A Publication for the Radio Amateur Worldwide

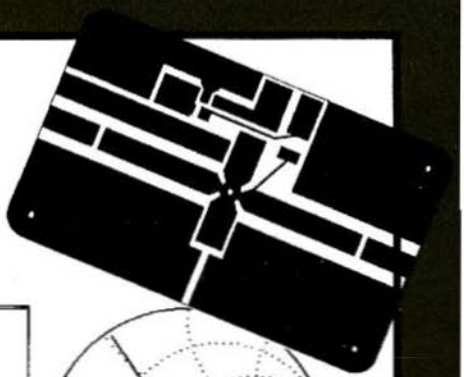
Especially Covering VHF, UHF and Microwaves

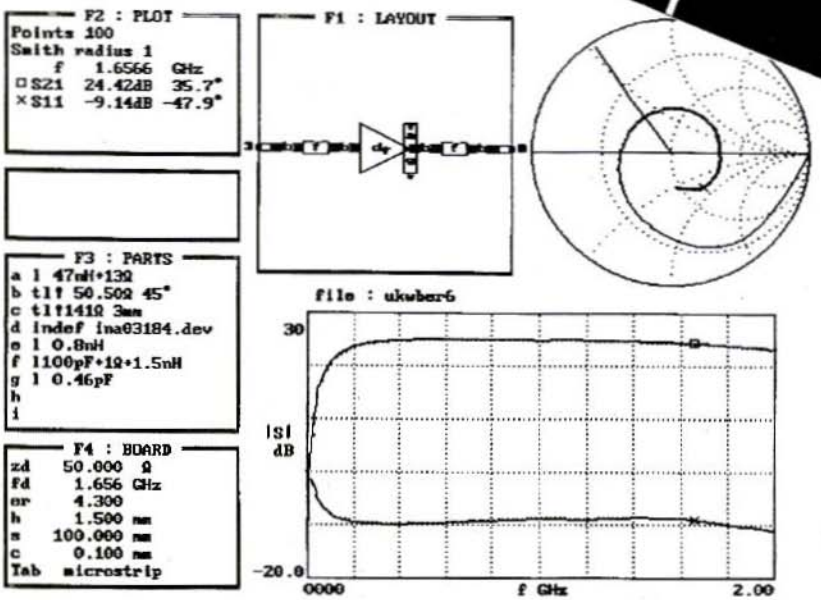
VHF COMMUNICATIONS

Volume No.29 . Winter . 4/1997 . £4.75

Gunthard Kraus, DG8GB

Design & Realisation of Microwave Circuits: 1.7 GHz MMIC Amplifier





BOOKS FOR ENGINEERS AND SERIOUS AMATEURS

Microwave Field-Effect Transistors

By Raymond S. Pengelly

703 pages of data on GaAs FET devices and circuits! Basic information on fabrication, packaging and performance is followed by circuit design data on amplifiers, oscillators, switches, multipliers, mixers, phase shifters and other functions. The most complete book for design using GaAs FETs. **Order NP-8 $89.00**

Transceiver System Design for Digital Communications

By By Scott R. Bullock

A thorough summary of radio system design for digital communications. Draws on developments driven by military systems, but presents them as fundamental information essential for commercial wireless communications system development. 432 pages. **Order NP-7 $54.00**

Transmission Line Transformers

By Jerry Sevick, W2FMI

Professionals and amateurs alike can use this unique resource, a modern classic that answers the questions, "How do they work?" and "How do I build them?" Guanella and Ruthroff configurations are described, with performance verified by hundreds of measurements on balanced, unbalanced and balun types. The best book ever on this subject! **Order NP-9 $34.00**

Oscillator Design and Computer Simulation

By Randall W. Rhea

Understand oscillator design and operation! This book presents a unified approach to design, using a wide variety of active elements and resonator types — bipolar transistors, FETs or MMICs with crystal, L-C, SAW, coaxial, microstrip or other resonators. Strong coverage of basic concepts with many design CAD models and circuit examples. **Order NP-1 $64.00**

HF Filter Design and Computer Simulation

By Randall W. Rhea

Filters for the practicing engineer who actually *builds* filters! Includes lumped element, coaxial, helical, dielectric resonator, stripline and microstrip types. Thorough review of filter principles with practical information on topologies, transformations, component effects and matching. Especially good for the design and realizaton of microstrip filters. **Order NP-2 $59.00**

Order by telephone, fax, mail or our Web site!

We accept payment by VISA, MasterCard and American Express or check drawn on a U.S bank. Shipping charges are – U.S.: $5.00 for the first item, $1.00 for each additional item; Canada: $10.00 first item, $2.00 each additional item; Other countries (Air express): $32.00 first item, $10.00 each additional item.

CRESTONE
TECHNICAL BOOKS

CRESTONE TECHNICAL BOOKS
Div. of Noble Publishing Corporation
2245 Dillard Street • Tucker, GA 30084 • USA
Tel: (770) 908-2320 • Fax: (770) 939-0157
www.noblepub.com

Contents

Wolfgang Schneider DJ8ES	VHF, UHF, and SHF Measuring Methods Using a PC Part-5: mW Meters, Wobblers and Spectrum Analysers	194 - 208
Gunthard Kraus DJ4BG	Design and Realisation of Microwave Circuits; Part-4	209 - 214
Lorenz Oelschlegel DL6NCI	DB1NV's Image Memory in Combination with the HP141 Spectrum Analyser	216 - 217
Denys Roussel F6IWF	An Exceptionally Low-Cost HF SSB/CW Transceiver Part-3: Description of Assembly	218 - 236
Dipl.-Ing Detlef Burchard	The Sine Converter	237 - 251
Matjaz Vidmar S53MV	GPS/GLONASS Receiver Hardware and Software Update #1	252 - 253

Krystyna and I would like to take this belated opportunity to wish you all a Happy and Prosperous 1998 and success in all that you do.

We would also like to thank you for your past and, hopefully, your continued support of VHF Communications Magazine - without you, of course, it would not exist.

Michael

KM Publications, 5 Ware Orchard, Barby, Rugby, CV23 8UF, UK

Telephone: (0)1788 890365; INT: +44 1788 890365; FAX: (0)1788 891883

Email: michael.j@vhfcomm.co.uk WWW: http://www.vhfcomm.co.uk

Wolfgang Schneider, DJ8ES

VHF, UHF and SHF Measuring Methods Using a PC

Part 5: mW Meters, Wobblers and Spectrum Analysers

5. mW METERS

The mW meter can be used to measure power levels between 100mW (+20dBm) and 10µW (-20dBm). For this purpose, two functional units are used from the hardware range available.

The power level - or a voltage corresponding to the power level - is measured using the diode detector (1SS99) with an integrated 5W moving load.

This voltage must be suitably prepared for the PC - say, be digitised. The A/D converter is used as the second assembly here.

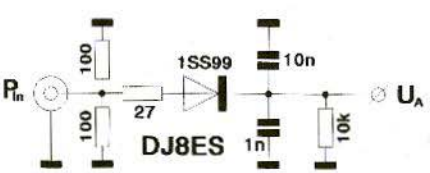


Fig.20: Diode Detector Circuit

5.1. Diode Detector

The diode detector has a dynamic range of over 50dB with a 1SS99. This high

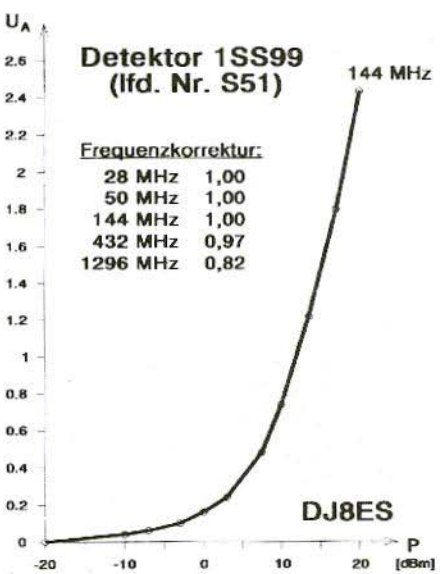


Fig.21: Output Voltage to Input Power Ratio;
Detektor **= Detector,** ***Lfd. Nr.*** **= Serial no.,** ***Frequenzkorrektur*** **= Frequency correction**

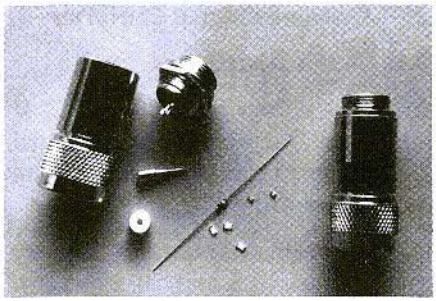


Fig.22: The BNC DetectorVersion

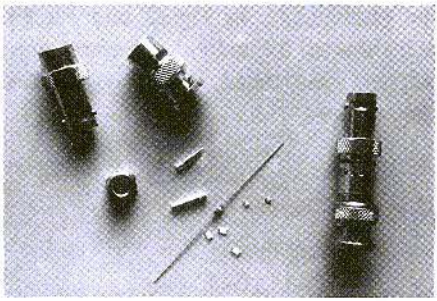


Fig.23: The N-Norm Detector

value naturally requires the circuit to be assembled to be suitable for radio frequencies! SMD components and short connecting wires where wired components are used are an absolute must.

Fig.20 shows the detector circuit; Fig.21 shows its output voltage, plotted against the measured input power in the range between 10mW (- 20dBm) and 100mW (+ 20dBm).

5.1.1. Assembly Instructions

The detector is inserted in the standard way, in the BNC version. As an air assembly, directly incorporated in a plug / jack, it obtains the best performance.

All the details for assembly have already been described - in Part 2 of this series of articles ("mW meters for short-wave to SHF"). Assembly in an N-plug is naturally also possible.

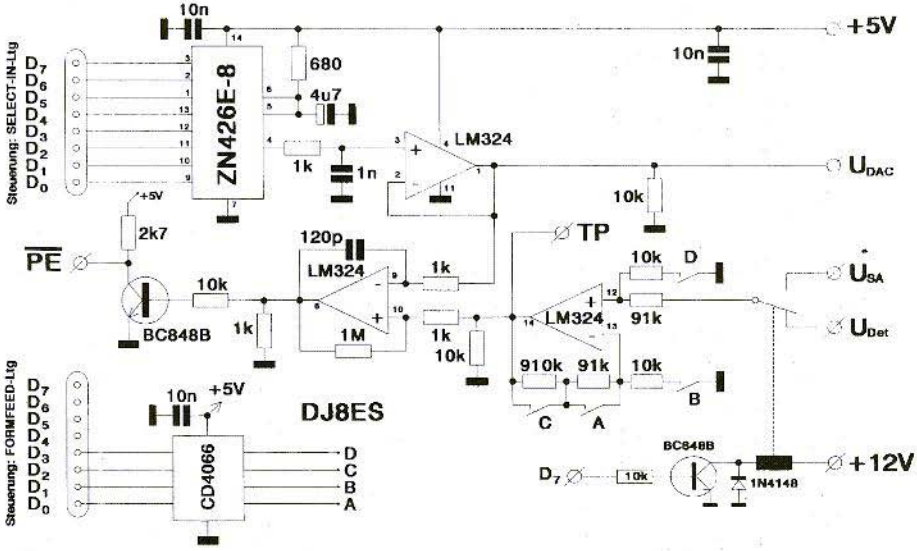


Fig.24: A / D Converter Circuit with Detector*Steurung* = Control

The photographs in Fig.22 and Fig.23 each show all the individual components and the associated, fully assembled detector head.

5.1.2. Component List

1 x 1SS99 diode
1 x UG 88/U BNC plug
(Alternative: 1 x N-plug (Aircom))
1 x UG 89/U BNC cable jack
(Alternative: 1 x micro-jack (2-pin))
1m. RG58U coax cable with suitable pin-and-socket connectors

SMD resistors and capacitors in 1206 format:

1 x 27Ω
2 x 100Ω
1 x 10kΩ
1 x 1nF
1 x 10nF

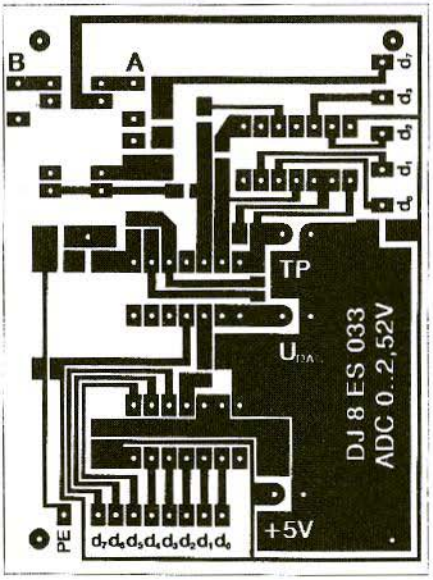


Fig.25: DJ8ES 033 PCB; A / D Converter

5.2. Analogue / Digital Converter

5.2.1. Circuit Description

A reference voltage between 0 and 2. V is generated in the analogue / digital integrated circuit from a pre-set 8-bit data word with a ZN426E (Fig.24).

The comparator compares this with the voltage to be measured. It gives the decision as to whether the reading is higher or lower than the reference voltage at the PE connection, in the form of a high or low level signal to the PC.

The method used to determine the measured voltage is that of successive approximation. In this way, the scoring logic in the program approaches the measurement result through a series of jumps.

So that usable measuring results can also be obtained in the lower, flat section of the detector diode characteristic curve, an operational amplifier is connected in series to the converter. This can be adjusted in stages, with regard to its amplification (x 0.1 / x 1 / x 10 / x 100). Depending on the

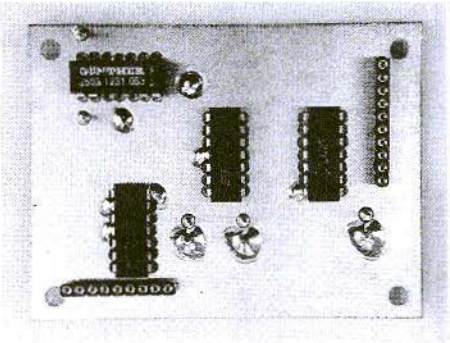


Fig.26: Components Side of Prototype DJ8ES 033

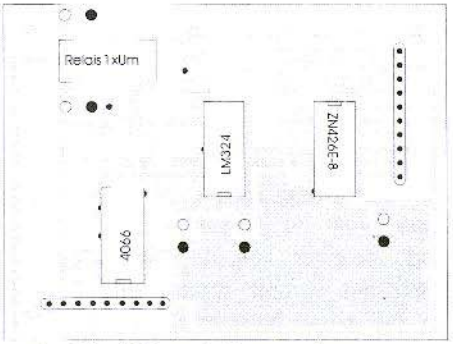


Fig.27: IC's and Connection Elements are mounted from the Earth Side; *Relais* = Relay

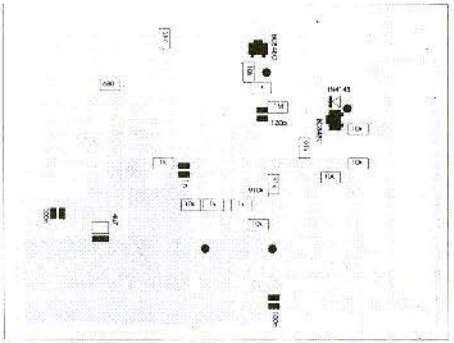


Fig.29: Tracks Side has only SMD Components

switching condition of the CD 4066, the input signal is thus always in the optimum range.

5.2.2. Assembly Instructions

The analogue / digital converter circuit is assembled on an epoxy board, copper-coated on both sides, with the dimensions 60 mm x 80 mm (Fig.25). IC's, relays and all the connection jacks and pins are mounted on the components side (Fig's.26 and 27). All earth connections are soldered onto this fully laminated side as well. The SMD components are soldered directly onto the individual tracks in accordance with Fig's.28 and 29.

