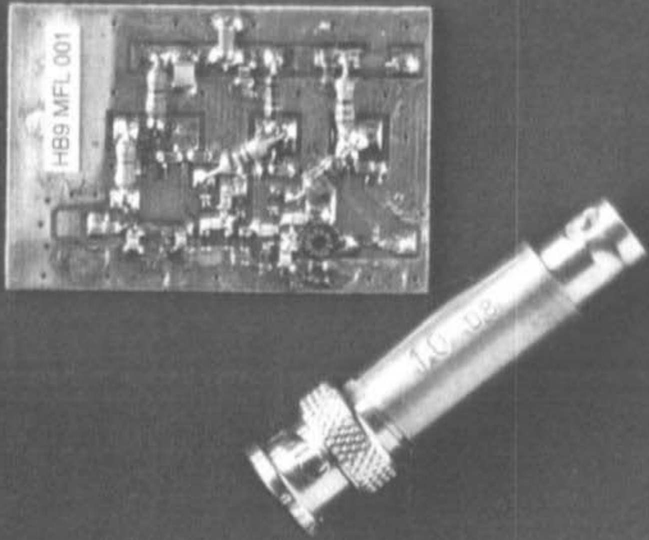
A Publication
for the Radio-Amateur
Especially Covering VHF,
UHF and Microwaves

VHF communications

Volume No. 20 · Spring · 1/1988 · DM 7.50

Using the SMD Technology for a 23 cm amplifier yields 200 mW output, 34 dB gain



VHF communications

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

Volume No. 20 · Spring · Edition 1/1988

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 1988 subscription price is DM 25.00 or national equivalent per year. Individual copies are available at DM 7.50 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
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 Michél, F 5 SM, SM Electronic,
20 bis, Avenue des Clairions, F-89000 AUXERRE
Tel. (86) 46 96 59

Finland
Peter Lytz, OH 2 AVP, Gesterbystingen 14 E 49
SF- 02410 Kyrkslätt, Tel.358/0/2 98 17 61
SRAT, PL 44
SF-00441 Helsinki, Tel. 358/0/5625973

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 Cortés 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

Konrad Hupfer, DJ 1 EE

Wideband Power-Divider / Combiner for the 2 m and 70 cm Bands

During the latest VHF meetings in Munich and Weinheim (1985/86), transistorised VHF/UHF amplifiers were introduced which were capable of amplifying broadband across both 2 m and 70 cm. Most of them employed push-pull circuits which, with suitable wideband transformers, enabled the generation of 100 W output signals in the range 144, 146, 430 - 440 MHz.

The object of this article is to describe the so-called power-divider or power-combiner, by means of which, two similar wideband power stages may be connected in parallel.

For frequencies above 100 MHz the so-called Wilkinson-dividers have been employed, together with the well-known 90°-3 dB directional coupler, for the division or addition of RF powers.

The simplest arrangement consists of two 70 Ω lines with an electrical length of λ/4 shown in **fig. 1**, (1,4).

Load R_{L1} = 50 Ω (e.g. power-amplifier input) is transformed through Z_{L1} at point 1 to R_{e1} = 100 Ω. In the same manner R_{L2} is transformed with the aid of Z_{L2}. As R_{e1} and R_{e2} are in parallel at point 1, the port impedance at R_e is 50 Ω.

The necessary de-coupling is carried out by means of the balancing resistance R_A. The de-coupling of the two branches R_{L1} and R_{L2} is required to prevent mutual coupling from causing instability. Assuming, for example, that there was an HF input at 2, this would cause a signal at 3 which had a 180° phase reversal when considering the path via Z_{L1} and Z_{L2}. The direct path 2 to 3 via R_A effects a 0° phase-shifted signal at 3 which results in a cancellation of the signals at point 3. In general, the following relationships shown in **fig. 2** apply.

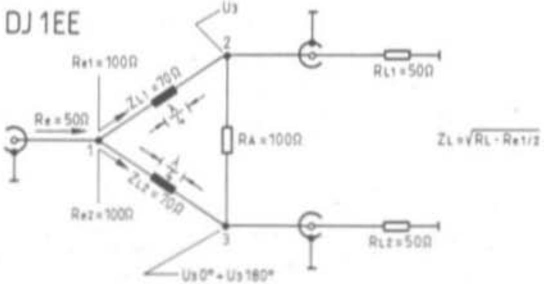


Fig. 1: Wilkinson power divider principle

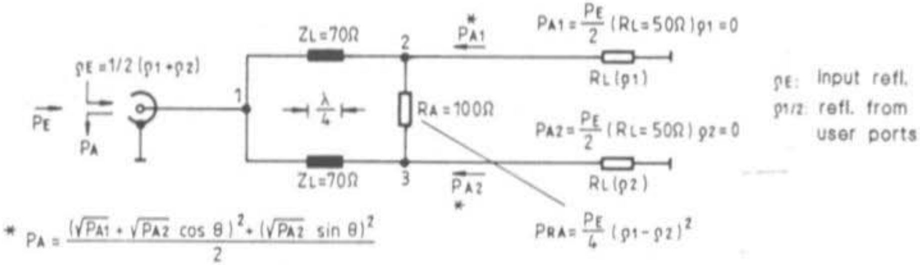


Fig. 2: Relationship between reflection coefficient and power at input and outputs
***) Valid also for operation as combiner**

The arrangement is, of course, reciprocal; two similar in-phase powers fed to P_{A1}, P_{A2} appear added together at 1 (P_A).

The same sort of divider/combiner is, in this simple arrangement, narrow-band (bandwidth ca. 30 %) and only suitable for either the 2 metre or the 70 cm band. If, however, several λ/4 segments are cascaded, then larger bandwidths become possible. For example, using seven λ/4 line pairs, the bandwidth ratio fH/fL = 10! (3).

For a bandwidth of 140 - 440 MHz (fH/fL = 432/144 = 3), only three λ/4 wave sections are required. A computer may be employed to optimise the performance in the ranges 140 - 150 MHz and 420 - 440 MHz as the device is not required for the frequencies inbetween.

A wideband power-divider with a single input and two outputs as in **fig. 4** can be formed by two single λ/4 wideband transformers (**fig. 3**) which have been connected together at one end.

According to (2,6) three λ/4 wave sections of the appropriate impedance are sufficient to obtain a wideband match (VSWR ~ 1.3) from 50 Ω to 100 Ω in the band 140 - 440 MHz.

Carrying out the principle of fig. 1, the two single transformers can now be connected together to form a power-divider/combiner as in **fig. 4**.

Without incurring any great penalties in bandwidth and matching, the simplification of **fig. 5** may be carried out.

Fig. 3:
Two single wideband transformers
50 Ω ⟷ 100 Ω from 3 x λ/4 sections

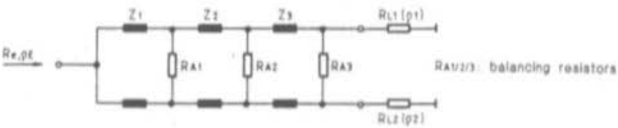


Fig. 4:
Wideband Wilkinson divider from two single wideband transformers

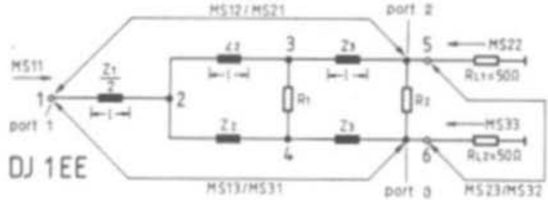


Fig. 5: Simplified (from 6 to 5 λ/4 sections) wideband Wilkinson divider

With the aid of a general purpose network computer program, called Super Compact PC (5), the circuit can be optimised to obtain the best matching and isolation in the two bands of interest. 140 - 150 MHz and 420 - 440 MHz.

It is not necessary to be a computer specialist to use this Super Compact PC program. As a demonstration, the circuit description (file) for the computer is reproduced in **table 1**. The circuit is allocated with nodes, between which, the individual components (**fig. 5**) are added.

Resistors: R1 = 100 Ω; R2 = 150 Ω
Stripline: Epoxy ε_r = 5.2;
Substrate 1.6 mm thick,
l_A to 135 mm calculated from the centre frequency and given to the computer as a "start value".
HU = ∞ (**fig. 6**)

The optimised values for the stripline lengths and widths are printed in lines 2 and 3 of **table 1**.

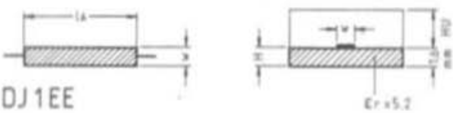


Fig. 6: Defining the stripline's construction. Fixed values for ε_r and substrate thickness H. HU: Information for the computer when constructing the stripline in the screened housing

```
   SUPER COMPACT PC  01/03/80

   *BREITBANDTEILER 145/435MHZ
   *FILE: HU\WTB2.CKT *

1  BLK
2  P1:?130.71MM?
3  W1:?4.1944MM?
4  W2:?1.3529MM?
5  W3:?2.3558MM?
6  R1:100
7  R2:150

8  TRL 1 2 W=W1 P=P1 TEF
   TRL 2 3 W=W2 P=P1 TEF
   TRL 2 4 W=W2 P=P1 TEF
   TRL 3 5 W=W3 P=P1 TEF
12 TRL 4 6 W=W3 P=P1 TEF
13 RES 3 4 R=R1
14 RES 5 6 R=R2
15 A:3POR 1 5 6
   END
17 FREQ
   STEP 140MHZ 440MHZ 20MHZ
   END
20 OUT
21 PRI A S
   END
23 OPT
   A
   F=140MHZ     MS11=-25DB LT
                MS33=-25DB LT
                MS32=-25DB LT
26 + F=440MHZ   MS11=-25DB LT
                MS33=-25DB LT
                MS32=-25DB LT
   END
28 DATA
29 TEF: MS H=1.6MM HU=10MM ER=5.2
   END
```

Table 1

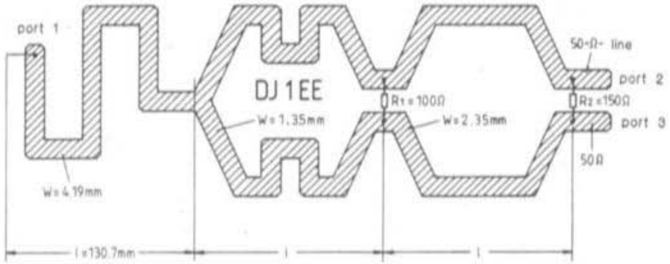


Fig. 7: Proposal for the layout form of the wideband Wilkinson divider I, w: The lengths and strip widths of the individual λ/4 striplines are given by the computer

The divider/combiner can be constructed in stripline form according to the arrangement shown in **fig. 7**, using double-sided epoxy board, 1.6 mm substrate thickness.

Now a few words about the balancing resistors R1 and R2. These may be obtained from the US firm KDI (Germany: Microscan) for 30 to 80 DM (!). They are high-power resistances on a berylium substrate and possess a very small stray capacity (< 1 pF). In normal operation the asymmetry (the unbalance of the parallelled amplifiers) is mostly so small that, for amateur purposes, small power resistors may be connected in parallel.

The results for the transfer functions MS 21/ MS 12, and the input and output reflections MS 11/MS 22/MS 33 are given in **table 2** and in **figs. 8 and 9**.

SUPER COMPACT PC 01/03/80

Freq MHz	MS11 dB A	MS12 dB A	MS13 dB A	MS21 dB A	MS22 dB A	MS23 dB A	MS31 dB A	MS32 dB A	MS33 dB A
140.000	-37.65	-3.01	-3.01	-3.01	-24.99	-25.55	-3.01	-25.55	-24.99
160.000	-26.12	-3.02	-3.02	-3.02	-32.94	-23.52	-3.02	-23.52	-32.94
180.000	-19.72	-3.06	-3.06	-3.06	-32.26	-20.16	-3.06	-20.16	-32.26
200.000	-16.70	-3.10	-3.10	-3.10	-22.66	-18.75	-3.10	-18.75	-22.66
220.000	-14.99	-3.15	-3.15	-3.15	-18.24	-18.84	-3.15	-18.84	-18.24
240.000	-14.00	-3.19	-3.19	-3.19	-15.66	-20.38	-3.19	-20.38	-15.66
260.000	-13.45	-3.21	-3.21	-3.21	-14.14	-24.00	-3.21	-24.00	-14.14
280.000	-13.21	-3.22	-3.22	-3.22	-13.42	-33.38	-3.22	-33.38	-13.42
300.000	-13.20	-3.22	-3.22	-3.22	-13.38	-32.25	-3.22	-32.25	-13.38
320.000	-13.42	-3.21	-3.21	-3.21	-14.02	-23.49	-3.21	-23.49	-14.02
340.000	-13.94	-3.19	-3.19	-3.19	-15.43	-19.91	-3.19	-19.91	-15.43
360.000	-14.88	-3.15	-3.15	-3.15	-17.85	-18.30	-3.15	-18.30	-17.85
380.000	-16.49	-3.11	-3.11	-3.11	-21.94	-18.04	-3.11	-18.04	-21.94
400.000	-19.31	-3.06	-3.06	-3.06	-29.83	-19.17	-3.06	-19.17	-29.83
420.000	-25.05	-3.02	-3.02	-3.02	-32.40	-22.14	-3.02	-22.14	-32.40
440.000	-41.80	-3.01	-3.01	-3.01	-25.01	-26.34	-3.01	-26.34	-25.01

End of table. Hit any key to continue ....

Table 2: Values of input and output reflections and the transfer resp. isolation between individual ports

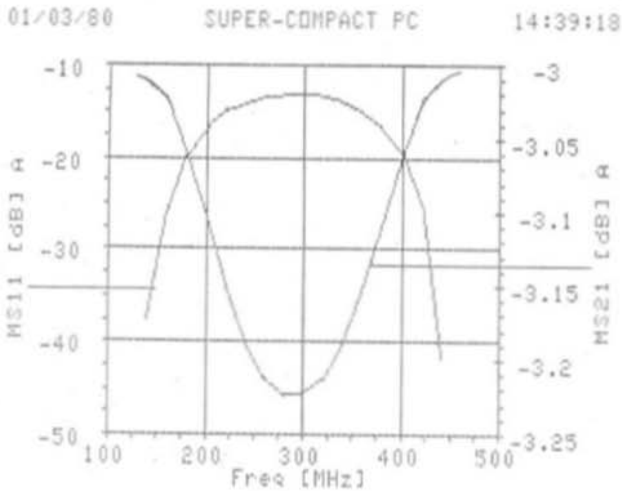


Fig. 8:
Transfer characteristic port 1 → port 2 (3 dB) and input reflection

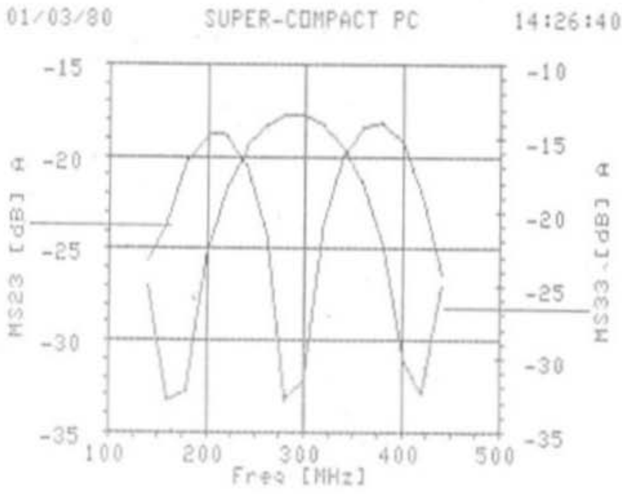


Fig. 9:
Isolation between ports 2 - 3; output reflection from port 3

Explanation of required "Computer file"

1) Name of computer block diagram with nodal points
2) P1: Length of stripline; between 2 question marks is entered the start length l_A, the optimal result is already given. The values between the two question marks mean for the computer that it may alter them in the optimisation process.
3) Width of line 1
4) Width of line 2

5) Width of line 3: also here the estimated start values are given from (2).
6) Balancing resistor R1 = 100 Ω = const.
7) Balancing resistor R2 = 150 Ω = const.
8) A line (TRL) between node 1 and 2 of width W1, length P1 and of electrical-mechanical construction given in the DATA lines.
12) Ditto
13) Resistor R1 = 100 Ω lies between node 3 and node 4.
14) Resistor R2 = 150 Ω lies between node 5 and node 6.
15) Circuit A is a 3-port with input 1 (port 1) and outputs 5 and 6 (ports 2 and 3).
17) The calculation is effected from 140 to 440 MHz in 20 MHz steps.
20, 21) The result given on screen or printer is the S-parameters (dB) for circuit A.
23, 26) The circuit A is optimised for both 140 MHz and 440 MHz; The reflection coefficient and isolation should be at least 25 dB.
28, 29) Data about the construction and material of the stripline

LITERATURE

(1) E. J. Wilkinson
"An N-Way Hybrid Power Divider"
IRE-Trans. MTT-8, Jan. 1960, p 116 - 118

(2) L.Young:
Tables for cascaded quarterwave transformers, IRE-Trans. Microwave Theory and Techniques, MTT 7, April 1959, p 233 - 238

(3) K. Mayer:
"Mehrstufige λ/4-Transformatoren",
AEÜ Bd. 21, 1967, H. 3

(4) CAD/CAE-Design of Wideband Wilk. Divider
Microwave Journal Nov. 1984

(5) Handbuch Super-Compact PC
Fa. TSS, K. Eichel, D-7012 Weissenhorn
Fa. Microscan, D-8045 Ismaning

(6) Mattaei, Young:
Microwave Filters, Impedance Matching Networks and Coupling Structures
McGraw Hill Book Co

Dr. med. Hans Schlüter, DJ 7 GK

Rear-Feed Dish Radiator with Corrugated Horn

The most widely employed feed horns for 10 GHz band, paraboloid antennas are the "dish radiator" (1)(2), the dipole radiator (3), the "cap radiator" (1) and the simplified cap radiator (4). The common item of these arrangements is the rear feed undertaken by a section of R 100 (WG-16) waveguide with a suitable reflector system which diverts the RF energy into the mirror aperature.

With the exception of the Cassegrain systems (which exhibit shadowing losses with relatively small dimensioned antennas), these focuses exhibit only a very mediocre illumination efficiency of maximum 55 %.

An increase in the illumination efficiency may be effected by the utilisation of a "grooved" or "corrugated horn" (5). The use of a rear-mounted feed system is not, however, usual with this type of antenna. A little consideration and a few experiments resulted in the finalised construction described below.

The waveguide feed (R 100, WG-16) is closed via a 1 λ diameter disc and also provided with $\lambda/2$ by $\lambda/20$ side slots in the same manner as the "dish radiator" suggested in (1) and (2). It is then provided with a flange as shown in **fig. 1**.

The original polar diagram, which resembled that of a dipole with reflector, is altered by the addition of the flange so that it is a nearby wedge-shaped reflected illumination. The improvement in the illumination efficiency can amount to up to 90 % according to (6) and (7) but here about 70 % can be relied upon. With the aid of a detector, I have measured a 66 % increase in field strength at a point only a few metres from the antenna. In open-air tests, DL 1 VX measured a signal strength increase of 2 to 3 dB at a distance of 20 km from the antenna.

A tapering of the wave guide end, as mentioned in (1) and (3), is not absolutely necessary in this arrangement. Any slight mis-matches may possibly be ironed-out by means of a three-screw tuner.

The flange was made adjustable by means of two pieces of copper tubing which telescoped into each other so that experiments with the plane of the horn edge flange could be carried out. Up to now, it has resulted in the same plane as the plate. At this point it may be mentioned that a parabolic reflector of D = 62 cm and f/d = 0.37 (F at 23 cm distance) was used.

