA 324 (I2 and I3) and together formed up to a fourth-order filter circuit. Through the use of an active all-pass filter, capacitors and resistors of awkward values can be avoided and the normal series values may be selected. The following, however, should be noted:

All resistors $R\nu\mu$ ($\nu = 0, 1, 2, 3; \mu = 1,2$) should be metal film and 1 % tolerance. All capacitors $C\nu\mu$ should be polystyrene or similar specification in order that the temperature drift characteristics are low. The resistors R0 can, in principle, be any value from about 10 kΩ to 470 kΩ but the important thing is that two resistors of exactly the same value are used on each individual amplifier.

N	ν	$\tau_{\nu 1}$ / µs	$\tau_{\nu 2}$ / µs	φ '	Rv1 / kΩ	Cv1 / nF	Rv2 / kΩ	Cv2 / nF	φ "
1	0	51,704	504,02	± 18,96 °	51	1	51	10	− 18 ° / + 20 °
2	0	22,924	1136,8	± 1,598 °	10	2,2	110	10	− 1,7 ° / + 3,6 °
	1	277,56	93,890		100	2,7	20	4,7	
3	0	14,938	1744,6	± 0,133 °	15	1	180	10	− 0,8 ° / + 1,1 °
	1	112,78	231,07		24	4,7	130	1,8	
	2	504,02	51,704		130	3,9	51	1	
4	0	11,114	2344,8	± 0,011 °	For N = 4 the E 12/ E 24 value steps are too great				
	1	70,524	369,52						
	2	211,11	123,44						
	3	720,43	36,173						

Table 1: The time constants $\tau_{\nu\mu}$ as a function of the number N of equipped all-pass filter-chain sections over a frequency range of 270 Hz to 3600 Hz. φ' represents the maximum phase departure from the idea over this freq.range. φ'' is the maximum phase departure to be expected if $\tau_{\nu\mu}$ is determined by $R\nu\mu$ chosen from the E 24 series and $C\nu\mu$ from the E 12 series of standard components (see text).

The values for $Rv\mu$ and $Cv\mu$ are dependent upon the number of all-pass stages which it has been decided to equip. The point to observe here is that each chain must be equipped with the same number of stages. The **table 1** gives a summary. It shows the calculated values of the time-constants $\tau v\mu = Rv\mu\ Cv\mu$ for the frequency range 270 to 3600 Hz as a function of the number of equipped stages. The table also shows the maximum phase deviation φ' from 90° reference occuring in that frequency band. It can be seen that as the number of stages N is increased φ' is clearly reduced. As the sideband suppression is reduced as φ' increases, it is advisable to use at least two stages i.e. $N = 2$. On the other hand, the choice of $N = 4$ is only meaningful when it is possible to check and adjust with high-precision test gear. This is because the tolerance in components becomes more critical as the phase difference becomes closer to reference. A compromise choice would therefore be a figure of $N = 3$.

The **figs. 10 and 11** show the comparison between the calculated and the attained phase differences at the outputs of both all-pass chains. Already it can be seen that the phase error, mainly brought about by the component tolerances, is greater than that theoretically predicted. This causes the skewed phase response but at ± 0.5° it is sufficiently small.

Table 1 also gives the component values for the time constant $\tau_{v\mu}$. Using capacitors of the E 12 series and E 24 resistors, the maximum phase deviation φ'' is given. If, however, an exact $\tau_{v\mu}$ is required, the $Rv\mu$ and the requisite $Cv\mu$ (digitally measured) can be determined exactly by $Rv\mu = \tau_{v\mu}/Cv\mu$. The exact values may be made up with added series or parallel components. In addition it may be mentioned that the bases of output emitter follower T6 and T7 (point b) is connected to the last of the equipped stages of the all-pass filter. Fig. 8 shows the case $N = 3$ (point b connected to point 3). The components R 31, R 32, C 31, C 32 together with 4 x R0 can therefore be dispensed with.

c)
The carrier phase-shift networks each consist of a high and low-pass first-order filter (R4, P4, C4 and

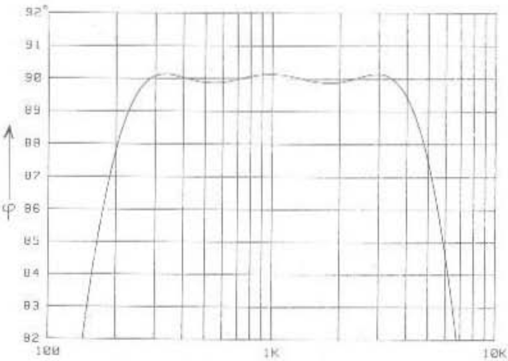


Fig. 10: Calculated all-pass output phase difference taken for N = 3

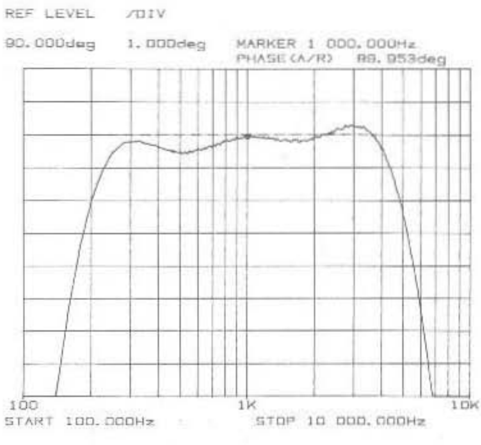


Fig. 11: As fig. 10 but measured values. Standard series components were used: resistors E 24, capacitors E 12.

R3, P3, C3). P3 and P4 are adjusted in order to achieve the relationship $(R3 + P3)C3 = (R4 + P4)C4 = 1/2\pi f_c$ where f_c = carrier frequency. The two networks will then have identical characteristics possessing equal amplitudes and a 90° phase difference. Metal film resistors of 1 % tolerance are required for R3 and R4, P3 and P4 should be cermet multi-turn potentiometers. C3 and C4 should be polystyrene or similar. The values are given for a carrier frequency of 9 MHz but may be appropriately scaled for other frequencies.

d)
Both mixers I9/I10 use the Siemens IC S 042 P which is both inexpensive and easily obtained. Both ICs receive at their inputs (pins 7/8) a 90° phase shifted carrier waveform. In the case of the SSB send condition, the other input (pin 11/13) is used for the 90° phase-shifted audio input to the mixer. The mixer outputs (pin 2) are summed via 1 nF capacitors at the 9 MHz tuned circuit L4/C5 and taken out via a coupling winding. The 1 mH choke, also connected to pin 2, serves to block the passage of RF and does not effect the circuit otherwise.

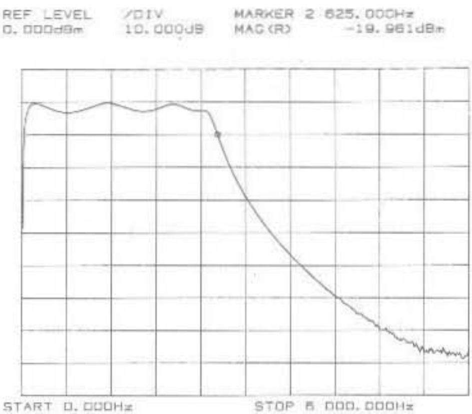


Fig. 12: Measured frequency response of the output AF low-pass 7th-order filter. Frequencies under 50 Hz are also suppressed.

In the receive condition, the IF signals are led to the inputs (pin 11/13) of the two mixers via two 470 pF capacitors. As each input works at both RF and AF, two decoupling capacitors are fitted to pin 13. The low-frequency base-band, present at each mixer (pin 2), is taken via a 1 mH RFC to output emitter-followers T4/T5. It is then taken to the inputs of the all-pass chains for phase-shifting.

The two cermet multi-turn pot'meters P1 and P2 serve to balance the two mixers exactly. The two 47 kΩ resistors connected to these pot'meters are 1 % metal-film, L4/C5 must be suitably adjusted if the carrier frequency departs from 9 MHz.

e)
The required selectivity is achieved by a 7th-order low-pass filter (L1, L2, L3, 2 x 470 nF, 2 x 680 nF) which exhibits a slope of 42 dB/octave (**fig. 12**).

The filter has a characteristic impedance of 330 Ω which should be borne in mind when testing it. The following audio amplifier should therefore not have an input impedance of less than 5 kΩ. This filter is only used for the receive mode where it determines the selectivity of the receiver. It plays no part in the working principle of the SSB mode. For the initial testing, it can readily be replaced by a simple RC audio frequency low-pass filter.

f)
The mode selection and send/receive switching are, with the exception of S1, carried out by the CMOS switches (I5, I6, I7, I8). Their function can be seen from **fig. 7** and described in the first part of this article.

S1, it may be mentioned, is used only in the SSB mode to carry the two 90° phase-shifted audio signals to the mixer. In the DSB and the AM mode, both mixers receive one and the same signal, therefore it makes no difference whether the switch S1 is in the USB or the LSB position. The resistor R5 serves to unbalance the mixer in the AM mode thus allowing the correct carrier amplitude to be modulated.

For the sake of completeness the CW mode has also been provided. The mixer does not receive an audio signal but the carrier is unbalanced with the sender key. The carrier is produced for calibration purposes in a similar fashion.

7. CONSTRUCTIONAL DETAILS

The complete circuit of **fig. 8** was accommodated on a 72 mm x 146 mm sized printed circuit board. It was designed to fit a tin-plate box of (74 x 148 x 30 mm). Only a single-sided layout was developed on the grounds of cost. The layout may be seen in **fig.13**. There are, however, relatively few wire bridges which had to be employed and also two through-contacts to the ground plane: the wipers of P1 and P2 and the ground connection of L4's output coupling loop. The ground connections of the 680 nF capacitors in the AF low-pass filter were accomplished by soldering directly to the tin-plate housing wall.

The PCB component side is not etched and serves as a ground plane. All connections, which are not earthed, passing through the PCB to the components, are prevented from shorting to earth by countersinking the holes with a 4 mm drill on the ground-plane side.

The construction is best effected using the following procedure:
First drill all holes for the feed-through capacitors and the inputs and outputs in the side walls of the housing. It is not necessary to emulate the example in **fig. 14** and put them all to one wall. They can, of course, be located on other walls in order to serve the interest of shorter connecting leads. Care should be taken however, that they do not foul other components by drilling the holes too close to the top surface of the PCB. Four 4 mm

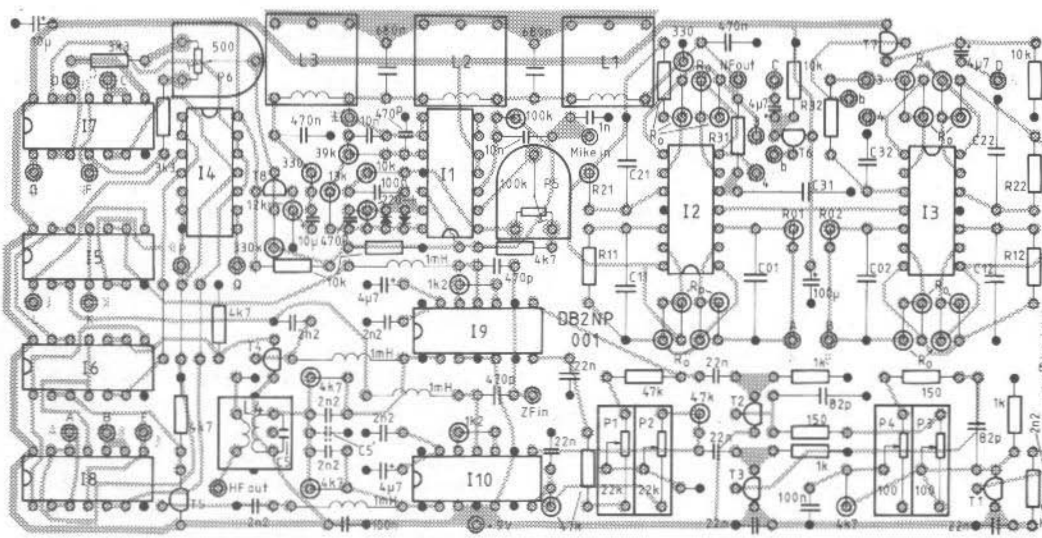


Fig. 13: Component plan of PCB DB 2 NP 001 for schematic of fig. 8.

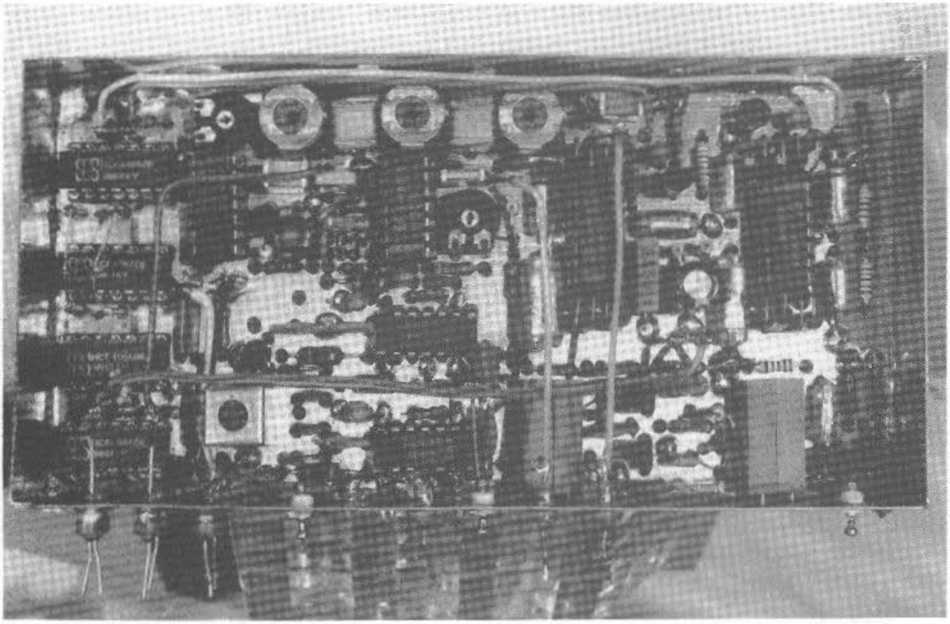


Fig. 14: Prototype of SSB/DSB/AM phasing method processor.

drillings are also required in the PCB for the trim-pot'meters P1 to P4. Their location may be obtained from the component plan of **fig. 13**.

The PCB can then be soldered into the tin-plate box about 5 mm from the lower edge. The edges of the ground-plane side should be completely soldered to the housing wall except in the vicinity of L1 to L3. In this spot, the ground edge on the PCB underside may be solderd to the tin-plate wall.

When equipping the board, ensure that all components which are soldered to the ground plane are inserted first, giving priority to those at the edge of the board. This will facilitate the free access of the soldering iron without damage to other components. All IC holders are then soldered-in but I9 and I10 which are directly soldered to the PCB, have their ground pins 1, 4, 6, 9 and 14 bent so that they do not pas through the board but are soldered directly to the ground plane.

Before the board is equipped, it must be decided which of the possible modes are actually desired and also if full Tx/Rx or only Tx is required. The following simplifications are possible:

a)
SSB-only operation with USB and LSB. No DSB or AM operation: Connect point F with K and G with L direct on the PCB. The external mode switch is not required.

b)
LSB-only or USB-only: IC 5 and IC 6 are not required. Wire bridges are necessary to establish the correct 'switch' connections.

c)
Tx-only: The ICs I7 and I8 are dispensed with and replaced by four wire bridges (instead of I7, pin 3 and pin 4 are bridged, also pin 8 bridged to pin 9.

Instead of I8, pin 1 is connected to pin 2 and pin 10 to pin 11). If, in addition, case b) above is considered, I4 is also superfluous. Also, all components of the output AF filter (P6, T8, L1 to L3 etc.) are not required.

8. ADJUSTMENT AND TEST RESULTS

The following adjustment procedure is based upon the usual assumption that a spectrum analyzer is not available. There should, of course, be on hand the usual run-of-the-mill items such as a low-frequency oscilloscope, an HF receiver with BFO capable of receiving at the relevant frequency or a simple detector receiver with a separate HF test oscillator, a Wien-bridge oscillator etc.

All tuning elements are so arranged that they can be accessed only from above (P5, P6, L4) or from a side wall (P1 to P4). The first task is to select SSB and Tx and balance both mixers. The HF receiver input is connected to the "HF out" by means of a coaxial cable and the "Mike in" is short-circuited. Using the receiver BFO, the residual 9 MHz carrier will be heard as an audible whistle. The tuned circuit L4/C5 is adjusted for maximum volume. If this circuit is made from 10.7 MHz filter components, the capacitor C5 may require a little padding.

Following that, the potentiometers P1 and P2 are iterated until a minimum volume has been achieved whereby the output tone is almost at circuit noise level. Should the tolerance spread of I9/I10 result in these potentiometers being near their end stops, the 47 kΩ series resistor may be adjusted accordingly.

The Hilbert transformer is then checked for correct functioning by injecting a 2 kHz sine ware into "Mike in" at a level of 20 mV_{pp} and setting P5 to its mid position. The outputs C and D of the Hilbert transformers are connected to the X and Y inputs of the oscilloscope. At equal amplitude signals at the scope deflection plates and a 90° phase shift, there should be a perfect circle displayed on the trace. This should also be true for any frequency of the input signal between 270 Hz and 3600 Hz.

When the audio input has a frequency of say 2 kHz, the "HF out" has two spectral components, one at 8998 kHz and another at 9002 kHz. The 9 MHz carrier is also present to some degree. Using the BFO, the weaker of these two signals is selected. If the USB is selected and at a BFO frequency of 8997 kHz, a weak 1 kHz tone and a stronger 5 kHz tone should be present. Now, P3 and P4 are iterated until the 1 kHz tone falls to a minimum into the noise. In the LSB position the conditions are reversed i.e. the 1 kHz tone remains strong and the 5 kHz tone is absent.

The LSB/USB tuning adjustment may also be carried out in the receive condition. In the position USB, a test signal to the "IF in", will deliver a varying (beat) tone at the "AF out" when it is 0.2 to 3 kHz higher than the carrier frequency. If the test oscillator is set to 1 kHz lower than the carrier, a weak, noise-plagued, 1 kHz signal will appear at the output. The sideband suppression can now be optimized in the receive condition with P6. This should result in a clear audio minimum in the middle of the potentiometer's track.

There is no adjustment required for the double-side-band (DSB) mode. The AM mode, however, may require an adjustment to the value of R5 dependent upon the tolerance spread of I9/I10. A broadband oscilloscope can display directly the "HF out" signal. Resistor R5 is so chosen that at a suitable audio level (adjust P5) a 90 % modulation factor is obtained. R5 can also be set by simply listening to one's own audio signal and selecting the best value.

It is to be observed that in "receive" the SSB demodulation is effected even when DSB or AM has been selected. This is because the audio filter has a high attenuation at 0 Hz in order that the carrier is suppressed (no carrier growl on AM). This SSB detection mode, having a re-inserted carrier, has certain advantages over AM using the conventional AM envelope detection.