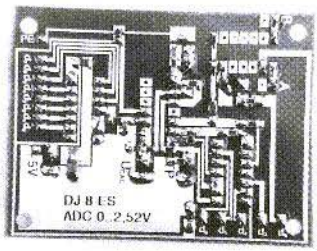


Fig.28: View of SMD Components on Tracks Side

5.2.3. Component List

1 x	LM 324 operational amplifier
1 x	ZN 426E-8 digital / analogue converter
1 x	4066 analogue switch
2 x	BC848 C, SMD transistors
1 x	LL4148, Mini Melf diode
1 x	Reed relay (1 x Um)
8 x	1.3mm terminal pins
2 x	10-pin socket terminal strips
1 x	4.7µF SMD electrolytic
1 x	DJ8ES 033 PCB

SMD resistors and capacitors are 0805 or 1206 format:

1 x	680Ω
4 x	1kΩ
1 x	2.7kΩ
5 x	10kΩ
2 x	91kΩ
1 x	910kΩ
1 x	1MΩ
1 x	120pF
1 x	1nF
1 x	10nF

5.2.4. Putting Into Operation

The data lines d0 to d7 for the A / D conversion are to be connected through a ribbon cable with the corresponding pins of the Selln-Latch on the interface card.

The PE pin is to be used to report back to the PC on the A / D conversion.

Lines d0 to d3 on the FormFeed-Latch are reserved for setting the amplification. Bit d7 of this port switches the analogue input over (A or B).

The diode detector is connected to input B.

The circuit should now be ready for operation, so its functioning can be tested using HFMESS software.

With a voltage of +5V, the current consumption of the analogue / digital converter assembly amounts to about 15 mA.

The final step is to measure the reference voltage of the A / D converter. The value should be 2.52V at pins 5 / 6, to the earth of the ZN426E. Any discrepancies have an influence on the correction factor, AU (cf. Section 10.13.).

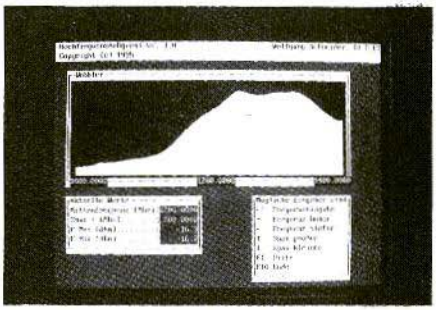


Fig.30: A Filter for ATV Applications at 23 cm as Test Object

6. WOBBLER

No additional circuit need be added for the functioning of the wobbler. The combination of frequency synthesiser, diode detector and A / D converter allows for wobble measurements over the full range between 10 and 14000 MHz.

The dynamic range of the wobbler thus results, on the one hand, from the output power of the synthesiser, and on the other hand, naturally, from the limiting sensitivity of the detector; in the given circuit, therefore, just 40dB.

The test object is merely looped in between the synthesiser and the diode detector for the measurement. The software does everything else.

Fig.30 shows the result of a wobble measurement in the range from 1000 to 1400 MHz. For this purpose, the entire range is traversed step by step, and for each frequency step the current level at the output of the test object is determined and plotted.

7. SPECTRUM ANALYSER

The only new assembly which comes up under the heading "Spectrum analyser" is a mixer with a logarithmic detector amplifier.

Only in combination with the frequency synthesiser does the complete spectrum

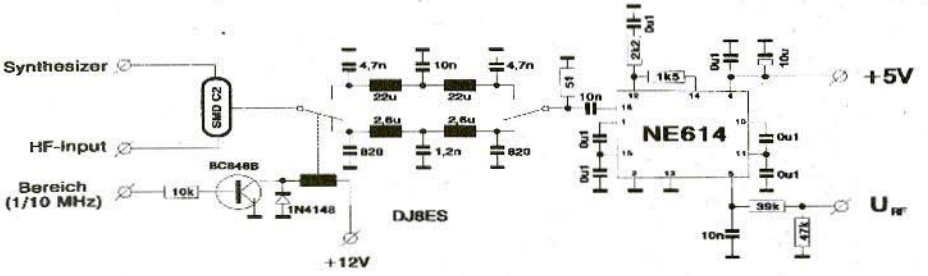


Fig.31: Wiring Diagram of Mixer with Logarithmic Display Amplifier; *Bereich* = Range

analyser emerge, for the frequency range from 10 to 1400 MHz.

7.1. Circuit Description

The mixer assembly with logarithmic detector amplifier operates on the direct mixing principle. The frequency range to be represented is traversed step by step and mixed down (Fig.31).

The two deep pass filters make two band widths possible: DC to 0.5 MHz, or broad-band DC to 5 MHz.

The logarithmic detector amplifier is created using an IC from FM technology (NE 614). Depending on the input level, the voltage meter output supplies a voltage between 1V and 4V; the dynamic range amounts to just 70dB.

The PC takes over this level information using the ADC assembly.

The voltage divider (39k / 47k) is provided at the output of the IC, to match the voltage range to the measuring range of the converter (0...2.5V).

7.2. Assembly Instructions

The mixer assembly with the logarithmic detector amplifier is also assembled on an epoxy board, copper-coated on both sides, with the dimensions 54 mm

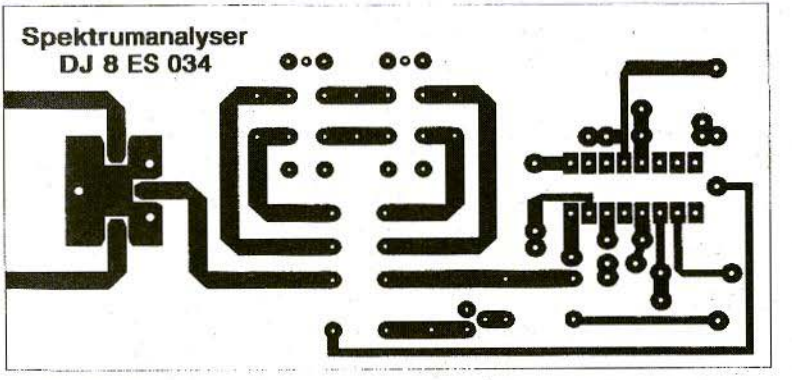


Fig.32: DJ8ES 034 Printed Circuit Board; Direct Mixer with Logarithmic Level Display; *Spektrumanalyser* = Spectrum Analyser

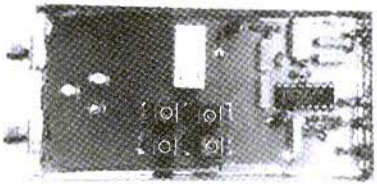


Fig.33: View of Prototype Mixer

x 108 mm (Fig.32). The circuit thus fits in a standard tinplate housing (55.5 mm x 111 mm x 30 mm) (Fig.33)

The coupling capacitors and the mixer are soldered on the foil side, all other components on the components side (Fig.34).

The earth feedthroughs are manufactured using silver-plated copper wire, with a diameter of 1.5 mm.

Operating voltages, the switch voltage for the filter conversion and the log. voltage output are connected by means of 1nF feedthrough capacitors. The radio-frequency connections are carried out using SMA connectors.

7.3. Component List

1 x NE 614
1 x BC549C transistor
1 x 1N4148 diode
1 x SMD C-2 mixer
1 x relay (2 x Um)
2 x BV5138 (2.6µH) Neosid ready-made filter
1 x BV5313 (22µH) Neosid ready made filter
4 x 1nF DF, solderable
6 x 0.1µF, RM 10, MKH
1 x 10µF / 35V tantalum
1 x tinplate housing, 55 x 111 x 30 mm
1 x DJ8ES 034 PCB

1/8 W resistors, 10 mm grid:

1 x 51Ω
1 x 1.5kΩ
1 x 2.2kΩ
1 x 10kΩ
1 x 39kΩ
1 x 47kΩ

Ceramic capacitors, 2.5 mm grid:

2 x 820pF
1 x 1.2nF
2 x 4.7nF
3 x 10nF

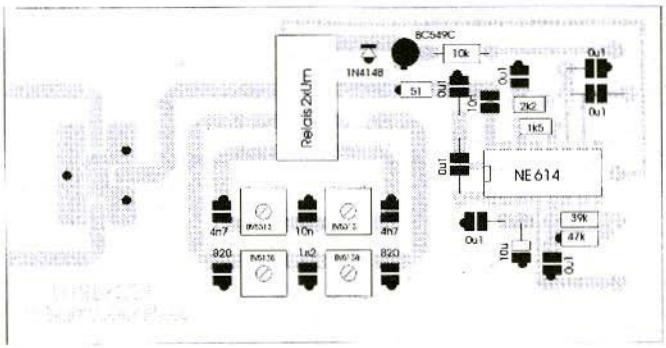


Fig.34: Components Plan; SMD Mixer and Coupling C's are mounted on Tracks Side;
***Relais* = Relay**

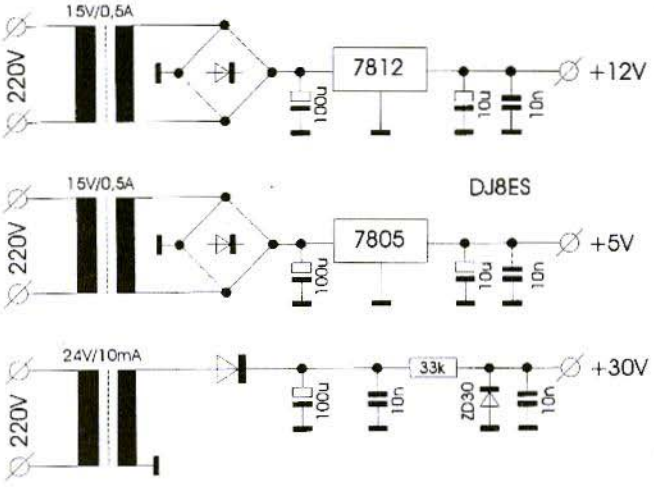


Fig.35: Example of Circuit for Power Supply

7.4. Putting Into Operation

The assembly requires two operating voltages: +5V / 3mA and, for the band width selection, +2V / 18mA.

The measurement output, U_{RF}, is connected to input A of the analogue / digital converter. The filter switching, 1 to 10 MHz, is determined by pin 4d on the FormFeed-Latch of the interface card.

The oscillator input of the mixer (LO) must not be saturated. The LO output can be held constant at +7dBm only from software version 3.0 onwards. Earlier versions require a 10dB attenuator between the synthesiser and the LO input.

Since the compression on the SMD C-2 mixer sets in as early as app. 100µW (-10dBm), a 10dB attenuator should also be connected in series to the radio-frequency input. The maximum usable input level thus rises to 1mW (0dBm).

The value MaxVoltSA for 0dBm for the software is set at 2.0V. The default value for MinVolt SA (-70dBm) is 0.4V.

Entries are to be matched in the file HFMESS.CFG if necessary (see Section 10.12.).

8. POWER SUPPLY

Three operating voltages are needed for the power supply (Fig.35):

±5V for digital section, PLL and A / D or D / A converter

±12V for switching relays, OP's, amplifier and transistor stages

±30V as tuning voltage for VCO

In the specimen unit, this power supply is assembled on a standard breadboard. Because the total current consumption is low, no heat sinks are required for the 78xx family of regulators.

9. INTER-CONNECTING INDIVIDUAL ASSEMBLIES

The individual assemblies are to be combined with one another depending on the application (mW meter, synthesiser, wobbler or spectrum analyser). E.g., simple switching relays are used here, as well as a coax relay, for the radio-frequency branch - users can decide the details for themselves.

For the prototype assembly, N-connectors were used for the external radio-frequency connections, and SMA connectors for the internal ones.

Other connections inside the equipment are carried out using single-row pin-and-socket connectors. Connections to the exterior are made using feedthrough capacitors (1nF).

10. SOFTWARE

It is frequently necessary to spend a lot on electronic components for operation and display in connection with measurement and test procedures. The mathematical preparation of the signals can also be very complicated.

The expense increases still further if interfaces also have to be provided with other equipment such as, for example, a printer.

Ideally, the computer can take over a whole series of tasks from a measuring instrument. It can be operated through the keyboard, and readings can be displayed numerically or even graphically on its screen.

Other equipment can be influenced through its interfaces - printers can be controlled, and in addition readings can be logged and put on paper.

Its capabilities for converting readings mathematically leave scarcely anything to be desired.

The various assemblies essentially use components which are highly compatible with a PC. Here, for example, there are I^2C bus circuits and µP-compatible AD / DA converters. The HFMESS software was developed so that this technology could be used as a whole.

The program controls the assemblies through the printer interface, which can be freely selected (LPT1, LPT2 or LPT3).

10.1. Operating Software

The program diskette contains the following files:

HFMESS.EXE is the main program itself. It can run on MS-DOS and IBM-DOS.

HFMESS.HLP contains help texts to guide the operator. They can be called up at almost all points during the running of the program, using the F1 function key.

HFMESS.CFG is used to pre-set necessary parameters (e.g. interface addresses). The data can be entered or changed using a standard text editor, or with EDIT.COM, supplied with MS-

DOS. The contents are individually listed in Section 10.12..

HFMESS.MNU is the list of the menu items available.

CALIBRI.BIN contains level settings from -10dB to +10dB, over the entire frequency range. This file is generated through the "Calibrate" routine (see main menu).

FEHLER.TXT shows the measured values per frequency in MHz for level discrepancies. This file is generated through the "Calibrate" routine (see main menu).

MWTABLE.TXT contains all the tables for the output calculation and calibration of the diode detector. This file is discussed again, in greater detail, in Section 10.13..

READ.ME contains current information on the software, and should be read before installation.

10.2. Hardware Requirements

The program requires a PC with a 286 processor, a 640-kbyte main memory and a VGA graphics card. However, a 386, or better still a 486, is preferable for the wobbler and spectrum analyser.

If a hard disc is available as well, if the computer is faster, and if more main memory is available, plus RAMDRIVE and a colour monitor, then naturally everything is even more splendid. Above all, everything can be done much faster!

10.3. Installation

To be able to use the **HFMESS** program, it is more advisable to copy it into a sub-directory of its own on the hard disc.

MD C:\HFMESS

COPY A:*.*C:\HFMESS

10.4. Program Start

In order to start the program, go into the directory on the hard disc and call the program up by its name, **HFMESS**, or start it through a batch file.

After a short loading time, the program briefly introduces itself, and then displays a menu with its functions:

Synthesiser
Wobbler
Milliwatt meter
Spectrum analyser

Calibration
Error list

End

The red bar above the individual menu items can be moved up and down using the cursor key. Press the "Enter" key to go to the selected function.

10.5. Synthesiser

The sub-program ***Synthesiser*** displays two windows on the operator screen. The top window displays the name "Synthesiser", the current frequency and the output level.

The frequency can be varied between 10 MHz and 1400 MHz. The output level lies in the range -10dBm to +10dBm.

The message "UNCAL" may be displayed in the output level field, due to tolerances in the hardware (PLL oscillator, mixer and level regulator). The inter-relationships are discussed in Section 10.9., "Calibration".

The bottom window on the screen indicates the significance of individual keys and their functions.

The frequency can be changed, step by step, by moving the cursor key left or right, and the level can be changed, step by step, by moving the cursor key up or down.

Use the "Return" ("Enter") key to go to an entry window, into which the desired frequency can be entered directly via the keyboard.

This entry procedure is terminated using "Enter".

Use F1 to view the "Help text", which depends on the context. Press "Enter" or "Esc" to leave this text.

Use F10 to leave the "Synthesiser" function and go back to the main menu.