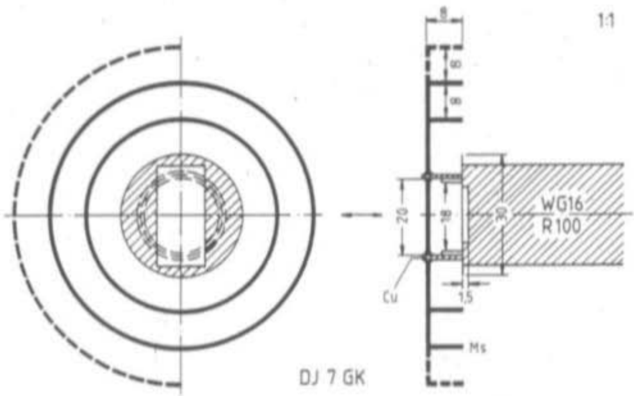


Fig. 1:
Dish radiator for 10 GHz with corrugated horn

At the SHF meeting in Notzing/BRD, on 27th June 1987, a series of exciters were tested on the same reflector (D = 60 cm, f/d = 0.37) at 10 GHz. The increase in gain, on account of the fitting of the flange to the **"dish radiator"**, was **2.2 dB**. This illumination was within measurements tolerances which may be considered to be the same as the cap radiator with tapered waveguide, as in (1) together with a rearfeed circular waveguide arrangement (20 mm internal dia.) with reflector disc by DK 2 RV. When the flange was fitted to **the latter**, an increase of gain by a **further 2 dB** was apparent. The **"simplified" cap radiator (without waveguide tapering)** was clearly inferior and gave the same results as the dish radiator without flange.

Conclusion

The widely used "dish radiator" can, by means of the introduced flange, be noticeably improved. If one has the opportunity to work with circular wave guides, the results may be even better, as a frontal feed is not then mandatory. This will be the starting point for further interesting experiments.

DK 2 RV is warmly thanked for the availability of the literature and DL 2 DO for the prompt acceptance of the idea and the carrying out of his own experiments leading to similar results.

REFERENCES

(1) Reithofer, J., DL 6 MH:
Praxis der Mikrowellen-Antennen
Verlag UKW-BERICHTE 1987

(2) Jessop, G. R.:
VHF/UHF Manual, 4th Edition; 9.73 - 9.77
Radio Society of Great Britain 1983

(3) Griek R. and Münich M.:
A 3 cm Primary Radiator for Parabolic Antennas
VHF COMMUNICATIONS Vol.11,
Ed 2/1979, P. 74 - 75

(4) Heidemann R., DC 3 QS:
A Simple Radiator for 3 cm Parabolic Dishes
VHF COMMUNICATIONS Vol. 11,
Ed. 3/1979, P. 151 - 153

(5) Vidmar M., YU 3 UMV:
TV Satellite Receive System, Part 1
VHF COMMUNICATIONS Vol. 18,
Ed. 4/1986, P. 194 - 213

(6) Hawkins, D. J.:
Electronic Letters, Vol. 8 No. 19, Aug. 1972

(7) Geyer, H.:
Runder Hornstrahler mit ringförmigen Sperrtöpfen zur gleichzeitigen Übertragung zweier polarisationsentkoppelter Wellen.
Frequenz 20 (1966) Nr. 1, S. 22 - 28

Microwave Power Transistors

The following article will review some silicon and gallium-arsenide power devices and also take a look at further developments.

Silicon Power Devices

Point-to point and Satellite Communication

Within the classical fields of radio communication in tropospheric scatter, satellite etc. there are power transistors available for the particular application. The narrow-band communication starts at about 700 MHz and, with FM modulation, continues on into the 3 GHz band and digital modulation rates of about 34 Mbit. Typical power classes of output transistors for this application are given in **table 1**.

For special communication bands, these power transistors can be internally matched at both input and output to suit the particular band in use. In this way, higher performance may be obtained (**table 2**).

The tables only show the output stage transistors but of course, the corresponding penultimate and pre-amplifier transistors are available well into the small signal range.

VHF Silicon Field Effect Transistors

These transistors find employment in the frequency range 200 to 500 MHz on board ships, in tropo-scatter and in ECM systems. High continuous power ratings at bandwidths of over one octave are required here.

Type	frequency range GHz	single transistor output pwr W	typ. transistor power W	efficiency %	remarks
MSC 89020	1	20	10	55	
MSC 82010	2	10	10	35	TV, DPCS*
MSC 83005	3	5	2 x 5	30	DPCS, FM
MSC 4003	4	2.5	4 x 2.5	25	DPCS

* DPCS: digital point-to-point comms. syst.

Table 1: Silicon power transistors for use of point-to-point communications

Type	Freq. range MHz	output power W	efficiency %
AM 80610-50	750 to 960	50	50
AM 81416-20	1400 to 1600	18	45
AM 81618-20	1600 to 1800	16	45
AM 81821-18	1800 to 2100	18	40
AM 82023-16	2000 to 2300	16	40
AM 82327-15	2500 to 2700	15	24

Table 2: Internally matched transistors for broadcast radio

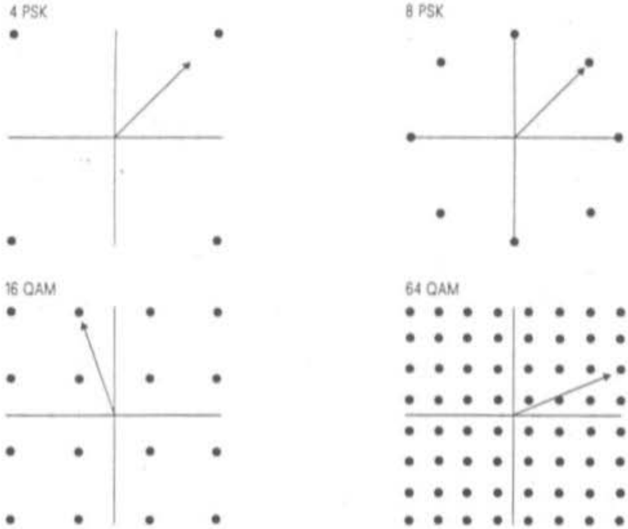
Type	freq. range MHz	power W	efficiency %
MSC 0204-150 F	225 to 400	150	50
MSC 0105-100 F	100 to 500	100	50

Table 3: Silicon FETs for the VHF range

Type	freq. range GHz	output power W	gain at 1 dB compression point dB
MSM 3442-10	3.4 to 4.2	10	7
MSM 5964-10	5.9 to 6.4	10	7
MSM 6472-10	6.4 to 7.2	10	7
MSM 7177-10	7.1 to 7.7	10	7
MSM 1112-10	10.7 to 11.7	10	6
MSM 1213- 2	12.7 to 13.3	2	5.5
MSM 1718- 1	17.7 to 18.7	1	5
MSM 1920- 1	18.7 to 19.7	1	5
MSM 2223- 1	22.4 to 23.6	1	5

Table 4: GaAs power transistors for the communication bands (intern. matched)

The field effect transistor, on account of its considerably higher impedance, is able to fulfil this requirement very adequately (as opposed to the bipolar transistor). Two representatives of this performance class are shown in **table 3**. Naturally the necessary penultimate and pre-amplifier transistors are also available.



Modulation techniques and vector diagrams

GaAs Power FET Transistors

Above 3 GHz silicon devices are out of the running on technical grounds. GaAs comes into its own here with its four-fold higher charge carrier mobility.

A new family of GaAs class A, power FETs was developed by MSC (Microwave Semiconductor Corporation: a Siemens AG concern) after a huge investment program. They are intended for the future vector modulation communication system at 4 PSK, 16 QAM, 64 QAM entailing extreme demands on phase and amplitude linearity. In order to achieve this, the transistors are driven some 5 to 10 dB below their saturated power condition. Data from a few power devices may be seen in **table 4**. Also, the full range of penultimate and pre-amplifier transistors are available for all bands.

These power transistors have structures of below 1 μm (gate length) whilst the active zone length can be as long as 10 mm (gate width). Frequencies above 25 GHz have structures of about 0.5 μm and under requiring production facilities such as electronic beam writing and X-ray illumination.

Conclusion

Silicon power transistors (bipolar and FET) have consolidated their position with, above all, higher continuous and pulse powers and together with integrated matching make the design work much easier.

GaAs power devices and wideband systems open up new perspectives in respect of both power and frequency. They are the key elements of future system families. Today, on a world-wide basis, 30, 60 and 90 GHz systems are under consideration. Systems, whose demands that may possibly outstrip the capabilities of GaAs and call for the technologies of Gunn or Impatt diodes of Indium Phosphide (InP).

Extract from: "Microwellen Leistungsbauelemente – Schluessel zur Nachrichtentechnik"
by Konrad Poehl,
in Siemens Components 25 (1987), edition 2

Armin Meier, DC 7 MA

The SDA 4212 – the 4211's Successor

Prompted by the article "The SDA 4211 – An Interesting UHF Prescaler" (1), I approached the manufacturers for data about it. I then found out that this integrated circuit has now a (better) successor: The SDA 4212 (order no. Q67000-A8049).

Comparison of the two data sheets revealed the following observations:

Both types are similar inasmuch that they:

- are pin-compatible
- have the same working voltage and current consumption
- have switchable division factors of 64 : 1 and 256 : 1.

The differences between them are:

- The ambient maximum working temperature is higher: 0 – 80°C for 4212 and 0 – 70°C for the 4211.

frequency MHz	**4211** dBm	mV	**4212** dBm	mV
70	– 26	11	– 26	11
80	– 27	10	– 27	10
120	– 30	7	– 30	7
250	– 32	5.5	– 32	5.5
600	– 27	10	– 27	10
1000	– 27	10	– 27	10
1100	– 27	10	– 22	18
1200	– 21	20	– 15	40
1300	– 15	40	– 9	80

Maximal input voltage: 3 dBm/315 mV

Table 1: Comparison of the SDA 4211/4212 sensitivities taken from the manufacturer's data sheets

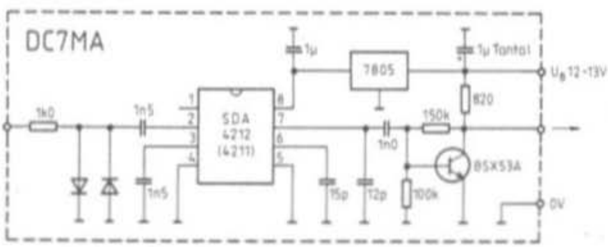


Fig. 1: UHF prescaler using the SDA 4212

- Smaller output voltage: 0.6 V_{PP} for 4212 and 1.0 V_{PP} for 4211.
- Better 3rd harmonic suppression: for 4212 30 to 35 dB and for 4211 13 – 20 dB.
- Lower sensitivity at the high end of the frequency range (**see table 1**).

The SDA 4212 was constructed in the test circuit of **fig.1**, the UHF prescaler being set to a division factor of 256 : 1. It is a part of a programmable frequency counter using MOS which I have developed to work at frequencies up to a maximum of 20 MHz.

It was found that frequencies down to 5 MHz could be counted with the same facility as the highest for my requirements: 4 x 435 MHz is 1.74 GHz. These Results were obtained with both the available examples and indicates that this prescaler could be capable of yet more performance. Unfortunately, owing to inadequate test equipment, levels could not be determined.

The prescaler boards were connected directly to the BNC input socket as shown in **fig. 2**.

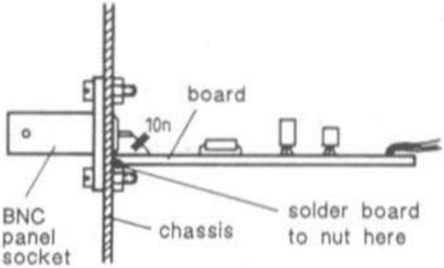


Fig. 2: Construction detail: Connection of input-BNC socket to printed circuit board

REFERENCES

(1) Jirmann J., DB 1 NV:
The SDA 4211 – An Interesting UHF Prescaler
VHF COMMUNICATIONS Vol. 19,
Ed. 2/1987, P. 109

Andreas Schaumburg, DF 7 ZW

Receiving METEOSAT with Yagis

This article describes the home construction of a Yagi antenna for the 17.7 cm band. The article is commenced with a theoretical consideration of the field strength to be expected.

1. SIGNAL-TO-NOISE MARGIN

According to ESOC, METEOSAT transmits with an output power of 5 W. The radiated power is 200 W (+ 53 dBm) with reference to an isotropic radiator. The path loss A_d is then

A_d = 32.4 dB + 20 log d + 20 log f
where d is in km
f is in MHz

With d = 40,000 km and f = 1700 MHz, the A_d = 189 dB.

A 3 metre long Yagi antenna can have a gain of 19 dBm i.e. 21 dBi at 1700 MHz (1).
The received power/voltage is then

$P_R = P_O - A_d + G_{ant}$ = 53 dBm − 189 dB + 21 dBi
= − 115 dBm/50 Ω ≙ 0.4 μV

The overall noise figure of the receiving installation is about 3 dB consisting of the following items: 2 dB converter noise figure (2), 0.5 dB cable attenuation (1 m RG-213) and an antenna noise temperature of approx. 35 Kelvin (pointed to cold sky). A 3 dB overall noise figure corresponds, according to (3), to a noise power of − 126.5 dBm at B = 30 kHz, Z = 50 Ω. The received level is − 115 dBm therefore the signal-to-noise margin is 11.5 dB.

At this value the FM threshold is exceeded, for the FM index used by METEOSAT, and therefore a noise-free picture can be expected.

Editor's note:
The author has assumed a typical METEOSAT radiated power which was available until May 1987. Since then (Sept. 87), it is somewhat smaller and can fall even lower. ESOC is obligated to exceed only the "worst-case" values. The worst-case EIRP is:
Ch. 1: 48.3 dBm; Ch. 2: 48.1 dBm.

Also, a value of 35 K for the antenna noise figure is a little optimistic. At least occasionally, the noise temperature is far higher.

2. ANTENNA CALCULATIONS

The antenna was dimensioned according to the article by DL 6 WU in (1). The wavelength is 17.7 cm

$$d/\lambda = 4\text{ mm}/177\text{ mm} = 0.0225$$

where d = director dia.

The director length was teken from fig. 2 in (1) and to this 2/3 of the boom diameter must be added (10 mm).

The elements spacings and the dipole were correspondingly reduced in size (17.7 cm/23 cm) from the 23 cm Yagi described in (1).

3. CONSTRUCTING THE ANTENNA

The overall arrangement is shown in **fig. 1**. **Table 1** contains all the dimensions. The 4 mm diameter aluminium rod can be cut to size with wire cutters and then trimmed to the exact size by means of a file. Using a vernier gauge, an accuracy of ± 0.25 mm may be easily achieved.

The holes for the directors, given in table 1, are marked out in line using a rule to an accuracy of ± 0.5 mm and drilled completely through using a 3 mm twist drill in a drill press. The holes are then drilled from both sides with a 3.9 mm drill. The elements can then be force-fitted – they must sit exactly central. A little care should be used here to ensure accuracy using frequent vernier-gauge measurements. If a few elements are a little loose fitting, they can be gripped tighter by centre-punching around the edge of the drilling in order to ensure a permanent fit.

The dipole is made from flat 4 mm x 2 mm copper stock bent into the required form. It is then fixed to the boom by soldering it to the head of a brass 4 mm screw which has been tapped into the boom at the appropriate position. This is shown in **fig. 2**.

The author soldered the balun together with the coaxial feed cable directly to the underside of the dipole as in **figs. 3 and 4**. This can, of course, be done with the aid of an N plug and socket as in (1).

No	length of elements	distance from reflector	No	length of elements	distance from reflector
–	106 x 106 mm	0	25	70.5 mm	1556 mm
–	Dipol	35 mm	26	70.5 mm	1626 mm
1	82.6 mm	49 mm	27	70.5 mm	1696 mm
2	80.8 mm	81 mm	28	70 mm	1767 mm
3	80 mm	120 mm	29	70 mm	1837 mm
4	79 mm	164 mm	30	70 mm	1908 mm
5	78 mm	214 mm	31	70 mm	1978 mm
6	77.3 mm	267 mm	32	70 mm	2048 mm
7	76.4 mm	323 mm	33	69 mm	2119 mm
8	75.8 mm	381 mm	34	69 mm	2189 mm
9	75.5 mm	442 mm	35	69 mm	2260 mm
10	75 mm	506 mm	36	69 mm	2330 mm
11	74.6 mm	572 mm	37	69 mm	2400 mm
12	73.7 mm	640 mm	38	68.5 mm	2471 mm
13	73.7 mm	711 mm	39	68.5 mm	2541 mm
14	73.7 mm	781 mm	40	68.5 mm	2612 mm
15	73.2 mm	852 mm	41	68.5 mm	2682 mm
16	72.3 mm	922 mm	42	68.5 mm	2752 mm
17	72.3 mm	992 mm	43	68 mm	2823 mm
18	72.3 mm	1063 mm	44	68 mm	2893 mm
19	72 mm	1133 mm	45	68 mm	2964 mm
20	72 mm	1204 mm	46	68 mm	3034 mm
21	72 mm	1274 mm	47	68 mm	3104 mm
22	71 mm	1344 mm	48	68 mm	3175 mm
23	71 mm	1415 mm			
24	71 mm	1485 mm			

Table 1: METEOSAT-Yagi antenna dimensions

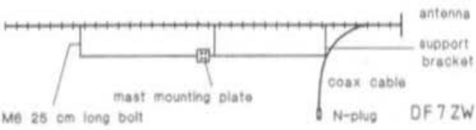


Fig. 1: The 3 metre long antenna with support bracket and cable

Finally, the boom support bracket, consisting of a length of 15 mm x 15 mm square section aluminium tube, is affixed to the boom (also of the same material) by means of three 24 cm x 6 cm rods tapped at both ends. The antenna is fastened to the top of the mast by means of the support bracket thereby ensuring a spacing of at least one wavelength i.e. 17.7 cm between mast and antenna elements.

The reflector is an aluminium plate, 106 mm x 106 mm x 1 mm.

4. RESULTS

The author built, first of all, a 3 metre long version of this Yagi antenna which had a gain of 19 dBd (1). It was used with a DK 1 VA converter (2) having a 2 dB noise figure with 1 metre of RG 213 connecting cable.

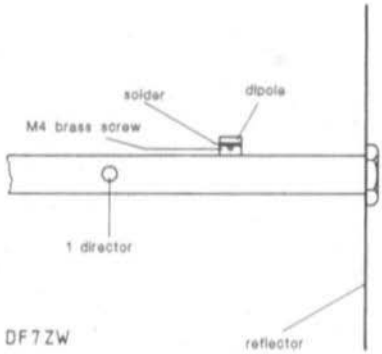


Fig. 2: Mounting the dipole element

The author was amazed how good, and with what reserve, the METEOSAT signal was received. In comparison with a 1.15 m diameter paraboloid reflector antenna, the gain was about 2 dB lower but it served to verify the 19 dB gain quoted above.

According to reference (1), the antenna has a logarithmic gain decrement profile which allows the boom to be broken at any desired length. The author then built a second antenna in order to determine the minimum length of antenna

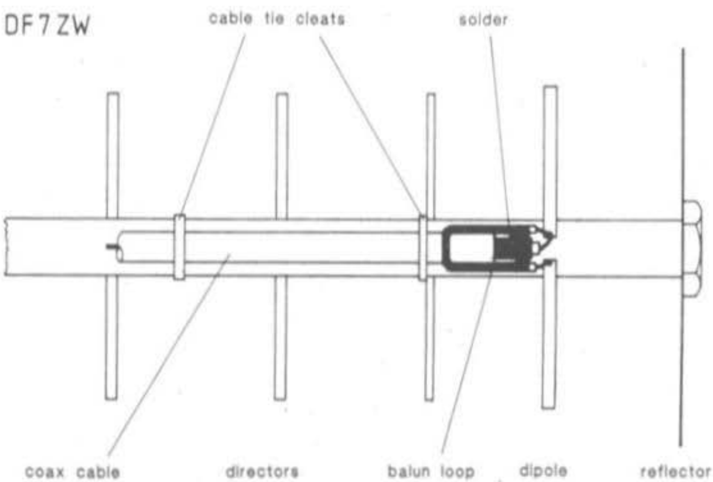


Fig. 3: The feed and balancing arrangements

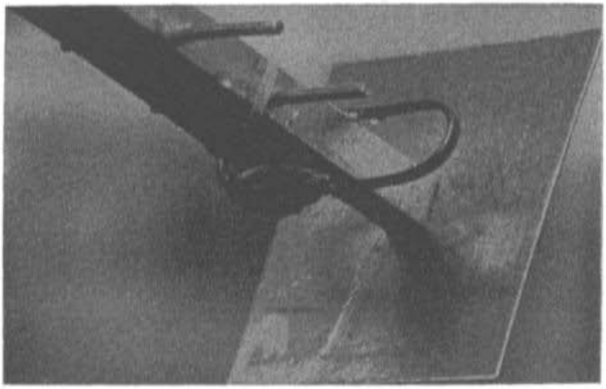


Fig. 4: The dipole feed showing the reflector

required to obtain a noise-free picture. An antenna length of 1 metre results in 15.5 dBd gain but unfortunately, slight noise was apparent on the picture. When the antenna length was increased to 1.5 metres, and thereby 16.5 dBd gain, the reception was as good as noise-free.