Following a careful tuning/adjustment procedure, the spectrum diagrams of figs. 15 - 18 where taken. Finally, if, owing to a wiring fault, the other sideband is being transmitted to that which was

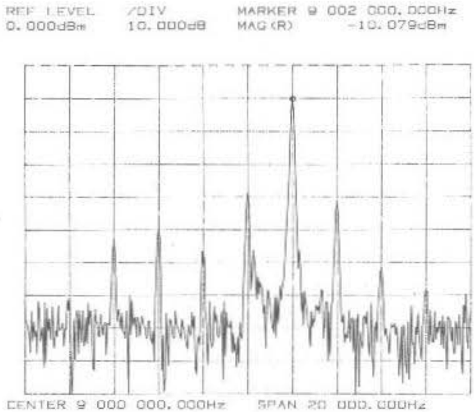


Fig. 15

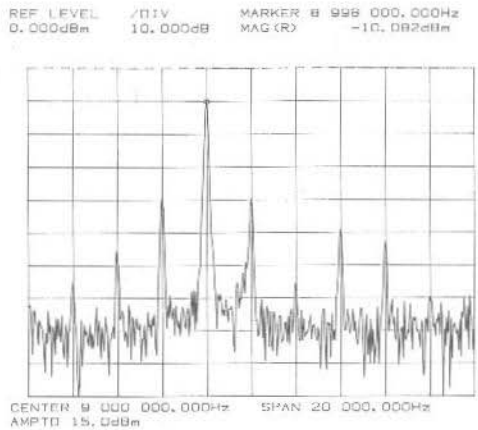


Fig. 16

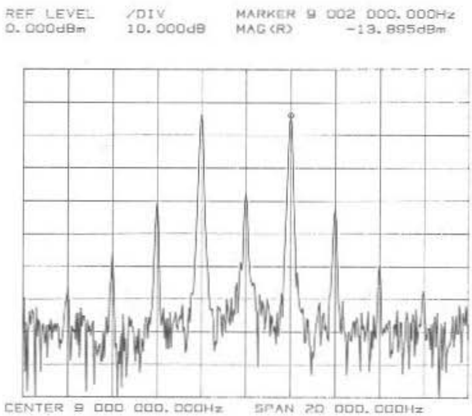


Fig. 17

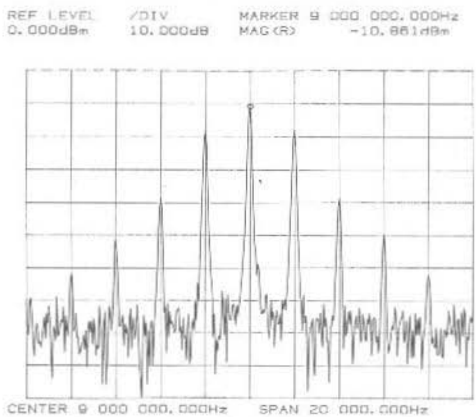


Fig. 18

Fig. 15 – 18: Measured output spectrum taken with 9 MHz carrier and modulated at 2 kHz for USB (fig. 15), LSB (fig.16), DSB (fig. 17) and AM (fig. 18).
The higher-order intermodulation signals are mainly generated in the mixer but they are sufficiently small compared to the peak sideband signal.

received, it is only necessary to change the inputs to the all-pass chains with each other (points A and B).

9. COMPONENTS

T1...T8:	BC 238 B or similar
I1...I3:	324 (various makes)
I4:	4011 (various makes, can be 4001)
I5...I8:	4066 (various makes)
I9, I10:	S 042 P (Siemens)
P1...P4:	Multi-turn pot'meter RM 5/7, 5/2,5
P5, P6:	Trimmer pot'meter RM 10/5 (Piher)
L1...L3:	Pot core 11 mm dia x 7 mm (Siemens) material N 48, A_L value 250 nH/turn^2, CuL wire 0.1 mm dia, L1, L3: 290 turns (21 mH) L2: 297 turns (22 mH)
L4/C5:	10.7 MHz IF filter green (TOKO) or Neosid BV 5138 with C5' approx. 100 pF

All RFC approx. 1 mH.
All resistors marked with * and also R0 and Rvμ of the all-pass network are 1 % metal-film.
C3, C4 together with all Cvμ of the all-pass filter chain capacitors:
polystyrene or similar quality.
All HF coupling and by-pass capacitors:
ceramic RM 5
All AF coupling and low-pass filter capacitors:
MKL types RM 5 or 7.5.
All electrolytics (except 100 μF): tantalum
All feed-through capacitors:
short solderable type 1 nF (approx.)
3 items teflon feed-throughs for "HF in", "HF out" and "IF in".
8 items IC holders, 14 pole.
1 item tin-plate housing 74 x 148 x 30 with two lids

Andreas Schaumburg, DF 7 ZW

Broadband HF Power Amplifiers

A broadband transistor power amplifier is now the norm in every modern HF transceiver. Over a frequency range from 1.5 to 30 MHz, the necessary impedance transformation is carried out by means of a ring-core transformer. This, together with the low-pass filter, the stabilization of the quiescent current and the directional coupler are the subjects of this, the first part, of the article. The second part will deal with the description of a 280 W power amplifier supplied with 13.8 VDC.

The construction of such a power amplifier is not, however, a project for the beginner. The construction of a transistor PA is not to be advised, especially if a 100 MHz oscilloscope is not available.

The author also has access to a supply of cheap HF power transistors in the series SD 1076, MRF 454, SC 2290 so that the experience of destroying many transistors during testing is not prohibitively expensive. As a warning it can also be mentioned that, without protection, a transistor can be instantly destroyed in the event of a mismatch occurring. Push-pull power amplifiers for short-waves are not nearly so robust as those for the UHF amateur bands where they are always narrow-band devices.

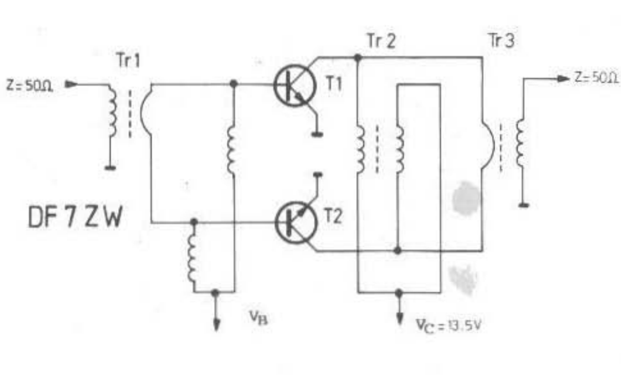


Fig. 1: Push-pull amplifier basic combiner

1. THE POWER AMPLIFIER MODULE

The PA employs AMIDON ferrite ring cores stacked together in a 2 x stackpole of 57-3238 elements. The basic diagram of the PA is shown in **fig. 1** and run-through of the quantitative characteristics is undertaken as follows:

$$R_c = \frac{(V_c - V_{sat})^2}{2\,P_o}$$

where V_{sat} ± 1 volt
V_c = 13.5 volts
P_o = 70 watts

with the above values $R_c = 1.1\ \Omega$

The output is transformed from 2 R_c to 50 Ω

$$n = \sqrt{\frac{Z\,out}{2\,R_c}} = \sqrt{\frac{50\ \Omega}{2.2\ \Omega}} = 4.7$$

For this PA a ratio n of 4 was chosen

$V_s = \sqrt{2\,P \cdot R}$ = peak value

The output power is 150 W at 50 Ω

$V_s = \sqrt{2 \times 150 \times 50}$ = 122,5 V peak

The magnetic flux B (max) can now be calculated:

$$B = \frac{V_s}{2\,\pi\,f\,AN}$$

where V_s = peak output voltage

N = number of turns
A = area of core in m^2 (here 1.1×10^{-4})
B = flux density in V_s/m^2 = Tesla (T)
(1 T = 10^4 Gauss)

Example

$$B = \frac{122.5\ V}{2\,\pi \times 1.5\ MHz \times 1.1 \times 10^{-4}\ m^2 \times 4} = 0.0296\ T$$

The core mix 61, according to AMIDON (5) has a saturation density of 2150 Gauss i.e. 0.215 T. It is now seen that the maximum magnetic field, produced by the PA, is well within the capabilities of the core.

If the PA is used below 3.5 MHz and a higher core loss can be tolerated, a core cross-sectional area of 0.3 cm^2 should be sufficient. By these formulae all ring-cores used in the PA are calculated.

2. THE POWER COMBINER

To bring together two powers of the same order a power combiner is required. This may be constructed as shown in **fig. 2**. For a summed power

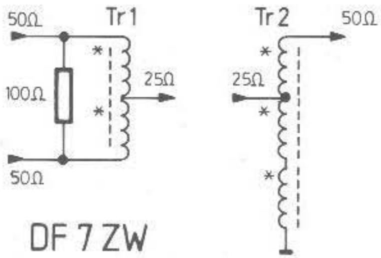


Fig. 2: Power combiner

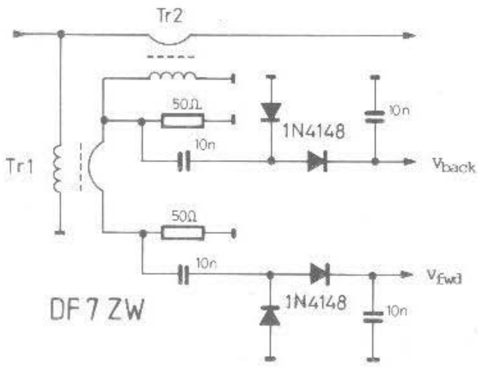


Fig. 3: 3 – 30 MHz directional coupler

of 500 W (max), the following core and winding details should be observed: –

Tr 1: 2 x FT 82/61 mix with 2 x 6 turns copper lacquered wire (CuL) 1.5 mm dia

Tr 2: 2 x FT 114/61 mix with 3 x 5 turns CuL wire 1.5 mm dia

The transformer Tr 1 combines the power of the two separate components of the push-pull amplifier, each with an output impedance of 50 Ω, to a combined output impedance of 25 Ω. The equalizing 100 Ω (**fig. 2**) resistance only absorbes power when the two input signals have unequal phase and equal amplitude wave forms. A wattage rating of 15 W is sufficient for a total output power of 300 W. Both windings of Tr must be carefully bifilar-wound in order to ensure a tight coupling. A loose coupling at the higher frequencies approaching 30 MHz would cause problems.

Of course, the arrangement of **fig. 2** can also be used for the splitting of an HF power where, perhaps, the power is much smaller and therefore smaller cores can be used.

The transformer Tr 2 transforms the impedance from 25 Ω to 50 Ω, or more exactly 56 Ω, as the square of the turns ratio 3 : 2 = 1.5 is 2.25. Again, the three windings are trifilar and tightly wound for good HF performance. The arrangement of fig. 2 for the combining of two equal HF powers into a single output of double the power can, of course, be used to split a single power into two equal components.

3. THE DIRECTIONAL COUPLER

The directional coupler has the important task of reducing the drive by means of the ALC (automatic level control) in the event of a severe mismatch in order to prevent the destruction of the transistors. The directional coupler circuit of **fig. 3** is somewhat more detailed than usual.

For a power of 300 W, the two similar transformers use the following components:

Tr 1, Tr 2
FT 50/61 mix
35 turns 0.4 mm dia
1 turn 1mm dia
lacquered copper wire (CuL)

In the case that the transformer Tr 1 is saturated and starts getting warm, the number of turns should be increased from 35 to 40 – 45. This increases, however, the coupling factor of the directional coupler.

4. THE LOW-PASS FILTER

A low-pass filter is required after the power amplifier in order to suppress the harmonics largely generated by the latter. According to the

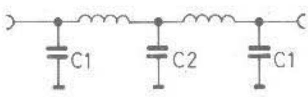


Fig. 4: Low-pass filter for 4 MHz, 8 MHz and 15 MHz (for values see table)

Table 1:	f_{limit} (MHz)	C 1 (pF)	C 2 (pF)	L (μH)	turns on T 94
80 m	4	920	1570	2.725	18 on / 2
40 m	8	460	780	1.36	13 on / 2
20 m	15	240	420	0.73	10 on / 6

band in use, the limit frequency should lie between 4 MHz (80 m) and 30 MHz (10 m). for the 4 octaves 2 – 4, 4 – 8, 8 – 15 and 15 – 30 MHz, four different low-pass filters were designed and built.

The three lower frequency low-pass filters possess 5 elements, as in **fig. 4**. They exhibit a 0.1 dB ripple corresponding to a VSWR of 1 to 1.36.

The calculated suppression attenuations are:

2 f : 35 dB
3 f : 54 dB
4 f : 67 dB

The dimensioning can be carried out according to the standard formulae:

$$C\,1 = \frac{3650\ \text{pF}}{\text{MHz}}\ ; \qquad C\,2 = \frac{6286\ \text{pF}}{\text{MHz}}\ ;$$

$$L = \frac{10{,}9\ \mu\text{H}}{\text{MHz}}$$

Table 1 contains the values for the lower three shortwave amateur bands.

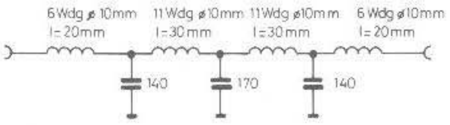


Fig. 5: 7 pole, 30 MHz low-pass filter, all inductors 1.3 mm CuL

The low-pass for the 10 metre band has seven elements and a limit frequency of 30 MHz. The component values are given in **fig. 5**. All inductors consist of lacquered copper wire 1.5 mm dia.

According to ref. (2) the filter components should have the following goodness factors and tolerances:

Low-pass with five elements and 0.1 dB ripple: max 5 % tol. Q = 65.

Low-pass with seven elements and 0.1 dB ripple: max 5 % tol. Q = 120.

According to AMIDON's data book, a far higher Q is to be expected from their ring-cores than 65.

If the capacitors are rounded-off to the standard series values, the completed low-pass should be examined with a sweep generator set-up and the turns on the cores adjusted for the required results.

The seven-pole low-pass with a limit frequency of 30 MHz should be connected directly to (and left connected even when working on other bands) the output of the power amplifier. This ensures that the harmonic radiation does not exceed – 50 dBc whereas the PA alone is only capable of a 20 dB suppression of the predominant (third) harmonic.

5. QUIESCENT CURRENT STABILIZATION

For this, the author resorted to a little more complication than would at first be thought necessary.

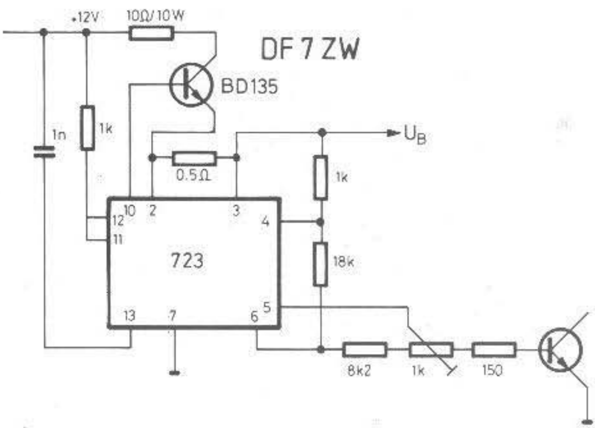


Fig. 6: Quiescent current stabilization

The circuit (**fig. 6**) was taken from ref. (1) with a change of two resistor values in order that the two PA elements could be accommodated.

The pin-out refers to the LM 723 14-pole DIL plastic package. The 1 kΩ potentiometer adjusts the quiescent current of the four transistors to approx. 4 A.

6. PA CONSTRUCTION

The construction of the PA was carried out in conformity with the Motorola application note 762. The only departure was that the ring-cores and

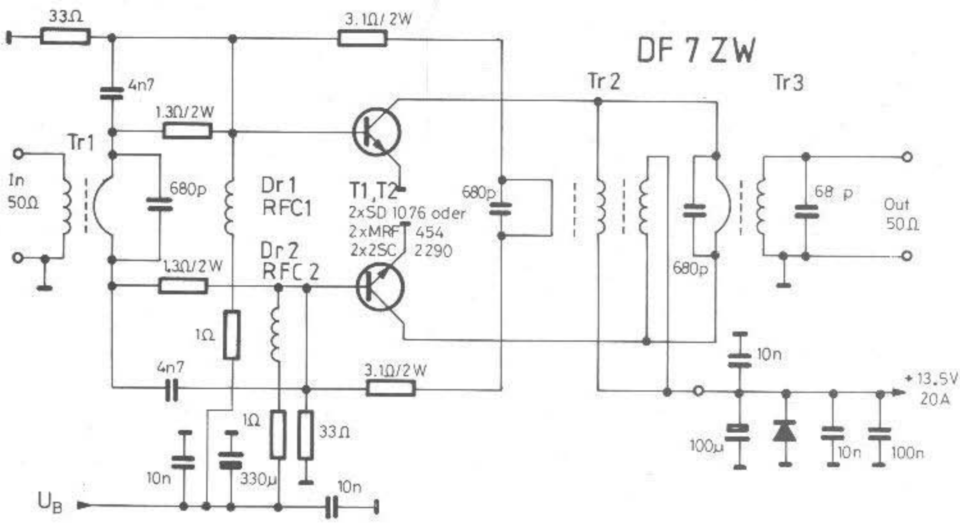


Fig. 7: 150 W broadband PA module

transistors were changed to types which were available in Germany. The circuit is shown in **fig. 7**.

The following data is given for the wound components:

Tr 1
Core 4 x FT 50/61 mix
4 turns – 1 turn (brass tube)
Impedance ratio 16 : 1

Tr 2
Core FT 82/61 mix
2 x 6 turns (bifilar)
1.5 mm dia CuL
feedback wdg.: 1 turn

Tr 3
Core 2 stack-pole 57-3238
or 2 x 8 FT 50/61 mix
4 turns – 1 turn (brass tube)

Choke 1, Ch 2: Valvo 6-hole bead

All capacitors which are not for decoupling purposes must be either mica or chip capacitors as these are capable of handling the heavy currents – disc ceramics are not.

The feedback consists of the 3.1 Ω and 1.3 Ω resistors, the 4.7 nF (or 5.7 nF) capacitors ensure a constancy of gain over the 1.5 – 30 MHz band.

The 50 Ω winding of Tr 3 should be carried out using 1.5 mm dia teflon insulated copper wire.

The capacitors across transformers Tr1 and Tr 3 have only approximate values. They should be experimented with as they will be dependent upon the type of transistors used, the characteristics of the wound components as well as the power output.

The author did not develop a PCB for the prototype but used a epoxy-glass board with a thick coating of copper. Islands were milled out of the copper surface to which the components were soldered. The grounding of the heat-sinks was effected by means of screws through to the ground plane.

The supply potential feed-in point via Tr 2 results in a better harmonic suppression than that provided by the direct connection to the centre point of Tr 3.

Fig. 8 shows the power frequency response taken at 4 W drive V = 13.5 volts and I < 19 A.

The input VSWR is 2.3 (max) below 6 MHz and above this frequency it is better than 2.0.

When one transistor is conducting, the other is biased off. In the blocked phase the peak collector voltage can reach 40 V. The data sheet specifies a collector-emitter breakdown voltage

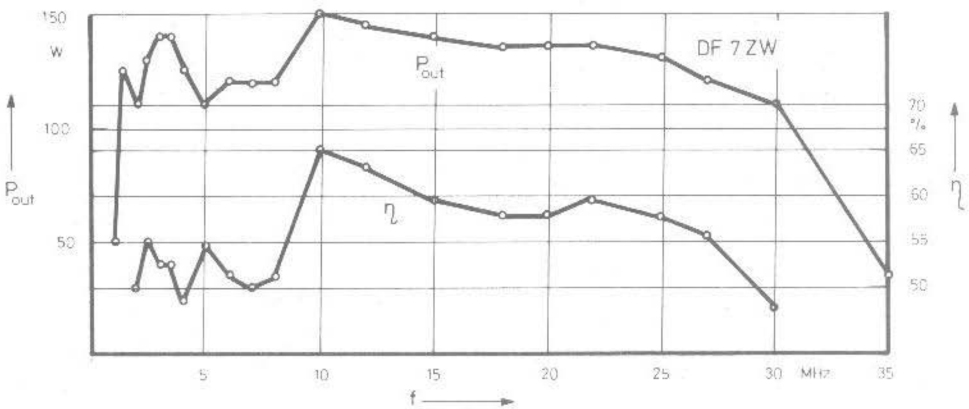


Fig. 8: Power output and efficiency with frequency of PA driven at 4 W and supplied with 13.5 VDC

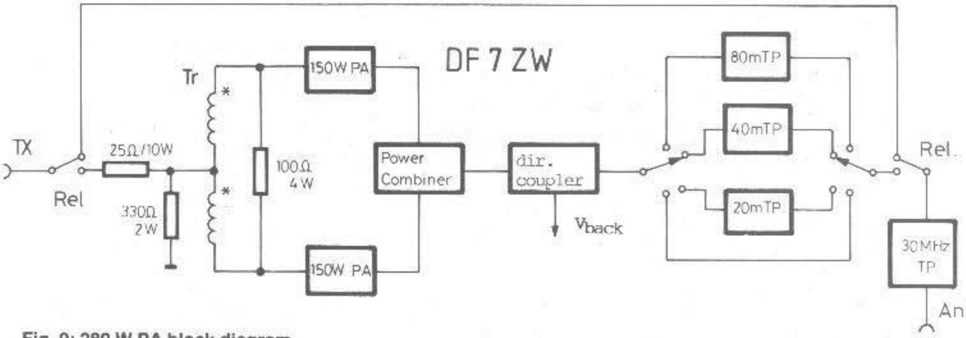


Fig. 9: 280 W PA block diagram

of 36 V which is over-stepping the mark a little. A mismatch, moreover, could send the peak voltage soaring to 70 V, leading to the instant destruction of the transistor. For this reason the employment of a protection circuit should be considered as mandatory. Valved power amplifiers can be operated without any form of protective device whereas shortwave amateur transceivers are comprehensively protected. Maybe the next generation, using VMOS-FETs, will also be as inherently robust as valves were/are. The power amplifier introduced in this article will, anyway, be protected against mismatch damage by means of a directional coupler.