10.6. Wobbler

When the ***Wobbler*** function is selected, the screen is set up in graphics mode. This preparation is directly dependent on the processor output and the VGA card speed.

A large, wide window in the top half of the screen shows the wobble curve, while the bottom left-hand window shows the current values for the mean frequency and the span in MHz. The bottom right-hand window, on the other hand, shows the control options through the keyboard.

The mean frequency can be changed step by step by moving the cursor key left or right (half span) and the span (wobble width) can be changed step by step by moving the cursor key up or down, from ±10 MHz to ±500 MHz.

Use the "Enter" key to go into an entry window where the mean frequency can be set through direct entry. This entry procedure is to be terminated using "Enter".

The program monitors the technical options and prevents the frequency limits from being exceeded.

Use F10 to leave the "Wobbler" function and return to the main menu.

10.7. The Milliwatt Meter

Three windows are displayed in the ***Milliwatt meter*** function. At the top right, you can see the measured output in milliwatts in particularly large characters, which can still be seen easily, even from some distance away. Below this is a horizontal yellow bar, which changes in line with the output for tuning purposes.

At the bottom left, the current readings are shown, such as milliwatts, output in dBm, voltage in Volts, and naturally, the frequency range set.

The bottom right window shows the operating options.

The frequency band can be selected by moving the cursor key up or down. Here

the program takes account of frequency-dependent non-linearities of the diode detector and converts the readings accordingly.

Use F1 to call up the context-dependent help text, and "Enter" or "Esc" to leave it again.

Use F10 to leave the "Milliwatt meter" function and return to the main menu.

10.8. The Spectrum Analyser

When the ***Spectrum Analyser*** function is selected, the screen is set up in graphics mode. This preparation is directly dependent on the processor output and the speed of the VGA card.

A large, wide window in the top half of the screen shows the measured frequency spectrum. The bottom left window shows the current values for the mean frequency and span in MHz.

The bottom right window, on the other hand, shows the control options through the keyboard.

Moving the cursor key to the left or right changes the mean frequency step by step (half span) and moving the cursor key up or down changes the span step by step from ±10 MHz to ±500 MHz (wobble width).

Use the "Enter" key to enter an entry window, where you can enter the mean frequency directly. Press "Enter" to terminate this entry procedure.

You can now use F2 to select the intermediate-frequency band width - band widths of 500kHz and 5MHz are possible.

The program monitors the technical options and prevents the frequency limits from being exceeded.

Use F10 to leave the "Spectrum Analyser" function and return to the main menu.

10.9. Calibration

The ***Calibration*** function runs through the entire available frequency spectrum, step by step. The maximum output power of the frequency synthesiser is measured every 1 MHz, and the level settings in the range from +10dBm to -10dBm are determined as well.

The test reading is graphically represented in the top half of the screen.

Carrying out a successful calibration is a pre-requisite for the functioning of the "Synthesiser" and "Wobbler" applications.

Possible tuning problems - e.g. maximum output level smaller than +10 dBm - are outputted in the form of an error list (cf. Section 10.10. "Error list"). When the calibration is complete, this table is displayed for the user in a window on the lower left-hand side.

The "Calibration" program item is time-intensive! In unfavourable circumstances, e.g. with slow computers or if the parameters have not been set optimally, the complete running time may amount to 10 - 15 minutes.

While the reading is being determined, the function can be interrupted at any time, just by pressing a key. In that event, the calibrated values are not updated (file CALIBRI.BIN) on the hard disc.

10.10. Error list

When the ***Error List*** menu item is called up, it is set up for graphical display. Implausible measurement values are shown in table form as frequencies, theoretical levels and actual values.

If this table takes up more than one screen page, the display can be changed through the cursor keys.

Use F1 to view the context-dependent help text and "Enter" or "Esc" to leave it again.

"Esc" terminates the error list display and returns you to the main menu.

10.11. End

With this selection, the **HFMESS** program is completed, and the screen returns to what it was before the call-up procedure.

After a brief farewell, you return to the DOS operating system level.

10.12. The Configuration File

In the ***HFMESS.CFG*** file, various parameters, essentially conditioned by the hardware, are added to the ***HFMESS.EXE*** program.

The individual values are briefly outlined below:

READINGS = MWTABLE
Conversion table for diode detector

AD-DELAY = 200
Time delay A/D converter (1..1000)

IQ-DELAY = 20
Transmission pulse time delay (1..1000)

PRNPORT = LPT1
Centronics interface (LPT1..3)

FREQUENCY = 1200 Mean frequency target (0..1400)

MENU = HFMESS
Description of option screen

SPAN = 2
Frequency span (1, 2...6)

WOBBLE STEP = 4
Step width for wobble measurement (1..16)

MAXVOLTSA = 2.0
Max. output spectrum of spectrum analyser

MINVOLTSA = 0.4
Min. output voltage (noise)

MAXOUTSYN = 17
Max. output level (dBm) of synthesiser

PRINTERDRIVER = ...
Hard copy function (carried out later)

PRINTERCHECK = ON
Printer test (carried out later)

TEMPDRIVE = C:
Description of RAM disc, if applicable

10.13. The Readings Table

MWTABLE.TXT

The electrical behaviour of the diode detector (Fig.20) is summarised in the MWTABLE.TXT table, including necessary correction factors for frequency and voltage. Any standard text editor allows the correction of this table and thus the capture of the current values for the detector in use.

– *Diode detector characteristic line*

For the diode detector, the output voltage should be measured in relation to the input power. This can be done in a satisfactory way with a digital voltmeter and a test transmitter for, e.g., 144 MHz. A transceiver with adjustable transmission power can be used as well, of course.

The program is designed for a precision (minimum step width) amounting to 1 dB. The voltage values measured are to be entered in the table MWTABLE.TXT in data fields MESSWERTTABELLE [-, 3].

– *Frequency correction*

At high frequencies in particular, the input impedance diverges, in reality, from the desired 50Ω. A correction factor in the data fields FREQQUENZTABELLE [- , 2] balances this out through a computer.

The factor should be measured for the frequency range 28, 50, 144, 432, 1296 and 2320 MHz, with a control power of - for example -1mW (0dBm).

Frequency correction =
Theoretical value at 144 MHz / Reading

– *Correction factor, ΔU*

Small tolerances for the components, or even for the D / A converter ZN 426E-8, result in divergences in the measuring voltage. The ëU factor acts as a computer correction. Its value is app. 0.0099. The reference voltage can be precisely measured at pin 5 / 6 of the D / A converter, using a digital gauge.

$$\Delta U = U_{Ref} / 256$$

10.14. HFMESS.MNU

The HFMESS.MNU file contains the individual items of the main menu, with a description and a screen number (e.g. synthesiser; 110).

The description can be changed, but not the screen number allocated. This datum is coded in the program and is thus always pre-determined.

11. Conclusion

As already stated at the start, the test rig just described represents a welcome and cost-effective gauging aid.

Naturally, with this relatively low expenditure, it is not possible to create a high-end apparatus, and so practical work with the radio-frequency rig demonstrates the limits of this technology sooner or later.

Certainly improvements could be made here, as with any circuit. It is important here that individual hardware changes do not also make a change in the software imperative.

Circuit modifications by users can possibly be published in this magazine.

12. Literature

[1] Carsten Vieland, DJ 4 GC; Diode Detectors in 50W Broad-Band Technology
VHF Communications 4/87, pp. 223 - 237

[2] Günter Sattler, DJ 4 LB; SDA 3202 - New PLL-IS up to 1.5 GHz
VHF Communications 4/85, pp. 205 - 209

[3] Günter Sattler, DJ 4 LB; Crystal-Stable VCO Frequencies for ATV Set-Ups
VHF Congress Weinheim 1991; Congress Proceedings, pp. 173 - 184

[4] Wolf-Henning Rech, DF 9 IC; Simple Gauging and Smoothing Aids for the UHF Range Part 5: A Simple Spectrum Analyser, 5..1500 MHz
VHF Congress Dorsten 1990; Congress Proceedings, Pp. 42 - 70

[5] Designing with the NE602/NE604; Robert J.Zavrel Jr.
Valvo Application Note, February 1987

[6] The New Professional's Book; Martin Althaus
Sybex Verlag GmbH

[7] Borland Pascal with Objects Programming and Reference Manual
Borland GmbH

Gunthard Kraus, DG8GB

Design and Realisation of Microwave Circuits Part 4

7. MMIC AMPLIFIER

7.1. Preliminary Observations

At two points in our "Meteosat Converter" application project, we need at least 20dB amplification with simultaneous tolerable noise background:

a. Directly behind the "low-noise preamplifier" referred to in Section 6.2. of Part-, for the significant increasing of the signal level before the subsequent 1,700 MHz band pass, with simultaneous damping of the passive mixer.

b. Directly behind the mixer, or in the downstream intermediate-frequency stages to have a signal level sufficient for a relatively long coaxial cable connection from the converter (near the antenna) to the evaluation circuit in the building.

The use of a "universal amplifier module" from 50 Ohm technology is on offer here; to this end, MMIC's (microwave monolithic integrated circuits) are available in the most varied formats, and with varying limiting frequencies.

After a look through the relevant literature with application circuits, the choice finally fell on type INA 03814 from Avantek/HP, because of the current price and delivery situation.

We find the following specification for this MMIC in the data sheet; 25dB amplification in the range up to 2 GHz, for noise factors of about 3dB.

The S-parameter can also be found, on the Avantek data diskette, which can be supplied for Puff.

Here you merely need to convert the "touchstone file" (INA03184.S2P) into Puff format (which gives INA03184.dev). As mentioned, the conversion program you need for this is supplied with the Puff diskette.

But don't forget to estimate the values for the "zero Hertz" frequency and add them to the *.dev file!

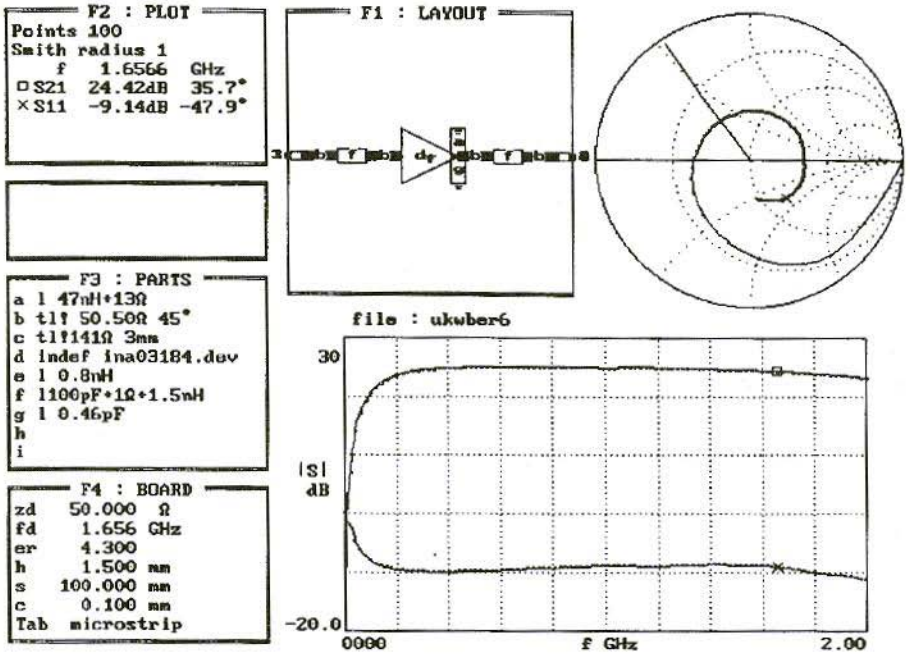


Fig.25: Ideal MMIC amplifier with INA 03185

7.2. Draft Circuit and Simulation

To judge from what many semi-conductor manufacturers say, drafting a microwave amplifier circuit is child's play, and is exceeded in simplicity and operational reliability only by the drafting of a power supply circuit with a half-wave rectifier.

The input and output pins are simply connected using 50 Ohm microstrips, into which are inserted the DC blocking capacitors. The power supply DC voltage is fed in through a 47nH SMD choke.

If, during the simulation, the same equivalent circuits are used for chokes and capacitors as for the low-noise preamplifier, you obtain a wonderfully smooth amplification curve between 1 and 2 GHz, with a peak value of app. 27dB. Fig.25 shows the MMIC amplifier simulation.

However, in reality the assembled circuit will initially oscillate at f = 1.4 GHz. Moreover, it is almost impossible to produce any attenuation by means of the well-known "calming measures" - e.g. using conducting foamed material, etc..

If we now make use of the experience gained in the drafting of the LNA and test the influence of the "earth feed-throughs", using Puff, we discover something really amazing:

Even a total inductance of 0.3nH, applied for the feed-through of the two earth surfaces of the IC to the bottom earth plane, is sufficient to make the

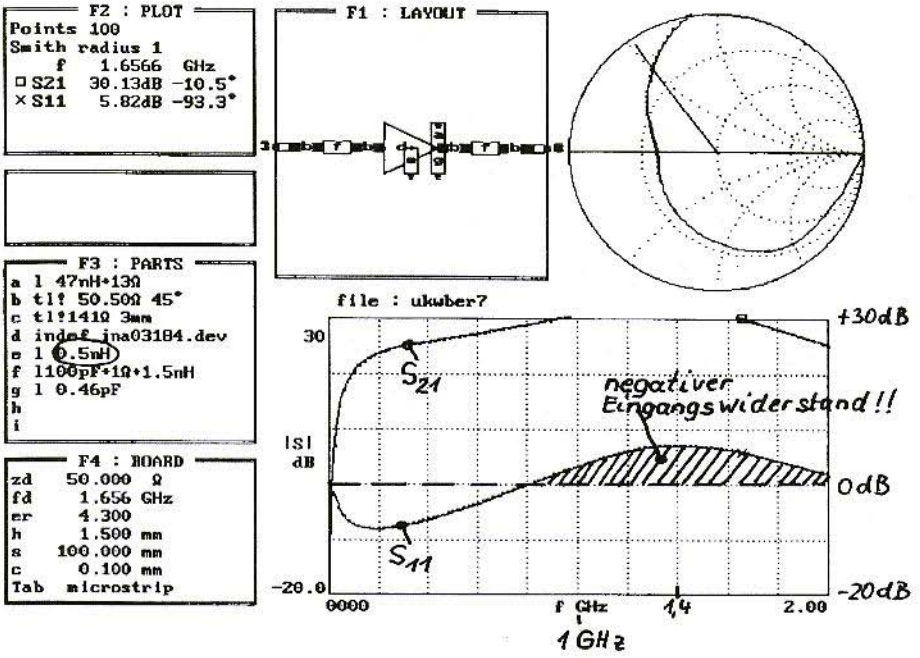


Fig.26: Simulation with Puff: MMIC amplifier with L = 0.5nH in earth branch

input reflection factor in the frequency range at 1.4 GHz markedly greater than 0dB (Fig.26 shows the ratios for 0.5nH).

The circuit then has a negative input resistance. In these circumstances, it can't help but oscillate!

We must therefore use all means to ensure that this inductance falls; Fig.27 shows how this is done.

a. A hole with the external diameter of the MMIC housing is drilled in the printed circuit board.

b. The top earth surfaces are now fed as a wedge into the layout until they are directly against the housing of the MMIC.

c. The feed-throughs are started as close as possible to the housing. 5 holes are drilled on each side.

It can probably be easily recognised from the diagram that the question of the insulated earth surfaces for input, earthing and output has again been taken to heart (it's just that putting in the many 0.8 mm. rivets, with a "watchmaker's power riveting machine", specially created for this, as an extra, that really is a brute of a job!)

The output and input microstrips require special consideration at 50 Ohms. They should have a "wedge-shaped point" , and then be reduced, right down to the width of the IC terminal lug, thus avoiding additional inductances and points of impact.