If the 1 metre long antenna had been used with a first-class GaAs-FET pre-amplifier, then noise-free results would also have been assumed. If at all possible, the three metre version should be built in order to ensure a 2.5 dB gain reserve.

One of the antennas was measured for the input VSWR, it was 1.3 at 1710 MHz and 1.8 at 1695 MHz. Further optimization was considered to be superfluous with these results.

The final operation was to weather-proof the antenna coaxial cable connection with paint or silicon. This point must not be forgotten, as an ingress of moisture, during a heavy rain shower, will completely ruin the reception, as the author found to his cost!

5. USED MATERIAL

- a 106 mm x 106 mm x 1 mm aluminium reflector plate
- a 62 mm (screening length) balun, 50 Ω semi-rigid cable (SR-3)
- dipole: length 84 mm, height 18 - 20 mm, 4 mm x 2 mm flat copper (as used for transformers) or 3 mm round copper wire
- a 3.20 m boom, 15 mm x 15 mm square section alu tube (2 mm thick)
- 4 mm dia aluminium rod elements, force-fitted in the boom

6. REFERENCES

(1) Hoch G., DL 6 WU:
Extremely Long Yagi Antennas
VHF COMMUNICATIONS Vol. 14,
Ed. 3/1982, P. 130 - 138

(2) Bartkowiak B., DK 1 VA:
A Low-Noise METEOSAT Converter with GaAs-FET Preamplifier and Mixer Stage
VHF COMMUNICATIONS Vol. 17,
Ed. 1/1985, P. 8 - 16

(3) Lentz R., DL 3 WR:
Noise in Receive Systems
VHF COMMUNICATIONS Vol. 7,
Ed. 4/1975, P. 217 - 235

Dr. Eng. Jochen Jirmann, DB 1 NV

A 12 Volt to 12 Volt Converter

The title of this article appears at first sight to be the introduction to an April Fool's joke but upon reflection, one of the following problems may have been experienced with your communication equipment.

– You are planning an extended mains-independent use of the field-day rig and you are looking for an optimum source of primary power. You have come to the conclusion, that the use of primary elements is much more advantageous than accumulators, an alkaline-magnesium mono-cell has a capacity of about 10 ampère-hours (a.h.), a nickel-cadmium (nicad), of the same physical size, only about 4 a.h. But that is'nt the whole story. When a dry battery is being discharged to complete exhaustion, the terminal voltage sinks gradually to about half its initial value. The re-chargeable battery, on the other hand, has a negligible voltage decline until at the limits of discharge, the voltage fails catastrophically. Will your battery transceiver work with only 6 V?. To take more battery cells is also no solution as the equipment runs the risk of over-voltage. What is required then, is a low-loss converter that, when fed with an input voltage of 10 to 20 volt, will deliver a stable voltage of 12 volt. If the transceiver has digital circuits on board then it would be convenient to have a 5 V regulated voltage available, as the normal supply method of dropping it via a series-pass transistor is not too economical on batteries.

– Say a computer is taken along on a field-day to be used as a log book. In order that the entire memory content is not lost in the event of a power-set failure (...who's forgotten to fill up the petrol tank again?), the computer is connected to the automobile battery. The required 5 volts is no problem and the quiescent battery voltage of 12 V can supply both the disc-drives and the screen monitor. But what happens when the motor is started to charge the batteries and the supply voltage to the computer equipment rises to some 14 to 15 volt complete with voltage surges?

In both instances, the solution is the employment of a converter which will work with an input voltage of between 10 and 20 volt and deliver an output at a constant 12 volt. The desirable characteristics of such a converter could be summarized as follows:

* **output voltage 1: 12 to 14 V adjustable, at 3 to 5 Amps.**
* **output voltage 2: 5 V at 2 A**
* **input: protected from both low and high voltages**
* **output: short-circuit proof**

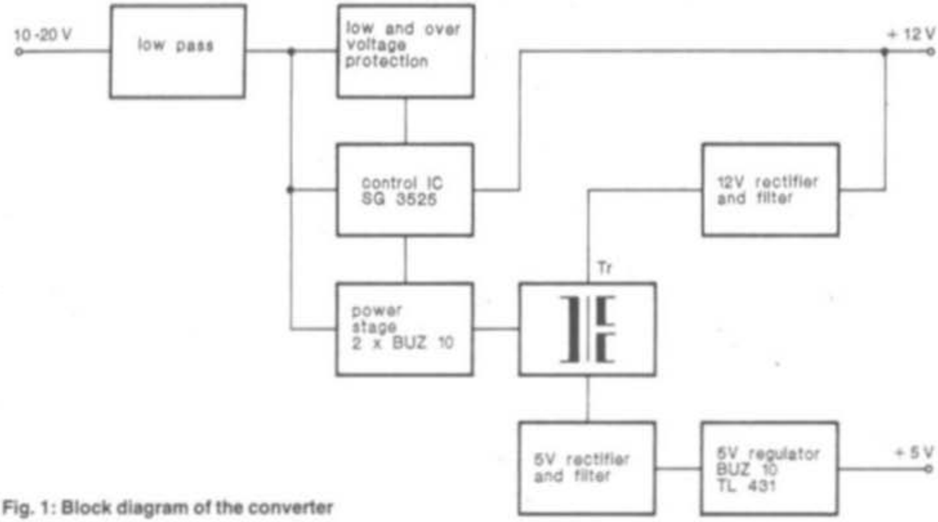


Fig. 1: Block diagram of the converter

1. THE CIRCUIT CONCEPTS

From the target data, outlined above, it may be deduced that the converter must deliver a power of 60 W. This corresponds to a primary supply current of a 6 A at the minimum input voltage of 10 V. From the literature, and e.g. (1), there are three basic circuits to choose from: the forward converter, the single-choke converter and the full-wave forward converter. The half and full bridges, used in 220 V switching power supplies, are unsuitable for low-voltage supplies as the switching current must pass through two series power semi-conductors and the potential difference lost across them is proportionately more at 10 V than at 220 V.

If the converter had been designed around a forward converter, then relatively high currents would have to be switched (this particularly applies to the blocking converter) as power is only delivered during up to 50 % of a period.

For this reason, the full-wave converter has to be the chosen concept. Since normal integrated control circuits have two differential outputs anyway, the external requirements are limited to another power transistor. The use of power MOSFET's, e.g. BUZ 10, with the control chip SG 3525 renders the use of intermediate power driver stages unnecessary.

One disadvantage of the push-pull converter should not, however, go unmentioned. It possesses a relatively poor "cross-coupling". When there are several secondary windings, a load variation on one of them is transmitted to other windings in a relatively strong manner. A blocking converter is much easier to control in this respect.

Using the Siemens application note (2), the mutually-coupled storage chokes of the secondary rectifier can be improved in order to combat this effect. In spite of this, it was necessary to include a series-pass linear regulator in the 5 V output line to prevent interaction between the two outputs – this was also necessary in (3). A MOSFET was used for the series-pass element, this reducing the PD loss to 0.2 V which represents a minimal power loss. The block diagram of the converter is shown in **fig. 1**.

2. THE CONVERTER CIRCUIT DETAILS

The detailed circuit diagram of the converter is shown in **fig. 2**:

The power stage of the converter consists of the transducer Tr and the two MOSFETs BUZ 10. The input voltage is taken via a current measuring circuit and a low-pass filter to the centre-tap of the transducer primary winding. The low-pass filter comprises a ring-cored choke of 220 μH and a 470 μF electrolytic and is included to prevent the contamination of the input supply line with switching frequency impulses and also to inhibit the introduction of pulsating DC when the battery is being charged by the car motor charging system.

The entire control circuitry is encapsulated in the SG 3525 integrated circuit with the exception of the oscillator frequency-determining components which are external, and connected via pins 5, 6 and 7. The oscillator is pulse-width modulated and controls the switching MOSFETs via an internal driver and pins 11 and 14. The 12 V output voltage is monitored at pin 1 after being suitably attenuated by a preset potentiometer P1 and two auxiliary resistors. Frequency compensation for this is provided by an RC circuit to pin 9. At pin 16, a 5 V reference voltage is available. The slow-start input is applied to pin 8 together with a low-voltage detector which blocks the pulse-width modulator when the input supply voltage falls below 8 V. The supply voltage to the SG 3525 is taken via an LED (which also serves as an "on" indicator) having a PD across it of

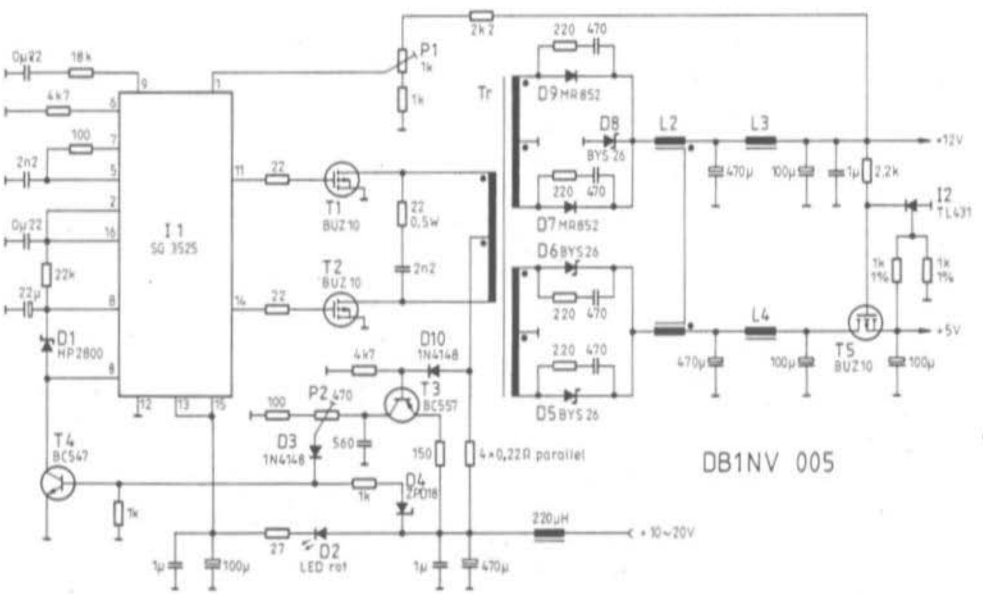


Fig. 2: Converter circuit schematic

2 V. The "low-voltage" threshold is thereby fixed at a 10 volt supply battery potential.

If pin 8 is earthed via the switching transistor T4, the output from the control chip is suppressed. This facility is utilised by the current limiter and over-voltage protection circuits. A current mirror T3 monitors the transducer primary current. If the voltage, as pre-set by P2, is exceeded, T4 is switched and the control chip is effectively switched off. If the supply battery voltage rises above 19 volt, T4 is likewise switched, this time via the 18 V zenner diode and, again, shutting down the converter.

The power transducer Tr has two inter-leaved secondary windings which each supply two full-wave rectifiers. Whilst the 5 V rectifier is equipped with a Schottky diode BYS 26, ordinary fast-switching diodes MR 852 are used in the 12 volt section owing to the high-voltage breakdown (70 V) requirements in this circuit. If high-voltage Schottky-diodes, such as the BYS 26-90, could be employed, an increase in efficiency would result.

Both rectifiers work into a two-stage filter circuit with choke input – both chokes being wound on a common CC 36 core with air-gap (A^L = 250). This mutual coupling of magnetic fields in the storage chokes improves the cross regulation considerably.

The second part of the filter chain comprises a rod-cored choke and a further electrolytic, by means of which, the switching frequency ripple is reduced to the order of a few millivolts. The residual value of the ripple is determined by the so-called ESR (equivalent series resistance) which is the HF internal resistance of the capacitor. This, of course, is dependent upon its quality!

The sample voltage is taken from the 12 volt output. The 5 V output consists of a simple series-regulator MOSFET (T5) and ancillary component I 2 – a variable zenner diode (TL 431). Even with full-lead output, this circuit functions with a minimum of voltage drop across the series MOSFET (about 0.2 V). A three-terminal rectifier would, on the other hand, exhibit a series PD of some 2 to 3 V at the same load current. An additional advantage of this circuit is, that a short-circuit across the 12 V output also results in the 5 V line being shut-down.

3. CONSTRUCTION

In order to facilitate the construction of this project, a printed crcuit board DB 1 NV 005 was developed. The dimensions are 155 x 70 mm and is also plated on the component side. The layout plan is shown in **fig. 3**.

The construction is best commenced with the fabrication of the wound components. The simplest are the two rod-cored chokes for the output circuit. Two rod-cores, type Valvo 4322 020 36810, are wound with 15 turns of 1 mm lacquered copper wire. One of the ends is threaded through the core so that the choke can be mounted upright on the PCB.

The storage choke L2 is wound on an air-gapped core CC 36 (A_L = 250) with two windings of 18 and 35 turns of HF Litz 60 x 0.1 mm. The beginnings and ends are marked so that the ends of the 35 turn winding can be identified and used for the 12 volt output circuit – the 18 winding choke is used in the 5 V portion of the circuit. Note the end polarities when fitting to the board!

The power transducer Tr comprises an ETD 34 core kit without an air-gap. In order to achieve a good winding symmetry, the two halves of the windings are bifilar wound and all the connections, including the centre-taps, are connected in series according to polarity. All windings are carried out with HF Litz wire 60 x 0.1 mm. The primary has 2 x 9 turns, the 12 V side has 2 x 15 turns and the 5 V side 2 x 8 turns. The ends of the HF Litz wire should be carefully stripped of insulation and wired, according to the component plan, to the solder pins on the body of the coil former.

Afterwards, all the components, with the exception of the converter transistors T1 and T2, can be mounted and soldered to the board. It should not be forgotten that power MOSFETs are sensitive to static charges and should be carefully handled in the appropriate manner.

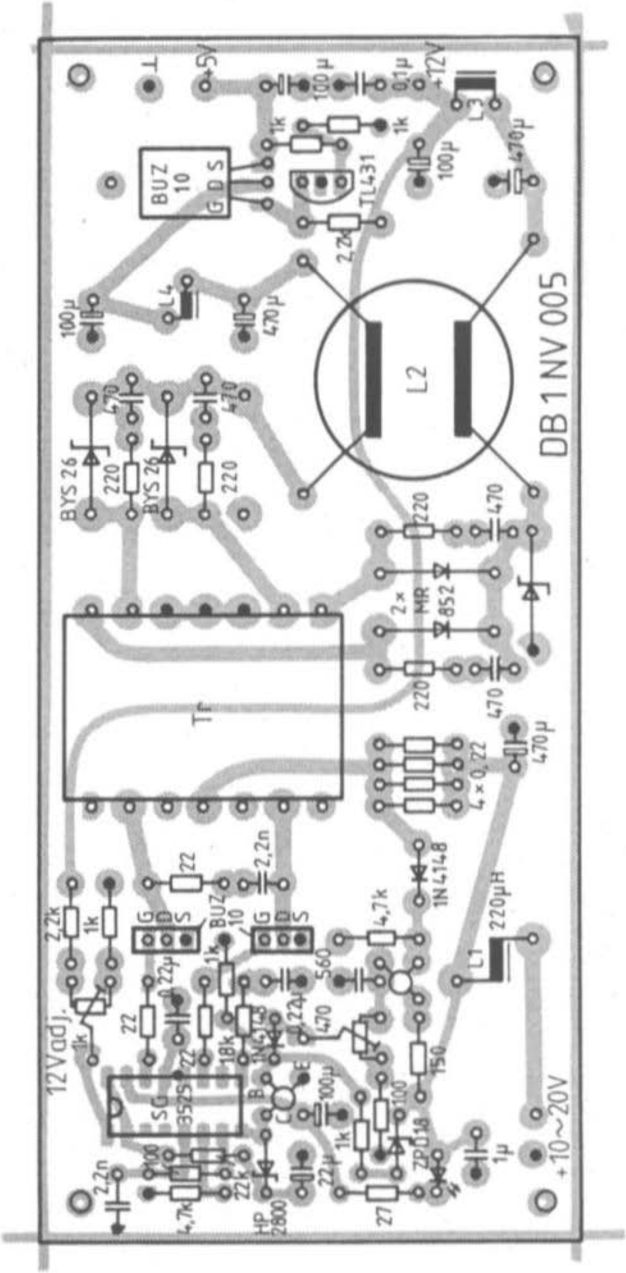


Fig. 3: Converter board component layout plan

3.1. Components

Semi-conductors:

I1:	SG 3525; Silicon General, Motorola, Texas Instruments
I2:	TL 431; Texas Instruments, Motorola
T1, T2, T5:	BUZ 10, BUZ 71; Siemens, Valvo
T3:	BC 557
T4:	BC 547
D1:	HP 2800, 1 N 6263; Hewlett-Packard
D2:	LED red
D3:	1 N 4148, BAW 76
D4:	ZPD 18
D5, D6, D8:	BYS 26 (Schottky diode 3 A/45 V); Siemens
D7, D9:	MR 851, MR 852 (fast diode 3 A/100 V); Motorola

Resistors:
Carbon or metal film RM 10
Preset pot'meter horizontal RM 5/10

Capacitors:
Ceramic and foil types RM 5
Electrolytics: Upright versions

Wound components:
Tnsfr.: Core kit ETD 34 without air-gap, primary 2 x 9 turns, secondary 2 x 8 turns (5 V winding) and 2 x 15 turns (12 V winding), HF Litz 60 x 0.1 mm (see text)

Storage choke: CC 36 Core with air-gap, $A^L = 250$, winding 1 with 35 turns, winding 2 with 18 turns, HF Litz 60 x 0.1 mm, polarity of windings important!

Input choke L1: Ring-core-suppression choke for Triac circuitry, 220 μH/6 A.

Suppression choke L3, L4: 15 turns on Valvo rod core 4322 020 36810 lacquered 1 mm copper wire.

Miscellaneous:
1 heat sink 40 x 40 x 40 x 1.5 mm (angled) aluminium strip for T1 and T2
1 U-formed heat sink 25 x 25 x 18 mm for T5
Insulation washer and mounting material for T1, T2 and T5
Solder pins
1 Brass or plastic screw M 4 x 40 with nut and washer to secure L2.

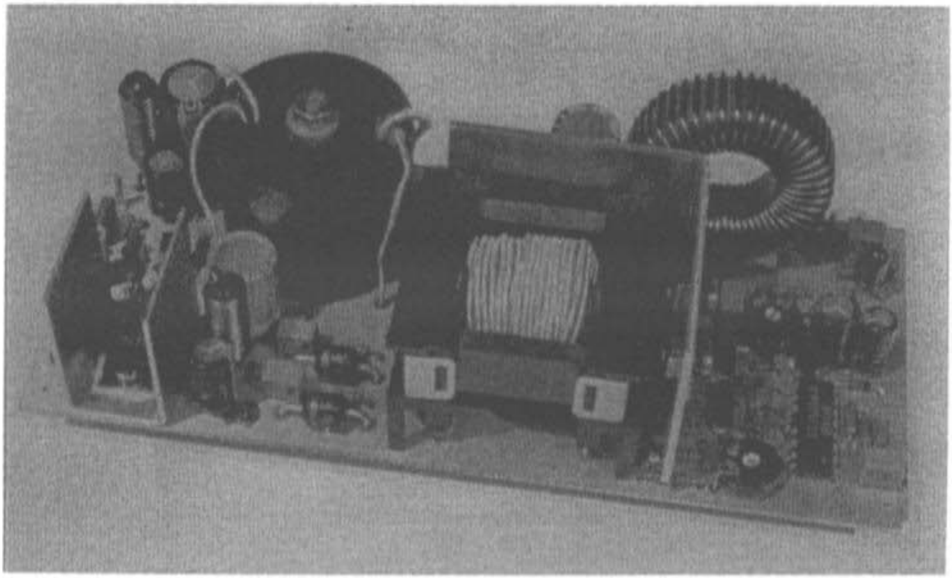
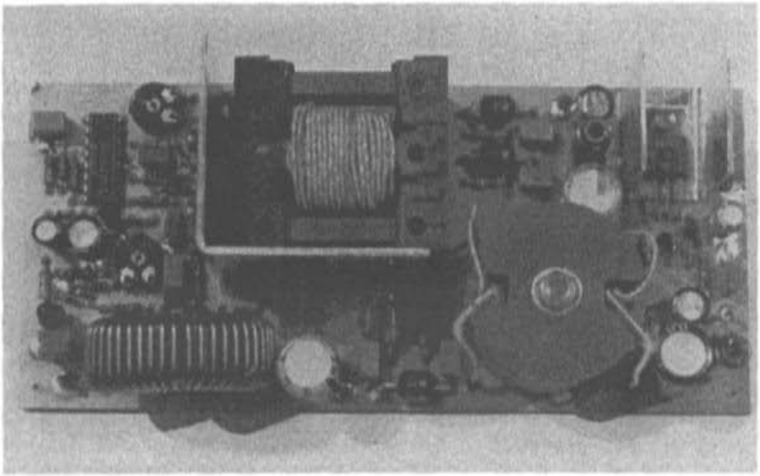


Fig. 4: The completed converter



4. COMMISSIONING

The two preset potentiometers P1 and P2 are first of all turned to their mid-range position and the input connected to a laboratory power supply. At a supply input of 12 V, the input current from the test supply should be about 20 mA.