Another possibility is now thrown in for consideration:
The PA can be built with 28 V transistors and driven to only 14 V in order that there is hardly a possibility of them being destroyed. A suitable type is the MRF 422 (28 V/150 W). In this PA circuit, the two transistors will deliver 120 – 130 watts output.

7. COMBINING THE MODULES FOR A 300 W PA

Following the description of the circuits, the construction according to the block diagram of **fig. 9** should present few problems.

The power splitter T comprises:

1 x FT 50/61 mix
2 x 6 turns bifilar
CuL 1 mm dia

The 30 MHz low-pass filter is connected in the antenna line solely for the 15 and 10 metre bands, and remains in circuit for the other bands too. The reasoning behind this is as follows:

Low-pass filters having ferrite ring-core inductors possess the unpleasent characteristic of core saturation, thereby causing to a greater or lesser extent the production of harmonics. The 30 MHz low-pass is, however, constructed with air-cored inductors. This will prevent harmonics from the lower band filters, which have been caused by core saturation, to be attenuated to negligible proportions in the critical region of 100 – 200 MHz. This measure limits the possibility of BCI or TVI to very low proportions.

Each of the PA modules requires a drive power of 4 W. In order that the 300 W amplifier can be driven with the normal 15 W, about 7.5 W must be dissipated as heat. This is done mainly by the 25 Ω 10 W resistor, forming an input attenuator, together with the 330 Ω 2 W resistor. Since the input impedance of the input power divider is also 25 Ω, the total input impedance to the PA is 50 Ω. The resistive attenuator serves to increase the real part of the input impedance and the VSWR is less than 1 : 2 over the entire working range.

Both 150 W modules must, of course, be absolutely identically built. This applies also to the

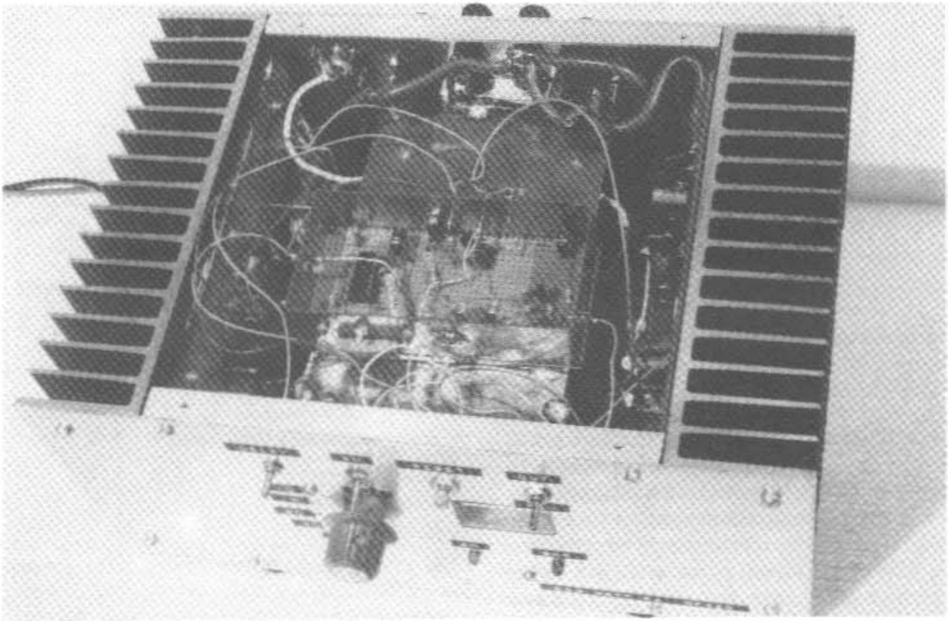


Fig. 10: The author's HF power amplifier

interconnecting cable lengths which must be short and symmetrical. Paired transistors were not available to the author but the output balancing resistor always remained cool. Should any assymmetry develop through a fault, or any other reason, this resistor would start showing signs of distress immediately. The 300 W PA has in fact proved more robust than the individual 150 W modules. The reason could lie in the fact that the power combiners, the low-pass filters and the connecting cables all contribute with their attenuators to the amelioration of the consequences of a mismatch. Nevertheless, a mismatch protective circuit is still necessary.

The input supply current can peak to 40 A at full drive thus necessitating a suitable connecting wire cross-sectional area. Supply inputs to both amplifier modules are fused at 20 amps.

When applying power for the first time, reduce the supply potential to some 9 V, at which voltage the effects of a fault may be withstood. Only when everything seems to be in order and the ALC has been adjusted can the full supply voltage of 13.8 V be applied.

The author is now planning the construction of a PA using 28 V transistors which can deliver a power in the order of 500 – 600 W. Whether or not a construction kit will come out of it is not yet decided. On the one hand such a project is not really suitable for the inexperienced and on the other hand valves are much more robust and simpler. They are also cheaper if they can be obtained at the flea-market.

Finally, the author would like to thank Oscar, DL 4 FA for the favourable terms under which the transistors were obtained. Without this help this project could not have been completed.

The author would also like to establish an exchange of information with any amateur who has built these, or similar, amplifiers.

8. REFERENCES

(1) Motorola Application Note AN 762
Linear Amplifiers for Mobile Operation
Jermyn Distribution, Bad Camberg,
Tel. 06437 - 23 - 0

(2) Red Eric: Arbeitsbuch für den HF-Techniker
Franzis-Verlag 19?? München

(3) TRW-Datenbuch RF Devices
European Ed.

(4) Thomson-CSF: RF & Microwave Power Transistors

(5) Amidon: Iron-powder and Ferrite Coil Forms
Elektronikladen Münster

(6) The ARRL Handbook for the Radio Amateur (1985)
Verlag UKW-BERICHTE

Additional Notes on the 70 cm Handheld DB 1 NV 004

The additional information has beed offered following the construction of several DB 1 NV 004 70 cm handheld transceivers described in edition 2/1986.

1. Synthesizer

Some examples of the MC 145152 synthesizer ICs have a spurious impulse from the modulus control output (pin 9) which causes errors in the counting of prescaler SP 8793. This results in the channel being switched to one lower than that which was selected. A ceramic capacitor of 10 – 50 pF, connected from pin 9 of the MC 145152 to ground, will eliminate this defect.

2. Control Amplifier

The foil capacitors used in this amplifier have specified a tolerance of 10 % or better. Normal ceramic capacitors, used for decoupling purposes, have a 50 % tolerance and are unsuitable for this particular application. The component values given in the circuit diagram represent the best compromise between the fast operation of the synthesizer and the loop stability. Too small a value engenders synthesizer microphony and deteriorates the cross-talk from adjacent channels.

3. BF 979 Buffer Amplifier

This stage tends to self-oscillate when the damping resistor in the BF 979's collector is physically too large. It should be replaced with a resistor of the 0204 series. It is also advantageous to slip a ferrite bead over the BF 979 collector lead, as in the tuner, in order to assist in damping the dual-gate FETs. This appears to be the most critical stage in the unit.

4. Send Amplifier

Some examples of the BFR 96 exhibit too little gain (under 20 dB) so that in spite of adequate drive (> 5 mW) the output power is not sufficient. The only recourse is to change the transistor.

5. IF Amplifier

The mixer crystal 20.945 MHz must be a series resonant mode type as indicated in the component list. Parallel types either oscillate on a parasitic or at 5 kHz too low. The 1st IF and the ceramic filter 2nd IF are then relatively displaced, leading to an unacceptably high distortion factor. Replace this crystal with the one specified.

6. Call-Tone Generator

Only a CMOS 555, i.e. ICM 7555 or TLC 555, should be used here. Normal NE 555s cause

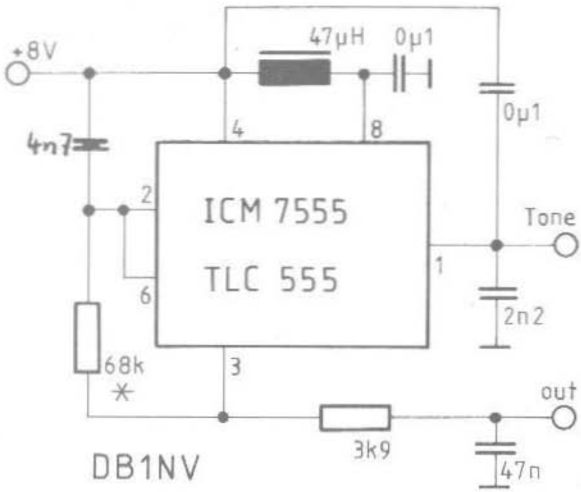


Fig. 1: Modified call-tone generator

needle pulses to contaminate the Vcc line leading to a call-tone frequency spectrum extending over several megahertz. As the adjustment of the call-tone with two resistors is not easy, the modification of **fig. 1** is suggested. The 68 kΩ resistor is so chosen that a call-tone frequency of 1750 +/− 30 Hz is produced. In addition, the supply potential of the TLC 555 is decoupled at pin 8 with an LC filter thus preventing the call-tone overmodulation altogether. If the long unscreened leads to the call switch picks up HF, they should be decoupled with a 2.2 nF capacitor and, if necessary, also an RFC.

7. Housing

The metal work of the housing and its contact to the PCB must be carried out thoroughly otherwise unpleasant scratching noises will be apparent as the unit is moved.

DL 5 NP/DB 1 NV

Jan-Martin Noeding, LA 8 AK

Improve the Oscillator Circuit in your old 2 m Converter like DL 6 HA 001 or Similar Designs

2m converters like the DL 6 HA 001 design (VHF COMMUNICATIONS 1/1970) are still in use in some radio amateur stations. In contrast to the operating techniques some 20 years ago, frequency accuracy is necessary today. The problem, experienced not just by myself, is that it is not possible to pull the crystal frequency of the above mentioned converter to exactly 116.000 MHz. This can be cured by relatively simple changes in the oscillator circuit. Since similar circuits are used in other converters, it seems worthwile to publish the modification.

1. MODIFICATION

My circuit does not require any modification to the PC-board DL 6 HA 001, in fact two capacitors and two resistors less are used than in the original circuit. The oscillator transistor T4 (**fig. 1**) is exchanged from a bipolar type to a JFET (**fig. 2**). The original crystal, delivered by the Verlag UKW-BERICHTE, is used.

Before commencing modifications, the injection level to the mixer stage should be measured, using any kind of (simple) RF probe. After the modification, the injection should be adjusted to the same level again.

Remove transistors T4 and T5 (BFY 37), as well as the capacitors C16, C17, C18, C19 and C20, and the resistors R8, R9, R10 and R12. It is important to use a smaller value for C20, since the original value of 15 pF causes too much load onto the oscillator; sufficient drive is obtained with only 2 pF.

Now connect the new components as shown in **fig. 2**. Solder a wire bridge across the crystal and check the resonant frequency of L8 using a dipmeter. It should be adjustable to 38.666 MHz, otherwise the coil has to be modified.

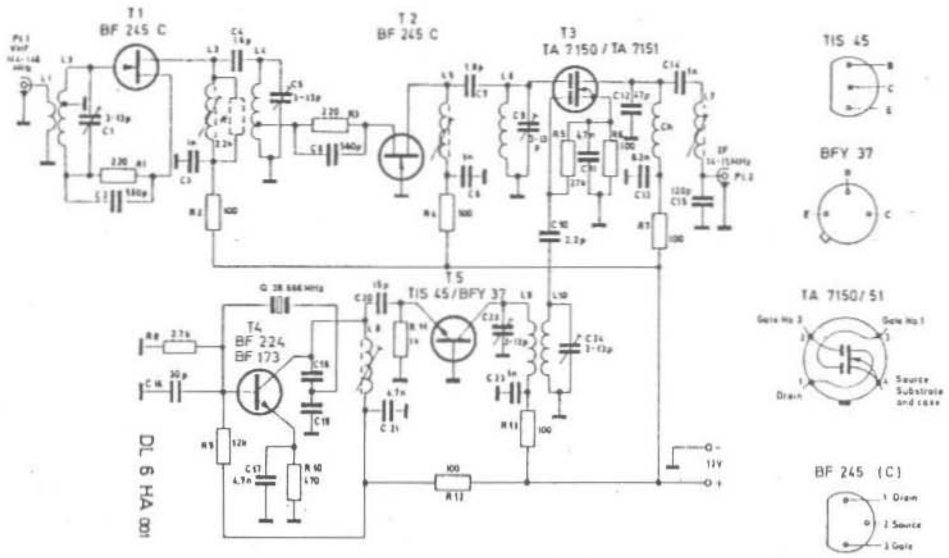


Fig. 1: The original circuit diagram of the DL 6 HA 2 m converter.

2. SAFETY MEASURES

In order to decouple the 12 V supply, 5 to 6 extra capacitors (around 1 nF) should be soldered to the supply line on the solder side. Also a ferrite radio-frequency choke (RFC) should be connected in series with the 12 V supply. These measures will help to avoid interference into the lower UHF TV channels. Additionally the stability is improved.

If a low-noise (high-gain) preamplifier is used at the antenna, it is advisable to remove the first amplifier transistor (T1) of the converter.

This modified converter is used extensively for 2 m dx work in SSB and CW by LA 5 MK and LA 8 AK.

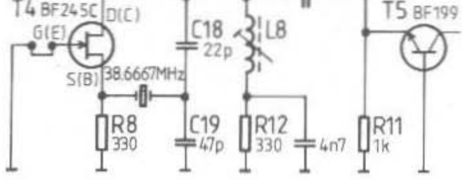


Fig. 2: The modified oscillator circuit with the original connections to the oscillator transistor T4 in parenthesis. A supply voltage variation between 9 and 15 V causes an oscillator drift of no more than +/− 200 Hz at 116 MHz.

Jochen Jirmann, DB 1 NV

A Spectrum Analyser for the Radio Amateur

Spectrum analysers enjoy a special place in the hearts and minds of amateur radio constructors. For one thing, they are able to display immediately the full output spectrum of a transmitter and the relative amplitudes. The other awe provoking thought which springs to mind is their enormous cost – now, almost as much as a small house. The possession of a spectrum analyser must surely be, for most amateur constructors, completely out of the question. Also, a perusal of a commercial analyser's specifications are enough to convince many that the amateur construction of such an instrument is fraught with unsurmountable difficulties. But, of course, one should'nt despair quite so easily because, for example, what radio amateur requires facilities such as a 10 Hz resolution at 20 GHz? In other words, if the facilities offered by a commercial spectrum analyser were pruned down to those required by the radio amateur then it is indeed entirely feasible to undertake the home construction of such an instrument.

The following article is intended to excite the reader to experiment with the project, those who require a "watertight" construction article complete with PCB layout patterns and a guarantee of sure-fire results will be disappointed. It must be appreciated that the detailed description of such a project would occupy all the pages of VHF COMMUNICATIONS for the whole year. The author is, however, prepared to give advice to various project-groups which might be formed.

In the almost continuously occupied frequency bands of today, a smooth succession of stations carrying various services can only be maximized within a given band if these stations observe the minimum demands concerning the radiation of extraneous energy. Even modern high-level receivers employing ring mixers fed with a high oscillator power in the region of 200 mW can, when poorly designed/constructed, cause a lot of trouble. Whilst one can be reasonably assured that a transceiver of a proprietry manufacturer will satisfy at least the minimum requirements of the radio regulatory authority, it will not be so certain that a home-constructed piece of equipment will offer the same freedom from spurious and unauthorized radiation. This can only be ascertained, in most cases, by a visit to the post office stand at a large ham-fest where the item may be subjected to the statutory tests.

If a complete survey of the harmonic and intermodulation content of a transmitter or oscillator is required then the spectrum analyser is the correct instrument to do it. It represents an electronically periodically tunable superheterodyne receiver which is able to display the level of the signal-under-test together with the relative levels of its modulation and spurious signals. A commercial instrument would cover, typically, a range from a few Hz right up to twenty GHz (with supplementary mixers 325 GHz) and cost from 20 to 300 thousand DM. Not many amateurs could afford to buy a new one and those offered for sale at flea-markets are ageing, obselete examples which have a restricted dynamic range. Those not having the luck to find a good second-hand analyser might consider the idea of making one for themselves.

In 1976, DL 8 ZX published one of the first home-brew concepts which covered a frequency range from 0 to 60 MHz and from 120 to 180 MHz. A cable TV tuner was introduced in 1980 which was used as the first down-converter. The present article will consider how a spectrum analyser with usable specifications can be realized with a tenable degree of constructional complexity.

1. CONCEPT

Before a circuit concept may be considered, it is necessary to define the facilities the instrument should offer in order that a sense of proportion is acquired and that no subliminal objectives should be striven for. A few minimal demands will therefore be set down as follows: –

Frequency range: 0 – 500 MHz and eventually 500 – 1500 MHz. This covers all the chief activities up to 70 cm with the basic unit. The additional down-converter covers the 23 cm band and all the important signal processing frequencies for the production of signals in the microwave bands.

Dynamic range: At least 60 dB. A harmonic and intermodulation capability of 60 dB is perfectly sufficient and is, indeed, better than that attained by some lower-priced commercial instruments. Valved power amplifiers, without output filtering, achieve only up to 40 dB harmonic and intermodulation specification and 2 m and 70 cm transistor PAs are not much better.

Resolution (analyser bandwidth): Switchable from 200 – 500 kHz for survey measurements and down to 1 kHz in order to identify third-order intermod. products and neighbouring synthesizer channels.

Stability: Short-term stability must be better than the smallest resolution bandwidth i.e. 1 kHz. Long-term stability: better than ± 3 dB both over the whole frequency and the whole dynamic range.

Sensitivity (10 dB s = N/N): better than – 100 dBm i.e. 12 μV/50 Ω at the smallest bandwidth.

LO spectral purity: Noise sidebands ± 25 kHz from carrier should be lower than – 60 dB in order to preserve the validity of the dynamic range specification.

The circuit should use easily obtainable components and require the minimum of tuning adjustment even if this might mean increasing the circuit complexity somewhat. Otherwise, the circuit might require the services of another spectrum analyser to align it – and that doesn't help anyone. A few modules will now be regarded a little more closely in order to evolve the various realization possibilities and at the same time identify a few potential trouble-spots in order to circumvent them. The most important step in this direction is the determination of the local oscillator and IF frequency plan.