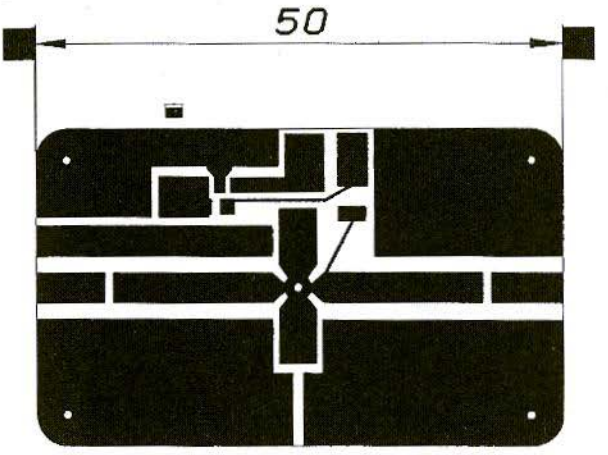


Fig.27:
Layout of MMIC amplifier for 1.7 GHz

The already well-known 47nH choke for feeding in the operating voltage is here too as well, together with its 6mm. long feed and its joint built-in capacitance of 0.46pF and the 13 Ohm losses in the equivalent circuit diagram.

The coupling capacitors in the input and output circuits are laid out, as usual, as two 0805-NP0 types in parallel, at 100pF. Firstly, this precisely fills out the width of the microstrips and thus reduces the reflections, and secondly, the

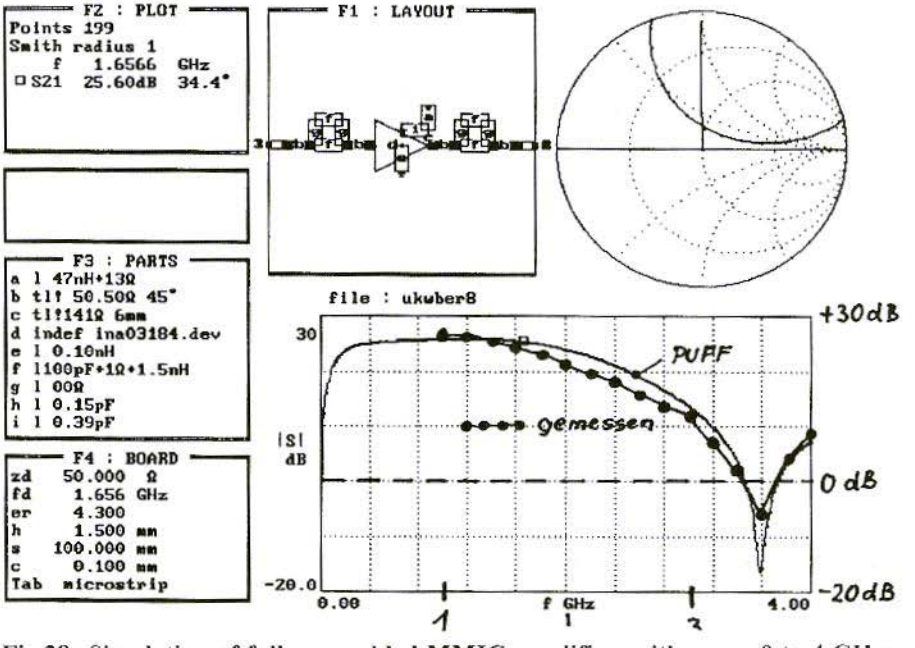


Fig.28: Simulation of fully assembled MMIC amplifier, with range 0 to 4 GHz; *Gemessen* = Measured

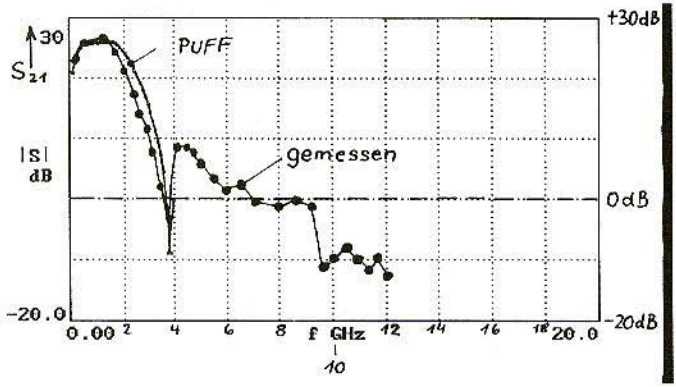


Fig.29: MMIC amplifier: gain slope |S21| from 0 to 12 GHz *Gemessen* = Measured

joint inductance of 1.5nH and the serial loss resistance of 1 Ohm are halved by using one component.

Fig.28 displays the full results of the simulation, including the measured results for up to 4 GHz. Of course, immediately after the measurement the simulation was adjusted to the reality.

The onset of amplification at 3.6 GHz is actually brought about by the 6mm long microstrip and the 47nH choke, with the (assumed) 0.46pF capacitor! Here a short-circuit is obviously transformed to the amplifier output, and we must reduce the 0.46pF to 0.39pF for the Puff simulation to supply the same resonance frequency. There is a quite simple reason why the minimum value in the actual assembled circuit is not as low as in the prediction. It is well-known that the losses for FR4 equipment increase more and more sharply from 1.5 to 2 GHz, and here we are always stimulating with a constant loss factor of 0.015. The assumed quality is therefore too high, and there is a discrepancy between reality and the Puff result.

Anyone who would like to have a more precise result would first have to measure the valid loss factor for this frequency, and then simulate for only a narrowly restricted frequency range for which this determined factor is approximately constant. For the same reason, the measured amplification in the range between 1.5 and 3.4 GHz is probably always somewhat lower than the theoretical progression, since - because of the poorer quality - as in every resonant circuit, the resonance curve of this "resonant circuit" must be flatter and broader.

One more little thing:

The S parameters of the INA 03184 are specified by the manufacturer only up to a maximum frequency of 4 GHz. But if, for example, an upper plot limit of 10 GHz is inadvertently entered, a bleeping noise is suddenly heard when the 4 GHz limit is exceeded, which in normal circumstances would be the acoustic signal for a computer fault. But the program has not crashed. It continues operating and bleeping, right up to the highest frequency entered, and you need only wait for the computer to quieten down.

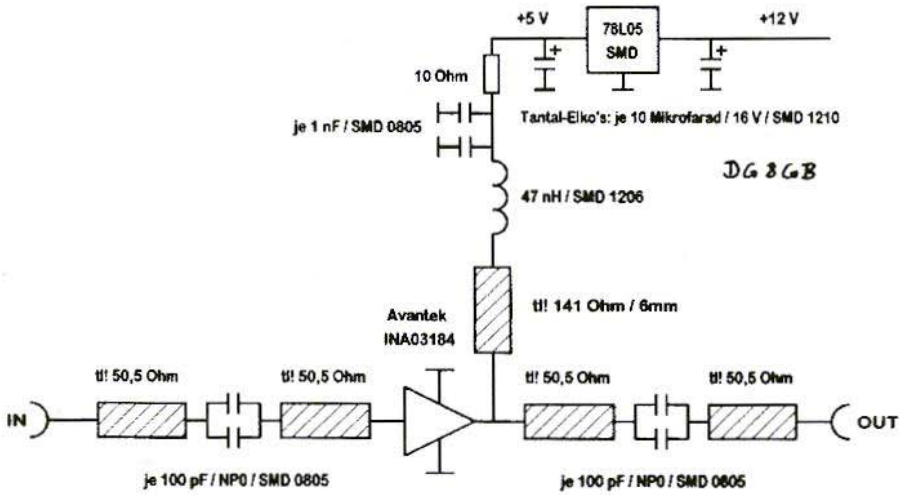
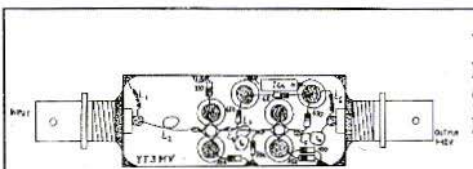


Fig.30: Wiring diagram of MMIC amplifier prepared for Puff;
***Tantal-Elko's* = Tantalum Electrolytic Capacitors,**
***Mikrofarad* = Microfarads, *Je* = Every**

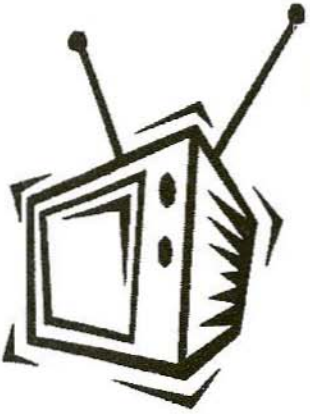
The diagram created using a network analyser (HP 8410A) shows the simulation behaviour of the circuit up to 12 GHz.

Finally, Fig.30 shows the wiring diagram for the layout, which can certainly be analysed without problems.

(To be continued)



Very low noise aerial amplifier for the L-band as per the YT3MV article on page 90 of VHF Communications 2/92. Kit complete with housing Art No. 6358 £36.55. Orders to KM Publications at the address shown on the inside cover, or to UKW-Berichte direct. Price includes p&p



Interested in the Technical Side of Television?

B.A.T.C.

Over four decades of Television Service

CQ-TV Magazine is for you, packed full of practical and easy to build circuits for the Radio Amateur and Video Enthusiast.

For details, contact the BATC at Grenehurst, Pinewood Road, High Wycombe, Bucks, HP12 4DD, England

Internet http://www.batc.org.uk

Lorenz Oelschlegel, DL6NCl

DB1NV's Image Memory in Combination with the HP141 Spectrum Analyser

The DB1NV-010 image memory assembly represents an ideal extension to the HP141 spectrum analyser. The only problem to be solved is the level matching.

The idea was to combine the HP141 from Hewlett-Packard, the use of which has recently become very widespread, even in amateur circles, with the image memory from DB1NV [1]. If possible, this should cause no interference either with the analyser or with the image memory circuit.

The signals from the Spectrum Analyser are picked up from the outputs of the intermediate-frequency assembly 8552B.

A normal X-Y oscilloscope acts as a check monitor for the image memory readings here (e.g. Hameg 203).

A small circuit was also positioned in front of the image memory assembly for level matching for the X and Y signals.

Since there are only a few components, a layout was dispensed with, and the circuit was assembled, using discrete components, on an SMD demonstration board.

The Y signal from the vertical output of the 8552B intermediate-frequency assembly (IF section) is fed to a uA741 operational amplifier. Here it is inverted and amplified, and it is then fed on to the Y input of the image memory.

The X signal is picked up from the Scan in/out jack. It is matched in level through a 50k trimmer, and is fed to the X input of the image memory through another uA741.

Smoothing is carried out using the existing trimmer potentiometer of the image memory assembly or that of the matching circuit.

In combination with the PC-Plot [2] software from Bernd Kaa, DG4RBF, we thus obtain a useful measuring system for documenting the values measured with the spectrum analyser.

LITERATURE

[1] Dr.-Ing. J.Jirmann, DB1NV; A Digital Image Memory for the Spectrum Analyser
VHF Communications,. 3/91, pp. 130-146

[2] Bernd Kaa, DG4RBF; Expansion of Software for DB1NV's Digital Image Memory
VHF Communications 3/96, pp. 173-180

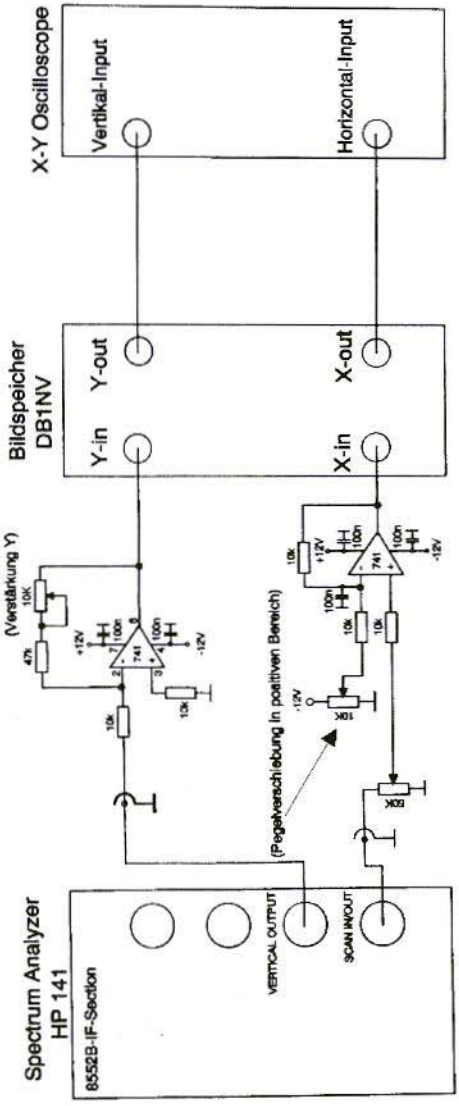


Fig.1: Level Matching: HP141/8552B - DB1NV's Image Memory

Denys Roussel, F6IWF

An Exceptionally Low-Cost HF-SSB/CW Transceiver Part 3: Assembly Instructions

A year and a half after the first prototype was put into operation (VHF communications 2 & 3/1996) a further transceiver has now been completed. The real work consisted of designing printed circuit boards, carrying out the mechanical assembly, and keeping an eye open for the possibilities for further expansion.

Some problems were recognised during the second assembly, and some necessary improvements were made which provided for a higher degree of reproducibility (Fig.1).

1. CHANGES TO THE BLOCK DIAGRAM

The single significant change is the additional stage at the power amplifier input. The reason for this is that the level at the mixer output was previously too low, and was insufficient to control the PA correctly and thus obtain a good inter-modulation interval, IM^3.

The MIC level can still be increased somewhat, which gives a sufficient level at the mixer output, which in turn improves the inter-modulation products. An increase of 6dB in the amplification keeps the IMD to an acceptable value.

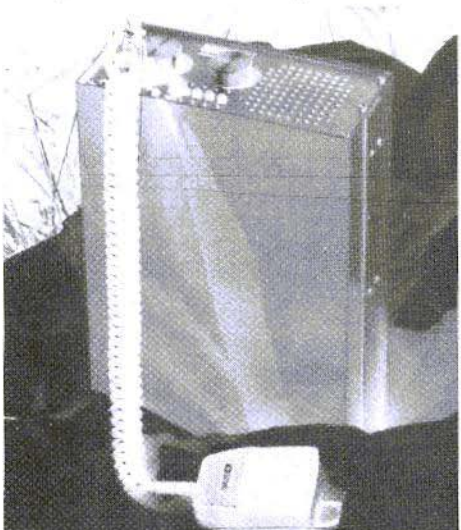


Fig.1: Second version of Short-Wave Transceiver ready to operate

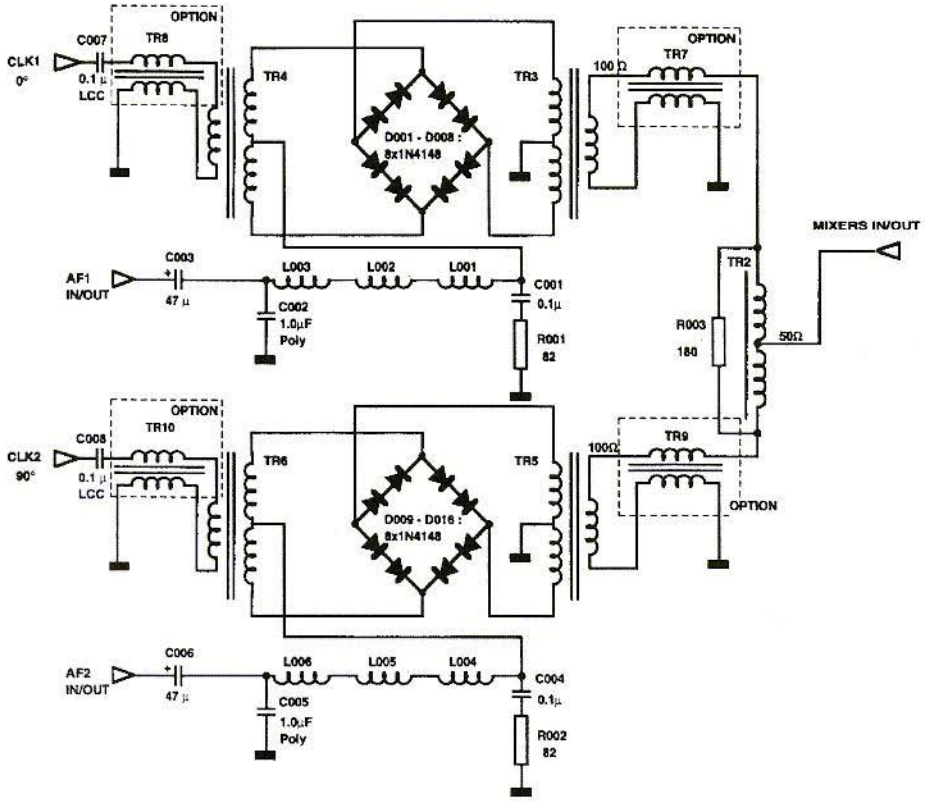


Fig.2: The Mixers, now with 100Ω matching

2. CIRCUIT CHANGES

The circuit changes are explained below in relation to the corresponding sections of Part-2 (issue 3/96). The corresponding circuit diagrams are shown with changes and component values.