An oscilloscope is then connected to the MOS-FET gate circuit and an approximately 60 kHz pulse train should be observed with a very nearly 50 % duty cycle and 10 V amplitude. Check that these pulses disappear when the input voltage is lower than about 10 V and also when it is higher than 19 V, thereby checking the over/under-voltage protection circuits.

The transistors T1 and T2 are now soldered into the board and the input voltage set to 12 V. A quiescent current of 30 to 50 mA should result. Adjust P1 for an output voltage of exactly 12 V.

The 5 V portion does'nt need any adjustment, normally, owing to the tight component tolerances of I2 but, if necessary, one of the two 1 kΩ 1 % resistors could be changed. A departure from nominal of more than 0.2 V would indicate a defective component or a circuit fault.

Now, the output can be connected to a load resistance and the regulation checked with both varying load and input voltage. This check must be carried out with a load current of at least 0.2 A at the 12 V output in order that the 5 V output can be loaded fully. The maximum load can be checked and the current-limit threshold set with P2.

The prototype converter is shown in the photographs of **fig. 4**.

5. REFERENCES

(1) Jirmann, J.:
Switched-Mode Power Supplies, Part 1
VHF COMMUNICATIONS Vol. 17,
Ed. 2/1985, P. 79 - 93

(2) Without author:
Siemens Schaltbeispiel Durchflusswandler-SNT mit mehreren Ausgangsspannungen (5 V/10 A, +/− 12 V/2 A), TDA 4718 und SIPMOS-FET, Order no. B/3030

(3) Jirmann/Krug:
Microcomputer System
Part 1: Switched-Mode Power Supply
VHF COMMUNICATIONS Vol. 18,
Ed. 2/1986, P. 108 - 120

Joachim Kestler, DK 1 OF

A 2 m / 70 cm Antenna Splitting Filter

The following article concerns mainly those mobile amateurs who operate in the 145 MHz and 435 MHz amateur bands. Of course, it is possible to have a dedicated antenna for each amateur band as well as for the broadcast receiver. This philosophy could also be applied to the mobile sphere of interest but an antenna park on the family-car roof doesn't do much for its re-sale prospects!

1. PRINCIPLE

All quarter-wave groundplane antennas conceived for 145 MHz can, in principle, also be used at 435 MHz. They are then excited on their third harmonic and present a real 50 Ω impedance at the base of the antenna (1). In addition, commercially designed multi-band antennas working at 2 m with a quarter-wave antenna length, represent also a 5/8 wavelength radiator having a 3 dB gain at 435 MHz (e.g. the UKW-Technik's MU3 and MHU3-Z). One could, of course, use a coaxial switch to turn the equipment in use, i.e. 2 m or 70 cm, to the antenna but this precludes the simultaneous use of both equipments to work duplex. For this facility an antenna diplexer (splitting filter) is required.

2. THE DESIGN

The circuit of such a diplexer is shown in **fig. 1**. The antenna is connected to port A, the 2 metre transceiver to port B and the 70 cm transceiver to port C.

The parallel tuned circuits L1/C1, L2/C2 and L3/C5 are resonant at 145 MHz and therefore present a very high impedance at this frequency. This prevents the power from branch A-B from

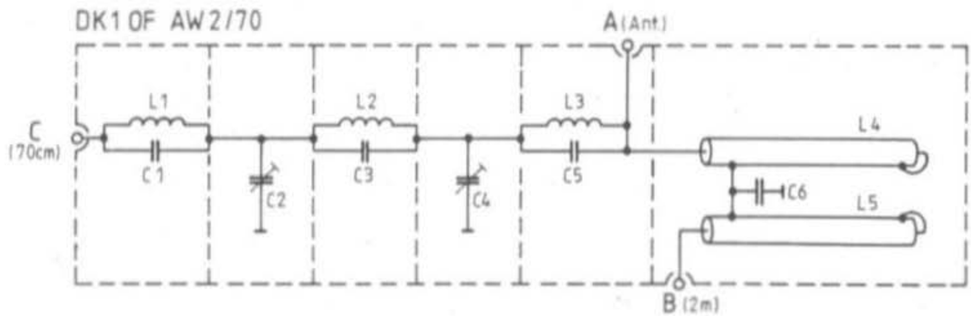


Fig. 1: 2 m band/70 cm band antenna diplexer

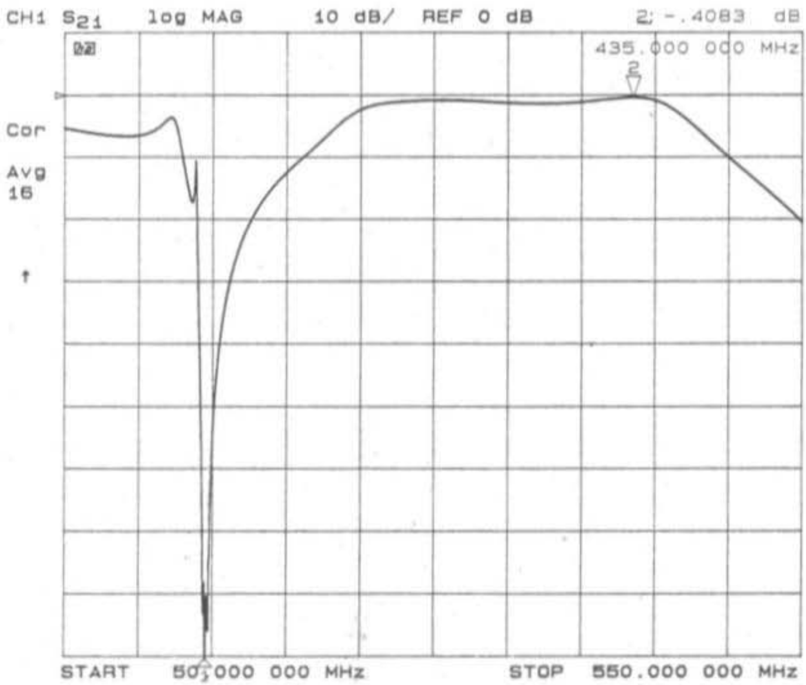


Fig. 2: Path A-C exhibits a stop attenuation at 145 MHz of 90 dB and an insertion loss at 435 MHz of 0.4 dB

appearing in C. For signals in the 70 cm band, the path A-C should, however, present the very lowest impedance possible. This is enabled by employing the parallel tuned circuits (they are effectively inductive reactances at 70 cm) as a T-form low-pass filter in conjunction with C2 and C4.

As **fig. 2** shows, the insertion loss for the 70 cm path A-C is only 0.4 dB whereas 2 m signals in this path suffer a 90 dB (approx.) attenuation.

The signal path A-B should – as may be imagined – behave in frequency completely opposite to the path A-C. That is, it should have the highest possible attenuation for 70 cm signals and at 2 metres have as low an attenuation as possible. As it is difficult at frequencies as high as 435 MHz to achieve reasonable values of Q with normal tuned circuits, i.e. coiled inductors with capacitors, line resonators comprising semi-rigid cable are employed, fig. 1. The coaxial tuned lines L4 and L5 are short-circuited at their ends and represents a quarter-wave tuned circuit at 435 MHz which is parallel tuned. Signals at frequencies around 145 MHz see L4 and L5 as inductive reactances, which with C6, are formed into a T-form, low-pass filter thus ensuring a minimum insertion loss in the 145 MHz band.

The characteristic attenuation/frequency for the path A-B is shown in **fig. 3**. Two-metre band signals are attenuated less than 0.07 dB whilst the isolation at 435 MHz reaches a value here of some 90 dB. Both the characteristics of fig. 2 and fig. 3 were obtained with the aid of an HP 8753 **Network Analyser**.

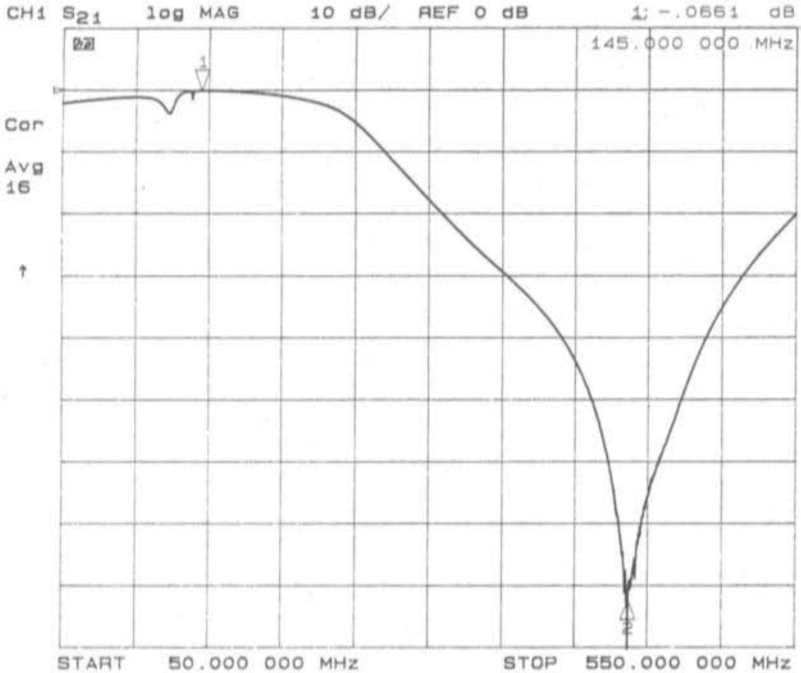


Fig. 3: Path A-B exhibits a de-coupling at 435 MHz of 90 dB and an insertion loss at 145 MHz of only 0.07 dB.

3. CONSTRUCTION

Now, a few notes concerning the construction. The diplexer, described here, is housed in a tin-plate box with the external dimensions 139 mm x 53 mm x 30 mm. **Figure 4** shows where the BNC sockets for ports A, B and C are located and the method of attachment – simply soldered around the flange edges to the tin-plate box.

The internal space of the housing is compartmented with tin-plate dividing walls having six holes through which pass the diplexer's conductors via soldered-in, teflon, feed-through insulators. The wall separating BNC socket A and socket B is necessary in order to prevent about – 40 dB of capacitive coupling.

Fig. 5 is a photograph of a constructed example of the diplexer and **fig. 6** gives the necessary dimensions for the fabrication of both coaxial resonators L4 and L5.

3.1. Components

L1, L2, L3: 3 turns silvered wire 1.5 mm dia, 7 mm int. dia
L4, L5: Semi-rigid cable SR-3 (Suhner) as drawn
C1, C3, C5: cer. disc 22 pF
C2, C4: tubular trimmer 0.5 – 3 pF
C6: cer. disc 10 pF
3 BNC panel sockets
1 tin-plate box 139 x 53 x 30 mm

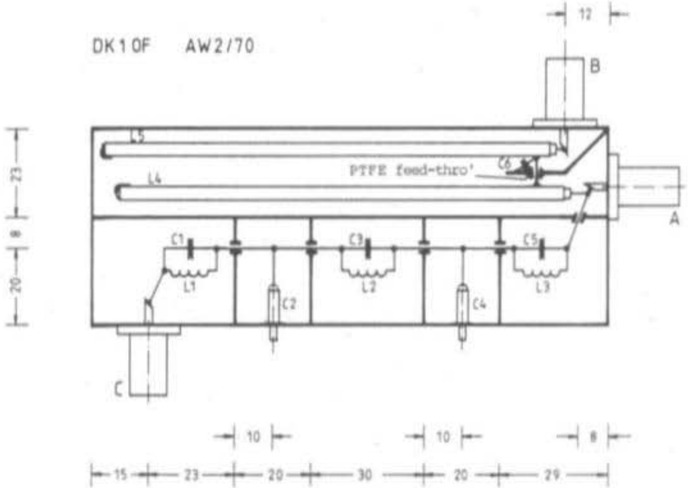


Fig. 4: Leading dimensions indicating placement of compartment walls, feed-through insulators and sockets

4. TUNING

The following hints concerning the diplexer's tuning are directed mainly to those amateurs who do not have a well-equipped HF test laboratory at their disposal. A (simple) SWR-meter should, however, be available; it is connected between socket A and the antenna.

The two metre transceiver is connected to socket C. Using a received signal from a local relay

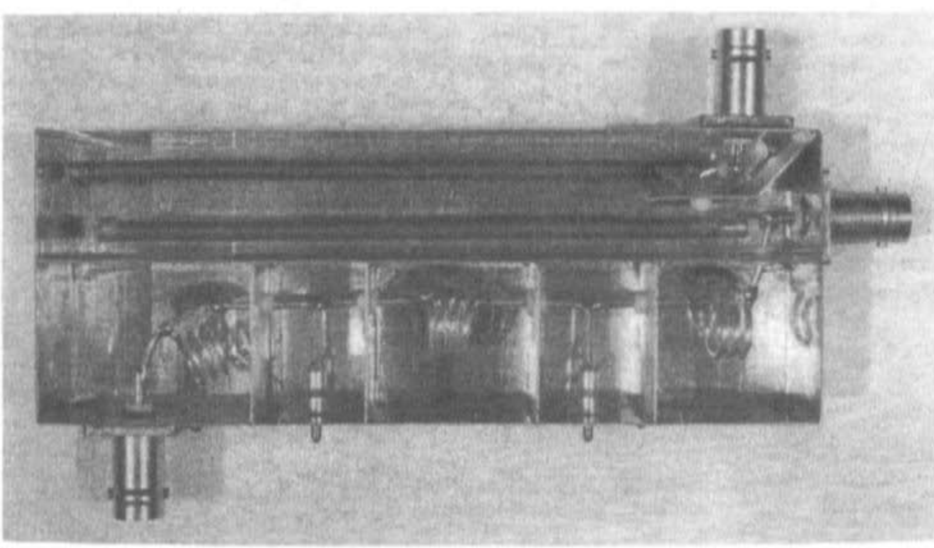


Fig. 5: Constructed example of 2m/70 cm diplexer

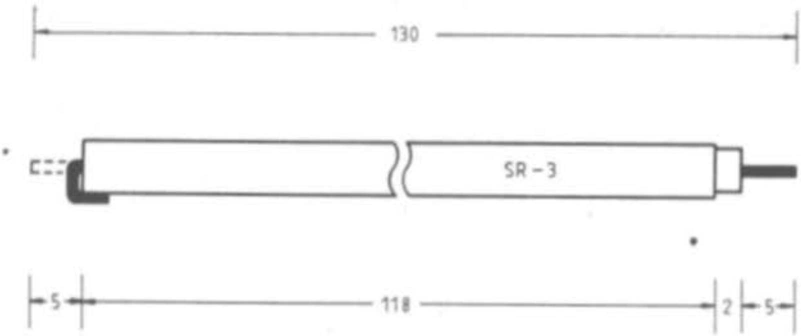


Fig. 6: Copper-clad, coaxial cable, quarter-wave resonators for L4 and L5

station or another amateur transmitter but in any case as strong as possible, L1, L2 and L3 are tuned for the greatest possible attenuation i.e. smallest S-meter reading. The tuning is carried out by carefully compressing and expanding the coiled inductors using an iteration tuning processer between the three inductors until the maximal attenuation has been achieved.

A 70 cm transceiver is connected to socket C and the equipment set to transmit in the middle of the amateur band. Capacitors C2 and C4 are then tuned, again iterating between them, in order to achieve a maximum in the SWR meter.

The tuning of branch A-B has been determined by the dimensions of the coaxial line resonator and does not therefore require adjustment. The following table summarises the technical specifications of the diplexer:

Path A-B:	Insertion loss at 145 MHz:	0.066 dB
	Stop attenuation at 435 MHz:	90 dB
Path A-C:	Insertion loss at 435 MHz:	0.41 dB
	Stop attenuation at 145 MHz:	90 dB

Impedance: All parts nominally 50 Ω
Power rating: 100 W continuous
400 W intermittant

Table of test results

5. OPERATION WITH MOBILE EQUIPMENT

As may be deduced from fig. 3, the loss in path A-B for frequencies around 100 MHz (VHF broadcast band) is only about 1 dB thus permitting the antenna diplexer described in (2) (i.e. broadcast/ 2 m diplexer) to be connected to port B. In this manner all three equipments, car radio, 70 cm transceiver and 2 m transceiver, can all share the same mobile antenna.

6. LITERATURE

(1) Meinke-Gundlach:
Taschenbuch der HF-Technik, 4. Auflage, S. N12 – N13, Springer-Verlag 1986

(2) Kestler J., DK 1 OF:
Antenna Splitting Filter for Broadcast and 144 MHz
VHF COMMUNICATIONS Vol.10, Ed. 1/1978, P.37 - 41

Index of Volume 19 (1987)

Article	Author	Edition	Pages
OSCILLATORS			
PLL Oscillators with Delay Lines Part 5: Digital Frequency Tuining	Joachim Kestler, DK 1 OF	1987/1	2 - 12
VCOs using Semi-Rigid Cable Tuned Circuits	Eugen Berberich, DL 8 ZX	1987/4	210 - 214
INTERESTING COMPONENTS			
Integrated Single-Stage Broadband 0 to 2 GHz Amplifier	Jochen Jirmann, DB 1 NV	1987/2	107 - 108
The SDA 4211 – An Interesting UHF Prescaler	Jochen Jirmann, DB 1 NV	1987/2	109
FILTERS			
Switched-Capacitor Audio Filter	Werner Rahe, DC 8 NR	1987/2	113 - 125
SHORTWAVE, IF-MODULES			
A 10 kHz-30 MHz Receiver Front End Part 1	Joachim Kestler, DK 1 OF	1987/1	13 - 26
A 10 kHz-30 MHz Receiver Front End Part 2 (Concluding)	Joachim Kestler, DK 1 OF	1987/2	99 - 106
Broadband HF Power Amplifiers	Andreas Schaumburg, DF 7 ZW	1987/3	141 - 149
2 m BAND			
Improve the Oscillator Circuit in your old 2 m Converter like DL 6 HA 001 or Similar Designs	Jan-Martin Noeding, LA 8 AK	1987/3	152 - 153
Converting the TELECAR TS 160 into a 2 m, 80-Channel Amateur FM Transceiver	Roland Barchet, DK 2 LT	1987/4	194 - 203
Low-Noise 144 MHz Pre-Amplifier using Helical Tuned Circuits	Dragoslav Dobričić, YU 1 AW	1987/4	243 - 252

Ralph Berres, DF 6 WU

Improved AFC Unit for the DJ 4 LB ATV-Transmitter Sound Carrier

The ATV transmitter, published in (1) by Guenter Sattler, DJ 4 LB, has been, in the course of the year, reproduced by many radio amateurs. The sound oscillator at 33.4 MHz is frequency-modulated by the microphone. The difference frequency to the vision oscillator, namely 5.5 MHz, was stabilised by means of the module published in (2) by Josef Grimm. Unsatisfactory operation forced me to search for a simple but effective alternative which also, must use low-cost and easily obtainable components (e.g. CB crystals). For this, a special frequency plan had to be evolved.