1.1. Frequency Plan

The spectrum analyser is a periodically, tunable receiver with an extremely large frequency range encompassing two or three decades. Signals within this frequency range must be converted into a fixed intermediate frequency in order that the selection and amplitude processing can be carried out. Every superhet. receiver has a principal spurious receive frequency, known as the image-frequency whose effects are normally rendered harmless, in a conventional receiver, by the preselector filter circuits. It goes without saying that this technique is not suitable for this application where the tuning range is completely continuous. The only solution therefore is to

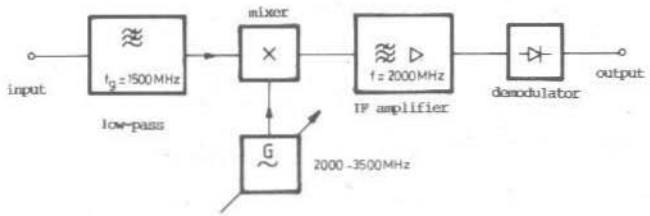


Fig. 1.1: Block diagram of 0 - 1500 MHz spectrum analyser

make the IF, as with many modern general-purpose receivers, lie above the highest receiver input frequency. The image frequency can then be identified and filtered out of the receiver input by means of a simple low-pass filter. The image frequency can also be used as a second input frequency and this is quite normal now in commercial instruments.

By using harmonics of the local oscillator for use in the mixer, further receive ranges can be arranged. It has to be borne in mind, however, that the conversion loss of the mixer is much greater at harmonic LO inputs. This is the usual practice for commercial analysers which use up to the 20th LO harmonic in order to provide coverage in the millimeter wave range.

The most obvious frequency plan for a receive range of 0 to 1500 MHz would be to locate the IF at above 1500 MHz, e.g. 2 GHz, and the oscillator tuning from 2.0 to 3.5 GHz as shown in **fig. 1.1**. The received input range is then fully covered and the image frequency can easily be eliminated with a simple low-pass input filter.

The snag with this solution is that a tuned oscillator range of 2.0 to 3.5 GHz would be required, which with amateur means, is hardly feasible, moreover, a mixer for these frequencies is relatively expensive.

The IF at 2 GHz presents no real problem but it should be borne in mind that a second conversion to a more amenable frequency for bandwidth and signal processing e.g. 10.7 MHz or 21.4 MHz will have to be carried out. Thus the receiver input range can remain unbroken with a high image rejection and the signal processing can be effectively carried out, all by means of the double-superheterodyne technique. If an IF of 500 MHz is chosen, two ranges of input frequency become available, 0 to 500 MHz and 1000 to 1500 MHz with the local oscillator tuning between 500 and 1000 MHz. The receive ranges may be separated with selectable high and low-pass filters. Diode

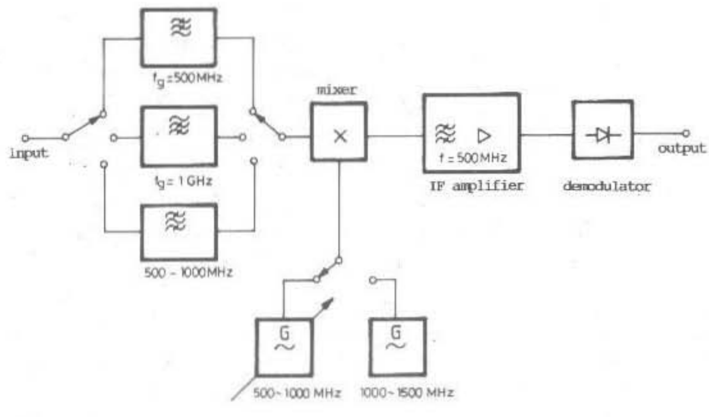


Fig. 1.2: A spectrum analyser tunable in 500 MHz bands

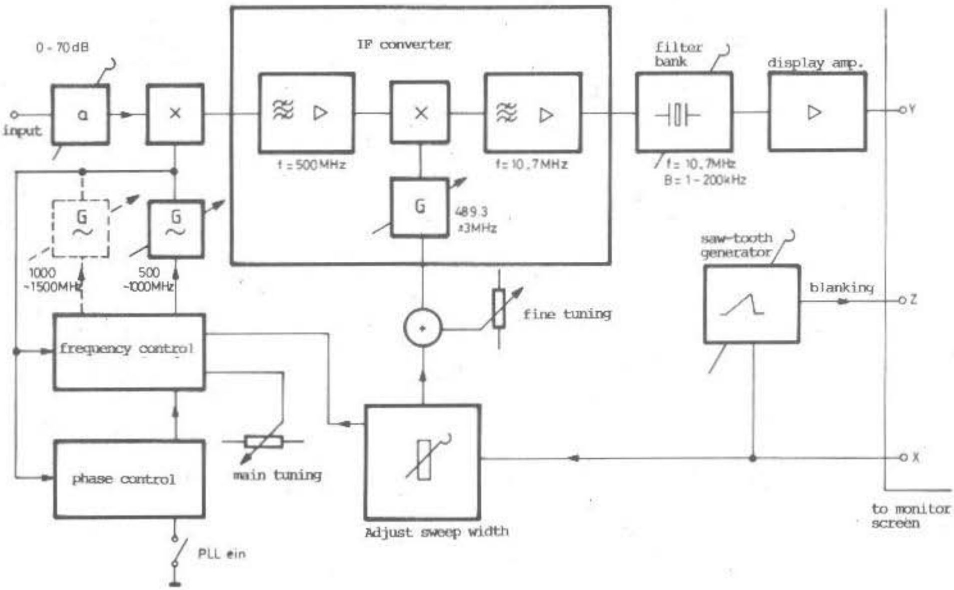


Fig. 1.3: Block diagram of a practical spectrum analyser

tuned VCO's are easily realized at these frequencies and the translation of 500 to 10.7 MHz may be carried out in a single step. The disadvantage of the missing input spectrum range of 500 to 1000 MHz may be overcome by employing an additional 1st LO covering 1000 to 1500 MHz. This further extends the received input range possibility from 1500 to 2000 MHz. The block diagram for this arrangement is shown in **fig. 1.2.**

A third possibility is represented by an IF of 1 GHz but this brings a rather more unfavourable receive range than the above cases as can easily be appreciated.

After settling, in general, that a frequency plan such as that of **fig. 1.2.** is in fact tenable, the block diagram of the spectrum analyser can be fleshed-out a little. This is shown in **fig. 1.3.** which includes the following modules:

- The input mixer: A proprietary mixer such as the SRA-220 may be employed here.
- The voltage-controlled-oscillator, (VCO) controlled by a PLL circuit for adequate frequency stability when using the higher resolutions.
- The IF 2nd mixer down-converting from 500 MHz to 10.7 MHz or 21.4 MHz.
- The main analyser filtering in the 2nd IF.
- The logarithmic display amplifier.
- The control circuits for the tuning and control of the oscillator.

The 2nd mixer, which brings the first IF down to a standard 10.7 MHz or 21.4 MHz, will be considered first.

2. THE 2ND MIXER

Following the plan as laid out in the previous chapter, the 500 MHz 1st IF signals are produced by received inputs being mixed with a first local

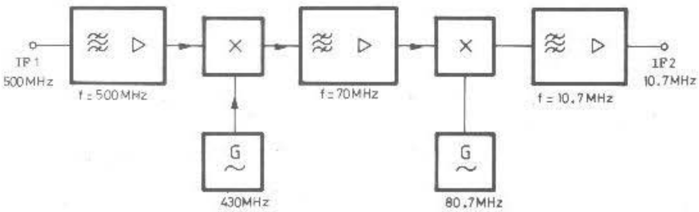


Fig. 2.1: An IF converter with two conversions

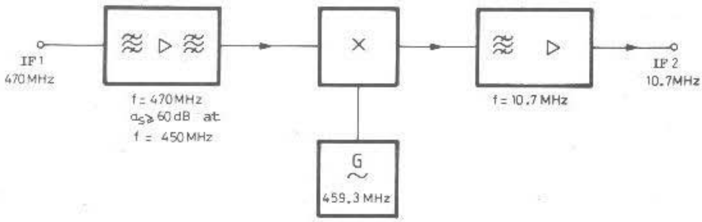


Fig. 2.2: An IF converter with a single conversion

oscillator with a tunable range of 500 MHz to 1000 MHz. The conversion of the signals to 10.7 MHz must be carried out with an image frequency suppresssion of at least 60 dB in order to reduce spurious inputs to the following circuits. In a normal receiver, this would be accomplished using a further IF at about 70 MHz in order to reduce the demands upon the mixer input filter for this level of image suppression. This scheme is shown in **fig. 2.1.**

The disadvantage of this technique is the greatly increased possibility of producing spurious signals from the total of three instrument LOs which would have to be employed. The author has therefore decided upon the direct conversion of 1st IF signals at 500 MHz to an IF of 10.7 MHz and this is outlined in the scheme of **fig. 2.2.** It will be observed that the image frequency is at 21.4 MHz removed from the 500 MHz input frequency and the filter must satisfy the 60 dB image suppression requirement.

This order of suppression can only be achieved with pot resonators or with helix resonators – the latter, on account of its smaller form, is to be preferred. The mini helix filter 10H3 can be obtained from Telequarz, for about DM 20,-. This is tunable from 440 MHz to 500 MHz and at 450 MHz it has a selectivity of some 40 dB. Cascading two of these filters ensures that the image specification of 60 dB suppression will easily be achieved. This entails fixing the first IF not at exactly 500 MHz but at 460 to 470 MHz in order to avoid direct interference from any local channel 21 to 25 television transmitter.

The satellite television receivers use a surface wave filter which has a suitable mid-frequency of 479.5 MHz but the insertion loss at 20 dB is far too high. Also its bandwidth is around 35 MHz with a selectivity of 50 dB. The latter specification would make two in cascade necessary. The author has therefore decided for the helical filter. The detailed schematic of **fig. 2.3** shows two such filters separated by an amplifying BFT 66 transistor. The net gain of this combination is 5 dB with a bandwidth of 6 MHz.

A ring mixer follows the second helical filter which supplies the second IF at 10.7 MHz. The 2nd local oscillator consists of a BF 247 Colpitts oscillator circuit working at 10.7 MHz below the first IF and supplying 10 mW output power. The os-

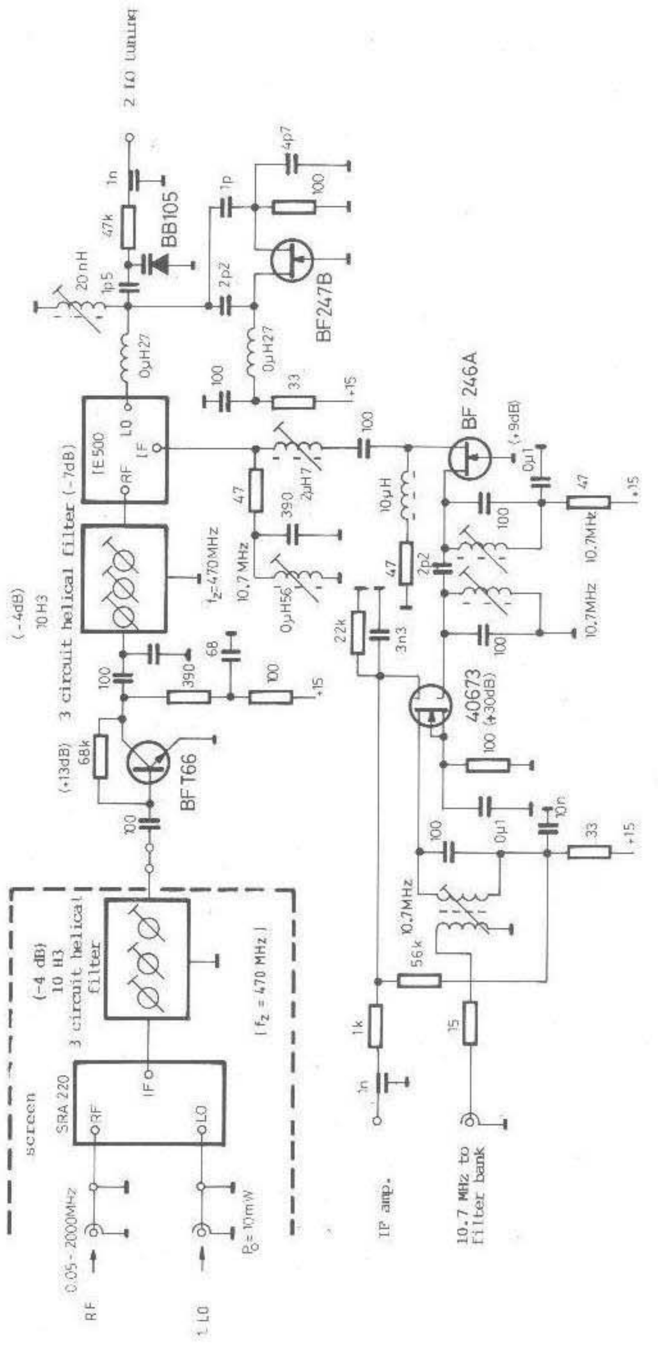


Fig. 2.3: The detailed circuit of the IF converter from 470 MHz to 10.7 MHz

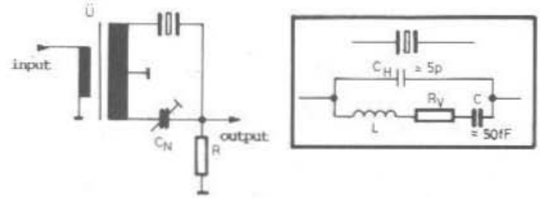


Fig. 3.1:
Basic circuit of a crystal filter

cillator frequency can be shifted a few MHz by means of a varicap diode.

It has been proved that small swept bandwidths, i.e. under 5 MHz, are better accomplished by fixing the 1st LO with a crystal and sweeping the frequency of the 2nd LO. The analyser fine-tuning between the synthesizer step frequencies can also be done here.

The ring-mixer IF output is terminated with a diplexer but it is not strictly necessary. A BF 246 transistor matching stage amplifies the signal by 10 dB. The signal is then filtered by a 10.7 MHz bandpass circuit followed by a further 30 dB of amplification in a MOSFET (e.g. 40673) amplifying stage. The gain of the converter can be controlled downwards from 30 dB by the application of a bias on the gate 2 of this MOSFET stage. The 10.7 MHz amplifying stage has a − 3 dB bandwidth of 500 kHz and a − 60 dB bandwidth of 5 MHz.

For the display of larger bandwidths, 100 to 500 MHz used for example in the investigation of harmonic and spurious signals, this resolution is optimal. The output signal of the IF converter can then be directly fed to the logarithmic display amplifier in order to present a dB-linear display. If a higher resolution is required then an appropriately dimensioned filter can be included in the signal path. The design of suitable filters will now be considered.

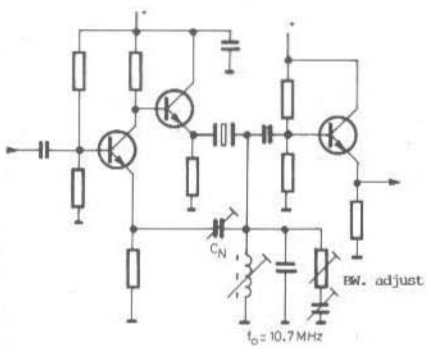


Fig. 3.2:
Crystal filter stage with variable bandwidth

3. THE FILTER BANK

The filter quality determines the spectrum analyser's resolution, i.e. the ability to separate two signals in juxtaposition. Whilst at 10.7 MHz, resolutions of 50 kHz may be obtained with LC circuits, only crystal filters can achieve the lower bandwidths. In fact, it may be stated that one way forward is to select suitable crystal filters from the extensive range offered by firms such as Tele-Quarz or KVG and switch them with relays or diodes into the signal path.

Apart from the fact that each filter costs around DM 150,- they are intended for communication receivers and as such are only of limited use for a spectrum analyser. This is because their steep-sided flanks cause the signal-under-examination to ring as it is swept through the filter. A more suitable passband is bell-shaped, possessing an exponentially falling response equally disposed about the centre frequency in the manner of the so-called Gaus filters. The realisation of such filters is really not so difficult when one is acquainted with the basic circuits of crystal filters. Older amateurs will recall the times when propriety crystal filters were a rarity and filters had to be home-made with surplus crystals.

The basic circuit of a quartz filter is shown in **fig. 3.1.** The input signal is fed in antiphase, from a transformer to a crystal and to a preset capacitor thus forming a bridge circuit. If it is remembered that a crystal is basically a high Q, series-tuned circuit having a parallel holder capacitance then the function of the bridge circuit becomes a little clearer. The capacitive preset arm neutralizes the stray parallel crystal holder capacitance leaving the bandwidth to be determined by the crystal Q and the terminating resistance R. If this preset is made variable, the filter may be tuned over a range of frequencies.

A practical circuit is shown in **fig. 3.2.** The transformer has been replaced by an amplifying stage with two low-impedance outputs and the termination by an LC circuit tuned to the crystal's nominal frequency. This circuit has the advantage that the ever-present unwanted crystal resonances are suppressed completely by this form of termination. An adjustable attenuation, or tuning of the tuned circuit, alters the loading on the crystal and thereby the circuit bandwidth. The output must be loaded by the next stage with a higher impedance.

One such filter stage offers a selectivity of around 20 dB and therefore several must be cascaded. The supply of suitable crystals presents no problems if a quartz filter from a 50 kHz step PLL (from an old NÖBL) transceiver is taken apart and identical crystals extracted. An experimental circuit with four cascaded filters of the type shown in **fig. 3.2.** yielded a 1 to 20 kHz (– 3 dB) tunable bandwidth and a form factor (BW (– 3 dB)/BW (– 60 dB)) of 10 – sufficient for most purposes.

The final version of this filter with four cascaded stages is shown in **fig.3.3.** complete with a by-pass line and final amplification. The bandwidth switching is carried out by means of diodes which switch in the various attenuation networks in parallel with the tuned circuit.

The adjustment for this filter is only possible by the use of a sweep generator test set-up. Most commercial sweepers are of little use for this purpose as they are not able to sweep with precision over such small deviations. The best solution is to make one from a VCO at 10.7 MHz, tuned by a varicap diode over a range of ± 100 kHz. The sweep frequency should also cause the Y amplifier of an oscilloscope to traverse the trace synchronously with it.

Each stage is tuned individually by adjusting the load tuned-circuit for maximum bandwidth. The neutralizing capacitor is then tuned so that the response curve is symmetrical about 10.7 MHz. The fine tuning will be carried out later in the finished analyser for a symmetrical overall response and highest-possible rejection.

The next module in the analyser's signal path to be considered is the logarithmic amplifier.

4. THE LOGARITHMIC DISPLAY AMPLIFIER

It is not normally consequential simply to amplify the IF signal at this stage and to linearly rectify it for presentation to the monitor screen. This would result in signals which have at the most a 20 dB amplitude difference being represented correctly on the same trace. Instead, an amplifier/rectifier arrangement is employed which delivers an output which is proportional to the logarithm of the input voltage. The accuracy of this conversion process determines the level of the instrument.