2.1. The Mixer (Section 6.1.)

The mixers were previously matched to 50Ω. The mixers now operate at 100Ω which firstly reduces the stress on the high-frequency phase shifter and secondly saves a transformer (Fig.2).

Space was provided on the board for an optional balun transformer and a balance potentiometer (22Ω).

This allows better carrier suppression to be achieved in transmission operation -50dB is possible.

These changes are not really needed for the second assembly, as carrier suppression of 35dB, which has already been achieved by other improvements, is sufficient.

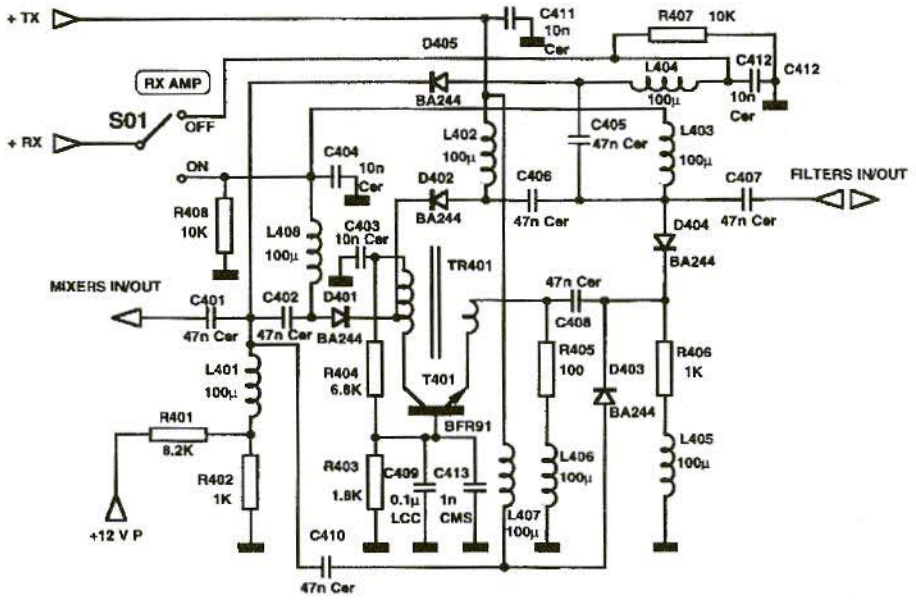


Fig.3: Circuit of High-Frequency Preamplifier with component values

2.2. The High-Frequency Pre-Amplifier (Section 6.2.)

Here the new assembly created instability. A tendency to oscillate arose in the 600 - 700 MHz range. The effect sounded like microphony if the C408 capacitor was touched.

The C413 SMD capacitor gives better de-coupling on the basis of T401 (Fig.3).

A ferrite bead can be provided for additional damping on the basic connection.

2.3. Low-Frequency Pre-Amplifier and Mixer-Driver (Section 6.3.)

There were no circuit changes here. The component values are shown in Fig.4.

2.4. VFO and High-Frequency Phase Shifters (Section 6.4.)

The VFO was completely re-designed, thanks to F5OYV. The circuit (Fig.5) became somewhat more universal, with an improved output level and additional options: RIT, tuning through Varicap, frequency drift monitoring, and an additional voltage regulator.

The cost of the additional stage was balanced out by the fact that a bipolar transistor was used in the oscillator instead of the J310.

It is not advisable to use Varicap tuning without a crystal-stabilised circuit for frequency drift compensation.

All components are housed on a small printed circuit board, only 36mm x 46mm.

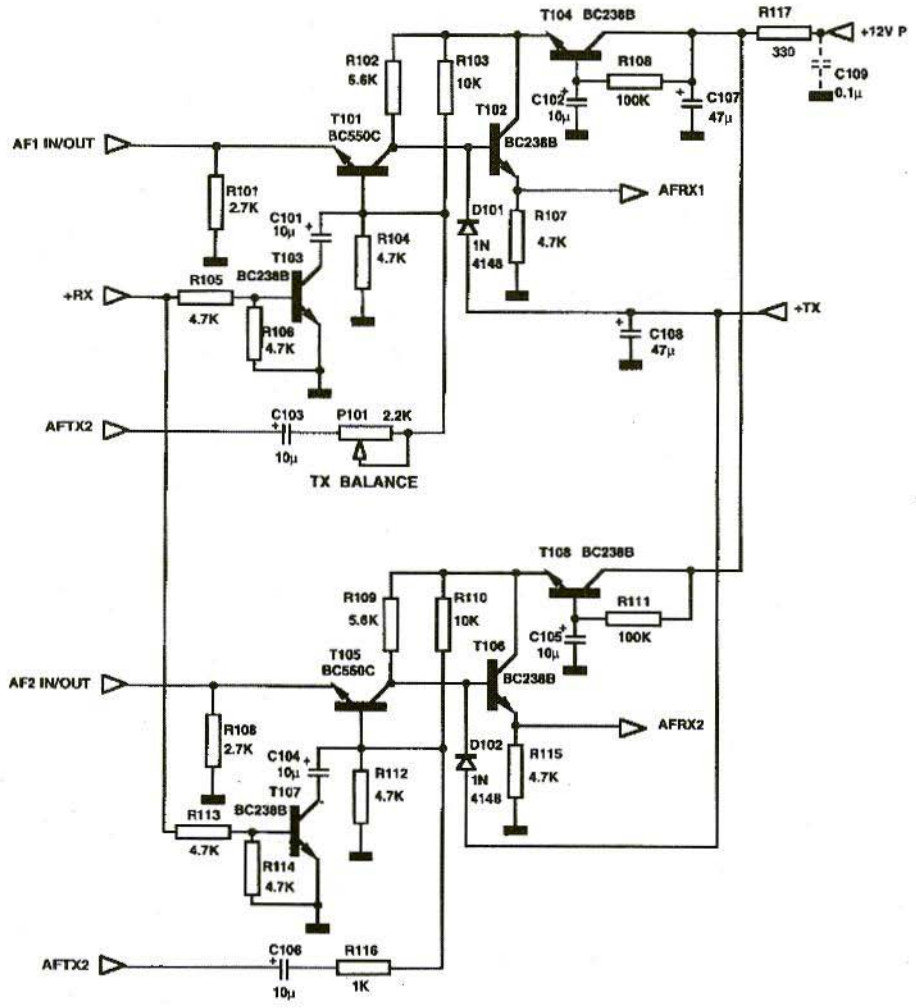


Fig.4: Circuit diagram of Low-Frequency Amplifier and Mixer Driver

The TTL-IC 7400 is replaced on the phase shifter by a type 2N2222 transistor. This increases the sensitivity and reduces the costs.

The outputs to the mixers were alternated between Q1 and Q2. These outputs are not fed back to their inputs in IC 901, which increases the power for the mixer.

In case of any changes in the sidebands, space was provided on the small printed circuit board so that the positions of the resistors R903 - R904 could be exchanged. The wiring shown corresponds to that for the lower sideband.

The high-frequency phase shifter is mounted on a plug, to provide for a possible expansion to a multi-band system.

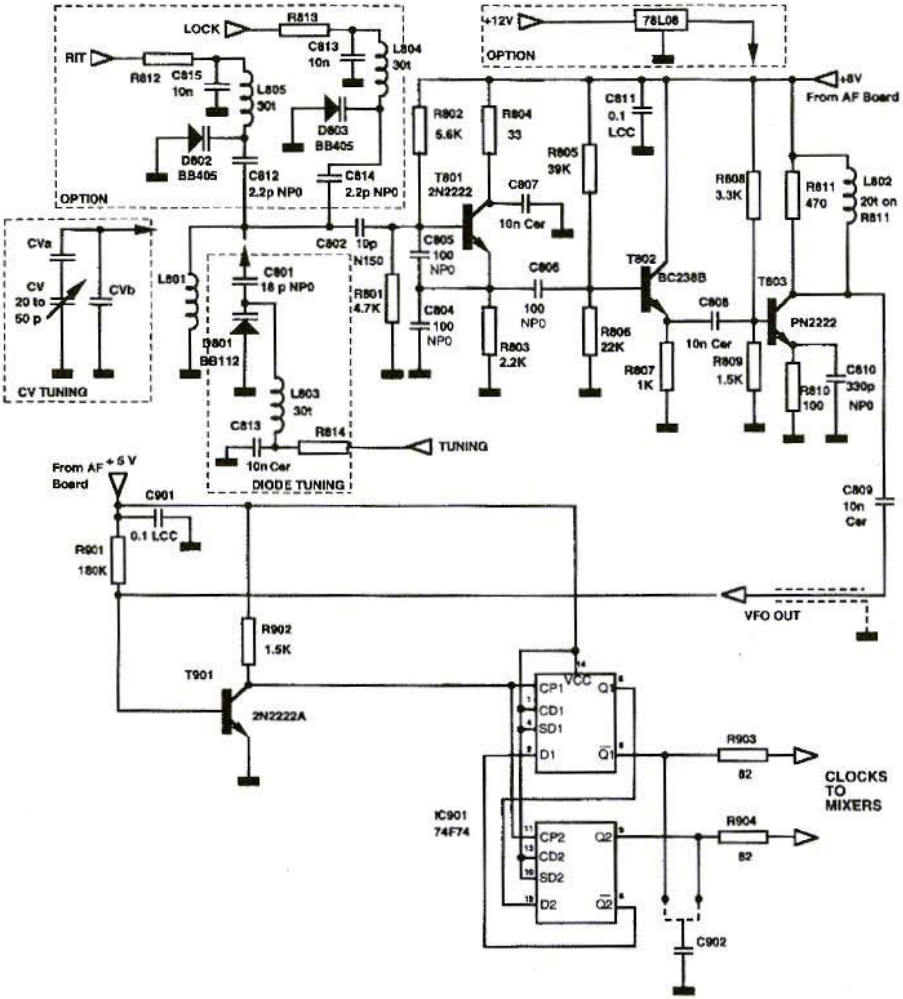


Fig.5: Circuit diagram of VFO and Radio-Frequency Phase Shifter

2.5. Low-Frequency Phase Shifter (Section 6.5.)

The circuit has been expanded by three potentiometers and a CMOS switch 4066, in preparation for a multi-band version (Fig.6).

These additional potentiometers are necessary, since the zero settings (image frequency suppression) are different for each band.

Since these functions are not required as yet, only P201 is fitted, and a bridge is placed between pins-8 and 9, instead of IC 205.

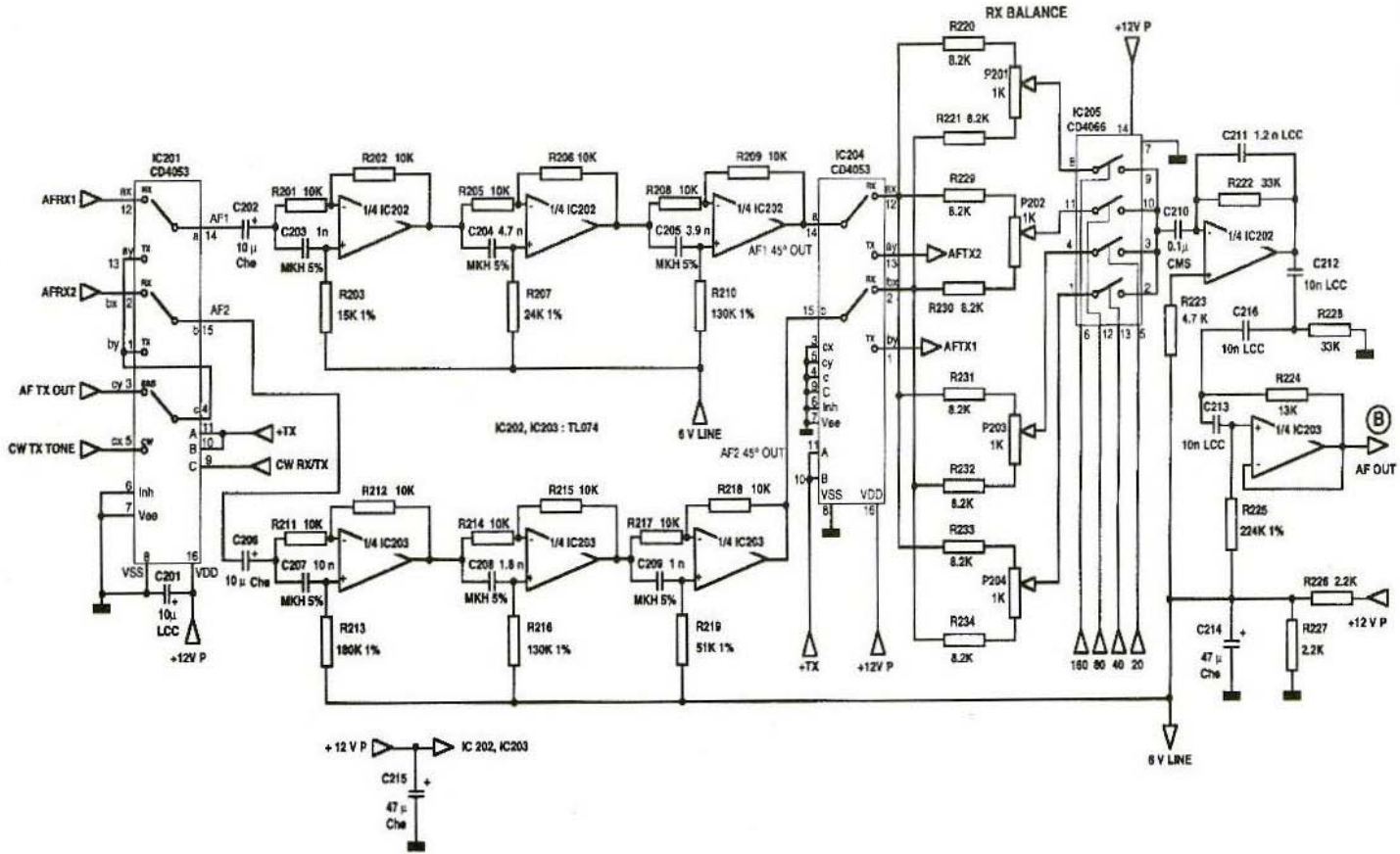


Fig.6: Low Frequency Phase Shifter with Switching

2.6. The Low-Frequency Section (Section 6.6.)

A voltage regulator IC 78L08 was used instead of the stabilisation circuit for +8.5 Volts (Fig.7).