1. CIRCUIT DESCRIPTION

The complete circuit schematic of the module is reproduced in **fig. 1**. The oscillator (T1) works on a CB frequency of 26.72 MHz. The next stage is an emitter-follower (T2) which buffers the oscillator from the following gate. At the input (9 and 10) of the 7400 is a preset pot'meter which serves to set the working point. The sinusoidal signal is converted by the TTL NAND gate into a square-

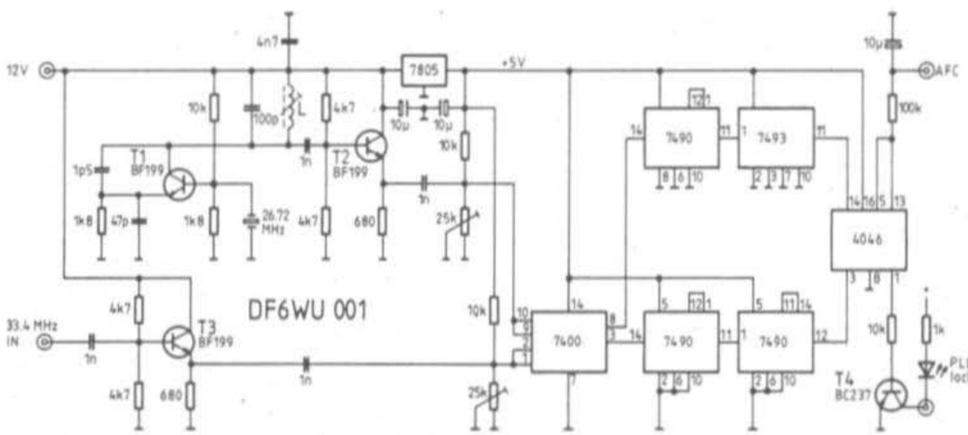


Fig. 1: PLL circuit using a CB crystal for the ATV sound oscillator

wave voltage. The TTL IC, connected to output pin 8, divides the frequency by a factor of 10 and then by 8 so that the output, derived from the crystal oscillator, is 334 kHz.

The signal from the sound-oscillator module (DJ 4 LB 002) is fed to the lower part of the circuit. Following the buffer stage (T3) and a preset pot'meter, is again a NAND gate and two TTL ICs. This combination divides by a total of 100, again resulting in a nominal frequency of 334 kHz. The integrated CMOS circuit 4046 is a combined frequency and phase comparator PLL. It delivers a control voltage at its output which is dependent upon the frequency or the phase difference between its two inputs. This control potential is then fed to a low-pass filter and then out to the sound-oscillator varicap.

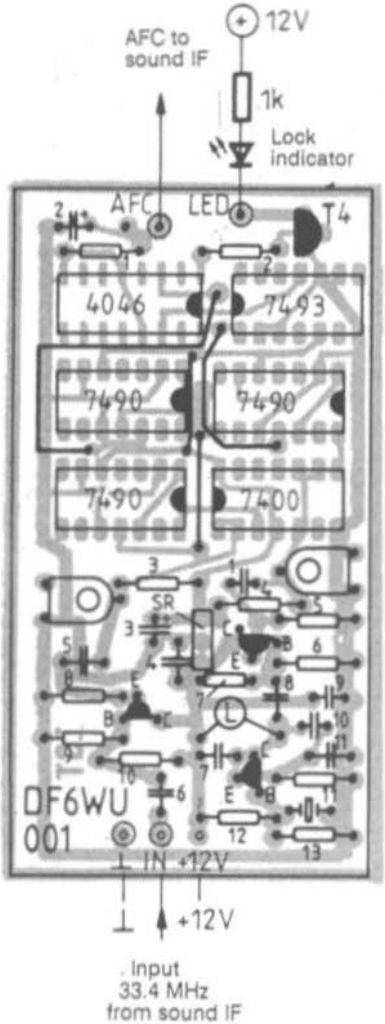


Fig. 2: The PCB DF 6 WU 001 is fully copper coated on the component side

2. CONSTRUCTION

A printed circuit board (**fig. 2**) was specially designed for the circuit of fig. 1. It is double-sided with a component layout plan and is 90 mm x 48 mm. The upper side is left completely unetched and therefore all holes passing leads which are not directly grounded must be countersunk on the ground-plane side in order to prevent a short-circuit to earth. This process has been mentioned many times – a 3 mm drill is most suitable for the operation. Unfortunately, the component layout plan has been printed with the BF 199 turned the other way round: **If the centre lead is bent towards the rear, the problem will alleviated.**

2.1. Components

Q: CB crystal 26.72 MHz
L: 14 turns silvered 1 mm copper wire wound on 5 mm former
Voltage regulator: 7805
T1...T3: BF 199 (BF 224)
T4: BC 107, BC 237
TTL-ICs: standard version, no LS types!
C-MOS: CD 4046 N
Preset: 25 kΩ, RM 10/5 mm
Resistors: 0207 series (RM 10 mm)
Capacitors:
small values: ceramic disc RM 2.5 mm
medium values: ceramic disc RM 5 mm
μF values: tantalum pearl RM 2.5

Resistor positions:	Capacitor positions:
1: 100 k	1: 1 n
2: 10 k	2: 10 μ
3: 10 k	3: 10 μ
4: 10 k	4: 10 μ
5: 680 Ω	5: 1 n
6: 4k7	6: 1 n
7: 4k7	7: 100 p
8: 680 Ω	8: 1 n
9: 4k7	9: 4n7
10: 4k7	10: 1p5
11: 1k8	11: 47 p
12: 10 k	
13: 1k8	

2.2. Housing Construction

The printed circuit board may be made to fit into a proprietary tin-plate box using only a file to trim it to an exact fit. I prefer to construct my own boxes for this purpose. The equipped board is then soldered on the component side to the housing in a continuous joint all the way around the walls. The supply voltage and the output AFC voltage are passed via 1 to 2 nF feed-through capacitors. The high frequency lead-ins use either a BNC or SMC connector for best results. The voltage regulator must be fitted with a heat-sink.

3. COMMISSIONING

After connecting the 12 V supply voltage, the 5 V output of the voltage regulator is checked at once, it must be between 4.9 V and 5.2 V.

A 50 MHz oscilloscope is required with a 10 : 1 probe together with a frequency counter. A signal generator covering 40 MHz would also be very useful but it must have a 1 V output (minimum). The setting-up proceeds as follows: –

* Put the oscilloscope probe at pin 9 of the 7400 and adjust the inductor core until the oscillator starts to oscillate. If it refuses to oscillate, unsolder the crystal and replace it with a 1 nF capacitor. The circuit should spring into oscillation, if not, something else is amiss. If all is in order, replace the oscilloscope probe with the counter on pin 9 or 10 and determine the frequency of oscillation. According to whether the frequency is too low or too high, turns are removed or added to the inductor L until the frequency of 26.72 MHz lies within the adjustment range. The test capacitor is then removed and the crystal soldered back into circuit. The inductor is again adjusted until the largest amplitude is to be seen on the oscilloscope screen.

Fig. 3: The output voltage waveform at pin 3 (8) of the 7400. Left: wrong, right: correct working point adjustment

* Apply the oscilloscope probe to pin 8 of the 7400 and adjust the adjoining preset pot'meter until a stable, jitter-free, square-wave signal is displayed on the screen (**fig. 3**).
* Connect the oscilloscope probe to pin 14 of the 4046 and check the divided-by-80 output. A clean square-wave trace must be obtained.
* If a signal generator is available, the lower branch can be checked and adjusted in isolation from the working signal of the sound oscillator in the module. Of course, the working signal will do just as well as a signal generator although perhaps not so convenient. In any event, a clean square-wave signal must be seen at pin 3 of the 4046, which has been divided by 100 with respect to the input signal.
* Now the AFC voltage should be seen at the output of the PLL integrated circuit. It must fall to 1 V when the frequency is too high and rise to 4 V when the frequency falls below the 33.4 MHz nominal.

4. INCORPORATING INTO THE ATV SENDER

The sound-oscillator module DJ 4 LB 002a (VHF COMMUNICATIONS 4/1977, Page 237) must

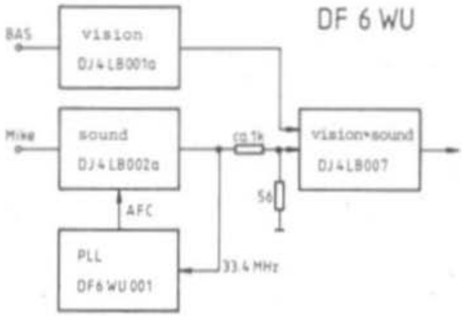


Fig. 4: The integration of the AFC unit into the parent ATV transmitter

first be slightly modified. This is done by omitting completely the 100 kΩ resistor R 220!

As **fig. 4** shows, the AFC output of the DF 6 WU 001 module is connected to the AFC input (Pt 206) of the DJ 4 LB 002a module. The output of the latter module Pt 202 (33.4 MHz) also, is connected to the corresponding input of the subject module. The 33.4 MHz signal should be adjusted for a maximum voltage as the subject module requires about 2 Vpp. Since this is too high for the (DJ 4 LB 007) picture-sound combiner, a voltage divider is provided, the attenuation of which must be determined experimentally.

The inductor L 203 in DJ 4 LB 002a is adjusted in order that the control voltage at the AFC input lies in the middle of the working range, that is 2 to 3 V.

Upon switching on the assembly, the "lock" condition takes about two to three seconds. This time is dependent upon the time constants of the RC network at the output of the 4046 PLL IC and should not be made much smaller than the one shown in the diagram. Otherwise, the lower frequencies will fall out of the control range.

5. REFERENCES

(1) Sattler G., DJ 4 LB:
A Modular ATV Transmitter
VHF COMMUNICATIONS Vol. 5, Ed 1/1973, P. 2 - 15 and
Ed. 2/1973, P. 66 - 80
A Modular ATV Transmitter with Video and Audio Modulation at IF Level
VHF COMMUNICATIONS Vol. 9, Ed. 4/1977, P. 233 - 246

(2) Grimm J., DJ 6 PI:
ATV Information
VHF COMMUNICATIONS Vol. 8, Ed. 1/1976, P. 19 - 23
A Vestigial Sideband Filter for ATV
VHF COMMUNICATIONS Vol. 9, Ed. 1/1977, P. 14 - 18

Hubertus Rathke, DC 1 OP

ATV-FM Driver for the 13 cm Band

The following article will describe an ATV-FM driver, the oscillator of which is stabilized by a conventional phase-locked-loop circuit. Its technical realisation has deliberately avoided the use of exotic specialities together with the need for expensive test equipment – the measurements will be drawn from available comparison quantities. The search for a suitable design also had to encompass the mitigation of the energy loss at these frequencies associated with the transmission between transmitter and antenna as well as a suitable frequency concept which allowed the employment of tried and tested circuits.

1. CONCEPT

If the characteristics of coaxial cables normally used in amateur practice are considered, it may be concluded that they exhibit an attenuation in the 13 cm band of between 20 and 40 dB / 100 metres. In order to improve the energy balance between transmitter and antenna in the transport medium – especially with FM working – a passive multiplication of the basic frequency, together with the baseband, in the immediate vicinity of the antenna was contemplated. The problem with this is, that sidebands, unavoidably produced in the multiplication procces, will also be multiplied along with the wanted signal thus causing a broad spectrum at the final output frequency.

It is certainly advantageous to transmit the RF power at a frequency other than in the UHF television band. The 24 cm band is suitable for this purpose. This is attractive owing to the availability of existing concepts for power amplification in the 23 cm band, and only a doubling process is required for the 13 cm band.

The next step is the production of a frequency-modulated signal from a crystal oscillator working at 1167.5 MHz. The problems start here, as standard components are to be used, the demands upon reproducibility and simplicity of adjustment are magnified. The latter is necessary in order to give the VCO regulated quantities for the positive operation of the phase-locked-loop.

As an alternative to the very high frequency oscillator, described above, a mixer concept consisting of a PLL oscillator and a crystal-controlled oscillator was examined. This configuration is shown in the block diagram of **fig. 1**. This concept was, however, dropped because of the multiplicity of frequencies involved and on top of that, the mixing process itself creates further

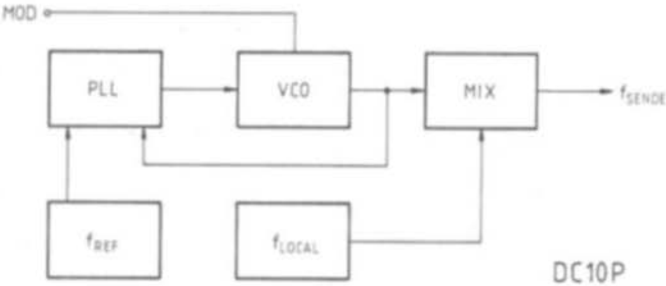


Fig. 1: The VCO signal mixed at a high frequency

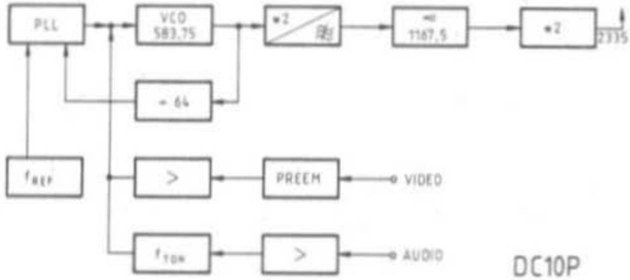


Fig. 2: The modulated VCO in the UHF range using two frequency doublers before power amplification

frequencies thus increasing the possibility of spurious products. Also, it does not allow a suitable frequency separation between the modulation frequency and the PLL reference frequency.

The search for a suitable concept within the frame-work of the given stipulations, was finally resolved with the adoption of a 583.75 MHz PLL and an active doubling of the produced power. A second, passive doubling follows, either at the transmitter, or at the antenna, according to requirements. The detailed block diagram is shown in **fig. 2**.

2. CIRCUIT DESCRIPTION

The complete circuit schematic is shown in **fig. 3**. The three-terminal oscillator is built in discrete form with a PNP transistor. The use of a PNP transistor here obviates the use of an RFC. The discrete construction also allows a concentric trimmer to be used, these being less prone to mechanically induced FM than air or foil types.

It is for this reason also, that the varicap RFC must be stuck to the board with UHU or a similar adhesive. The varicap is connected to the tuned circuit via a trimmer in order that a defined sweep width of 3 MHz per volt, with reference to the oscillator frequency, may be adjusted by iteration with other trimmers. The feedback conditions are fulfilled by a capacitive potentiometer in the frequency determining circuit.

The output from the VCO is taken from a tap on the tuned circuit and the output to the PLL divider is taken by means of an inductive coupling loop. The oscillator itself has a 9.3 V stabilized supply and is only driven moderately in order that the varicap diode is working well within its limits. The varicap tuning voltage is derived directly from the VCO's supply voltage and lies in the middle between about 2.5 and 4.5 V.

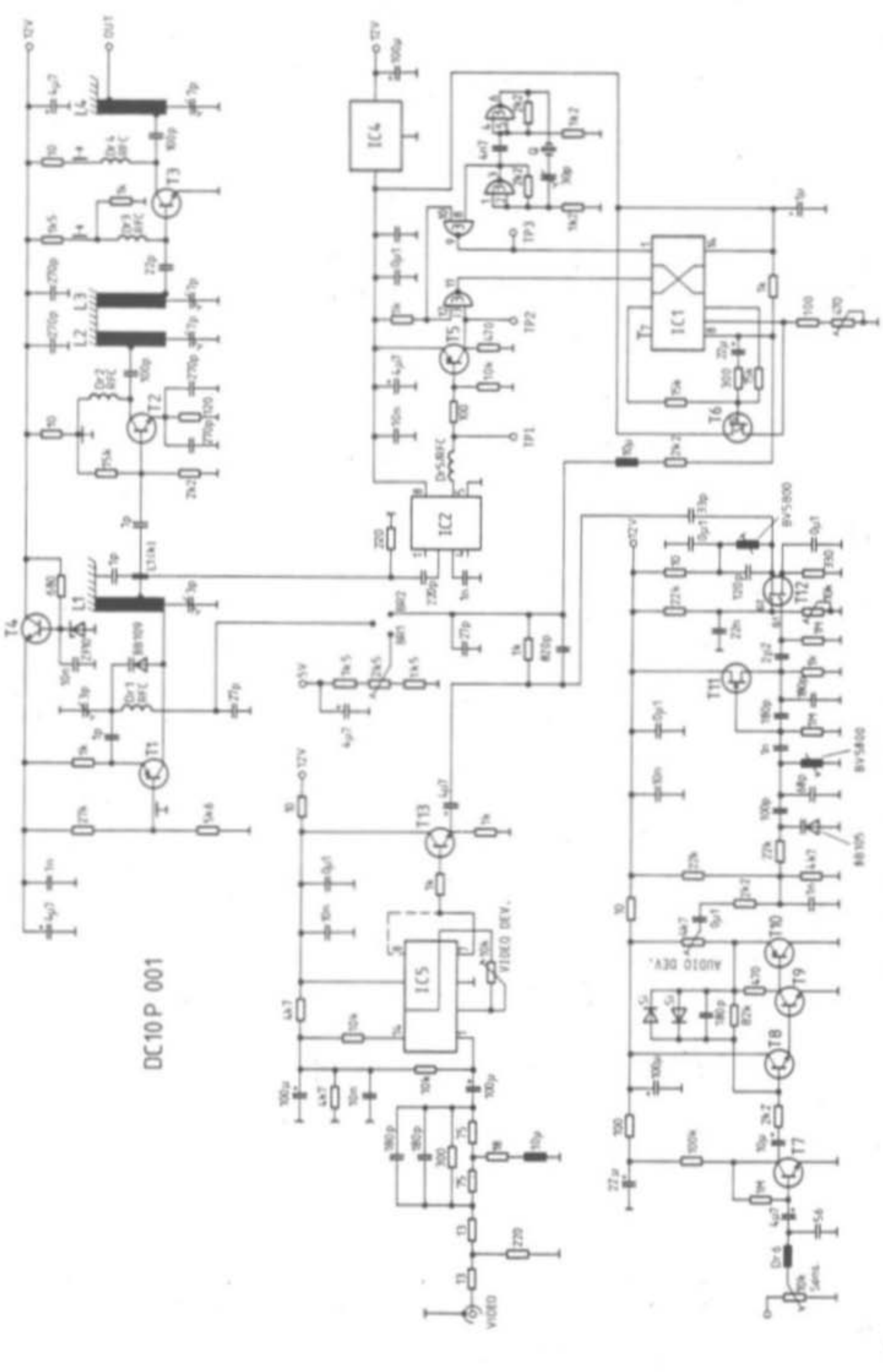


Fig. 3: The complete circuit schematic of the PLL FM-ATV 23 and 13 cm driver

The oscillator drives an active doubler stage using a BFR 34 transistor and a bandfilter. The signal, which is now available at half the final frequency, is then amplified on the board to 12 - 15 mW. Both these stages are fairly conventional, consisting of standard circuits and dimensions.

2.1. The Divide by 64 Stage

This stage divides the signal derived from the oscillator by 64 and supplies it to the PLL at 9.121 MHz. The circuit conforms to that of the chip manufacturer's application note. The upright portion at pin 6 of the divided differential signal is taken to a fast PNP switching transistor to bring it to TTL level. The following inverter also serves as a buffer.

2.2. Reference Oscillator

A crystal controlled TTL oscillator using 1/2 SN7400 was employed as the reference oscillator. The following inverter serves, as in the case of the divider, as a buffer. Note that this oscillator employes 2.2 kΩ feedback resistors to the gates which serve to increase the input impedance thus constituting a lower load on the crystal.

The output frequency of the oscillator may be trimmed to nominal by the series trimmer capacitor in the crystal circuit. The degree by which the crystal may be pulled is dependent upon the type of crystal. Fundamental mode CB crystals may be used, if no great importance is attached to an extreme degree of stability.

2.3. The Phase-Locked-Loop

Strictly speaking, this comprises the following function blocks (see fig. 2): oscillator divider, reference oscillator, phase/frequency comparator and low-pass filter. The IC used in this concept combines the functions of the phase/frequency comparator and the low-pass filter. As it has been identified in the literature as a PLL-IC then it will be so referred to here.

The block diagram shows a mixer at the output of the low-pass filter and at the control voltage input where the control signal, namely the diode tuning voltage, is superimposed upon the video and audio modulation signals. If these signals had not been attenuated by the low-pass filter, they would be seen as extraneous signals by the PLL and cancelled out.

To check the functioning of the low-pass filter – an important element in determining the behaviour of the PLL – a few factors will have to be considered: The lowest frequency to be transmitted in a video signal is the frame signal at 50 Hz. According to theory, however, that is not quite true as there is also a DC component, constituting the black pedestal, present in the video signal. This is only mentioned for the sake of completeness, as this part of the signal is re-inserted, following detection, by a DC restorer. The highest discrete frequency to be found in the video signal is the colour pilot carrier at 4.43 MHz. For the video signal then, a bandwidth of some 4.5 MHz, starting at 50 Hz, can be reckoned with in the spectrum.