The actual functioning of this type of chain-rectifier/amplifier will not be considered here as

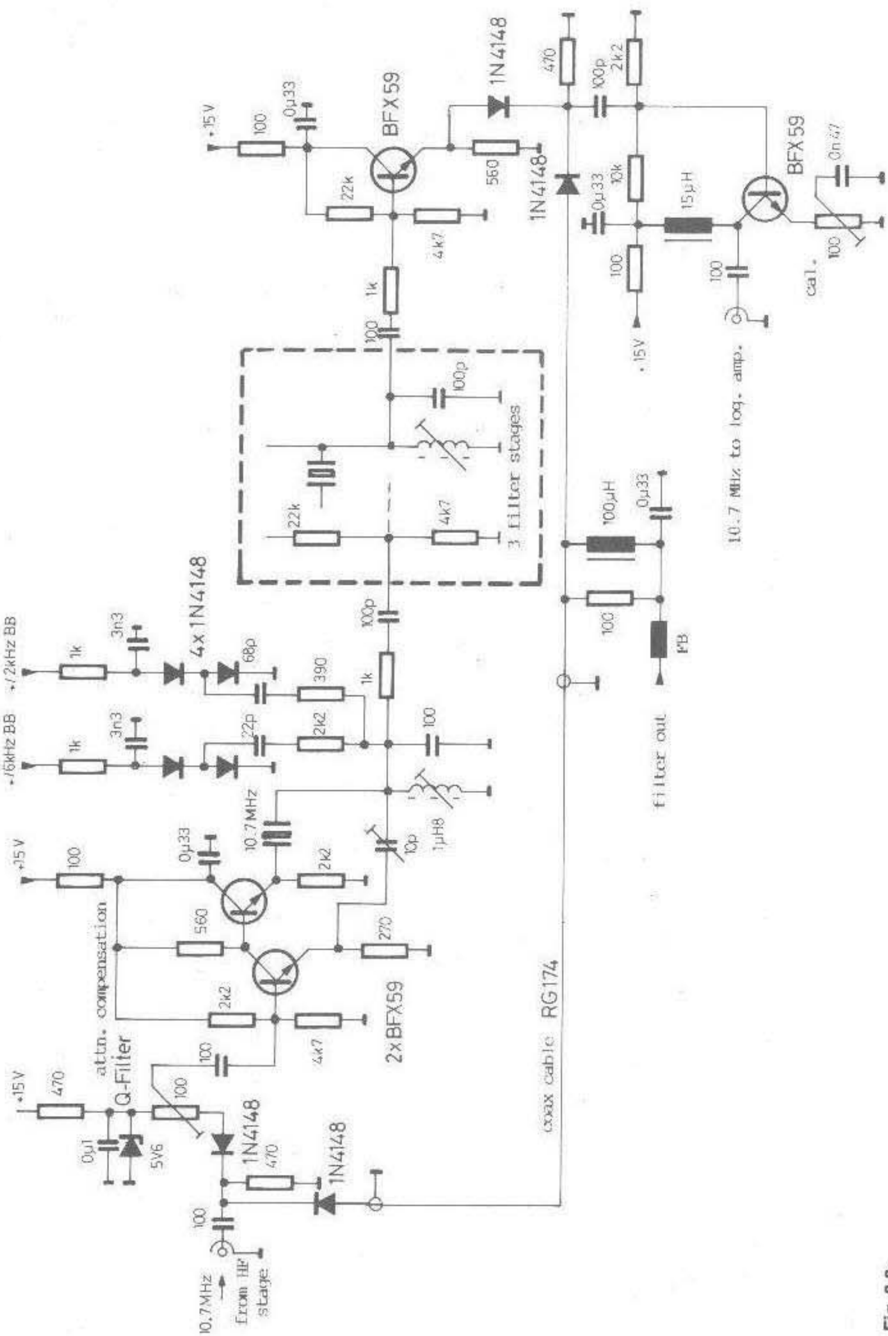


Fig. 3.3:
Spectrum-analyser crystal filter module having four cascaded filters

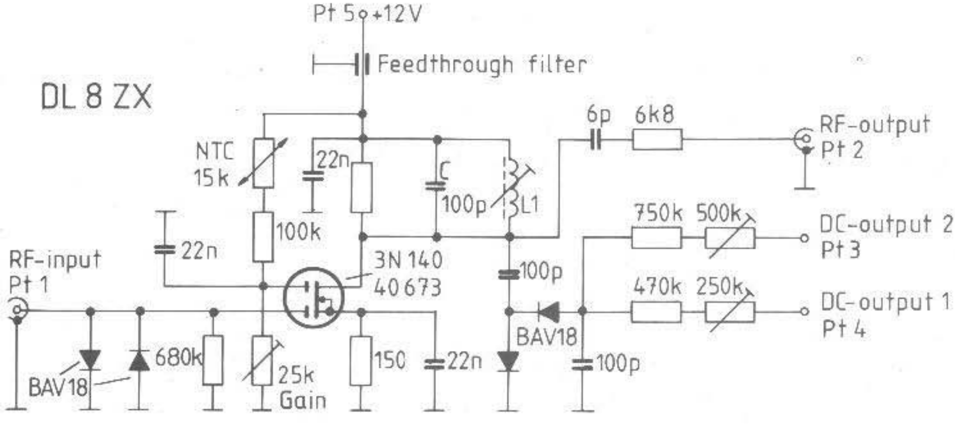


Fig. 4.1: One stage of a chain of rectifier/amplifier (DL 8 ZX)

it has been extensively treated in VHF COMMUNICATIONS 2/1977 by DL 8 ZX. The circuit considered in that article is reproduced in **fig. 4.1**. It is distinguished by its simplicity and easy reproduction. The main disadvantage is that each amplifier has four adjustment points which makes the calibration a bit of a head-ache. Today, semiconductors are much cheaper than they were a decade ago and therefore a trade-in can be made of a little more complexity for ease of calibration. The author obtained his inspiration from an industrial (Hewlett-Packard) logarithmic amplifier and modified it with the following results:–

- The amplifier stages consist of bipolar differential amplifiers whose gain is fixed by resistors and is therefore reproducible and does not need provision for adjustment. No tuning is required as the circuit has no tuned elements so, of course, the selectivity of the display amplifier is minimal.
- Instead of summing the outputs of the instrument rectifiers of each stage the HF voltages of each stage are combined and presented to one instrument rectifier. This obviates any necessity to match individual stage rectifiers. Any deviation from the required demodulator characteristics can be corrected by an adjustment to the summing resistances.
- The linear/dB presentation of the spectrum is effected with only a single rectifier.

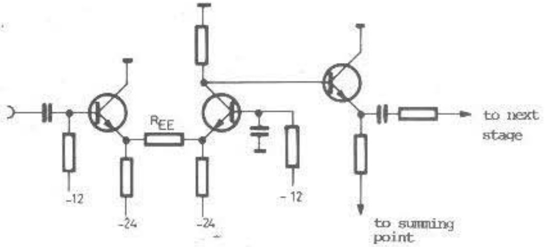


Fig. 4.2: One stage of a logarithmic display amplifier

Fig. 4.2 shows a single differential amplifier stage and **fig. 4.3** the practical circuit having a 70 dB dynamic range. Actually, each differential

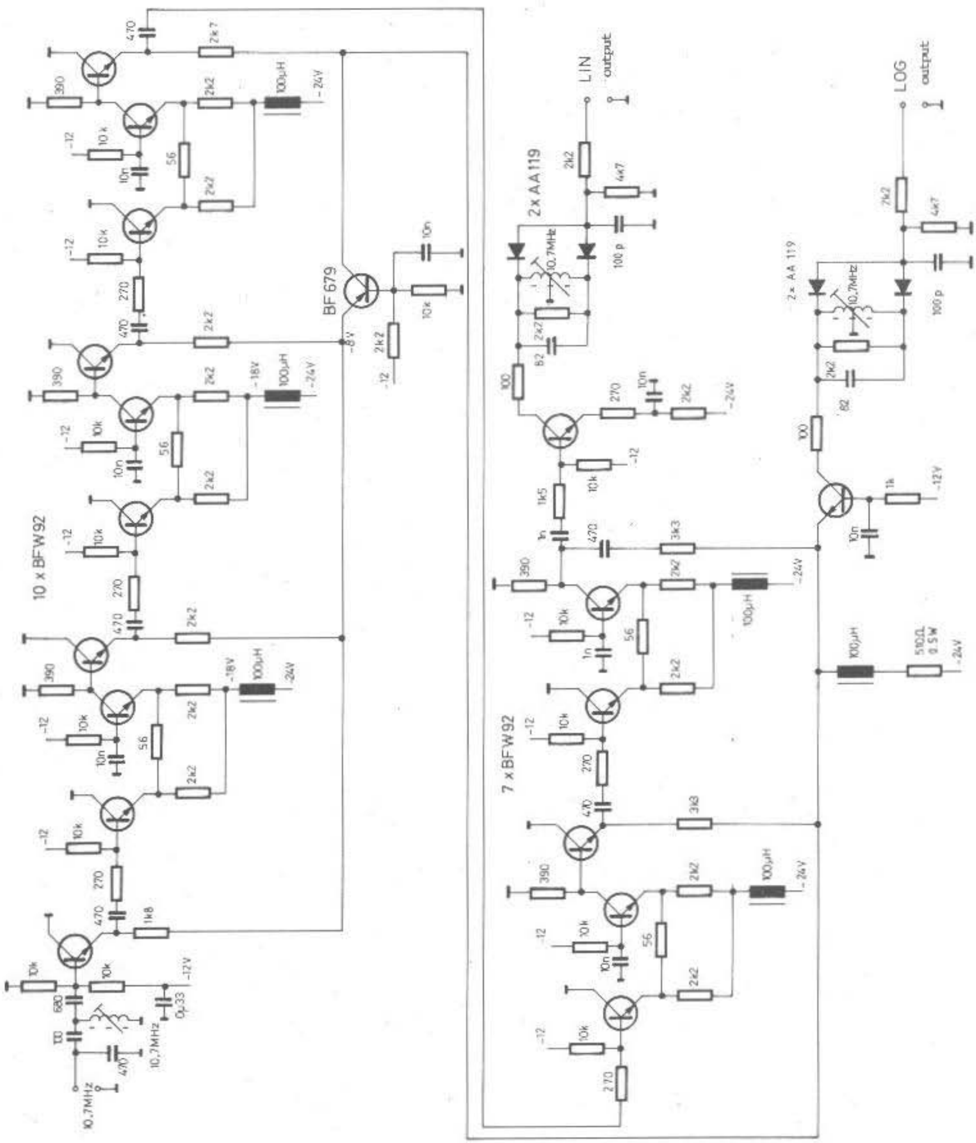


Fig. 4.3:
The logarithmic display amplifier

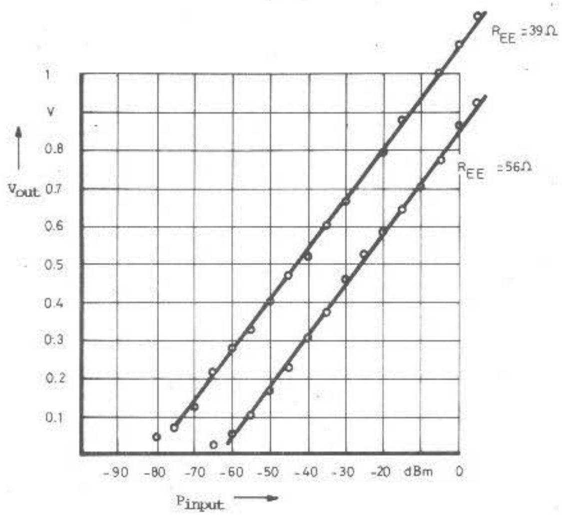


Fig. 4.4: The characteristic curve of display amplifier of fig. 4.3

amplifier stage should have a constant current supply arrangement in its emitter but it has been found that a single high resistance will suffice. The negative rail, however, must be relatively high – in this case – 24 V. Apart from the increased dissipation, this practice has no particular disadvantage.

The level accuracy of such an amplifier lies in the order of 1 to 2 dB absolute, which is considered to

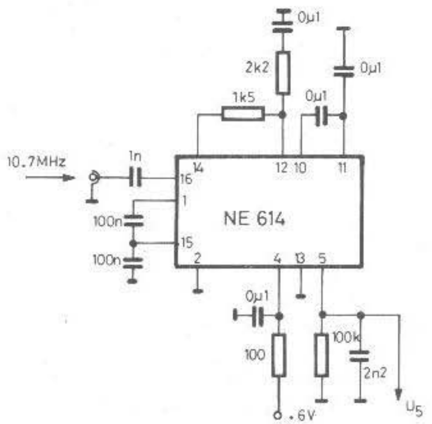


Fig. 4.5a: The NE 614 as a logaritmic display amplifier

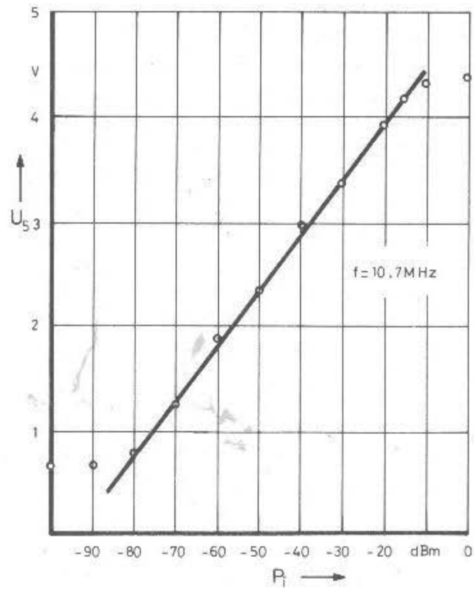


Fig. 4.5b: The characteristic curve of the NE 614 amplifier

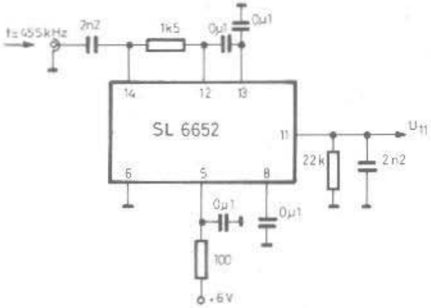


Fig. 4.6a: The SL 6652 as a logarithmic display amplifier

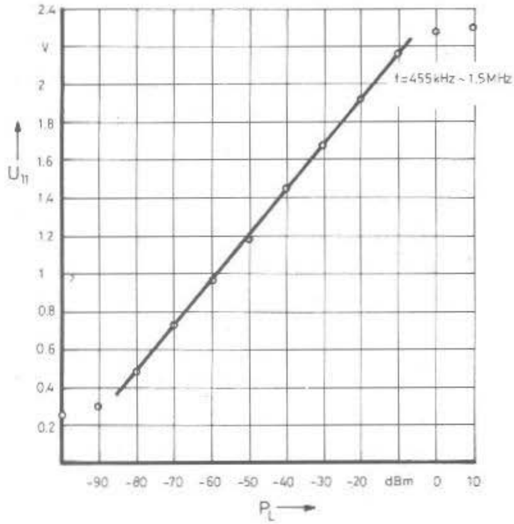


Fig. 4.6b: The characteristic of the SL 6652 log. amp.

be sufficient. The characteristic shown in **fig. 4.4** is that of the prototype amplifier shown in **fig. 4.3.** The accuracy can be improved by decreasing the resistance between each stage and thereby reducing the amplification but there is also a reduction in the dynamic range. If this resistor is dropped from 56 Ω to 39 Ω, for example, over 70 dB of dynamic range can be obtained but the maximum error also increases by ± 1 dB.

A supplementary demodulator, connected to the last amplifying stage, supplies a linear output from an IF input signal. This latter amplifier stage may be connected to the monitor deflection amplifiers and is very useful for more accurate minimal or maximal tuning adjustments as a linear change in level is easier to observe.

It can be seen that both modules can be employed as logarithmic IF amplifier/detectors if the demands upon dynamic range and linearity are not too great.

A further possibility for the realization of a logarithmic IF amplifier is the employment of the purpose-built Plessey differential amplifier series SL 521, SL 523, SL 1521, SL 1523, or SL 1613. All suffer from the disadvantage that they have a large (150 MHz) bandwidth. The wideband noise could be held within bounds by placing tuned circuits between each stage but this would increase the complexity. This circuit technique was not therefore persuad further.

If so much complexity is to be avoided, there are many FM/IF chips which can be used which have a dB-linear characteristic. The VALVO NE 614 is an IF/demodulator chip for frequencies to 15 MHz and the Plessey SL 6652 has an IF limit of only 1.5 MHz but it does have an internal mixer/oscillator circuit at the designer's disposal. The experimental circuit for the NE 614 is shown in **fig. 4.5a** and the IF (10.7 MHz) input signal (dB) versus output voltage is shown in **figs. 4.6a and 4.6b** respectively, but the input frequency in this case was 455 kHz. It can be seen that the SL 6652 has a more linear characteristic but a frequency translation is required from 10.7 MHz to under 1.5 MHz.

Manfred Muehlbacher, DB 9 SB

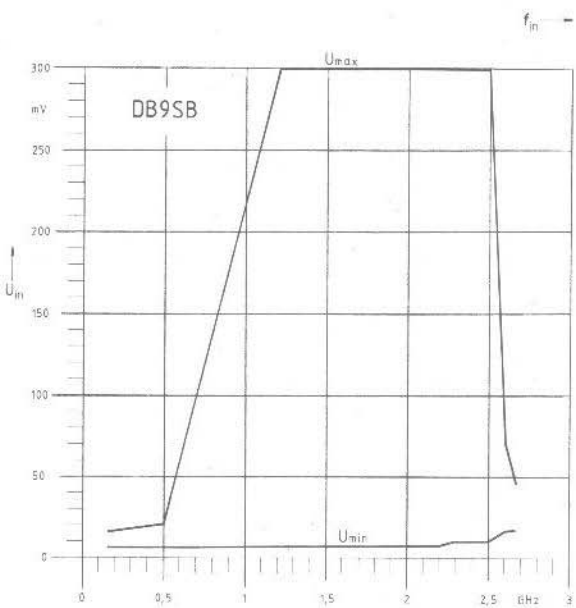
More about the 2.3 GHz Divide-by-100 Scaler

By way of supplementary information to the article in VHF COMMUNICATIONS 1/1985 about the 2.3 GHz divide-by-100 scaler, a new and better version will now be briefly described.

Telefunken has announced that the frequency divider U 822 BS has been replaced by the later type U 862 BS (available at the same price!). The U 862 BS is pin-compatible with the U 822 BS and it divides frequencies of up to, typically, 2.6 GHz, by a factor of 2. It will therefore fit directly on to the printed circuit board DB 9 SB 001 without any modifications.

The author obtained an upper frequency limit of 2690 MHz and a lower limit of 140 MHz. **Fig. 1** shows the dynamic range. It is not recommended that the scaler be used below about 430 MHz as the dynamic range rapidly deteriorates: about 5 - 15 mV in the 2 m band and that is too small for most purposes.

Telefunken also says that the U 862 BS tends to oscillate at approximately 2.1 GHz but a resistance of 12 kΩ connected from the input pin to + 12 V will eliminate the trouble.



Hans Oppermann, Quickborn

PC Interface for the YU 3 UMV Weather Picture Store

The reception of weather satellite pictures with the YU 3 UMV store, described in VHF COMMUNICATIONS 4/1982, 1/1983 and 2/1983, is carried out, in general, such that the signals are fed directly into the store and read out in parallel to tape on cassette machines. If these pictures are required to be viewed later from the tape, a relatively large time is required (about 4 mins.) until a 256 line picture has been completed.

As this store possesses an internal serial interface, it is feasible to control this externally and to connect it via a suitable interface to a personal computer (by personal computer it is meant the IBM PC/XT/AT or compatible computers of other manufacturers). The digitalized signals can be stored on the PC's data carrier, (floppy or hard disc) and recalled at will to the YU store again. This operation can be carried out at a much higher speed requiring only 20 seconds to form a picture.