The D307 diode creates a better insulation between the microphone and the 301C IC during receive.

In order to prevent problems in the low-frequency section during switching from transmit to receive, R312 and R338 were provided, which are there in order to retain the potentials of the capacitors, C305 and C330 respectively.

The CW filter has been replaced by a lower-noise circuit version.

A favourably priced 200µA type meter movement can now be used as an S-meter.

An 8.2Ω resistance has been inserted into the power supply circuit of the TBA820M and suppresses the low-frequency feedback to the 12Volt supplies..

2.7. CW Oscillator (Section 6.7.1.)

The oscillator is now fed +8V directly from the low-frequency printed circuit board (Fig.8).

2.8. Transmit / Receive Switching (Section 6.7.2.)

The transceiver now has reverse battery protection, by means of a P600 diode and a 10A fuse (Fig.8).

Switching from battery operation to an external power supply is now done manually through a switch.

PTT-Out has been provided as an additional output.

2.9. Band-Pass Filter (Section 6.8.1.)

The filter is now combined with a 6dB amplifier (Fig.9). Only five additional components are required for the assembly and switching of the stages.

D502 is switched from the transistor current. The emitter resistance has been inserted into the earth connection of TR502.

The transistor is a favourably priced BC238B in a plastic housing, adequate for the short-wave range.

The ferrite bead, as the basic connection for the damping, should not be forgotten.

2.10. Power Amplifier (Section 6.8.2.)

SMD components are now used for power supply switching and for the input stage instead of the wired components previously provided.

Moreover, D603/5 and D606/8 are available as protection against a B/E breakthrough at an extreme V_{cb} value (Fig.9).

Two additional transistors have also been provided in the PA, in order to increase the power by app. 20%.

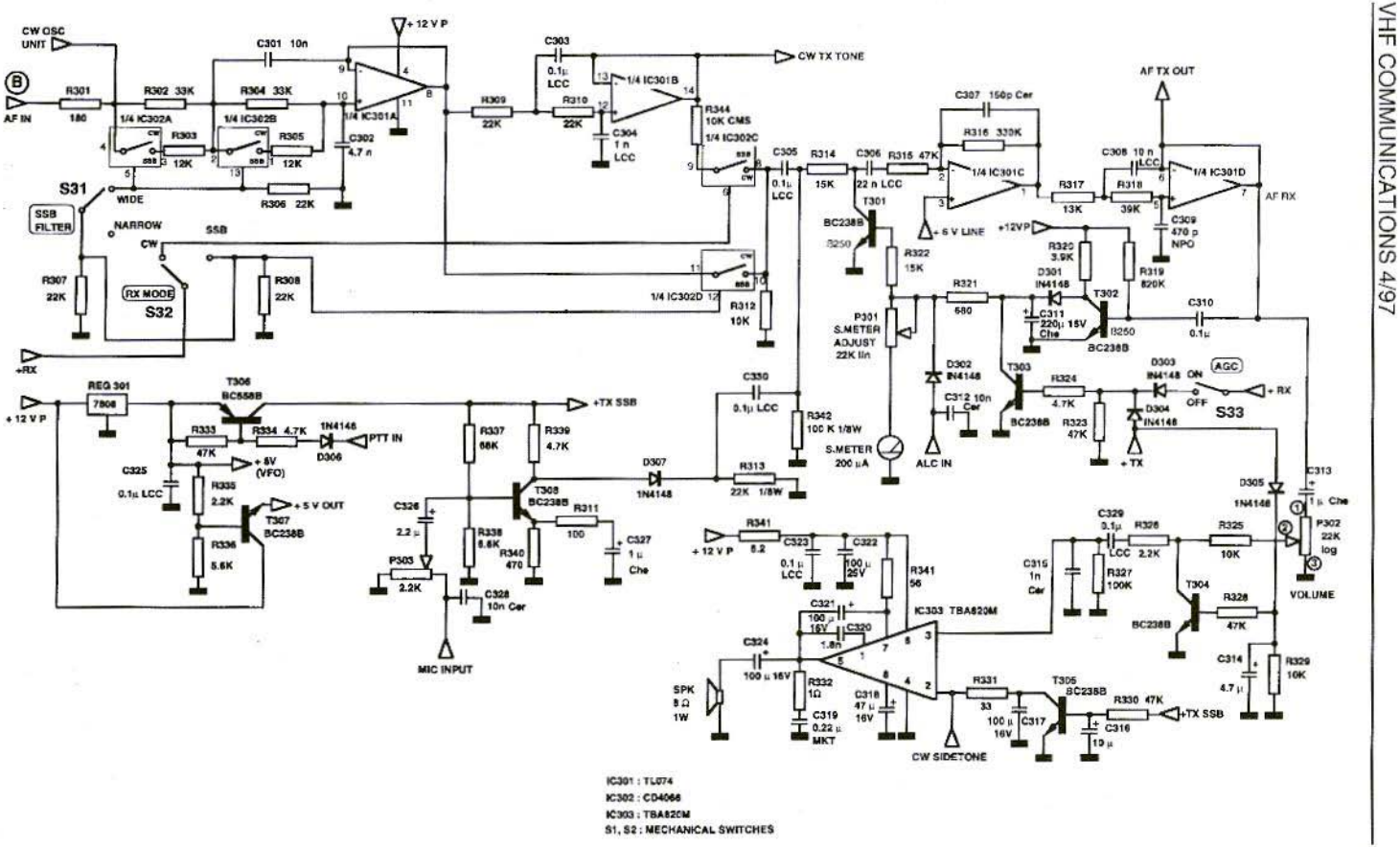


Fig.7: The complicated Low-Frequency section of the Phase Shifter

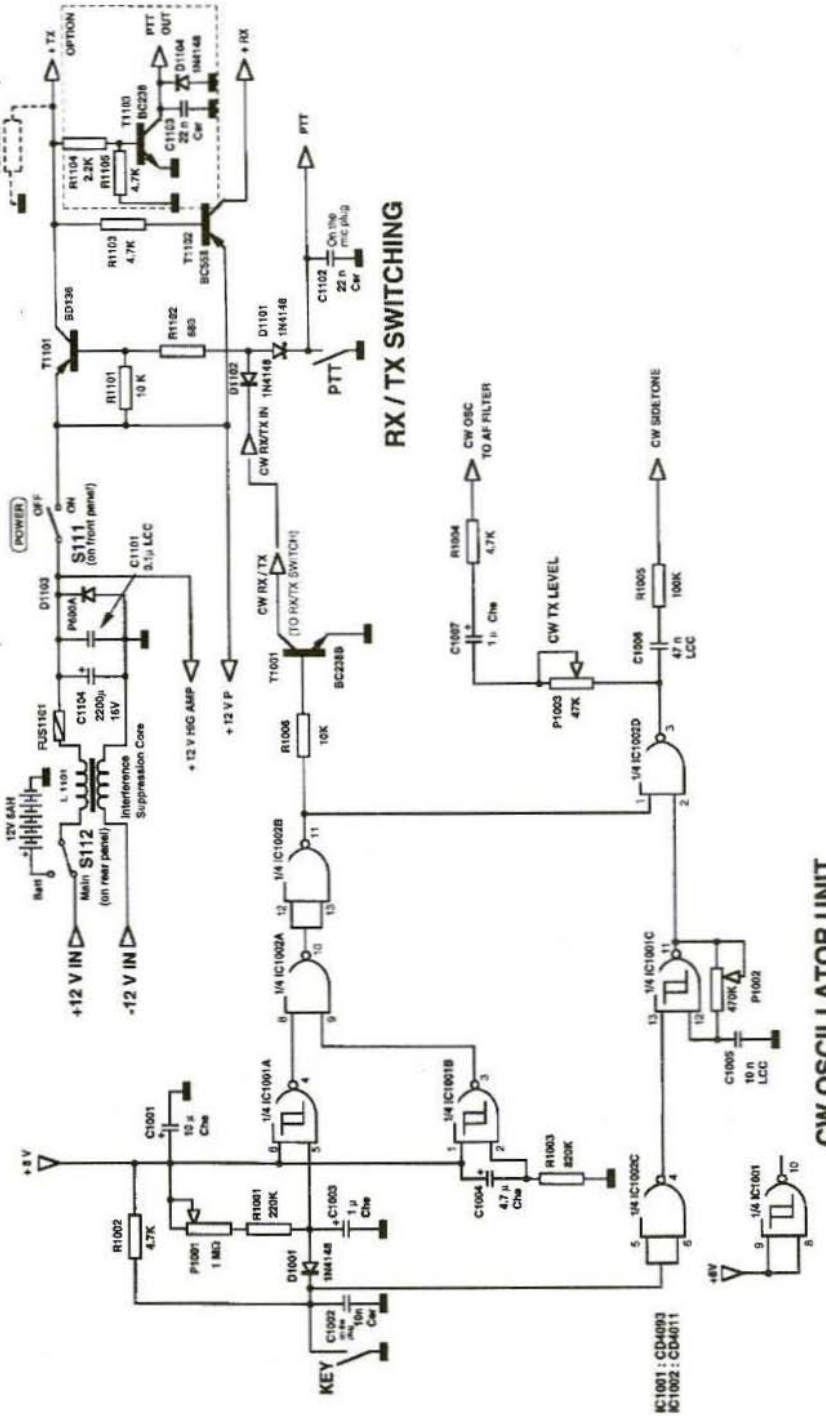


Fig.8: CW Module and Transmit / Receive Switching

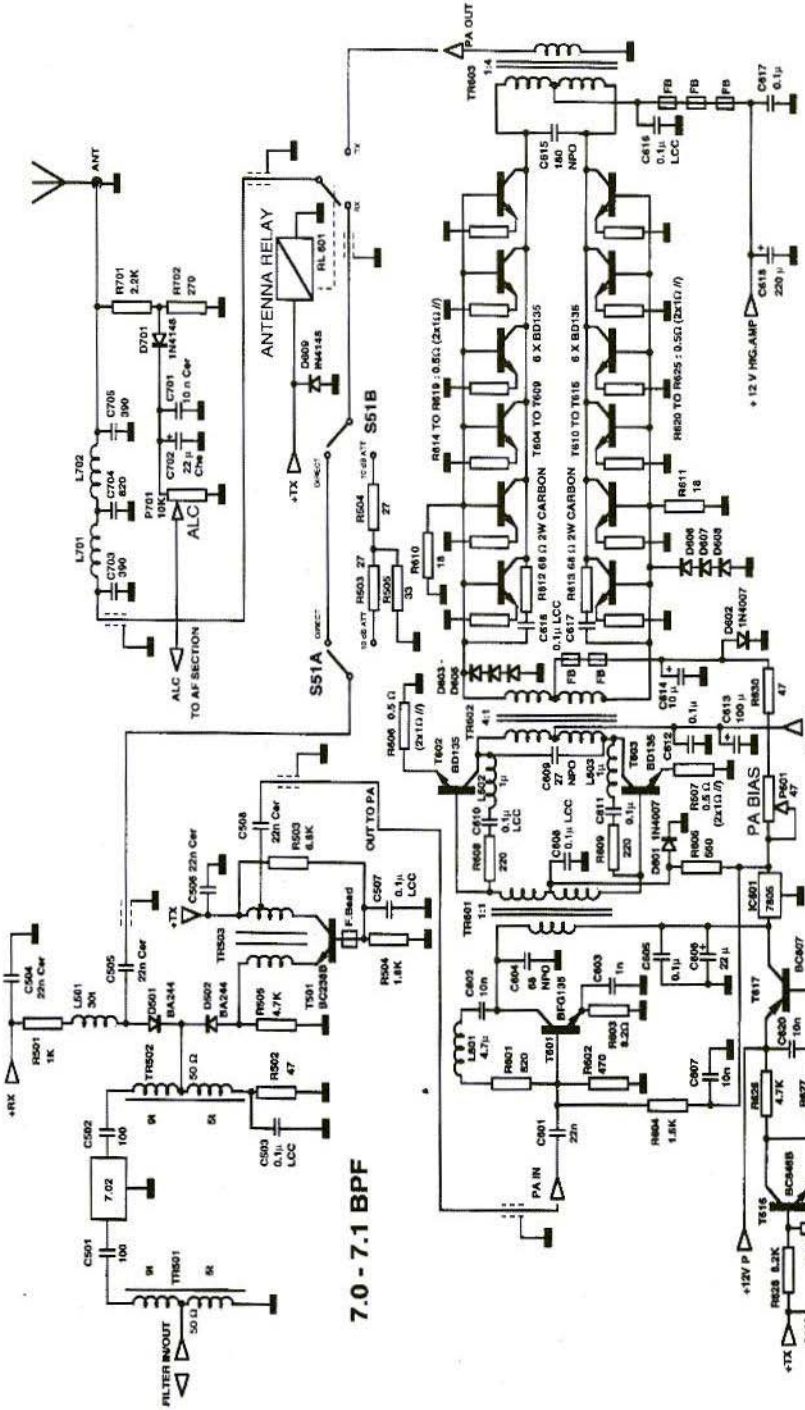


Fig.9: Band-Pass Filter, Transmit Amplifier, Deep-Pass Filter and ALC

3. ASSEMBLY

The transceiver is divided into eight modules for assembly:

- Low-frequency board as core of TRX
- VFO
- High-frequency phase shifter
- Mixer
- Band pass filter
- High-level stage
- Low pass filter
- Transmit / receive switching

3.1. Low-Frequency Board

This is the most complicated sub-assembly. Mounted on the 176mm x 76mm printed circuit board (Fig.10) are low-frequency phase shifters and low-frequency phase switching (Fig.5), the low-frequency section (Fig.6) and the CW oscillator. The tight squeeze is necessary to keep the equipment as small as possible (Fig.11).

It was the cost angle which led to the decision to disentangle the printed circuit board on one side only, instead of using a double-sided printed circuit board with feedthroughs, which would have cost 3 to 5 times as much to produce.

In addition, the one-sided version requires considerably more development work, and a quantity of unattractive wire bridges has to be accepted. So when equipping, we first begin by inserting the wire bridges, which are created using 0.5mm tin-plated copper wire.

In the IC 201/204 range, it is recommended that insulated wire of 0.3 to 0.4mm be used, to avoid short circuits. Don't forget the bridges under the IC's!

Next, the diodes, resistors, IC sockets, trimmers and finally capacitors and transistors are soldered in. For the IC's, it is recommended that sockets be built in. The sole exception is IC 303, the low-frequency amplifier, which is adequately cooled only if mounted directly onto the PCB.

C210 and R337 are SMD components and are soldered on last, on the tracks side. There was no longer enough room left for any other form of construction.

In addition, three insulated connections also had to be inserted, for which there would otherwise have been no room left:

8V In had to be linked to +8V Out

CW Out had to be linked to CW In

Tone Out had to be linked to Tone In.

Particular attention has to be paid to the earth connections with this construction. As already mentioned in the theoretical section, it is particularly difficult to suppress low-frequency feedback / coupling loops in receivers with direct switching and built-in loudspeakers.

The following methods have proved to be particularly suitable:

The printed circuit board must basically be connected to the earth. To do this, the board is screwed down only in the vicinity of the IC 303, with two screws, about 1.5cm apart, and earthed.