The audio signal is quite separate, being a frequency modulated signal on a pilot carrier at 5.5 MHz, i.e. it lies 1 MHz higher than the highest video component. The lowest PLL low-pass filter frequency, to ensure fidelity of video transmission and to filter noise components, is then fixed at 50 Hz.

In order that the VCO spectrum carries only the wanted signals, a further filtering of the control voltage is carried out by an additional LP filter located immediately before its entry into the oscillator. This filter has a limit of 6.5 MHz. It contributes towards the suppression of the reference frequency which has been introduced via capacitive coupling in the PLL-IC and appears at the PLL low-pass output. The PLL itself, is a digital type 4 (1) whose advantage is that it possesses, theoretically, an infinate capture and lock range.

2.4. Video Amplifiers using Pre-emphasis

The video amplifier circuit, complete with pre-emphasis, was taken as a complete unit from (4). The video signal is taken through a T-network for impedance matching to the following pre-

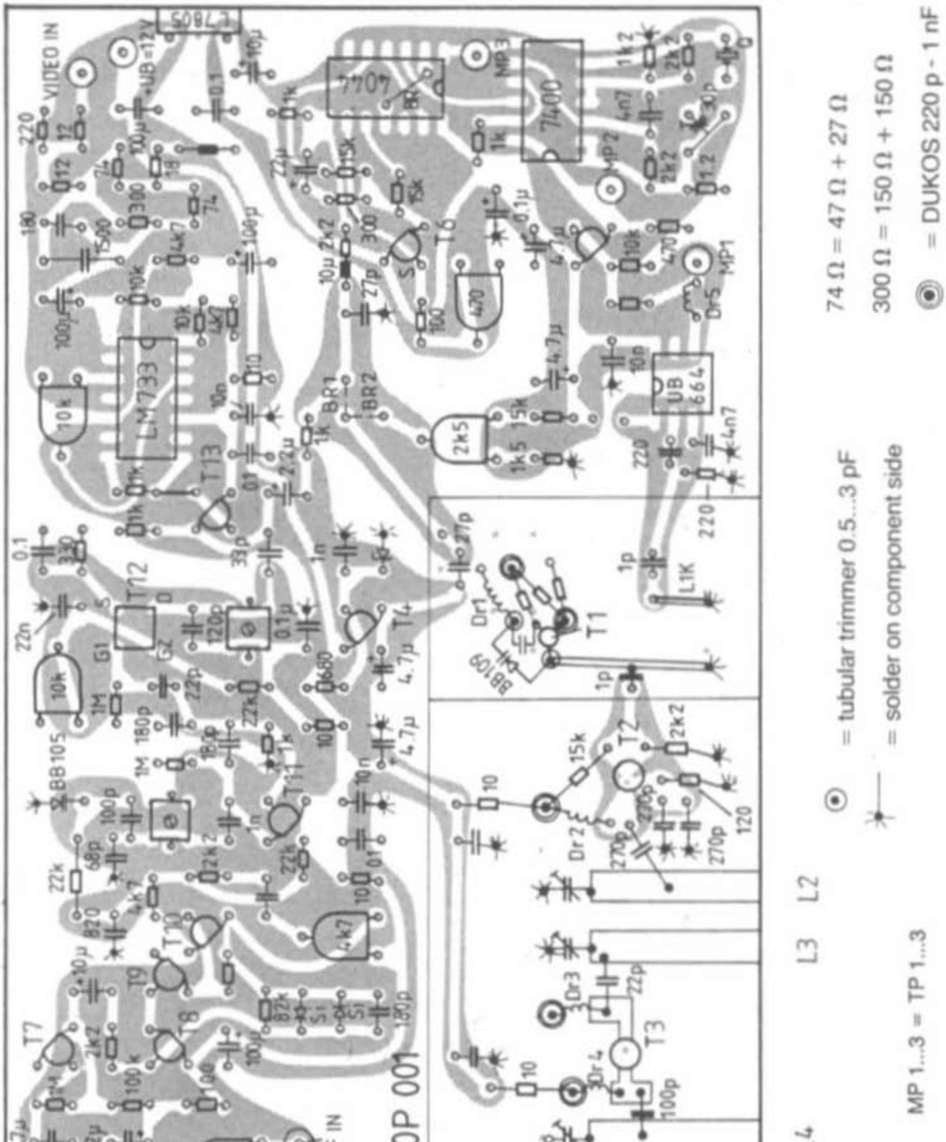


Fig. 4: The DC 1 OP 001 printed circuit board

emphasis network. The P.E. network enhances the high frequency components of the composite signal and attenuates the lower part of the spectrum. The level loss, caused by this operation, is compensated for by the following video amplifier. Thus a suitably balanced video signal is available to modulate the VCO varicap diode as well as buffering the functional circuit block. The AF frequency deviation is controlled by the 10 kΩ preset potentiometer.

2.5. Audio Amplifier and Sound-Carrier Oscillator

The above circuit blocks have also been taken directly from (4). The sound signal is taken first of all to the microphone preamplifier and then on to the modulation amplifier which has a limiter incorporated. A preset potentiometer sets the deviation of the 5.5 MHz sound carrier before the signal is taken to the sound-VCO's varicap diode. After the oscillator is a buffer stage, the output of which can be varied by means of the preset variable voltage provided on gate 2 of the dual-gate MOSFET. This adjustment determines the deviation of the sound carrier in the composite FM-TV signal.

3. CONSTRUCTION

The circuit schematic of **fig. 3** is constructed using a double-sided epoxy-glass printed circuit board in the Europa format. The top side of the board is left unetched as a ground-plane. Contact between the two sides of the board is effected by means of the component leads which are soldered at both the top and bottom sides of the board. A 30 mm strip of tin-plate, encompassing all four sides of the PCB, also serves this purpose.

Sound, video and PLL are all equipped with components with no particular sequence. Normal wiring is used to connect the HF parts, i.e. oscillator, doubler and buffer. This is particularly the case for the oscillator which is completely wired between the two trimmers and the feed-through capacitor which passes through the board.

The component placement diagram of **fig. 4** indicates the position of all components. It may be seen, that the video pre-emphasis filter in the input of the video amplifier is not equipped and the HF circuitry is screened from the rest of the circuit by means of a 20 mm shielding wall.

The oscillator compartment is completely screened by fitting it with a cover.

The 5 V regulator is affixed to the tin-plate frame and the supply leads can be taken to both video and sound circuits without the need for feed-through capacitors. In order that the centre pin of the BNC RF output socket sits on the etched tuned circuit where it can be properly soldered, the positioning of the hole for the socket body must be carefully marked out and drilled before soldering the surrounding wall to the PCB.

3.1. Constructional Notes

The printed circuit board is soldered into its tin-plate frame before it has been equipped. The oscillator components are then mounted and the screening compartment constructed. The rest of the components may then be soldered in any desired order. The coupling resistors in the audio and video portions are left out of circuit so that only the HF and the PLL circuits are operational for the initial commissioning later on.

3.2. Parts List

Semi-Conductors:

I1:	MC 4044 (Motorola)
T6:	2N5457 (Motorola) or BF 245 C
T5:	MPS 3640 (Motorola) or fast PNP-Si-switching-transistor
I2:	U 664 B, BS
T1:	BF 479 (Tfk)
T2, T3:	BFR 34 (Tfk, Valvo)
I3:	SN 7400 (not LS!)
I4:	L 7805
I5:	NE 592, LM 733
T12:	BF 900 (TI or equiv.)
T11:	BF 245
T7, T8, T9, T13, T4:	BC 183, BC 238 or equiv.
T10:	BC 213 or equiv.

BB 105 G (Siemens)
BB 109 G (Siemens)
1N4148 or equiv. switching diode

Conductors:

L 1:	silvered-wire loop, d = 2 mm
L 1 (k):	silvered-wire loop, d = 1 mm 5 mm from PCB
RFC 1:	14 turns 0.5 CuL on d = 2.5 mm
L 2, L 3:	etched
RFC 2:	4 turns 0.1 CuL, d = 2.5 mm
RFC 3:	2 turns o.1 CuL thro' 5 mm long ferrite bead
RFC 4:	4 turns 0.1 CuL, d = 2.5 mm
RFC 5:	5 turns 0.5 CuL, d = 4 mm
RFC 6:	ferrite bead
L 5, L 6:	BV 5800 (Neosid)

Crystal:
The 9.121 MHz crystal for the driver output frequency, 1167.5 MHz, is half the send frequency of 2335 MHz in the 13 cm band. Other transmit frequencies are calculated as follows:

$$f_Q = \frac{f_{tx}}{256}$$

For the VCO frequency:

$$f_{VCO} = f_Q \times 64$$

Note: The divide-by-64 IC has two types available, both of which may be used. The difference is, that the U 664 B does not generate a spurious in the event of no input signal, and the U 664 BS invariably does.

4. TUNING

For commissioning and tuning, a very high impedance multimeter is required along with a 30 MHz frequency counter, one oscilloscope and an HF indicator (e.g. absorbtion wavemeter).

1. Set all trimmers and potentiometers to the middle of their range. The 470 Ω presets in the PLL should be set to maximum resistance. Bridge 1 is made.
2. Switch on the supply voltage 12 - 13.5 V and check the 5 V and 9.3 V (VCO) lines. About 9 MHz should be measured at TP 1 and TP 2.
3. Connect the voltmeter to bridge 1 and adjust to 2.5 V with the 2.5 kΩ preset. By means of the trimmer, adjust the VCO's frequency so that it is about the same as that of the reference oscillator's frequency.
4. Adjust the doubler and buffer stage for maximum output signal at 1.1 GHz adjusting (3) above, if necessary.
5. Check the VCO's deviation: by means of the series trimmer to the VCO varicap, the frequency deviation is adjusted to ± 3 MHz with a ±1 volt change at bridge 1. Frequently check and adjust the nominal frequency (64 x f_Q) in the course of the deviation setting. Owing to the non-linearity of the varicap, the deviation will not be exactly symmetrical about the nominal frequency. The following table applies: –
6.

V (bridge)	f (TP2)	df (MHz)	f (VCO)
1.5 V	9.071	– 3.4	580.6
2.5 V	9.124	0	584.0x
3.5 V	9.171	+ 3.0	587

7. Check control loop voltage of the still open loop. It should be as near as possible to 2.5 V in the range 2 to 3 volt.
8. Open bridge 1 and close bridge 2. The control voltage of the closed loop should now lie between 2 and 3 volt. By carefully (!) adjusting the VCO tuned circuit, bring the control voltage to exactly 2.5 V.
9. The counter connected to TP 2 shows a slight jitter on the VCO frequency. By a steady adjustment of the 470 Ω preset in the control loop, a point will be found at which the PLL has "locked-in" completely.
10. Switching the supply voltage on and off, the frequency of the VCO should return and remain stable.
11. The tuning of the sound oscillator to 5.5 MHz follows using the frequency counter, the preset in the G2 circuit of the BF 900 is set to zero resistance. It is then adjusted so that

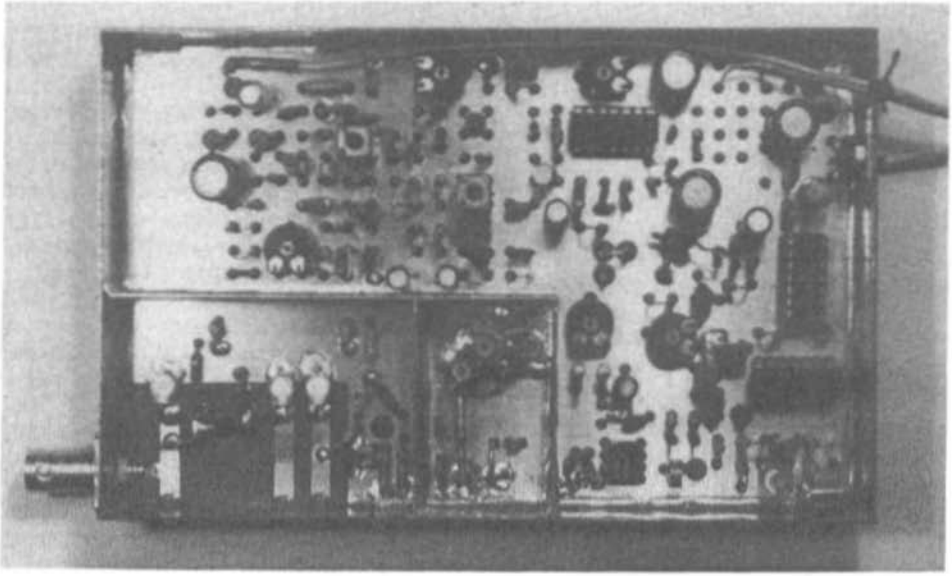


Fig. 5: A completed example of the circuit DC 1 OP 001

the deviation of the sound carrier does not cause horizontal bars in the monitor screen.

12. The video deviation in the 13 cm band should be about 6 MHz which corresponds to a video voltage of 500 mVpp at BR2.

This completes the driver adjustments. If the unit is built as a separate item for inclusion into an integrated system, then the tuning may have to be optimised when that system has been completed. The same applies if the unit is to be fitted with a cover, although this is not strictly necessary.

The most favourable values for video and audio deviations may be determined during actual operational conditions. The sound deviation (preset at G2 BF 900) should be so optimised that when a noise-free picture is received there should be barely noise audible in the sound channel and also no sound interference on the picture.

Other important specifications are not known by the author – they depend upon the receiving system. Definate values can, however, be achieved if the amplitude of the sound carrier at BR2 is measured in the absence of a video signal. It should be as that in point 6.

5. TRANSMIT-DRIVER DATA

The following data was collected from the prototype using a supply voltage of 12.5 V and without audio or video modulation: –

P_{outT} ± 12 mW at 1167.5 MHz

f (VCO)	– 35 dB
3f (VCO)	– 38 dB
4f (VCO)	– 38 dB
f (ref)	– 52 dB

Sideband suppression approx. 60 dB

PLL-lock-in time approx. 2.1 s

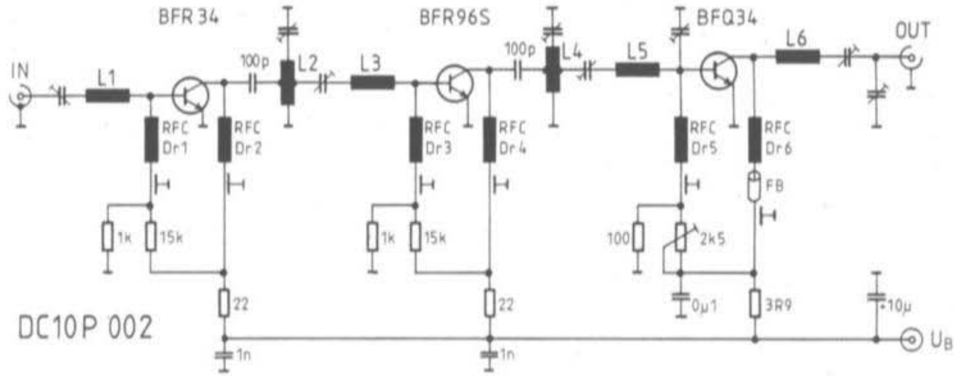


Fig. 6: Three-stage 24 cm-band transmit amplifier

6. TRANSMIT AMPLIFIER

A power amplifier, to produce the necessary power to drive a varactor doubler, was designed and constructed. With a supply voltage of 13.5 V and driven by the subject driver, it delivered an output power of 1 W at 1.1 GHz. The circuit schematic of this amplifier is shown in **fig. 6** and the component placement is shown in **fig. 7**.

6.1. Component List

Semi-conductors: BFR 34, BFR 96 S, BFQ 34

Inductors:
L 1 to L 6 etched

RFC 1, 3:	5 turns 0.2 CuL, length 5 mm	d = 2.5 mm,
RFC 2, 4:	3 turns 0.2 CuL, horizontal	d = 2.5 mm,
RFC 5:	4 turns 0.5 CuL, length 5 mm	d = 2.5 mm,
RFC 6:	2 turns 0.8 silvered, horizontal	d = 5 mm,

Trimmer:
VALVO foil trimmer (grey) 0.7...5 pF, RM 5 or SKY (green) 0.7...5 pF, RM 5

An inspection of the circuit diagram will reveal that although the amplifier has been designed for FM signals, it is nevertheless a linear device. The

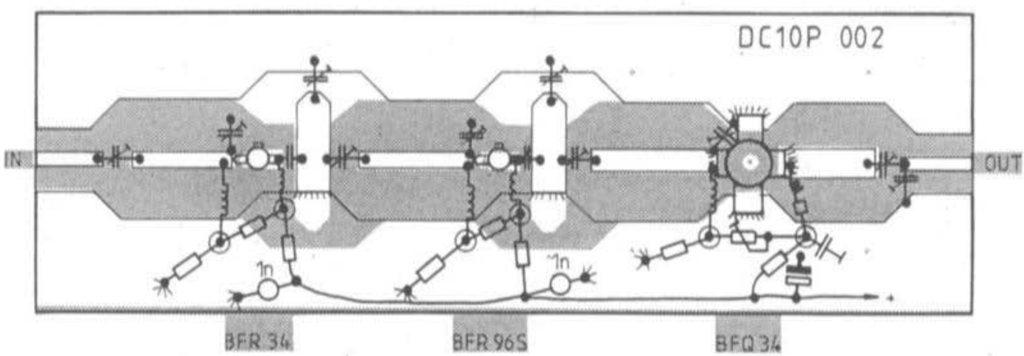


Fig. 7: The printed circuit board for the circuit in fig. 6

reason for this lies in the tendency to self-oscillation of the available transistors (irrespective of manufacturer) when being driven in class C. The quiescent current of the first two transistors are small, being in the order of 1 to 3 mA, and constant. The BFQ 34 has a standing current of 50 mA.

6.2. Construction of Transmit-Amplifier

The circuit is constructed on a double-sided, glass-epoxy printed circuit board, one side remaining unetched and serving as a ground-plane. The other side contains the components and the etched out portion.

At the cold ends of L 2 and L 4, slots are cut through the board in order that both sides can be bonded together with short copper strips. This is also carried out for the emitter leads of the output transistors. The rotors of the trimmers, also, are soldered at both sides of the board.

The 100 pF decoupling capacitors serve as soldering support pillars on the board. The final transistor is mounted upon an aluminium U-shaped heat-sink, size 40 x 70 mm, which is screwed to the PCB before the transistor is affixed to it.

The entire board can then be soldered into a tinplate or copper frame; both sides making continuous contact with the walls. Care must be taken with the location of the input and output BNC panel mounting sockets to ensure that the centre pin sits nicely on the appropriate pads provided on the etched side of the board.

6.3. Commissioning and Tuning

Following a visual inspection, turning the 2.5 kΩ preset to its maximum value and the trimmers to 30 % of maximum capacity, the supply voltage can be applied. The quiescent currents of both the input transistors are checked and that of the output transistor BFQ 34 set to its nominal value by the preset.

A drive voltage is applied to the input which normally results in a weak output signal. The trimmers are all adjusted for maximum output power. Normally, the trimmers of the first two stages will be near minimum capacitance and those of the output stage are at half capacity. The shunt output trimmer is nearly at maximum capacity. The driven current in the BFQ 34 should be in the region of 130 to 150 mA.

The tuning and testing was carried out using an absorption circuit.

7. THE VARACTOR DOUBLER

The prototype used a proprietary doubler, recommended by the literature (5), from 23 to 13 cm. Instead of the usual BXY 27 (VALVO), a DH 110 (Thompson) was employed, it being just as effective but cheaper. In this application, both diodes exhibit the same performance characteristics.

8. OPERATION

In order that there is no time delay during "break" operation, the PLL is continuously in operation even during "standby", thus ensuring the reception of a complete picture at all times. In this condition, the power amplifier has no supply voltage and, in conjunction with the antenna cable, acts as an attenuator between signal source and antenna. This technique works well in ATV relay working but when using single frequency operation, on the whole, it is better to switch off the PLL as well during "receive".