As the pictures are stored in the PC's data carrier in file form, the possibilities of digital processing are almost unlimited (**fig. 1**). It can, for example, give reproductions via the colour-graphics adapter (CGA) with 4 grey levels or colours, or via the extended graphics adapter (EGA) with 16 colours. In addition, the reproduction in monochrome graphic cards, such as for example the Hercules Graphic Card (HGC) is possible if the requisite processing of the grey levels is undertaken. The most impressive reproduction is, however, that of the weather picture on a matrix

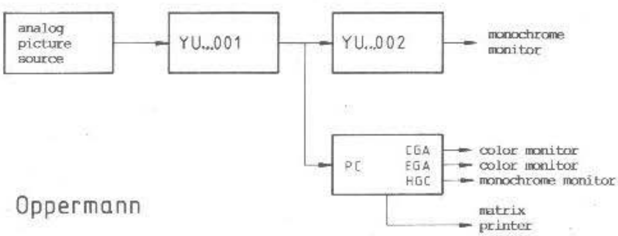


Fig. 1:
Block diagram of the connections and the output possibilities

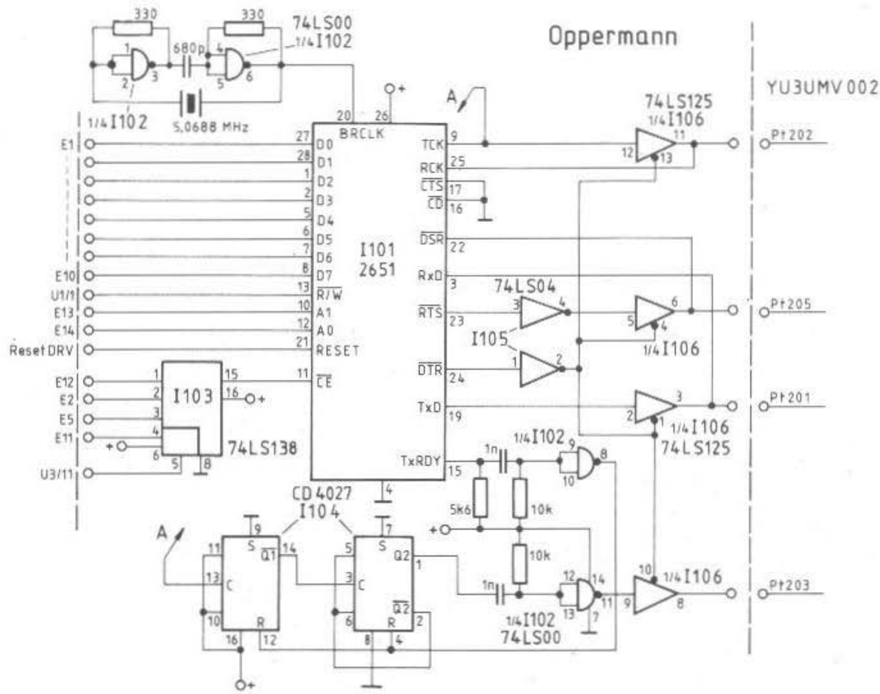


Fig. 2: PC Interface for the YU 3 UMV weather picture store

printer. Of course, pictures produced on software with the characteristics of weather pictures can be transmitted from the YU 3 UMV memory.

1. YU-MEMORY MODIFICATIONS

As already mentioned, the YU memory has a serial interface and is intended for connection to PCBs 001 and 002. A glance at fig. 9 of the YU article (1) might give the impression that a USART was to be added. Closer inspection revealed, however, that the serial data does not have a start bit but has two stop bits instead. As it is not possible to employ USART without a start bit (in the asynchronous mode) and only one stop bit is required, it should be possible to convert the redundant stop bit into a start bit. This is easily carried out by taking I 108 pin 4 to ground.
The normal functioning of the YU store will not be compromised by this measure.

2. PC INTERFACE

The PC interface (**fig. 2**) is connected between PCB 001 and 002 on the serial interface. The following table shows the signals appearing at the appropriate connection points:–

Pt 201	–	Data	I/O
Pt 202	–	Bit-clock	I/O
Pt 203	–	Pixel-clock	I/O
Pt 205	–	Line-clock	O

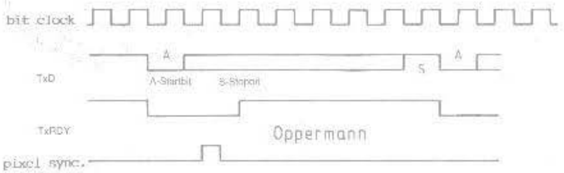


Fig. 3: Output impulse diagram

The designation input/output (I/O) may be regarded as valid as far as the interface is concerned.

The interface functioning is based mainly upon the USART (universal synchronous/asynchronous receiver/transmitter) 2651 (I 101) by Signetics or Valvo. Its extensive application possibilities will go unmentioned here. It is sufficient to regard it in its use here as a programmable serial/parallel converter.

The required clock rate for I 101 is produced by the circuit of I 102. I 103 forms the address decoder for I 101 with the port address H'300' – H'308'.

2.1. Input

In order to be able to read the digitalized signals with the USART, signals from Pt 201, Pt 202 and Pt 205 are necessary. The baud rate of the USART is controlled with the signal on Pt 202 whilst the actual picture information appears at Pt 201.

The line sync. at Pt 205 synchronizes the software with the beginning of each line.

2.2. Output

More complexity is necessary in order to make the digitalized picture signals to be released from the YU store as the latter does not operate without the supplementary signals from Pt 203 and Pt 205.

The basis for the production of the pix-sync. for Pt 203 forms the signal TxRDY. This signal goes low when a pixel at I 101 from the PC is delivered. This co-incides with the negative flank of the start bit (TxD) (**fig. 3**). TxRDY goes high when I 101 is again ready to receive a pixel. The TxRDY cycle is then the best suited for the pixel clock but the timing is not what the YU...002 requires. TxRDY appears exactly 1.5 bit-sync. too early. I 104 effects the necessary delay and consists of two JK flip-flops.

The line-sync. for Pt 205 is produced by the software under the control of RTS.

The bus-driver I 106 is controlled by the DTR signal from I 101. It becomes active when the YU 3 UMV store is read-out and inactive when data from YU is read into the PC. The appropriate control by I 111 on the YU...001 ensures that no crash of outputs occur.

3. CONSTRUCTION

The interface is realized upon an IBM-prototype-card (3). This is available at every IBM outlet for about DM 100.–. It must, however, be equipped with the integrated bus driver. The board already has the printed circuitry for it, so one is spared the exasperation and fiddle of wiring it in on Veroboard.

The actual circuit is, however, wired on Veroboard and placed on a free spot on the card. The number of ICs is not great and the work should proceed without problems.

4. PC SOFTWARE

The drive to the PC interface requires a few modules which are concerned with the read-in and read-out arrangements. With these modules a complete program with any required function, as indicated earlier, can be achieved.

These modules are written in Pascal and serve merely as an example as, of course, other languages can be used. The language should possess, however, the optimal speed characteristics, as the read-in and read-out of the picture elements is time critical.

The procedures are self-explanatory during the course of the commentary.

If the digitalized pictures are not read-out of the YU store but retained for further processing, a pixel conversion table (CONV) will be necessary. The YU store delivers the "most significant bit" (MSB) in the "least significant bit" position.

5. OPERATIONAL EXPERIENCE

The interface has been in use some six months now and it fulfills its purpose admirably.

During this period, a program (see appendix) with the following functions has been developed:

- Reading in and out pictures with the YU store
- Read-out via color-graphics adapter
- Read-out via the Hercules card (HGC) for monochrom half-tone pictures
- Read-out on the matrix printer

More attention should, however, be devoted to the last item. It is particularly problematical to reproduce grey levels on a matrix printer. Reference (2) proposes several solutions to the problem but only two of them are relevant, i.e.:

1. Grey-tone pictures via 4 x 4 matrices
2. Grey-tone pictures with the multi expansion technique

The latter technique has the advantage of relatively fast computer processing as well as the 1 : 1 relationship between read-in and read-out. This means, that every input element corresponds to an output element. The picture quality is, however, somewhat mediocre and I use it only for picture monitoring purposes.

The first technique entails the representation of every grey level on a 4 x 4 matrix – black being produced by 16 black elements and white by a matrix containing no elements. The other 14 values are represented by a number of elements corresponding to the grey value. The position of these elements is determined by a co-incidence generator to avoid patterning. In this manner, 16 grey levels can be presented with a really good picture quality.

Even better results may be obtained if 32 grey levels are used. This can be done by double printing in the following manner:–

The grey values for a line are first of all divided by 4 and the corresponding 4 x 4 matrices formed and fed to the printer. The printer head is then positioned at the start of this line. For all grey values whose halved original grey value is less than 15 (0 = black) are then given the same 4 x 4 marices again.

The processing time required for this technique on the PC/XT is about 20 minutes for a picture of 200 x 240 elements but the time spent waiting is really worth it!

6. REFERENCES

(1) Matjaz Vidmar, YU 3 UMV:
A Digital Storage and Scan Converter for Weather Satellite Images
VHF COMMUNICATIONS: Editions 4/1982, 1/1983 and 2/1983

(2) Peter Haberäcker:
Digitale Bildverarbeitung
Carl-Hanser-Verlag, München/Wien

(3) IBM Personal-Computer XT
Technisches Handbuch

Please turn over

7. APPENDIX · EXAMPLE OF A PROGRAM

```
program picio;
label loop;
const
  mr1 = $302;                    (* mode register 1           *)
  mr2 = $302;                    (* mode register 2           *)
  cr  = $303;                    (* command register          *)
  sr  = $301;                    (* status  register          *)
  dr  = $300;                    (* data register             *)
                                     (* look up table             *)
  conv: array[0per.63] of byte = ($0,$20,$10,$30,$8,$28,$18,$38,
                                  $4,$24,$14,$34,$c,$2c,$1c,$3c,
                                  $2,$22,$12,$32,$a,$2a,$1a,$3a,
                                  $6,$26,$16,$36,$e,$2e,$1e,$3e,
                                  $1,$21,$11,$31,$9,$29,$19,$39,
                                  $5,$25,$15,$35,$d,$2d,$1d,$3d,
                                  $3,$23,$13,$33,$b,$2b,$1b,$3b,
                                  $7,$27,$17,$37,$f,$2f,$1f,$3f);

type
   val =   array[1..32767] of byte;

var
  I,J,K,P:   Integer;
  a,B,O  :   byte;
  DataFile:  File;
  val1:      array[1..2] of ^val;
  filnam:    string[14];

procedure Init2651 (read: boolean);
begin                            (* Initialize 2651 (I101)    *)
if not read then
  begin                          (* Init for Output           *)
    port[mr1] := $46;
    port[mr2] := $2f;
    port[cr] :=  $33;
  end
else
  begin                          (* Init for Input            *)
    port[mr1] := $45;
    port[mr2] := $00;
    port[cr] :=  $34;
    B := $34;
 end
end;

procedure NewLine;
begin                            (* pulse RTS for newline     *)
  port[cr] := $33;
  port[cr] := $13;
end;

procedure PixelOut;
begin                         (* Output pixel in "O" to Interface *)
  a := 0;
  while a <> i do
```

```
  begin                               (* look for 2651 is ready          *)
    a := port[sr]: a := a and 1;
  end;
  port[dr] := O;                  (* output pixel                   *)
end;

procedure PixelIn;
begin
  a := 0;
  while a <> 2 do
  begin                               (* wait for received pixel         *)
    a := port[sr]: a := a and 2;
  end;
  if (port[sr] and $20) = 0 then (* test framing error                *)
  begin
    O := port[dr];                (* read in pixel                  *)
    B := B xor $20; port[cr] := B; (* pulse DTR to test loop time*)
  end else
  begin
    port[cr] := $14;                  (* reset overrun error             *)
    write('Error');
  end;
end;

procedure LinStart;
begin                                 (* wait for start of line          *)
  a := $80;
  while a <> 0 do                 (* search for DSR = 1 (line-clock) *)
  begin
    a := port[sr]; a := a and $80;
  end;
  while a <> $80 do                      (* search for DSR = 0 *)
  begin
    a := port[sr]; a := a and $80;
  end;
  port[cr] := $14;                    (* start of line found, reset      *)
end;                                  (* overrun error                   *)

procedure Grayscale;
begin                             (* example for usage of look up table*)
  repeat
  NewLine;
  for i := 0 to 63 do begin
     for j := 1 to 4 do begin
     O := conv[i];
     PixelOut;
  end; end;
  until keypressed ;
end;

(*******************  M A I N  ************************************)
var reply:  char;
begin
New(val1[1]);                     (* get storage for pixel buffer (32K)*)
New(val1[2]);                     (* get storage for pixel buffer (32K)*)
writeln('-- START -- ');
loop:
port[cr] :=  $34;                     (* set input direction             *)
```

```
writeln('Picture-input (I) / -Output (O) ? ');
read(kbd,reply);
case reply of
(*-------- P I C  -  O U T P U T ----------------------------------------*)
'o': begin
      writeln(' ---- Output ---- ');
      Init2651(false);          (* Init for output                        *)
      writeln('Enter Filename: ');
      readln(filnam);
      if length(filnam) = 0 then filnam := 'weather.pic'
              else filnam := filnam + '.pic';
      Assign(DataFile, filnam);
      Reset(DataFile);
      BlockRead(DataFile,vali[1]^[1],256 * 2);(*read 32K from disk*)
      BlockRead(DataFile,vali[2]^[1],256 * 2);(*read 32K from disk*)
      Close(DataFile);
      for p := 1 to 2 do begin
        for i := 0  to 128 do begin
          NewLine;
          for j := 1 to 256 do begin
            O := vali[p]^[i*256 + j];
            PixelOut;
        end; end; (* i=0 to 128 *)
      end; (* p=1 to 2 *)
     end;
(*-----------P I C  -  I N P U T ----------------------------------------*)
'i': begin
      writeln(' ---- Input  ---- ');
      Init2651(true);           (* Init for input                         *)
      for p := 1 to 2 do begin
        for i := 0 to 128 do begin
          LinStart;
          for j := 1 to 256 do begin
            PixelIn;
            vali[p]^[i*256 + j] := O;
        end; end; (* i=0 to 128 *)
      end; (* p=1 to 2 *)
      sound(440);                  (* signal "input is ready"        *)
      Delay(500);
      NoSound;
      writeln('Enter Filename: ');
      readln(filnam);
      if length(filnam) = 0 then filnam := 'weather.pic'
          else filnam := filnam + '.pic';
      Assign(DataFile, filnam);
      Rewrite(DataFile);
      BlockWrite(DataFile,vali[1]^[1],256 * 2);(* wrt 32K to disk*)
      BlockWrite(DataFile,vali[2]^[1],256 * 2);(* wrt 32K to disk*)
      Close(DataFile);
     end;
else writeln('End of Program');
     halt;
end;
goto loop;
end.
```

Andreas Claar, DF 9 CP

Electronically-Switched Attenuators

Attenuators, whether fixed or variable, have a very definate place in the realm of radio techniques. They render vital service in such items as signal generators, selective level meters and in radio communications systems.

In order to determine the optimal functioning of a radio item, the input signal levels must lie between certain fixed limits. In radio receivers especially, the input signal is so varied that it is difficult to come to any sort of prognosis upon it. The receiver must therefore be in the position to automatically adjust itself according to the prevailing signal conditions. This implies that the receiver must present all modules with a signal with which they can handle efficiently and without distortion. Too much signal power leads to inter-modulation distortion and also to the falsification of the S-meter indication owing to saturation effects – the needle remaining practically jammed at full-scale deflection and no meaningful signal-level reading can be carried out.

An attenuator, placed at a suitable point in the signal chain, does help to bring the signal down to a more workable level. Even better would be a continuously variable attenuator with a calibrated scale such as the one described in ref. (1). Often, however, an attenuator variable in known fixed steps is desired which allows a better appreciation of the input signal strength, or indeed, its measurement.

For many years now, step attenuators have been available commercially whose attenuation may be stepped by a fixed amount either higher or lower. The attenuator range, its insertion loss and input impedance together with its working frequency range are all important specifications in its evaluation and eventual selection. The attenuation adjustment is mostly carried by control potentials at TTL voltages in order that a convenient control circuitry can be developed. A user-friendly step attenuator could have the following criteria specified:–

- Attenuation continuously adjustable in 1 dB steps over the range 0 - 100 decibels.
- Wideband frequency range with negligible VSWR.
- The attenuation is effected by means of UP/DOWN buttons or a decade switch, the set attenuation being indicated on a digital display.

To buy such an item would cost a mint of money but precision would be assured. The radio amateur with a ready access to components can, how-

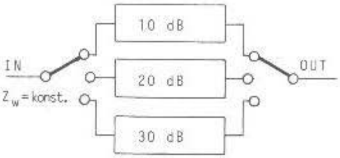


Fig. 1: Various attenuator four-poles with multi-way switch

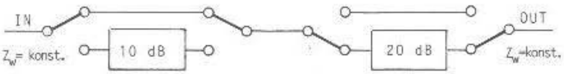


Fig. 2: Series attenuator pads which are either bypassed or inserted to make up the desired total attenuation.

ever, build a step attenuator for relatively little outlay which would entirely suffice for hobby purposes. No exotic parts are required and the technical specification of the finished instrument is worthy of note. The construction of the instrument to be described later in this article can be employed in the VHF range, 2 m and 70 cm bands and in the higher UHF ranges.

1. PRINCIPLE

A step switchable attenuator can be realized in one of the following ways:

1st Method

a) An attenuator four pole with the various desired values of attenuation can be selected with a multi-way switch as in **fig. 1**.

b) A specially dimensioned series of attenuators can be either inserted or bypassed in the chain by switches as in **fig. 2**.

If many attenuation steps are required to be selectable then method a) is unsuitable as one four-pole is selected for each value of attenuation required.

The method b) is to be favoured as the individual elements are more fully utilised. From n various independent four-pole elements, 2^n different attenuation values may be obtained. For example: If a 20 dB and a 10 dB attenuator pad were connected in series, the following combinations are possible:

0 dB – the two pads bypassed
10 dB – only the 20 dB pad bypassed
20 dB – only the 10 dB pad bypassed
30 dB – both pads connected in series

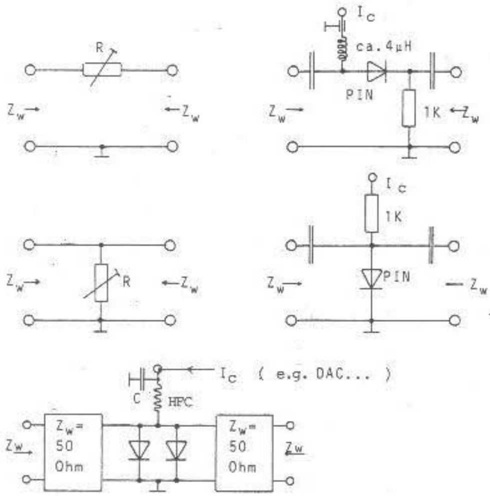


Fig. 3: a) **Series attenuator:**
No reflection-free termination
$Z = f(R) = (50 - \infty)\ \Omega$

b) **Parallel attenuator:**
No reflection-free termination
$Z = f(R) = (0 - 50)\ \Omega$

c) **Attenuator with matching circuit**
$Z = \text{constant (e.g. } 50\ \Omega)$

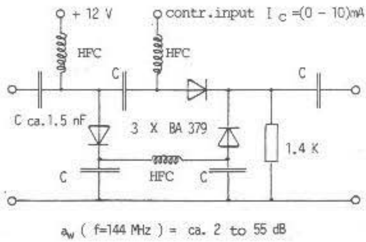


Fig. 4a: A continuously variable attenuator achieved by a trick circuit (ref. 1). The input and output resistance remains independent of the selected value of attenuation at exactly 50 Ω.

A step attenuator from 0 to 30 dB switchable in 10 dB steps, using only two elemental pads, is the outcome.

2nd Method

The four-pole attenuators of method 1 all possess a constant, defined input and output resistance of e.g. 50 Ω – as long as they are constructed in either a 'T' or a 'Pi' network. A definate and a constant-with-frequency characteristic impedance is very important. The controllable PIN-diode attenuator elements from (1) achieve these conditions by means of a circuit trick – but there are also other simpler methode of signal attenuation.