The earth pin near R316 (identified as "G") must be connected to the earth

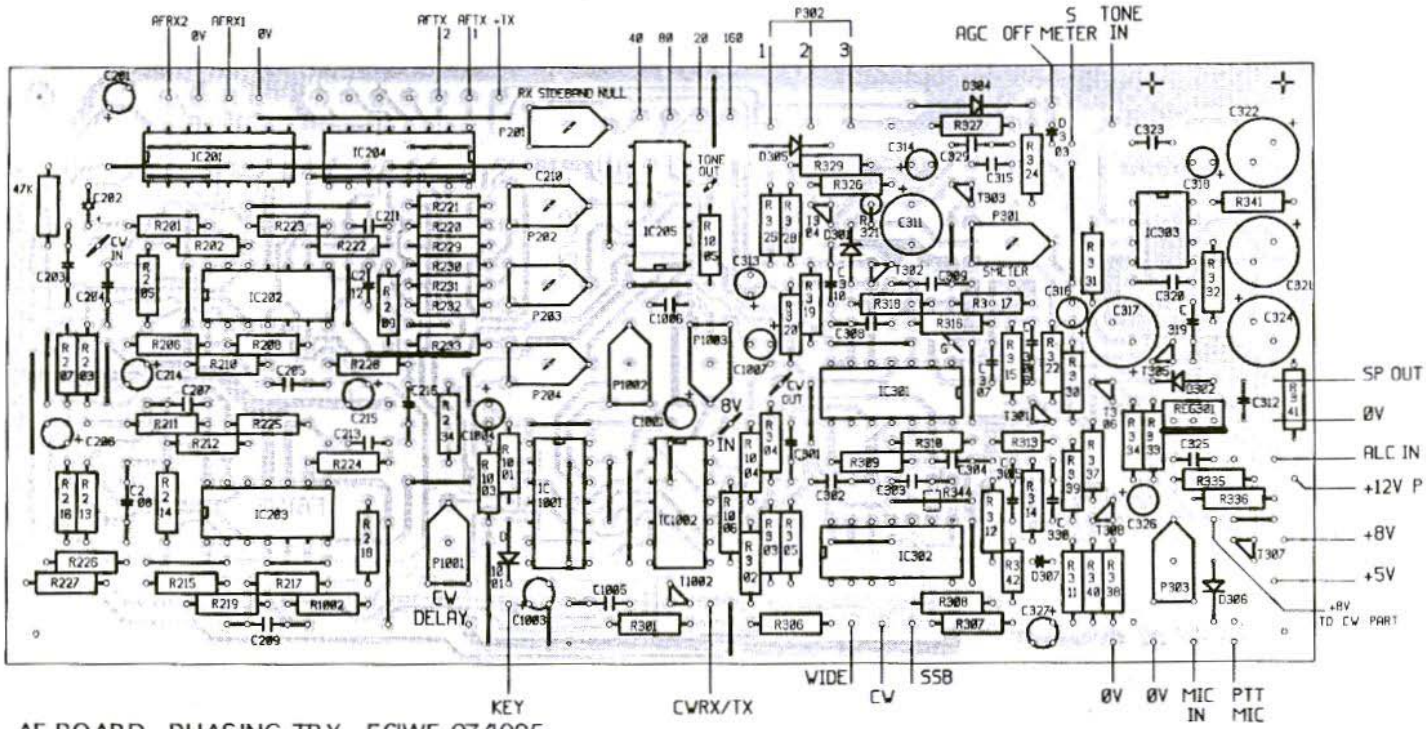


Fig.10: Low-Frequency PCB Component Overlay

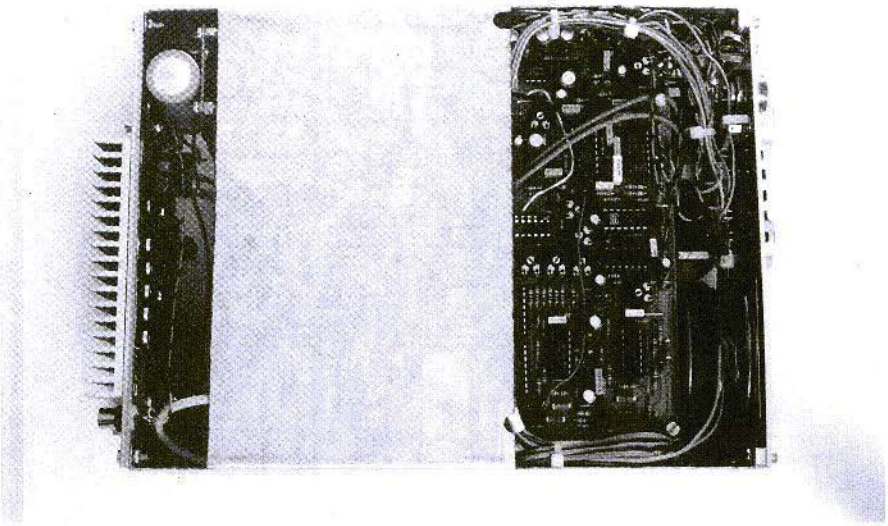


Fig.11: Bottom face of equipment, with view of Low-Frequency board

screw near C323 by the shortest route, and with an adequate cross-section (app. 1.2 to 1.5mm^2).

Other options were also tested, but naturally gave only unsatisfactory results.

3.2. VFO

This caused the greatest difficulties in the mechanical assembly. If tuning using Varicap is selected, D801, D802 and corresponding components do not need to be fitted (Fig.12). L802 is formed onto R811.

The frequency range of the VFO lies between 28.0 MHz and 30.4 MHz which, when divided by 8, gives the range 3.5 to 3.8 MHz. The frequency drift lies at about 500 Hz/h at 28 MHz, depending on the quality of the capacitors used.

3.3. High-Frequency Phase Shifter Module

This assembly takes the form of a plug-in module. This route was chosen in preparation for optional expansion to the multi-band version. However, the module can also be connected to the mixer assembly directly, using short wires.

IC 901 should be soldered in without a base, to avoid any problems which might arise from base capacities (Fig.13).

3.4. Mixers

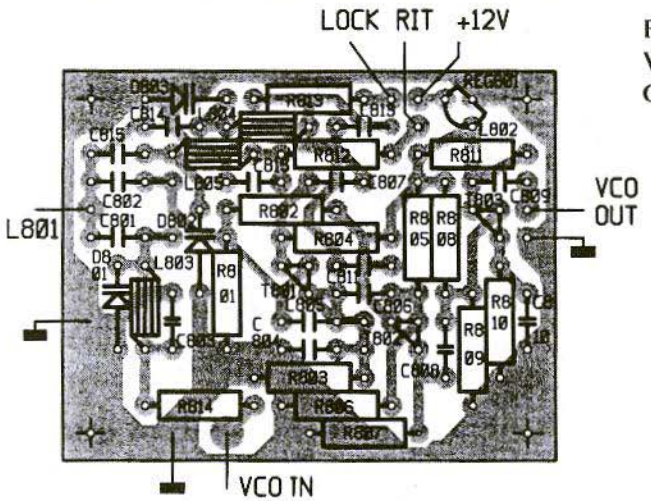


Fig.12: VFO assembly Component Overlay

This assembly includes the two mixers, the high-frequency pre-amplifier and the two pre-selector stages.

The mounting sequence is as described for the low-frequency board.

The components plan (Fig.14) shows the basic version of the mixer, with the options as per Section 2.1.

We should start by wiring up both mixers on the top face - e.g. with bridges instead of the balun transformers

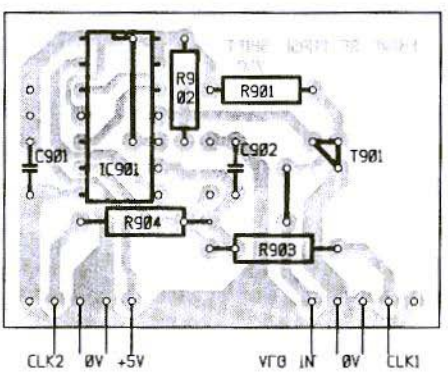


Fig.13: Components of Phase Shifter group

and potentiometers in the vicinity of R002.

The SMD component under C409 should not be forgotten.

When the assemblies have been successfully tested, the coils and radio-frequency transformers must subsequently be secured using "radio-frequency wax".

During assembly, make sure you obtain perfect symmetry for the mixers. To this end, the transformers must have exactly the same number of windings, and the material used must be identical.

SMD coils are fitted in the vicinity of the radio-frequency pre-amplifier (Fig.15).

3.5. The Radio-Frequency Filter Module

Pay attention to the direction of winding of the transformer during assembly.

The ferrite bead on T 501 should not be forgotten.

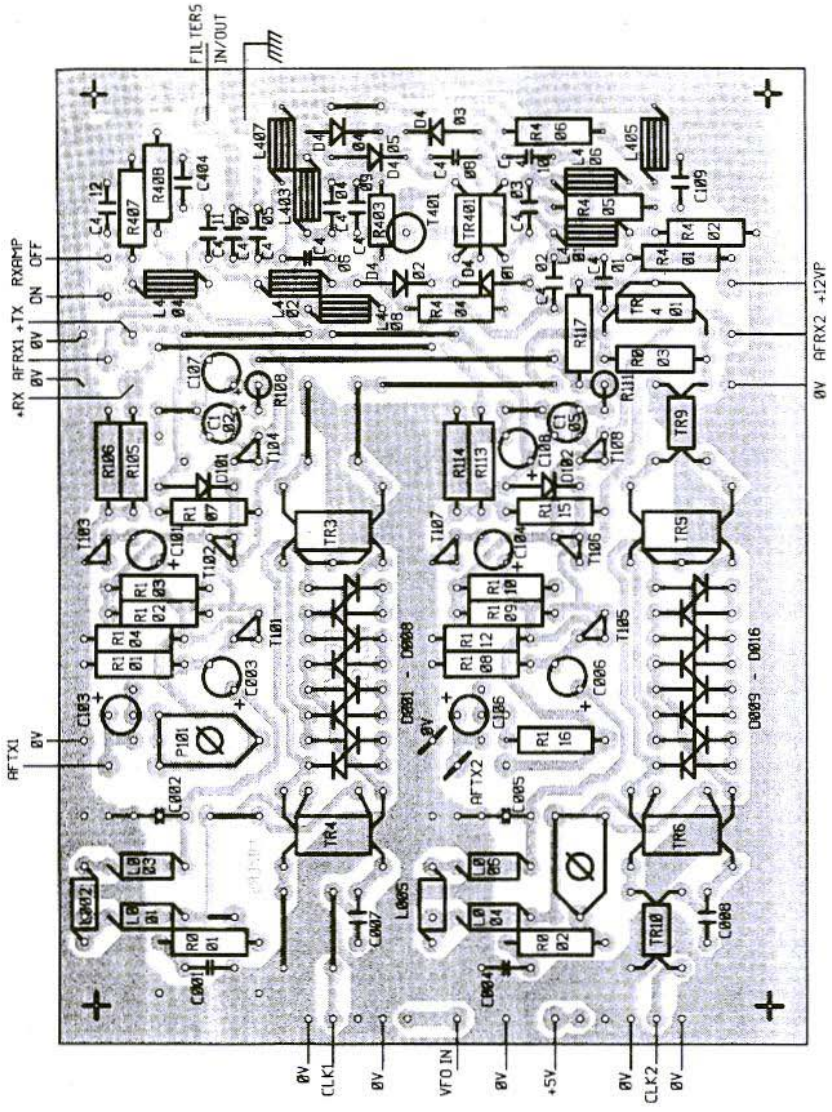


Fig.14: Component Overlay of printed circuit board, Mixer and Radio-Frequency Preamplifier

Following a successful test, C508 can be replaced by a bridge (Fig.16).

3.6. The Transmit Amplifier

You must naturally think about a special mechanical solution for cooling the transistors.

One very simple, but effective solution, is to mount the printed circuit board directly onto the cooling body. This requires a flat bottom face - i.e. all components, including the SMD components, are soldered on the top face (Fig.17).

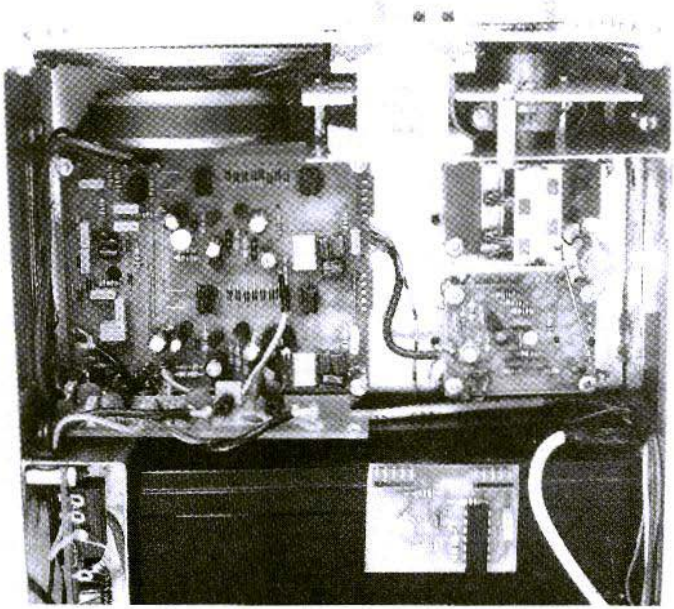


Fig.15: Mixer assembly (centre right) - the Radio-Frequency Phase Shifter is on the battery

The printed circuit board was designed for SMD components, which naturally also brings other advantages:

Thus, for example, the low-inductance feedback resistors, with 2W R612 and R613, which are very difficult to obtain, can each be replaced by 12 normal 22Ω SMD resistors.

Another advantage is the small space requirement, so that the PA assembly can now turn out to be considerably smaller.

Now for the assembly:-

First, all SMD components are firmly soldered on (Fig.18).

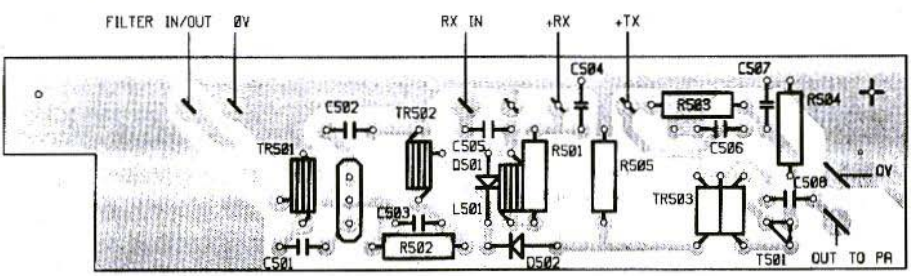


Fig.16: Filter assembly Component Overlay

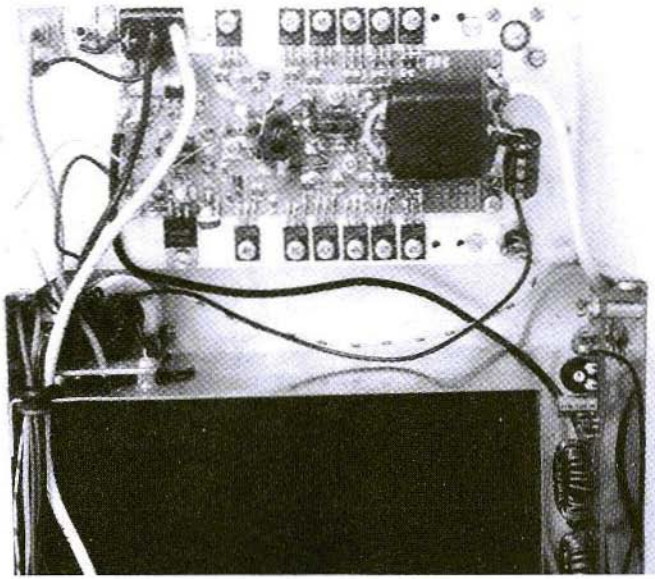


Fig.17: View of Amplifier assembly

During wiring, the connections should be continuously tested, using an ohmmeter, to avoid short circuits, which could have serious consequences.