In accordance with the considerations, outlined in the beginning of this article, the varactor doubler was installed in a weather-proof container directly behind the antenna. The first experiments showed that the doubler had to be tuned to match the transmitter (antenna feeder cable). An attempt was made to do the tuning more conveniently by using an identical length of cable but this was not successful and it became

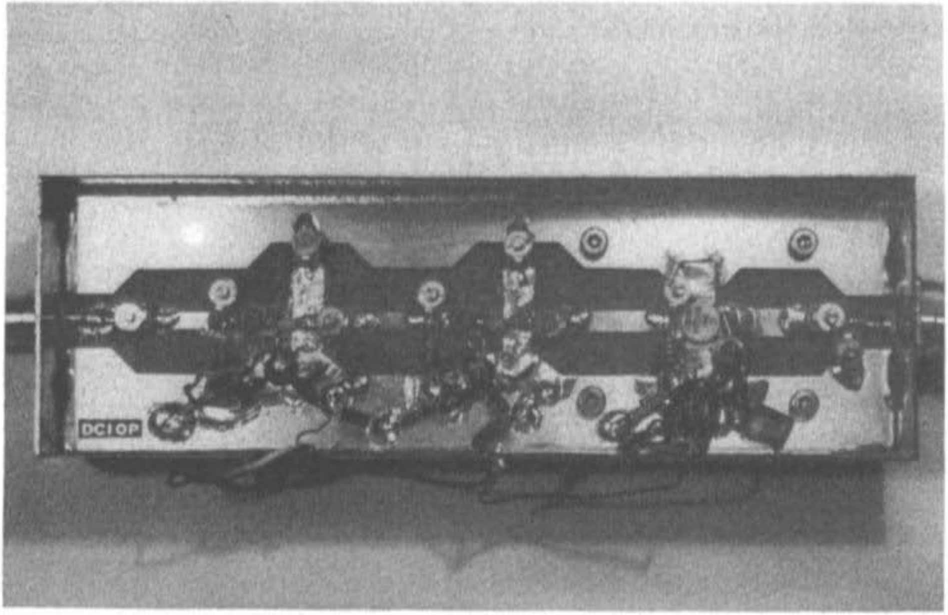


Fig. 8: A completed example of the DC 1 OP 002 transmit amplifier

clear that the tuning had to be carried out at the antenna. The power difference amounted to some 3 dB.

This concept was therefore abandoned and the doubler was installed in the transmitter housing. In order to compensate for the increased feeder losses at the transmit frequency, a further half-frequency power amplifier was used before the doubler.

9. REFERENCES

(1) Best, Dr. Roland:
Theorie und Anwendungen des Phase-Locked-Loops
AT Fachverlag GmbH, Stuttgart

(2) Motorola Datenblatt MC 4344/MC 4044, April 1970

(3) AEG-Telefunken
Halbleiterübersicht Integrierte Schaltungen, Ausgabe 1983

(4) Hirschelmann, Klaus H.:
Eine Amateurfunkfernsehstation im 10-GHz-Band
Abschnitte 2.2 und 2.3
TV-Amateur 45/1982

(5) Heidemann, Rolf:
SSB Transmit Mixers for the SHF Bands. Part 1: 13 cm Band
VHF COMMUNICATIONS, Vol. 11, Ed. 2/1979, P. 86 - 96

(6) Beckmann, Udo and Dahms, Juergen:
A Transistorized Linear Amplifier for the 23 cm Band
VHF COMMUNICATIONS, Vol. 11, Ed. 1/1979, P. 17 - 26

(7) Hupfer, Konrad:
An SHF Wavemeter
VHF COMMUNICATIONS, Vol. 7, Ed. 2/1975, P. 90 - 92

Wolfgang Schneider, DD 2 EK

70 cm Converter using GaAs-FET CF 300

Converters for the 70 cm amateur band are now an old topic. This article, however, will describe an amateur state-of-the-art, 70 cm converter which translates into 144 MHz, 28 MHz or into channel 2 (47 to 54 MHz) for amateur TV.

The version described here, translates the 432 MHz frequency band into 144 MHz. This is then used for a basis to describe other versions which may have other receiving frequencies or other output frequencies according to the modification table 1. **Fig. 1** shows that this converter only requires four transistors.

Portable operation has, of course, not been forgotten inasmuch that the converter works with a 9 volt regulator (supplying all four transistor stages) which may be supplied by a voltage in the range 10 to 15 volt.

1. CIRCUIT DESCRIPTION

The detailed circuit diagram of the 70 cm converter is shown in **fig. 2**. An established local oscillator chain was used, the oscillator working at 96 MHz with a three fold-multiplication to 288 MHz. The oscillator frequency (**see table 1**) multiplication must be increased to four-fold for the 28 MHz and ATV-Converter versions.

The received signal in the 70 cm band is taken via a λ/4 transformer to gate 1 of the dual-gate MES-FET T1. The signal is then amplified by T1 and coupled to gate 1 of the mixer via a stripline circuit. The local oscillator signal at 288 MHz is intro-

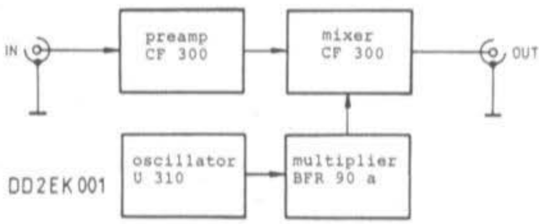


Fig. 1:
The converter comprises only four stages

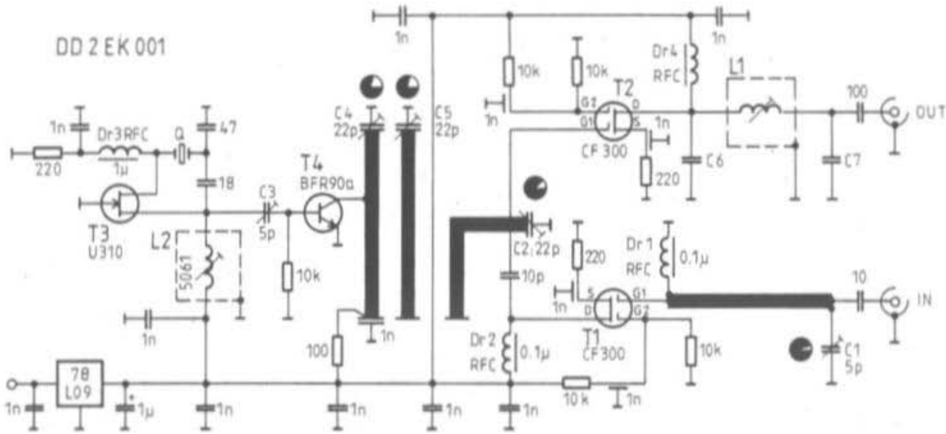


Fig. 2: Circuit schematic of the 70 cm converter. The IF dependent components are given in table 1

duced via a bandpass filter. A dual-gate MESFET is also used for the mixer. The IF at 144 to 146 MHz is taken to the output via a low-pass filter.

An on-board voltage regulator (9 V) enables the module to be supplied from voltage fluctuations in the range 10 to 15 V.

2. CONSTRUCTION

The printed circuit board, shown in **fig. 3**, is double-sided, copper-coated, 1.5 mm thick, epoxy material with dimensions of 53.5 mm x 108

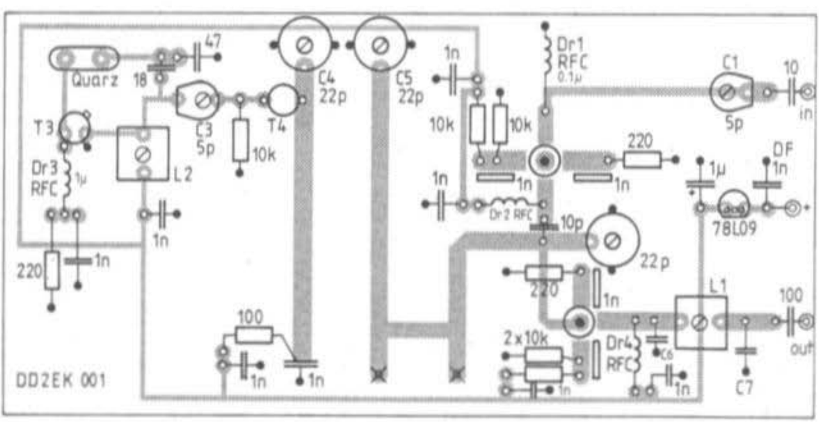


Fig. 3: The printed circuit board (PCB) has a full copper ground plane on the component side

mm. It will fit into a proprietary-made, tin-plate box 55.5 mm x 111 mm x 30 mm. All the resonant circuits take the stripline form being etched out onto the board. This serves to guarantee a high degree of reproducibility and also an extraordinary degree of selectivity is achieved. Detuning effects, caused by fitting or removing the housing cover, could not be detected in the prototype version.

After cutting and preparing the PCB, it is, first of all, rubbed down with a silver-coating chemical and then drilled. The silvering process is not absolutely necessary. For the stripline transistors suitable holes are to be drilled, the transistors then lie in the board. The chip capacitors, also, fit into slots cut out with a jig-saw. The bandpass filter contacts are effected with 1.5 mm silvered copper wire. Recesses of 1 mm x 8 mm are provided at the appropriate places at the board edges for the fitting, and eventual soldering, of the BNC centre connecting pins.

After drilling, the holes for the transistors, capacitors, crystal, coils etc. are then chamfered out on the component side with a 2.5 mm drill so that the copper (only) around the component drilling is removed in order that components are not shorted to earth by their leads. The slots for the chip capacitors are also chamfered, on the side that is not soldered to the board, for the same reason.

When this work has been accomplished, the board may be treated with solder spray and the chip capacitors and the feed-through contacts for the band-filter circuits installed. The rest of the components are mounted after the PCB has been soldered into the tin-plate housing.

The BNC panel-mounting sockets must lie with their flanges flush with the edge of the PCB so that the centre pin contacts the surface of the board along its entire length. If necessary, the teflon insulation on the BNC socket must be partly removed to achieve this. Check that after the components have been installed, particularly the connectors and the provisional fitting of the band-filter, that the housing cover can still be fitted securely.

Following the soldering of the PCB into the housing, the rest of the components may be soldered into the board.

2.1. Components

T1, T2:	CF 300 (Telefunken)
T3:	U 310 (Siliconix)
T4:	BFR90A (Valvo)
1 9 V-regulator	78L09
L1:	Inductor (Neosid) see table 1
L2:	Inductor (Neosid BV 5061: be/bn)
RFC 1, RFC 2:	0.1 μH (Neosid)
RFC 3:	1 μH (Neosid)
RFC 4:	(Neosid) see table 1
C1, C3:	Foil trimmer 5 pF (SKY, green)
C2, C4, C5:	Foil trimmer 22 pF (Valvo, green)

in	out	C 6	C 7	L 1	ch 4	crystal
438 MHz	144 MHz	22 pF	22 pF	BV5061 be/bn	1 μH	98 MHz
435 MHz	144 MHz	22 pF	22 pF	BV5061 be/bn	1 μH	97 MHz
432 MHz	144 MHz	22 pF	22 pF	BV5061 be/bn	1 μH	96 MHz
434 MHz	48 MHz	56 pF	56 pF	BV5049 yw/we	4.7 μH	96.5 MHz
438 MHz	28 MHz	82 pF	82 pF	BV5049 yw/we	10 μH	102.5 MHz
435 MHz	28 MHz	82 pF	82 pF	BV5049 yw/we	10 μH	101.75 MHz
432 MHz	28 MHz	82 pF	82 pF	BV5049 yw/we	10 μH	101 MHz

Table 1: Working frequency dependent components

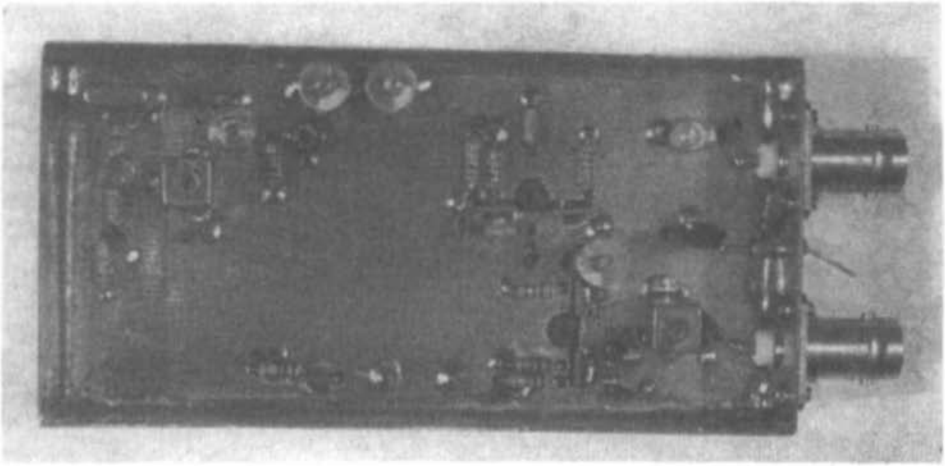


Fig. 4: This example is equipped for a 144 MHz IF

C6, C7: ceramic disc cap. RM 2.5; values see table 1

Further disc capacitors. RM 2,5:
2 x 10 pF, 1 x 18 pF, 1 x 47 pF, 1 x 100 pF, 9 x 1 nF
5 chip capacitors ca. 1 nF
1 tant. perl cap. 1 μF/16 V
1 ceramic feed-through cap. ca. 1 nF (solder-in)
All resistors: 0207 (for 10 mm grid)
Crystal: HC-18 U or HC-25U, frequencies see table 1
2 BNC panel sockets UG-290 A/U
1 tin-plate box, 111 x 55.5 x 30 mm

3. TUNING

The following test equipment should be available:

- Multimeter
- 500 MHz frequency counter
- suitable diode probe
- 70 cm signal source which can be introduced to the converter input via the antenna

First the frequency processing circuits of the local oscillator are aligned. The oscillator is tuned to the working frequency of 96 MHz with tuning inductor L2 and then the tripler to an output frequency of 288 MHz. The appropriate trimmers are shown in the schematic of **fig. 2**.

The IF tuned circuit is tuned for maximum noise and then a local 70 cm signal tuned in. The three tuned circuits between the pre-amplifier, multiplier and mixer are carefully adjusted with a plastic trimming tool for the best signal strength.

Finally, the input trimmer of the pre-amplifier is adjusted for best signal-to-noise ratio.

3.1. Observations

Several examples of the described converter were constructed and technically tested. All submitted similar data. The noise figure (NF) is 2 dB at a gain of 16 dB. The current consumption of the 70 cm converter is ca. 45 mA.

The outstanding specifications serve to indicate that, even with simple means, home-constructed equipment can achieve respectable results.

Armin Roesch, HB 9 MFL

A 1296 MHz, 200 mW Driver using SMD Technology

There are also advantages for radio amateur constructors in the employment of the SMD (Surface Mounted Devices) used to an increasing extent in industry. Using the SMD technology, high-performance SHF circuits using miniature components can be realised, thus tending to make equipment smaller and handier for portable use.

An amplifier for the SSB portion of the 23 cm band is described in this article which, thanks to CAE (EESOF Touchstone), can be built and optimised without the need for trimmers. The amplifier delivers an output power of 200 mW, at the 1 dB compression point, which is sufficient to drive the Mitsubishi M57762 linear, power amplifier. The driver's gain is approximately 34 dB, so that just about 100 µW should be available from the send mixer to drive the system to full output power.

1. THE CIRCUIT AND ITS EMPLOYMENT

The circuit in **fig. 1** may be observed to be a three-stage amplifier with a pi-section input attenuator.

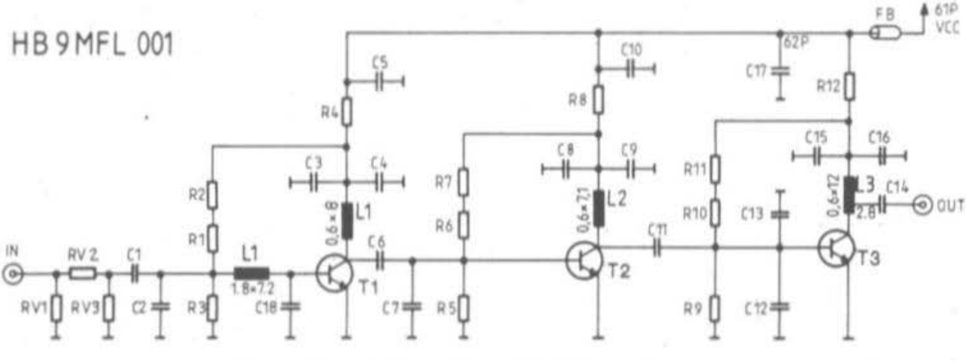


Fig. 1: Three-stage 1296 MHz amplifier with stripline dimensions indicated

attenuation dB	RV1/RV3 Ω	RV2 Ω
1,0	470	10
2,0	470	12
3,0	270	18
4,0	220	22
6,0	150	39
8,0	120	56
10,0	100	82

Table 1: Resistor values for various input attenuator requirements

This pad is used for matching the sender mixer/filter to the amplifier. The resistance values of the pads for the required degree of attenuation may be obtained from **table 1**.

The three stages are connected in the common-emitter arrangement and have a gain of 34 dB at 1297 MHz. The emitters are connected directly to ground in the interests of circuit stability. The second series resistance in the base circuit bias potentiometer chain, serves to control the quiescent current to the required value in the case of all stages. The working voltage must be stabilised at 12 V. For use with a lead-acid battery or NICADs, a 10 V stabilisor is a better choice as the battery capacity is then more efficiently utilised but, of course, the output is then a little lower.

The curve of **fig. 2** indicates the nearly constant output of the response between 1200 and 1400 MHz, the vertical graduations being 10 dB per box. Expanding the gain in the region of the hump in the response to 1 dB/box, more detail may be seen as shown in **fig. 3**.

An output power of 200 mW is obtained when the supply voltage is 12 V but the gain is 1 dB lower than at smaller input drive levels (1 dB compression). The amplifier should never be driven beyond this point as intermodulation will be produced, giving rise to the dreaded "splatter" effect on the transmit output signal.

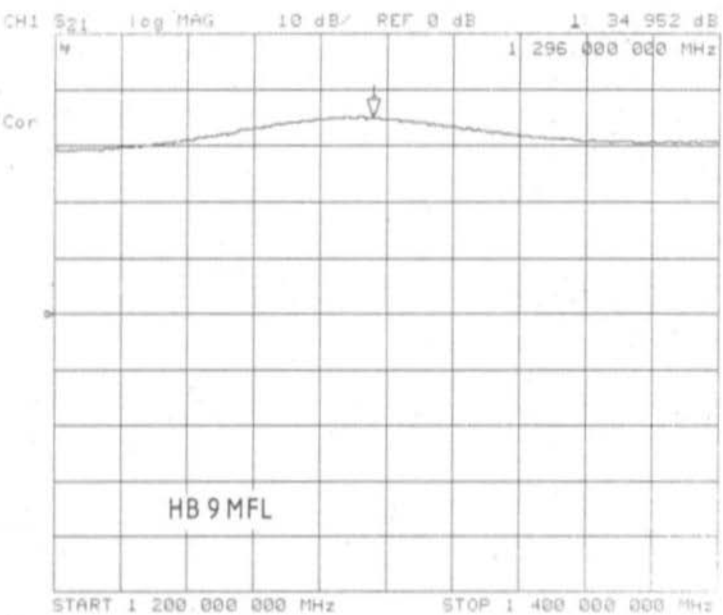


Fig. 2: The SMD technology amplifier has a very broad band frequency response

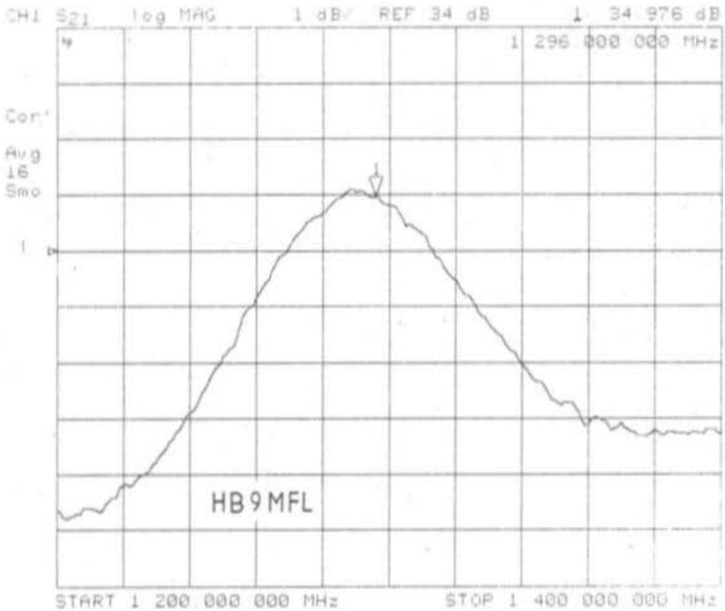


Fig. 3: Only with a magnification to 1 dB/box is the 3 dB bandwidth seen as being 90 MHz

As mentioned earlier, this linear driver can be employed with the Mitsubishi module M57762 power amplifier. Tests on several examples of the M57762 have shown that, at the 1 dB compression point, 12.5 W at a gain of 17 to 18 dB, is obtainable. A power of 20 W can be obtained but only at totally saturated drives and usable only with FM or telegraphy with suitable low-pass filtering.