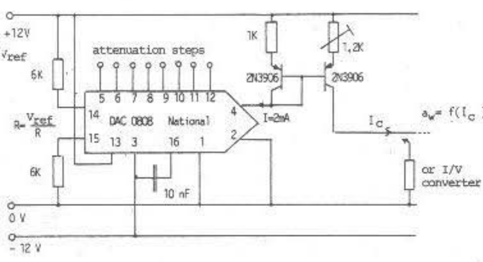


Fig. 4b: The continuously variable adjustment of the PIN diode attenuator above may be carried out by means of an analogue-to-digital converter. The pre-set potentiometer adjusts the control current to the desired value.

a) The principle of parallel and series attenuators is shown in **fig. 3**. The variable resistor here can be a resistor capable of working at HF or a PIN diode with the appropriate control circuitry.

This arrangement has the disadvantage that the input/output impedance varies over quite a large range, giving rise to energy reflections. There are circumstances in which a constant characteristic impedance is not absolutely essential. Sometimes a matching circuit must be made in order to keep the input characteristic constant, e.g. with a directional coupler. For frequencies under 1 GHz the complexity is too high. For such frequencies the following solutions are available:

b)The continuously variable attenuators of **fig. 4a** mentioned in (1), both possess a constant input impedance over the frequency range 50 to 800 MHz and at all attenuator settings.

The attenuator setting may be accomplished by means of a DC potentiometer circuit or with the assistance of a digital to analogue converter (DAC). For every combination of bits presented to the DAC (**fig. 4b**), a different value of attenuation is set-up and for which a different value of control current flows.

With the combination of an EPROM and an UP/DOWN switch it is possible to control the attenuation with any degree of required resolution.

The attenuation versus control-current characteristic is not a linear function. The appropriate function: a_w = f(cont.) must therefore be linearised with the help of an EPROM. The EPROM may be read e.g. with a forwards/backwards counter.

2. THE PIN DIODE AS AN HF SWITCH

Modern HF systems are no longer switched by means of relays except for the most special cases. RF coaxial relays are easy in their application but very uneconomical. In other respects too they fall short of the ideal, being slow in operation and they are prone to mechanical wear and tear. They do, however, offer low insertion im-

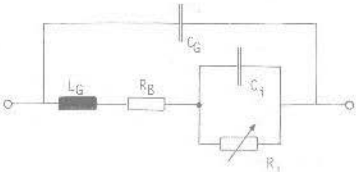


Fig. 5a: HF PIN-diode equivalent circuit

C_G = Capacity to ground
L_G = lead inductance
R_B = PCB track resistance
C_i = I-Zone capacitance
R_i = HF resistance (dependent upon control current)

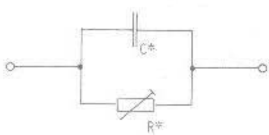


Fig. 5b: Simplified PIN-diode equivalent circuit
C* = Total capacitance
R* = Total resistance

pedance and high open-circuit isolation but nevertheless, they are being gradually replaced by the PIN-diode switch. How PIN diodes are able to switch HF energy will now be briefly explained.

2.1. What is a PIN-Diode?

PIN dides are basically silicon diodes. Their mode of operation, as opposed to other types of diodes, is quite easy to understand. In between the P and the N junction, normal to a silicon diode, an additional layer known as the I-zone is introduced. The I stands for "Intrinsic" (hence PIN diode) which means inherently conducting, in this context. The exact physical explanation of how they work is explained elsewhere (2). The PIN-diode characteristics at low frequencies and at DC, are indistinguisable from those of a normal silicon diode, that is, the transfer characteristics are similar.

As the frequency of operation is raised, the PIN-diode characteristics become evident. This happens at limit frequencies between 1 and 5 MHz. The rectifier action gradually disappears and the device assumes the character of an HF resistor. This HF resistor comprises several components, but in the main they can be compared to an adjustable ohmic resistance. The actual resistance is determined by the amplitude of a control current which is passed through the diode together with the signal to be attenuated. The HF resistance can be varied by these means from a few ohms to several thousand ohms. Because of these characteristics, the PIN diode may be said to be an HF resistor or switch capable of being conrolled by a direct current.

2.2. PIN-Diode HF Equivalent Circuit

The equivalent circuit of a PIN diode is shown in **fig. 5a**. Examination of this circuit reveals that there exists a definate limit to the frequency which can be handled by the device. These limits vary with the individual types of PIN diodes. Some of the undesirable properties are the parasitic inductance L_g (approx. 0.7 nH), the ohmic resistance R_b (approx. 1 Ω) and the capacitance C_i across the I zone. The resistance R_i represents the desirable property of the device which can be varied between 5 Ω and 10 kΩ (approx.). This is the component about which everything else revolves.

The parallel stray capacitance C_g tends to nullify the resistance R_i when the resistance is set to the higher values and also with increasing frequency, thus setting an HF limit to its full range operation.

It is, then, typical for every PIN diode to have a frequency specification for which the main resistance band remains within tolerance. The undesirable diode reactances should therefore be kept as small as possible. The characteristic of **fig. 6** shows the dependence of the high-frequency resistance (r_f) of the PIN diode upon the controlling direct current I_F.

2.3. The PIN-Diode Switch Function

When the PIN diode is employed as a switch, there are only two functions which are desired. In one condition the diode should have a lower resistance when it is closed and in the other condition, open, it should possess extremely an high resistance. In other words exactly the same

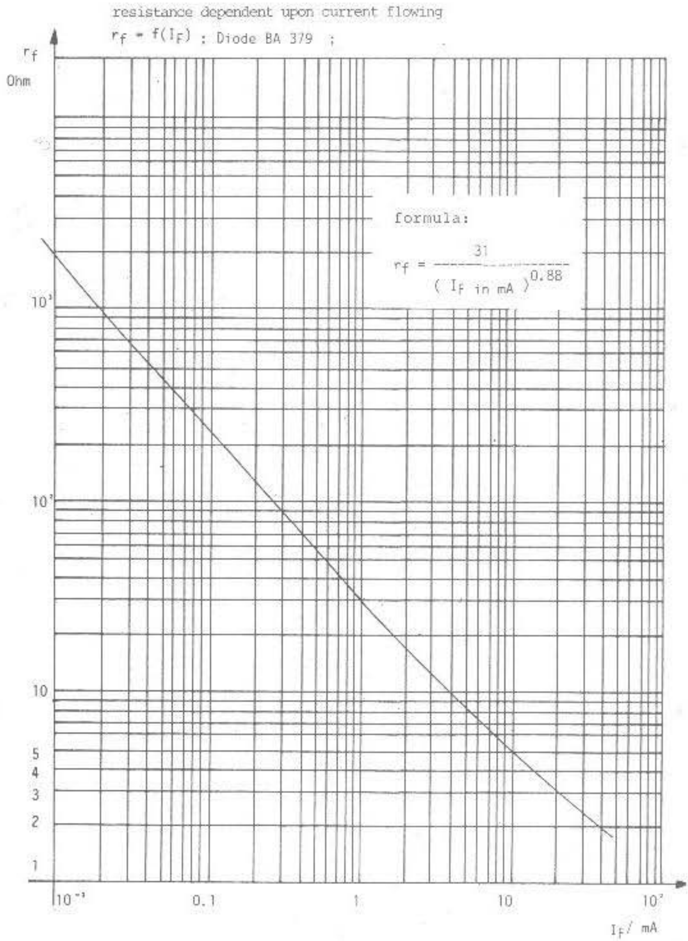


Fig. 6: Resistance of BA 379 as a function of current

properties as one would expect from a mechanically operated electrical switch.

In order to obtain a low-ohmic resistance, the largest permissable control current should flow through the diode – the resistance then is only a few ohms. When this current is removed then the diode reverts to its maximal resistance for that frequency. This high value of resistance can be maximised by placing a bias across the barrier direction but in general, no great benefits can be expected for the extra complexity involved.

In the 'open' condition the diode resistance is a few thousand ohms but this is frequency dependent and the total circuit resistance varies with both the characteristics of the individual diode and upon the external circuitry involved.

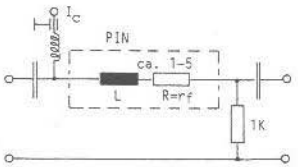


Fig. 7a: HF equivalent circuit of PIN-diode showing attenuation in the direction of conduction

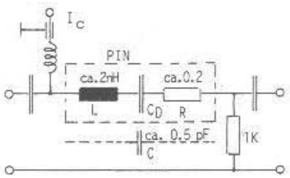


Fig. 7b: HF equivalent circuit of PIN-diode showing attenuation in the blocked direction

The equivalent circuit of a PIN-diode switch is shown in **fig. 7**. The undesired diode reactances may be partially compensated for by the external connection of both L and C which effectively tunes them out. This measure, of course, is only effective at or near a certain frequency and since most PIN-diode circuits are required to be wide band, resonance effects are normally to be avoided. Later on, a few techniques will be explained which are commonly used in HF circuitry:

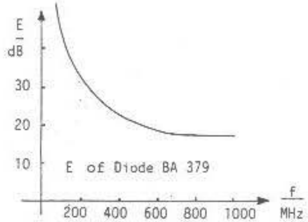


Fig. 8: The isolation as a function of frequency

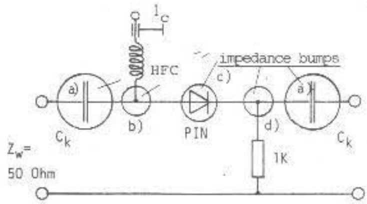


Fig. 9a: The inclusion of components in a 50 Ω system gives rise to impedance bumps

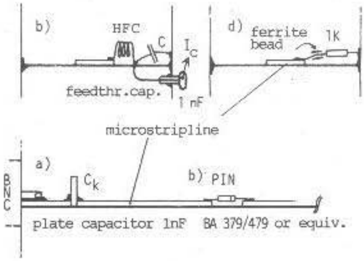


Fig. 9b: PIN-diode switches constructed in the stripline technique

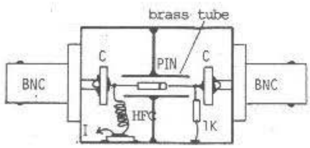
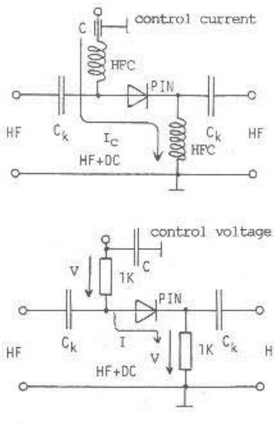


Fig. 9c: Suggested form of PIN-diode switch in coaxial form



Figs. 10a/b: Low-pass filter feed for the control current

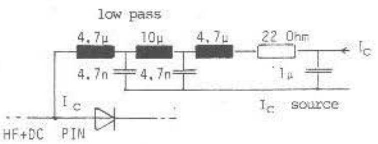


Fig. 10c: Method of control-current feed taken from a Hewlett-Packard circuit. Frequency range: 10 to 1000 MHz

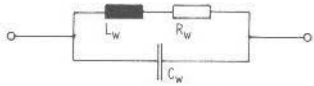


Fig. 11a: Equivalent circuit of one turn on a coil

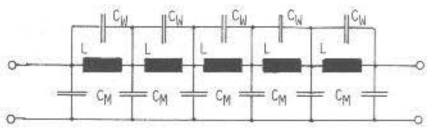


Fig. 11b: L_W = Inductance of a turn
C_M = Capacitance to ground
C_W = Inter-turn capacitance
At higher frequencies the ohmic component may be neglected but not the self-capacitance

a) Insertion loss: The inclusion of a PIN diode in a circuit always intriducs a residual attenuation (**fig. 7a**).

b) Isolation: The PIN diode in the biased-off state always lets a little RF energy flow through the circuit. The isolation I is strongly dependent upon the frequency and is expressed in dB. The higher the isolation (dB) the better the 'switch-off' state will be. This is of particular importance for change-over switches because the switched out portion of the circuit should have no influence upon that which is working (**fig. 7b**). The isolation I in dB is shown in **fig. 8**, plotted against frequency.

2.4. The Control of PIN-Diodes

The control current must be applied carefully as the PIN diode is included normally in a 50 Ω network. The control current is introduced via means of low-pass filters and is superimposed upon the HF signal to be controlled. These low-pass filters must not be allowed to interfere with the characteristics of the HF circuit. Every additional component, introduced into the 50 Ω system, brings a potential reflection point with it (**fig. 9**). In order that the control current be fed in and out of the PIN diode without affecting its operation, the following possibilities may be considered (**fig. 10**):

a) The DC feed-in is effected by means of wideband high-frequency chokes which present a high impedance to the working range of frequencies. The direct control current can, of course, flow unhindered through the chokes. This method is used preferably in a pure current control, for example, by a digital-to-analog converter.

b) The feed-in is effected by HF resistors. These, resistors in relationship to the characteristic impedance of the PIN-diode system, should be correspondingly high – a 50 Ω system using feed-in resistors of 800 to 1000 ohms. The loading effects upon the PIN-diode system are minimal and therefore the system's return-loss is unimpared. In order to fully control the diode, 10 to 100 mA are required which necessitates having a rather large control circuit potential – 10 mA through a 1 kΩ resistance is already a potential difference of 10 V. This means that heat loss in the feed resistors is unavoidable.

The best solution is probably a combination of both method a) and b). The first method has the disadvantage that care must be exercised during testing not to destroy the PIN diode. The second method's disadvantage is that a high-control supply voltage is required.

High-frequency chokes also have their problems. Commercial HCFs normally consist of many windings of fine wire which create quite a large capacitance across the component. The HFC's self-capacitance then resonates with the rated inductance at some frequency. In the interests of a predictable wideband system performance, this resonant frequency should not lie within the band of interest. Also, without resonance effects, a large HFC self-capacitance represents a capacitance directly across the 50 ΩPIN-diode system thus tending to shunt away the RF signal energy in an uncontrolled manner. High self-capacity chokes cannot therefore be employed in wideband attenuator systems. Genuine wideband HFCs covering HF to UHF are very rare components.

The classical VALVO, six-hole ferrite choke core represents an acceptable alternative but in order to improve its performance at high frequencies, a 4-turn, air-cored coil should be used in series with it at the 'hot' end!

Perhaps the best way of getting a suitable HFC is to make one. This can be done by simply winding 8 turns of copper wire around a 3 mm former thus making a self-supporting air-cored inductor. Several HFCs of this kind used in a system may lead to resonance effects but perhaps with luck the frequencies at which resonance occurs will not be used by the system.

The inductance of an air-cored coil may be increased considerably by the use of a suitable ferrite core. The number of required turns will be

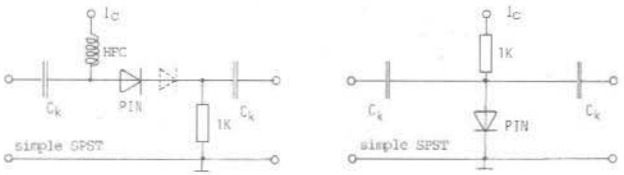


Fig. 12a:
Simple switch SPST

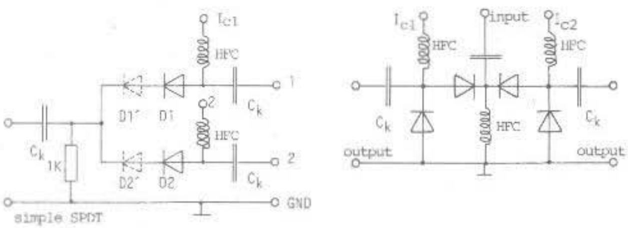


Fig. 12b:
Simple change-over switch SPDT (left). The free output is automatically short-circuited. The residual RF is reflected (right).

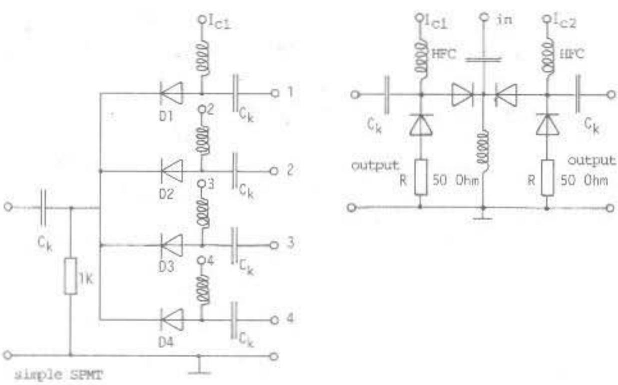


Fig. 12c:
Simple multi-way switch SPMT (left). The free outputs must be terminated in 50 Ω (right). The free output can be shorted without affecting the input VSWR

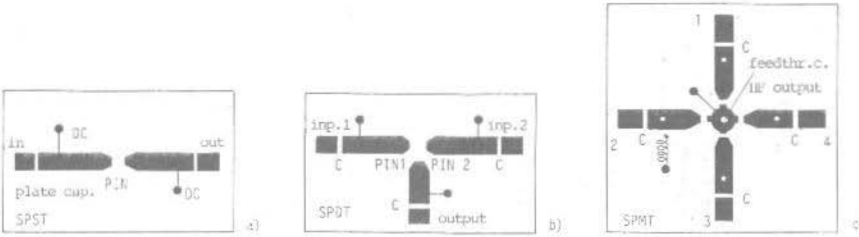


Fig. 13: Main constructional points in stripline technique

smaller and the self-resonant frequency higher. The simplified equivalent circuit of a high-frequency choke is given in **fig. 11**.

The HFC problem is only troublesome with wide-band systems, i.e. say 40 to 900 MHz. If only one band of interest is contemplated then the chokes described above can be used without any problems.

The cold end of the choke must be effectively RF-earthed with one, or preferably several, good-quality capacitors. Normal feed-thro' capacitors alone are usually not adequate for the job but a small value soldered in parallel should do the trick. A more complicated low-pass feed originating from Hewlett Packard is shown in **fig. 10c**.

3. PIN-DIODE SWITCH BASIC CIRCUITS

PIN diodes allow HF circuits to be realized with the same facility of those at AF. **Figure 12** gives a rough review of the possibilities. In principle, these possibilities are only limited by the imagination. The following circuits may be recognized:

a) Normal ON/OFF i.e. SPST switch (Single Pole Single Throw)
b) Change-over SPDT switch (Single Pole Double Throw)
c) Multipole SPMT switch (Single Pole Multi Throw)

In order to improve the decoupling between 'throws', the unused PIN-diode circuits may be readily shorted to ground by means of another switch wafer. This almost guarantees that there will be no RF at the output of the unused switch-ways. Short-circuiting the unused outputs, however, means that the input resistance of the appropriate PIN-diode switch-way will tend to zero. This affects the input resistance to the switch, making it deviate from 50 Ω and thus giving rise to power reflections. This can be undesirable, especially in systems where the modular input and output impedances require a 50 Ω termination. In such a case it will be necessary to include a 50 Ω resistance in the short-circuit path of the diode. Such devices are known as non-reflective switches.

In order to improve upon the PIN-diode switching performance, several of them may be connected in series. This greatly increases the isolation in the off-condition but the residual attenuation is, of course, increased. This is dealt with in ref. 3. The diagram of **fig. 13** shows the practical form of some PIN switches. The stripline form is particularly well-suited for this type of switch because the characteristic impedance is very well defined at all parts. Reflective points caused by the long, thin connecting wires of the PIN diode are avoided.

Even when the conductor track width of the stripline is not exactly 50 Ω, it is far better than a freely-wired circuit. A meaningful and effective alternative to etching the copper to form tracks is to cut them. The copper-sided board is cut into 3 mm wide strips and glued onto a conducting surface (tin plate or PCB stock). This method is very suitable for experimentation and when the testing and adjustment has been optimized, a final PCB may be etched out.