A power of 12.5 W is 41 dBm rel. 50 Ω. Of this, 18 dB must be subtracted to give a required drive of 23 dBm, i.e. 200 mW – exactly the power which is available from the subject linear driver.

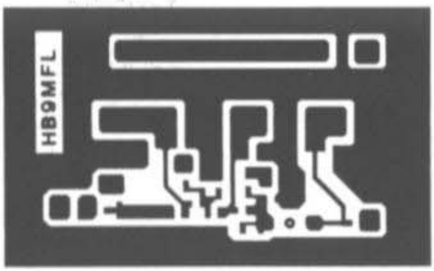


Fig. 4: The PCB HB 9 MFL 001 is constructed from normal glass-epoxy board

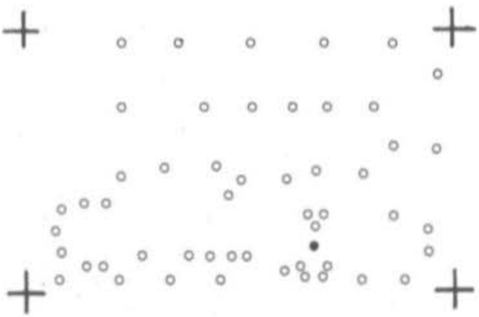


Fig. 5: All the through-plated poles are made with a 0.6 mm drill. A 5 mm hole is required for T3

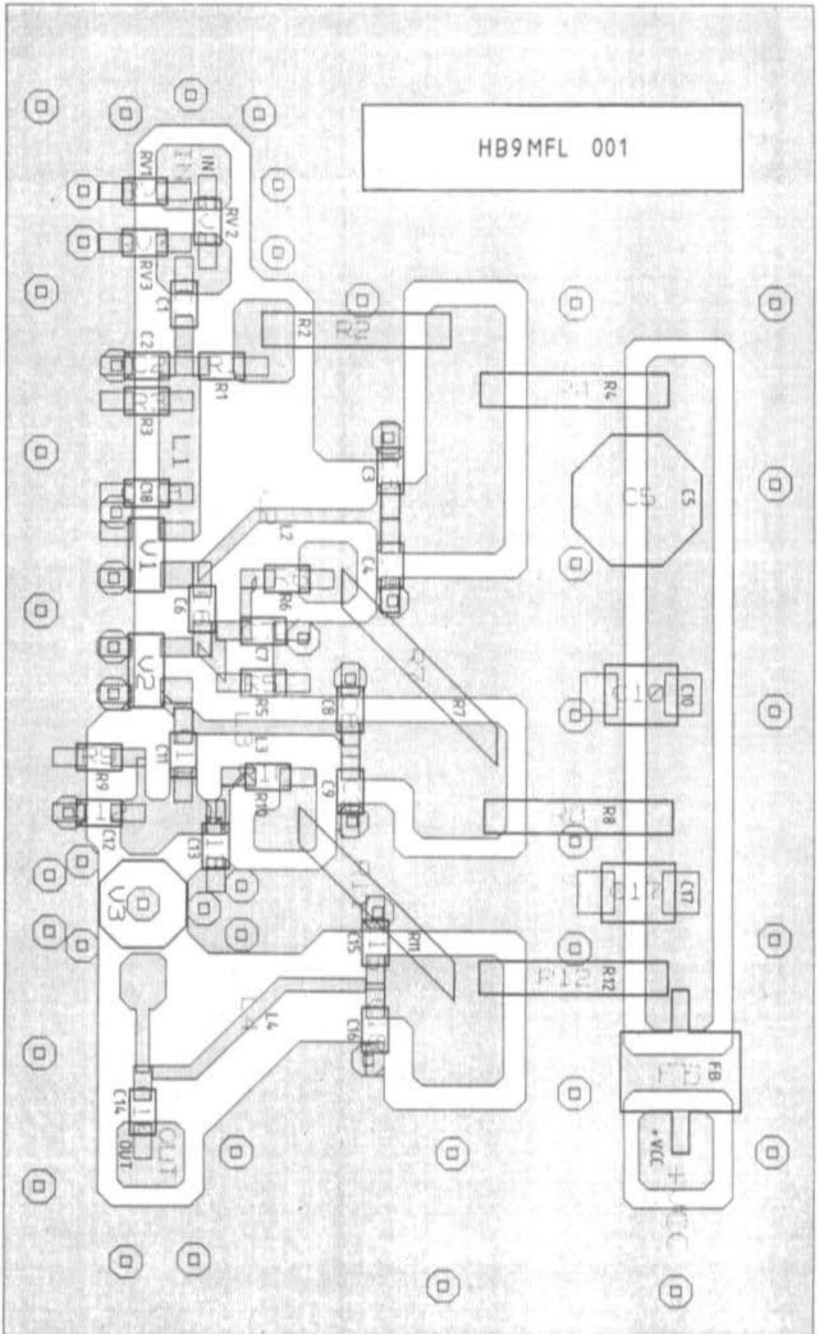


Fig. 6: The enlarged printed circuit board HB 9 MFL 001

2. CONSTRUCTION

The printed circuit board, developed for this project, is shown in **fig. 4**. The board is 1.6 mm thick, of glass-epoxy material, is double-sided and through-contacted. Its designation is HB 9 MFL 001 and is dimensioned 34 mm x 52 mm in order that it can be soldered into a 37 x 74 x 30 mm tin-plate enclosure.

To the printed circuit board belongs a drilling plan, shown in **fig. 5**, and a component layout plan, shown in **fig. 6**. It is possible to detect the "hand of the computer" here, and it may also be seen that the input matching attenuator is also accommodated on the small board.

The board, with its many through-plated connections, facilitates the construction of the driver module. No difficulties can be anticipated – except perhaps, that caused by the lack of hands-on experience with SMD components. They are sensitive to heat and therefore the soldering iron should be of modest size and not dwell too long on the joint. A very thin solder with a silver content is also advisable.

Whilst the first two amplifying stages are fitted with SMD transistors, the output stage has a "normal" BFG 34 stripline transistor. A 5 mm drilling is required to accommodate it. To obtain a low-resistance emitter-ground connection here, the upper and lower sides are connected together with a small copper band. The transistor is laid in the hole so that its emitter connector lies flat on the board.

2.1. Component list

V1:	BFG 92 A	(Philips/Valvo, designation P8, form SOT 143)
V2:	BFG 93 A	(Philips/Valvo, designation R8, form SOT 143)
V3:	BFG 34	(Philips/Valvo, form SOT 103)

C1, C3, C9, C14, C15:	22 pF chip cap. 0805, NPO
C2:	5.6 pF chip cap. 0805, NPO
C4, C8, C16, C17:	1 nF chip cap. 0805, XR7
C5:	10 μF tant. perl 16 V electrolytic
C6:	1.8 pF chip cap. 0805, NPO

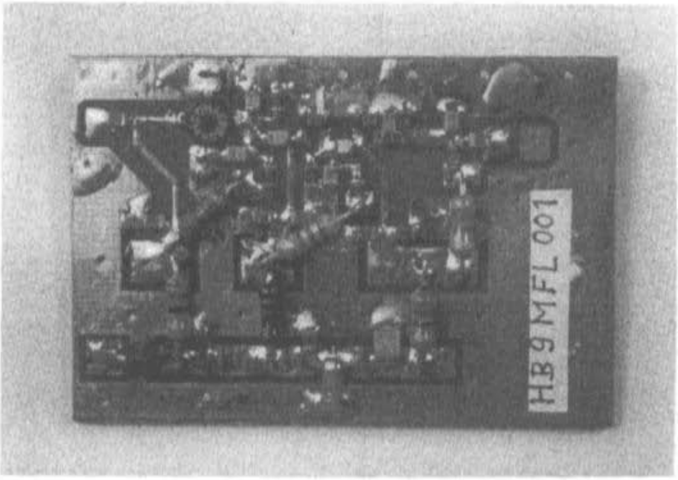


Fig. 7: The author's prototype

C7, C8:	3.3 pF chip cap. 0805, NPO
C10:	100 nF chip cap. 1210, XR7
C11:	6.8 pF chip cap. 0805, NPO
C12:	4.7 pF chip cap. 0805, NPO
C13:	3.9 pF chip cap. 0805, NPO

R1, R6:	8.2 kΩ chip resistor 0805 65 mW
R2, R7, R11:	"normal" resistors 0207, (100 Ω - 5.6 kΩ) value to be determined in operation.
R3, R5:	1.0 kΩ chip resistors 0805, 65 mW
R4:	82 Ω"normal" resistor 0207
R8:	120 Ω, as R4
R9:	470 Ω chip resistor 0805, 65 mW
R10:	3.3 kΩ chip resistor 0805, 65 mW
R12:	22 Ω "normal" resistor 0207

3. SETTING-UP

The completed board is checked thoroughly and compared against the photograph of **fig. 7**. The supply is then connected but without HF drive.

R4 in the first stage is then measured for a 1 V potential difference across it which indicates that the collector current of T1 is 15 mA. If it is not, change the value of R2.

In the same manner, R7 is changed until the PD across R8 is 3 V. T3 then has a collector current of 25 mA.

Finally, R11 is changed so that R12 has a PD of 1.6 V across it yielding a T3 quiescent current of 75 mA.

This is all that is necessary to bring the 1296 MHz SMD amplifier into an operational condition.

BRIEFLY SPEAKING...

Chip Capacitors

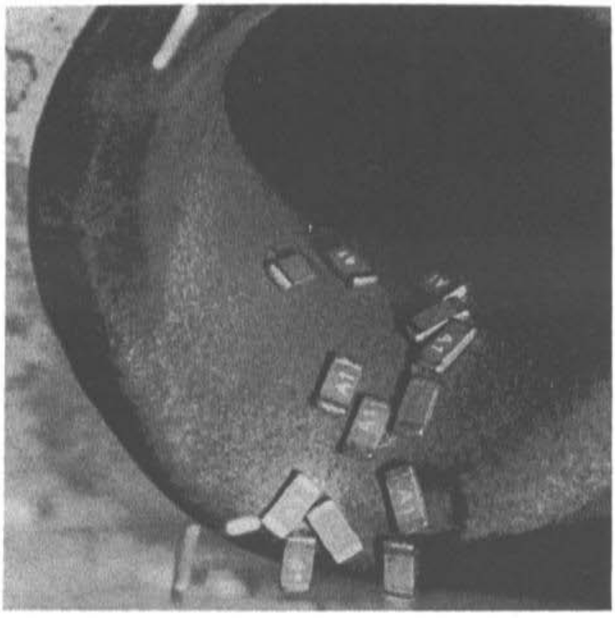
These multi-layer capacitors (B 37940/871) were conceived by Siemens for use in tuners and are generally accepted as being the fore-runners in the surface-mounted device (SMD) technology. The most frequent sort of capacitor, demanded for tuner use, is still the ceramic type but after the disc and trapezoidal form had lost their prominence, the multi-layer type became the preferred type. The absence of connecting leads has increased its HF properties and now a new ceramic with a particularly low dielectric constant has reduced the loss factor by some 30 %, measured at 500 MHz.

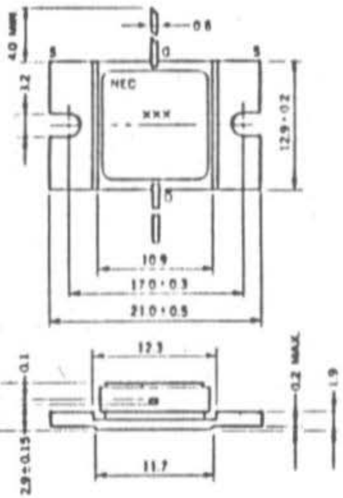
Siemens press picture

NEC C-Band Power GaAs FETs

The power FET series NEZxxxx-3 and NEZxxxx-6 have been replaced by the new series **NEZxxxx-3A, 6A and 8A**. The latest series have shorter gate structures (smaller than 1 μm) by means of which higher linear amplification can be achieved as well as lower thermal resistance enabled by the plated heat sink and the "hole structure". A Ti-Pt-Au metalising, together with a tempering process using SiN and SiO_2 guarantee a high reliability and stability. The series has already been qualified for applications in space.

The transistors are matched to the radio industry norm of 50 Ω. The 3 Watt transistors are





specified with a $P_{1\,dB}$ = 34.8 dBm and possess a linear amplification characteristic from 11 dB at 4 GHz to 7.5 dB at 8.4 GHz.

The 6 W transistors have a $P_{1\,dB}$ = 37.3 dBm (Version M) or a $P_{1\,dB}$ = 37.8 (Version L) and 10 dB (4 GHz) to 8 dB (7.2 GHz) specified amplification.

For the bands 3.7 to 4.2 GHz and from 5.9 to 6.4 GHz, 8 Watt transistors are available. Typical output power: 39 dBm at 36 % efficiency.

The transistors for most frequency bands are available at short notice.

Microscan GmbH,
Tel. 089-96163

New NEC Wideband Amplifiers

NEC now offer two new wideband amplifiers packaged in an SMD mountable housing. Both input and output impedances are the standard 50 Ω.

New GaAs FETs

NEC has introduced two new low-noise GaAs FETs: the **NE 202** is an "AlGaAs/GaAs-hetero-junction FET".

The electrons are excited to higher mobilities owing to the hetero-junction which has the effect of improving the HF characteristics.

The NE 202 is available in chip (NE 20200) or in a housing (NE 20283) packaging. Typical noise figures are 1.0 dB (12 GHz) or 1.5 dB (18 GHz) and typical amplifications are 10 dB (12 GHz) or 8 dB (18 GHz).

The NE 20283 A is available from stock. The second new introduction is the Dual GaAs FET **NE 25000** which is only available in **chip form**. Typical applications are e.g. active mixers for the X-band, or AGC amplifiers which exhibit a dynamic range of 35 dB (12 GHz).

Microscan GmbH,
Tel. 089-96163

Typical data	μPC1675	μPC1676
Frequency range (−3 dB)	1900 MHz	1200 MHz
Power gain (S21)	12 dB	22 dB
Noise figure	5.5 dB	4.0 dB
Isolation (S12)	24.5 dB	28 dB
Output power	4.5 dBm	7 dBm

Microsan GmbH

World Map for Radio Amateurs

Format 86 x 61 cm;
Verlag fuer Technik und Handwerk, Baden-Baden

This map is based upon the official international country listings of the ITU and those of the world authorititive DXCC lists of the ARRL. It may be taken for granted that the country identification call-sign prefixes are absolutely up-to date.

The map is drawn in the Mercator projection and distance scales facilitate the conversion between miles and kilometres.

It represents a useful addition and a work-horse for every ham-shack.

Radio Locator Maps

Central Europe;
Scale: 1 : 1.5 Million
Format: 66 x 84 cm
Verlag fuer Technik und Handwerk, Baden-Baden

The radio hobby magazine "Funk" offers this new map as a working aid in a handy format.

Countries **fully** covered by the map are **Germany, Switzerland** and **Austria**, together with the bordering areas of surrounding countries.

Displayed on the map itself is an explanation which will enable the user to quickly learn to exploit its facilities.

The locator grid is based upon the IARU recommendations which have found world-wide acceptance.

MATERIAL PRICE LIST OF EQUIPMENT

described in edition 1/1988 of VHF COMMUNICATIONS

DB1NV	**DC/DC Converter**		**Art.No.**	**Ed. 1/1988**
PCB	DB1NV 005	2-sided, drilled, without layout plan	6033	DM 34.–
Components	DB1NV 005	all semi-conductors, all inductors (unwound), Litz wire, enamelled copper wire, 1 brass screw, 5 solder tags, 25 resistors (inc. 2 MF), 2 trimpotis, 5 ceramic-, 7 foil- 8 electrolytic caps.		
Kit	**DB1NV 005**	**with all above parts**	**6035**	**DM 160.–**
DD2EK	**70 cm Converter (state band segment/IF)**			**Ed. 1/1988**
PCB	DD2EK 001	2-sided, (one side etched), silvered	6691	DM 29.–
Components	DD2EK 001	1 tin-plate box, 2 BNC skts., 1 voltage reg., 4 transistors, 2 filters, 4 RFCs, 5 trimmers, 1 feed-thro'-, 16 ceramic caps., 1 tant.cap., 5 chip caps., 9 resistors, 1 crystal	6692	DM 100.–

Please note: The kit/components contain crystals and specific inductance and capacitive values according to the frequency pairings specified. Please state by ordering. Details of the frequency dependent components are given in **Table 1**.

Kit	**DD2EK 001**	**PCB + components for given frequencies**	**6693**	**DM 125.–**
HB9MFL	**A 1296 MHz, 200 mW Driver in SMD Technology**			**Ed. 1/1988**
PCB	HB9MFL 001	2-sided, through-plated	6985	DM 22.–
DC1OP	**ATV-FM Transmit Driver for the 23 cm Band**			**Ed. 1/1988**
PCB	DC1OP 001 and 002			upon request
DK1OF	**A 2 m/70 cm Antenna Splitting Filter**			**Ed.1/1988**
Kit	DK1OF AW 2/70		6009	DM 55.–
Kit	**ready-to-operate DK1OF AW 2/70**		**6008**	**DM 125.–**

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, 2/1971, 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

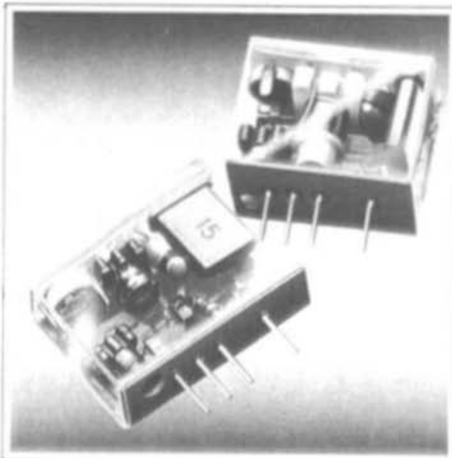
You should know what's behind our sign

We are the only European manufacturers of these

Miniature TCXO's

CCO 102, CCO 103, CCO 104, CCO 152
modulable table

- higher stability than a quartz crystal: less than ± 3 ppm over the temperature range –30 to +60°C. (types B)
- low ageing rate: less than 1 ppm per year.
- wide frequency range: 10 MHz to 80 MHz
- low supply voltage: +5 V
- low current consumption: 3 mA max. (series CCO 102)
- small outlines: CCO 104 = 2,6 cm³, CCO 102/152 = 3,3 cm³, CCO 103 = 4,0 cm³
- widespread applications e.g. as channel elements or reference oscillators in UHF radios (450 and 900 MHz range)



Our R + D engineers are constantly working with new technology to develop new products. We can offer technical advice for your new projects or manufacture against your specification.

Quartz crystal units in the frequency range from 800 kHz to 360 MHz Microprocessor oscillators (TCXO's, VCXO's, OCXO's) crystal components according to customer's specifications

Types	CCO 102	CCO 103	CCO 104
	A | B | F	A | B | F	A | B | F
Freq. range	10 - 80 MHz	6.4 - 25 MHz	10 - 80 MHz
stability vs temp. range	-30 to +60°C	-30 to +60°C	-30 to +60°C
Current consumption	max. 3 mA at UB = +5 V	max. 10 mA at UB = +5 V	max. 10 mA at UB = +5 V
input signal	-10 dB/50 Ohm	TTL-compatible (Fan-out 2)	0 dB/50 Ohm

CCO 152 A + B
same size as CCO 102 A + B
modulation input:
deviation: typ. 1 kHz/V
mod. frequency: DC to 10 kHz
impedance: 20 k Ohm

TELE QUARZ

... Your precise and reliable source

TELE-QUARZ GMBH · D-6924 Neckarbischofsheim 2
Telefon 07268/801-0 · Telex 782359 tq d · Telefax 07268/1435