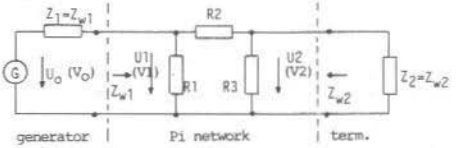


Fig. 14: Pi-attenuator with source and load resistances

4. FIXED ATTENUATORS IN THE PI-TECHNIQUE

In order to realize an attenuator network which can be stepped, the individual elements must be independent of frequency. The Pi circuit offers the best solution in the UHF/VHF range as a rule. The formulae, used to calculate the elemental resistance, are given below for the sake of completeness. The Pi circuit of **fig. 14** applies:–

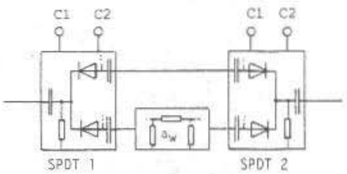


Fig. 15a: Switched four-pole with simple SPDT switch

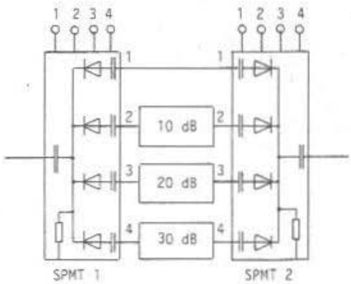


Fig. 15b: With the SPMT switch several attenuator pads may be selected. The poor isolation, however, causes problems

a) Unsymmetrical

$$R2 = \frac{n^2 - V}{2n} \cdot Z_{w2}$$

$$R1 = \frac{n^2 - V}{\dfrac{n^2 + V}{Z_{w1}} - \dfrac{2 \cdot n}{Z_{w2}}}$$

$$V = Z_{w1} / Z_{w2}$$

$$n = U1 / U2$$

$$n = 10^{aw/20} \cdot \sqrt{V}$$

b) Symmetrical

$$R1 = \frac{n^2 - 1}{2n} \cdot Z_w$$

$$R2 = \frac{n+1}{n-1} \cdot Z_w$$

$$V = Z_{w1} / Z_{w2} = 1$$

$$R1 = R3$$

$$n = U1 / U2$$

$$n = 10^{aw/20}$$

The impatient practical constructor should, however, look up the following table. It gives attenuation values up to 20 dB. If larger values of total attenuation are required, they should be realised by connecting several values in series to make up the total value i.e. no single pad should have a greater attenuation than 20 dB, e.g. a total value of 35 dB could be made of a 15 dB and a 20 dB pad. At single attenuations above 20 dB, the resistors R1 and R3 are tending to low-value resistances and R2 is tending towards being a high-value resistance. This means that the reactance of the self-capacitance across R2 tends to the same order of R2 itself thus tending to leak the RF over the attenuator and eventually nullifying its effect as the frequency increases to around 1 GHz.

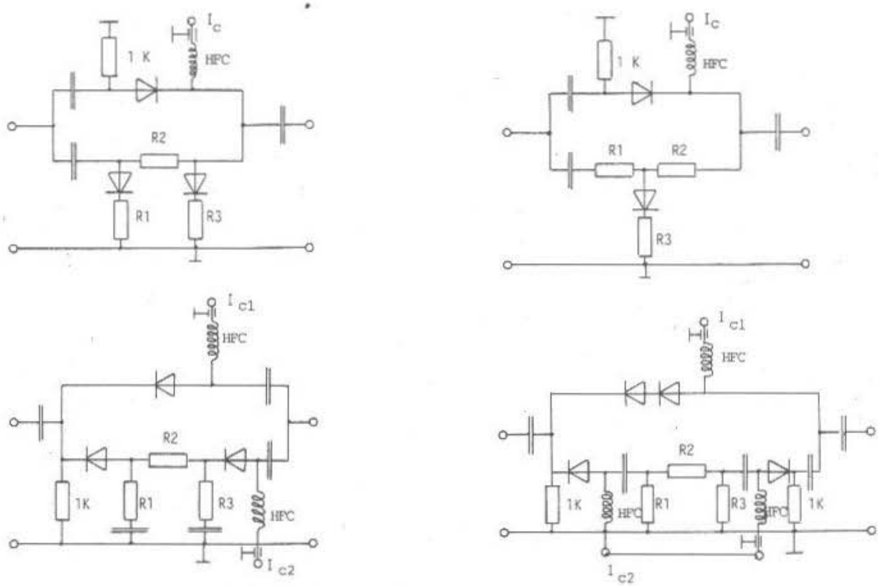


Fig. 16: Four simple methods of switching a four-pole network

aw / dB	R1, R3 / Ω	R2 / Ω
1 dB	896.5 Ω	5.8 Ω
2 dB	436.2 Ω	11.6 Ω
3 dB	292.4 Ω	17.6 Ω
4 dB	221.0 Ω	23.8 Ω
5 dB	178.5 Ω	30.4 Ω
6 dB	150.5 Ω	37.4 Ω
7 dB	130.7 Ω	44.8 Ω
8 dB	116.1 Ω	52.8 Ω
9 dB	105.0 Ω	61.6 Ω
10 dB	96.2 Ω	71.2 Ω
11 dB	89.2 Ω	81.7 Ω
12 dB	83.5 Ω	93.2 Ω
13 dB	78.8 Ω	106.1 Ω
14 dB	74.9 Ω	120.3 Ω
15 dB	71.6 Ω	136.1 Ω
16 dB	68.8 Ω	153.8 Ω
17 dB	66.4 Ω	173.5 Ω
18 dB	64.4 Ω	195.4 Ω
19 dB	62.6 Ω	220.0 Ω
20 dB	61.1 Ω	247.5 Ω

Table 1: Determination of the resistors for a 50 Ω pi-attenuator

$Z_{W1} = Z_{W2} = 50\ \Omega$

5. REALIZING A SWITCHED ATTENUATOR

By means of two change-over switches – SPDT or SPMT – a four-pole network may be switched in quite a simple fashion.
If the attenuators are switched in series then the SPDT switch of **fig. 15a** is required. .

Parallel switching of four-pole networks makes use of the SPMT switches as in **fig. 15b**.

In order to switch a four-pole network on or off two SPDT switches are required. It is a simpler process when one of the switch functions is combined with the four-pole network itself. The principle is always the same, namely:

- Input and output of the network are switched at high impedance.
- The high-impedance network can be short-circuited.

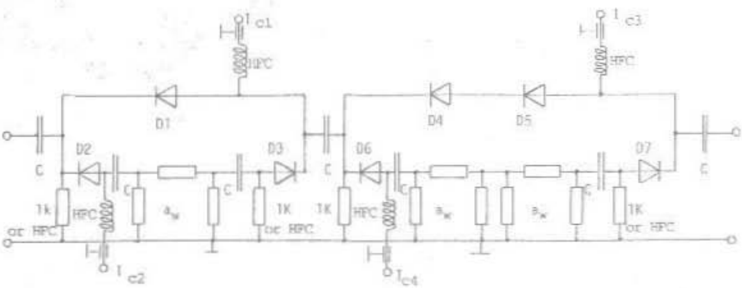


Fig. 17: Switched attenuator 0/20/40 dB (2 x 20 dB attenuators)

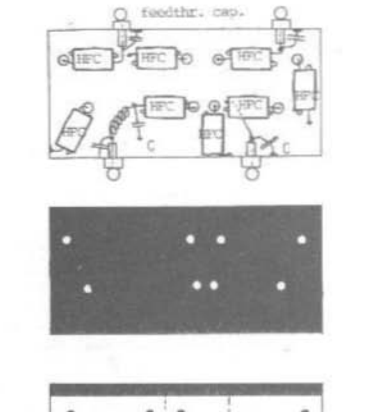


Fig. 18: Suggested construction for an electrically-switched attenuator in strip-line technique
1. Unit: a to 20 dB
2. Unit: a to 40 dB (i.e. 2 x 20 dB)

The review of **fig. 16** shows a few of the possibilities. **Fig. 16a and 16b**: These depict the introduction of the control source to effect the switching. This is accomplished by changing the polarity of the control current I_c, the attenuator being either activated or short-circuited. The control-current source must be able to reverse the polarity of the output current.

Fig. 16c and 16d: The polarity here plays no part in the switching, it is done simply by the presence or otherwise of the input to be switched.

The circuit of **fig. 17** has been used for some time now. The resistances of the Pi-network are galvanically separated in order that no DC can flow, otherwise the resistors would be overheated. The method of operation of this 20 dB attenuator is as follows:–

a) No attenuation: A 12 V potential is applied to the control input and diode D1 conducts. Diodes D2/D3 then receive a reverse bias. The attenuator input is now high impedance. Owing to a negative bias on control input 2, diode D2 switches to a slightly higher impedance. The lower the impedance of D1 and the higher the impedance of diodes D2/D3, the better the change-over action.

b) Attenuation selected: The control voltage potential is changed. Control input 1 now receives – 12 V, diode D1 is blocked and diode D3 conducts. A potential of + 12 V at control input 2 allows D2 to become conductive and the four-pole network is switched into circuit. The desired value of attenuation is now dependent upon the negative bias on diode D1.

Both of the 1 kΩ HF resistors used for the control-current feed can be replaced by high-frequency chokes. The diode current can then be increased thus achieving the maximal performance.

The isolation of the shunt diode is not able to bridge attenuations of more than about 30 dB. This was determined by experiments with the BA 379/479. Fixed attenuators up to 20 dB are able to be shunted with a PIN diode but higher values, say up to 40 dB may be bridged with two PIN diodes in series in order to increase the isolation.

A switchable 20/40 dB attenuator is shown in **fig. 17**. The 40 dB is realized by 2 x 20 dB elements for the reasons expounded earlier. A metal screening wall divides the two four-pole elements thus preventing mutual interference. This measure enables the full 40 dB to be held constant over the whole of the frequency range.

A step attenuator may be formed by the series of switchable fixed elements.

In order to adjust the attenuation to the desired value, the necessary electronics must be developed – but more of that later.

6. A PRACTICAL PIN-DIODE ATTENUATOR

Stripline techniques are used for a practical attenuator. The stripline's characteristic impedance depends upon the basic PCB material and from the width of the copper track. Using normal 1.5 mm thick epoxy, the following values may be taken:

Characteristic Impedance Z_w	PCB track-width
75 Ω	1.3 mm
60 Ω	1.9 mm
50 Ω	2.7 mm

Fig. 18 shows a possible layout, again in stripline technique. The printed circuit board fits into a proprietary tin-plate housing. The layout and the supply wiring are not quite optimal for the extended frequency range, but nevertheless, the module gives very good results.

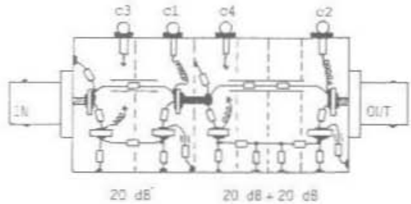


Fig. 19: Suggested construction for a step-switched attenuator in coaxial form. Not usually employed for wideband operation

If it is preferred to construct the PIN switch in a tin-plate box using free wiring, it is best to use the coaxial form of construction – the PIN diode running coaxially in a copper tube as in **fig. 19**. The diode lead wires must not be left too long as this will adversely effect the characteristic imped-

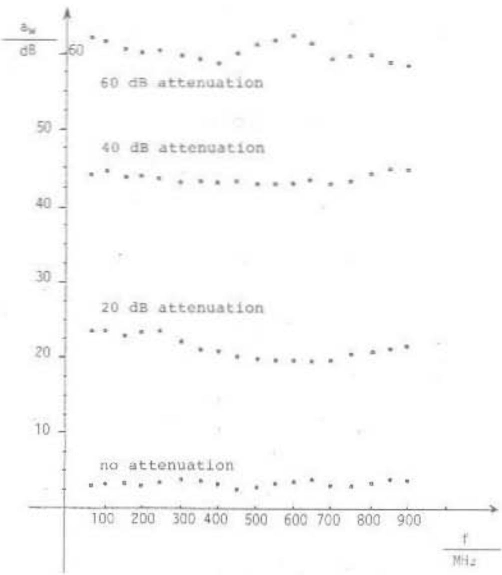


Fig. 20: Test results for an experimentally-constructed switched attenuator as described in text. (20 mA control current)

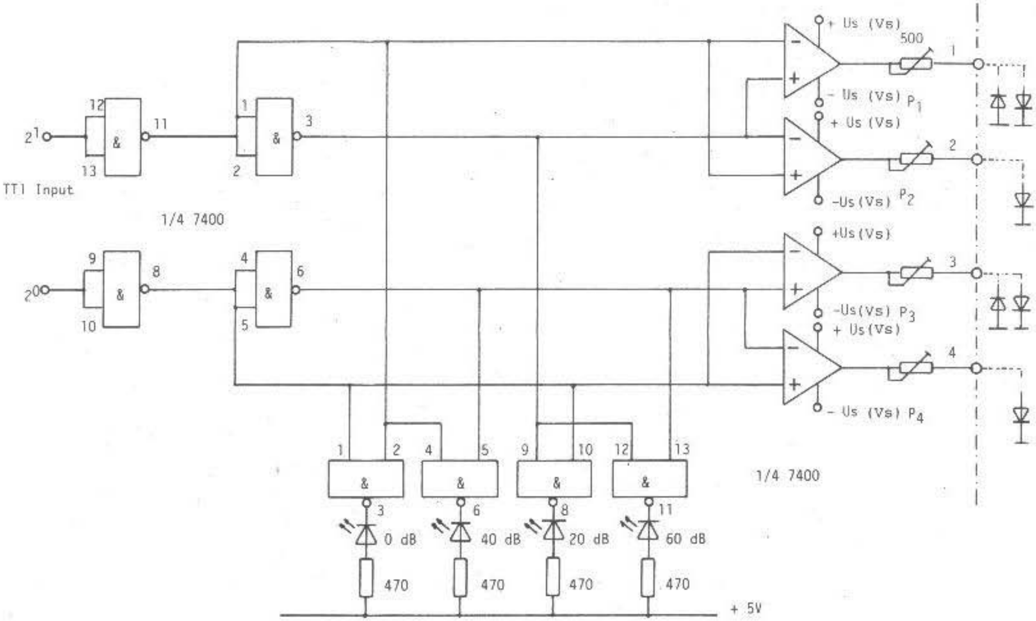


Fig. 21: Simple Interface for the PIN-diode switch
Supply voltage for 7400: + 5 V
for LM 324: ± 12 V
for LEDs: + 5 V

ance. As well as the danger of HF reflection points, mentioned earlier, there is also skin-effect losses to be considered. The higher the working frequency, the more these effects become prominent.

The great advantage of the stripline approach is, that at all points along the line, the characteristic impedance is closely defined. Only the components can cause trouble in this respect but the amateur can do nothing about it normally. The special chip components which harmonize with the stripline technique and homogenize the construction, are not universaly available. The test results show, however, that even without special components a good attenuator can be constructed.

6.1. The Printed Circuit Board

The copper under the Pi elements is not, in this version, etched away. The capacitance to ground tends to linearize the attenuation curve. All DC supply lines are effected via HFCs and the feed-in is carried out on the solder side by means of small air-supported chokes (4 turns wound on a 3.2 mm drill). The feed-through capacitors must be shunted by small high-quality capacitors in order to improve the low-pass effect – the value is uncritical, but in the nF region.

1 nF plate capacitors are used for coupling purposes. The 1/4 W resistors of the Pi network and also the PIN diodes are soldered directly on the top of the board's surface, no holes being necessary for the leads. All the cold ends of the Pi-networks are however, connected through to the ground plane via drilled holes. The screening wall is important in the interests of a good attenuation characteristic.

6.2. Test Results

The step-switched attenuator was constructed according to the diagrams of **figs. 17 and 18** and supplied the following results:

The precision, i.e. the linearity of attenuation over the entire frequency range as well as its stable input return-loss is dependent upon the method of construction and the components utilized. The most important single factor is the method of introducing the DC control current.

Using the recommended construction, the test results depend heavily upon the manner in which the DC was fed in. The test results for the given construction can be seen in **fig. 20**. It indicates the attenuation versus frequency for various system attenuations. When zero attenuation is selected, the residual attenuation is approximately 2 dB.

6.3. Parts List

All the low-pass filters allow a control current of 50 mA to be achieved.

1 Tin-plate box 74 x 37 x 30 mm
2 N-panel mounting connectors
1 Printed circuit board, two-sided etched as in fig. 18
Screening walls, as in fig. 18
8 Valvo 6-hole wideband VHF chokes
8 Self-supporting 4 turn 3 mm dia chokes
7 Plate capacitors, approx. 1.5 nF
7 PIN diodes BA 479 G; $I_{max} = 50$ mA (only 25 mA used here)
4 Feed-through capacitors approx. 1 nF or more
9 Resistors (3 x 3 resistors of the 20 dB Pi-networks of table 1)

Other items: Additional chokes, ferrite beads and capacitors as required in order to improve the low-pass filters.

7. SIMPLE CONTROL CIRCUITS

The electronically-switched attenuator requires an additional supply which is able to switch the polarity under the influence of TTL potentials.

The simplest solution is to use an operational amplifier in a comparator circuit. According to the input signal level, the output delivers a voltage which is either positive or negative.

A simple interface is shown in **fig. 21** which utilizes the 4 x operational amplifier LM 324. This is able to deliver the required 20 mA, or more, current required but if a power op-amp is available, so much the better.

The two TTL inputs of the interface allow four possibilities. The selected attenuation is indicated by an LED. By means of a pre-set potentiometer in the operational amplifier output, the required control current can be set as desired. The four outputs are connected with the appropriate control inputs of the attenuator module.

8. CONCLUSION

The electronically-switched attenuators are also usable in the high-frequency range. The control current must, however, be tailored to match the working frequency. The PIN-diode data sheet must be consulted to determine the lowest-permissable frequency of operation.

This article cannot represent all that one needs to know on the subject of PIN diodes for amateur purposes but it might at least serve to stimulate the contructor into practical experimentation. If the version described here is faithfully copied, a fully functional instrument will be aquired – and with good specifications.

9. REFERENCES

(1) A. Claar:
Regelbares PIN-Dioden-Dämpfungsglied
cq DL Heft 11/85

(2) H. Toll:
Bauelemente der Halbleiterelektronik, Teil 1,
Teubner Verlag

(3) Erich Renz,
PIN- und Schottky-Dioden,
Hüthig Verlag

(4) National Semiconductor,
Linear Databook: DAC 08xx

BRIEFLY SPEAKING...

High Performance Variable Capacitors are not yet obsolete. It is, thanks to the efforts of DL 3 FM, that an on-going source for these components is still available.

A series of air-spaced and ball-bearing variable capacitors is available in many forms, single to four-gang, from less than 10 pF to greater than 500 pF, as well as trimmers and slow-motion drives.

HATRA-Elektronic,
8124 Seeshaupt,
West-Germany,
Tel.: 08801 - 1269

Alois Pendl, OE 6 AP,
Co-founder of UKW-Berichte, a regular exponent of the higher frequencies and through EME, as well as numerous contact firsts, is renowned far outside the Austrian borders.

Following the drawing of the guide-lines for the awarding of ÖVSV (Austrian national radio society) honours and the newly-introduced performance shield at the 1987 AGM, it was decided to give **OE 6 AP** the shield 01 for his outstanding performance in the VHF, UHF and SHF bands. The shield 02 was awarded to club-station **OE 5 XXL** for an above average performance over the year in VHF, UHF and SHF contests. Only a maximum of three performance awards are given per year.

Congratulations from
VHF COMMUNICATIONS/
UKW Berichte!

Plastic Binders for VHF COMMUNICATIONS

- Attractive plastic covered in VHF blue
- Accepts up to 12 editions (three volumes)
- Allows any required copy to be found easily
- Keeps the XYL happy and contented
- Will be sent anywhere in the world for DM 9.00 including surface mail

Order your binder via the national representative or directly from the publishers of VHF COMMUNICATIONS (see below)

Reduced Prices for VHF COMMUNICATIONS !!!

	Volume	Individual copy
Subscription 1988	**DM 25.00**	**each DM 7.50**
VHF COMMUNICATIONS 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, 2/1971, 1/1972, 2/1972, 4/1972,	each DM 2.00
2/1973, 4/1973, 1/9174, 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!

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