The printed board assembly is directly screwed to the corresponding prepared rear wall.

The rear wall itself is made of 4mm thick aluminium and acts both as a cooling plate and as a mounting plate for the PA.

The board is drilled and provided with M3 threads - seven screws for the printed circuit board and one M3 x 5 screw with insulating sections for each PA transistor.

When the 7805 voltage regulator and all the transistors have been soldered on, check the resistance between collector connections and earth, and between base connections and earth.

The D601/601 diodes must be in thermal contact with the rear wall (heat conducting paste). Two breaches must be provided in the printed circuit board for this purpose.

Finally, the radio-frequency transformer is incorporated. The capacitors C604, C609 and C615 are soldered into the assembly directly. C614 is soldered to D602.

Up to 14 type BD135 transistors can be mounted on the printed circuit board.

The origin or manufacturer of the transistors is important in connection with use in the HFPA. I have experi-

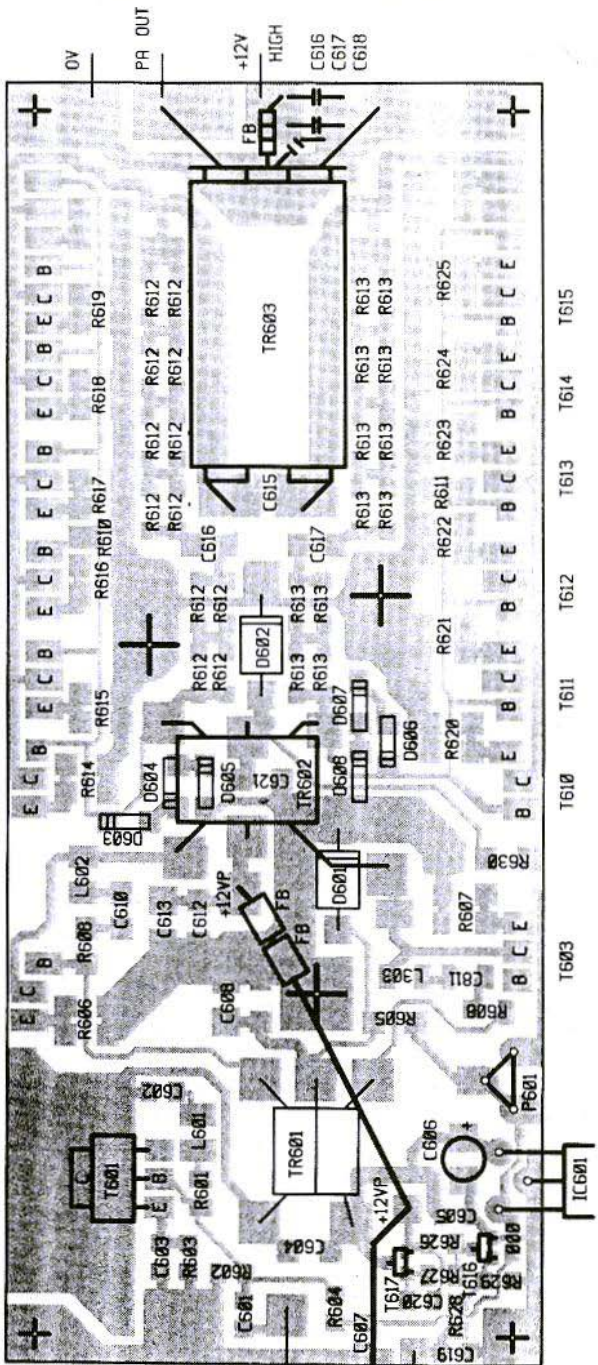


Fig.18: Transmit Amplifier Component Overlay (track side = component side!)

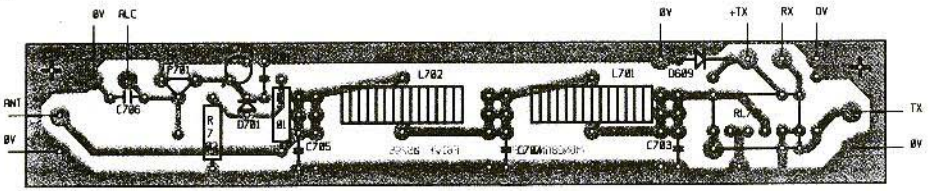


Fig.19: Low-Pass Filter strips

enced the best results using BD135-10's from SGS-Thomson.

To keep specimen scatter values as low as possible, the transistors should all be from the same production series / batch.

It certainly makes sense to test transistors from various manufacturers.

3.7. Low-Pass Filter

The printed circuit board is prepared for two different types of relay (Fig.19).

Should other relays be selected, the pin configuration should be suited to the printed circuit board.

3.8. Transmit / Receive Switching

The PTT option can be fitted if desired.

Wires with different colours should be used for L1101. The direction of winding at the output of L1101 is important.

The value for C1104 can be increased up to 4,700µF if the TRX is mainly intended for battery operation.

Fig.20 shows the components plan.

(To be continued)

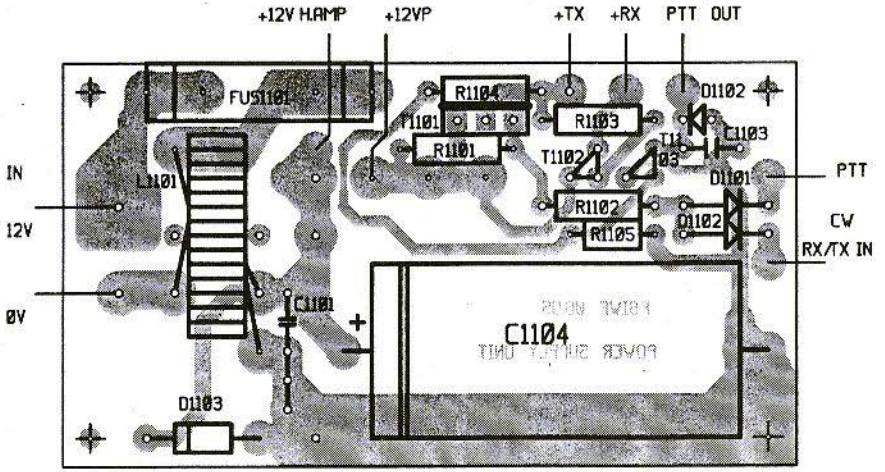


Fig.20: Power Supply assembly Component Overlay

Dipl.-Ing. Detlef Burchard, Box 14426, Nairobi, Kenya

The Sine Converter

To convert DC voltage into another value, or else galvanically insulate it (avoidance of earth circuits), an AC voltage must initially be generated. The flux and blocking-oscillator type converters predominantly used generate square wave voltages, but naturally sinusoidal voltage can be used as well.

1. INTRODUCTION

The principle of the sine converter is explained below, a model is dimensioned, and an example is given for those who would like to assemble and test something practical.

This is simultaneously a little training-ground for work on the magnetic field - which is not greatly loved by many design engineers, although the essentially still more complicated electromagnetic field provides them with their daily bread or with their hobby.

The calculations cannot be given like a cookery recipe. Even for something which seems as simple as a mains transformer, where there are cookery recipes [3], we have to start again from the induction law if we want to use a core made of air, cast iron or ferrite, which is possible - even if it is not practical.

An unconventional but practical representation of the magnetic field can be found in [1], which would allow calculations of the peculiar mains transformer.

The AC voltage of the sine converter can help with acoustic signals, call tones or clock frequencies. It can also be the sole objective of expense and trouble in itself, so that apparatus built for 50 Hz mains connection (or 400 Hz) can be operated with an accumulator, or radio-frequency power can be generated for coagulation in the operating theatre or for heating metals for hardening.

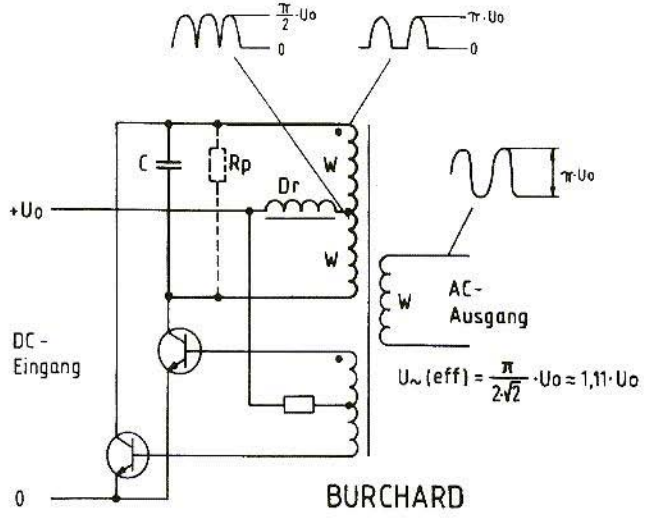


Fig.1:
Basic Circuit of Sinc Converter;
***Eingang* = Input,**
***Ausgang* = Output**

The principle is not new. It has even played a role in push-pull power amplifiers in amateur radio transmitters, and is regularly used in power inverters. A single literature reference [4] goes to show that it has not been completely forgotten in electronics.

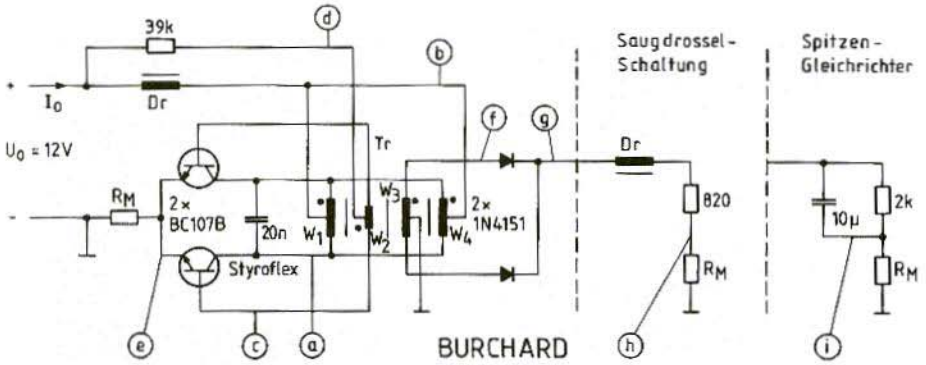
Here the range is restricted to converters of this type with low-power to medium-power ferrite coils, which use audio-frequency, or frequencies approximately 1 decade higher.

2. THE PRINCIPLE

The basic circuit, Fig.1, looks very similar to a push-pull rectangular converter. The differences are located in the capacitor, C, the choke, Dr, and the dimensioning of the transformer, Tr. The circuit is self-controlled and starts automatically. Following the start-up phase, the transistors work in pure switch mode.

In contrast to the forward or back type converter, they switch at a time when the load is extremely low, namely when their collector AC voltage passes through zero. The choke provides for a largely constant current from the DC source, which is automatically set to suit the load, and which also corresponds to the collector current in the transistors, except for a slight ripple effect. The resistor in the base circuit must provide sufficient base current for this collector current and must be correspondingly dimensioned. But it can also be (almost) as much higher as you like. Degrees of efficiency exceeding 80% are comparatively easy to attain.

The frequency is pre-set through the capacity, C, and the primary inductance of Tr. A load can be thought of as Rp, transformed in this resonant circuit. This must not be represented by an ohmic resistance. Due to the flywheel effect of the resonant circuit, the energy output



Dr
18ø×14 Al400
220 / 0.27

Tr
18ø×14 Al400
W_1 89+89/0.15
W_2 5+5/0.10
W_3 89+89/0.15
W_4 89+89/0.15

◯ Meßpunkte für Bild 3...5
R_M Strommeßwiderstand z.B. 10Ω 1%

Fig.2: Model Converter Circuit; ***Saugdrossel-Schaltung*** **= Drainage coil circuit,** ***Spitzen-Gleichrichter*** **= Peak rectifier,** ***Meßpunkte für Bild 3...5*** **= Measuring points for Figs. 3...5,** ***Strommeßwiderstand*** **= Current precision resistor,** ***Z.B.*** **= For example**

may not be uniformly distributed over the period. Half-wave, full-wave, bridge and drainage coil rectifiers, Delon rectifiers and Greinacher rectifiers and phase-angle controls can be connected. From the secondary side of Tr, things continue as we are used to seeing with a normal mains transformer. Only here the "flywheel" is mounted on the primary side, which we have to imagine for electrical distribution mains as the sum of all moving masses in the power stations.

Pulsatory energy extraction during the period thus does not cause a surge going right into the power station, but ends at the choke, Dr. This also ensures that bounces do not spread into the energy source during commutation (< 100mV). The primary RFI suppression of this converter is made considerably easier by this.

3. MODEL CONVERTER

The model converter in Fig.2 has been designed with a 1:1 transmission ratio in Tr. A drainage coil with 820Ω load or a charging capacitor with 2kΩ load can be connected to the full-wave rectifier. The load is transformed into the resonant circuit as

$$R = 0.5 \cdot \pi^2 \cdot R_L$$

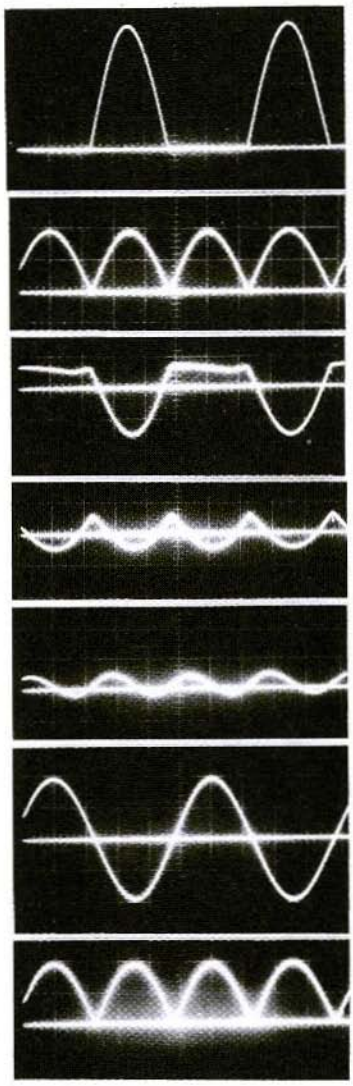


Fig.3: No Load;
Kurvenformen im Modellwandler =
Waveforms in model converter

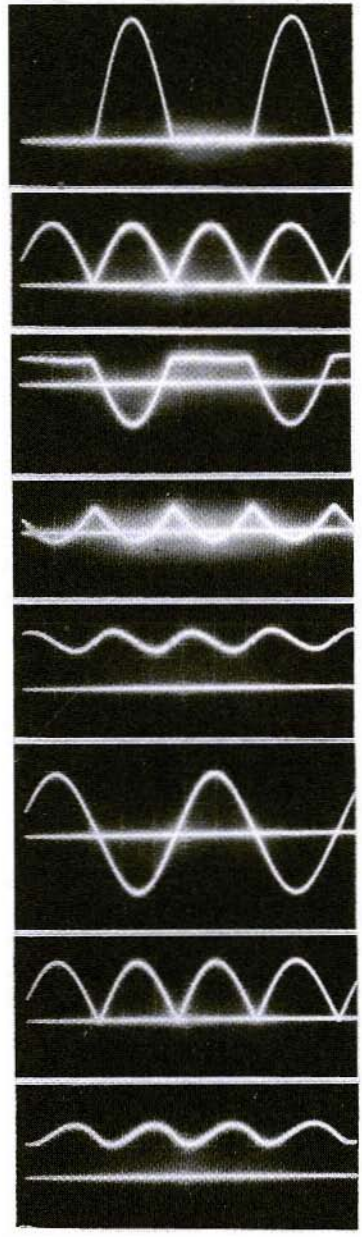


Fig.4: Nominal Load, Drainage Coil Rectifier

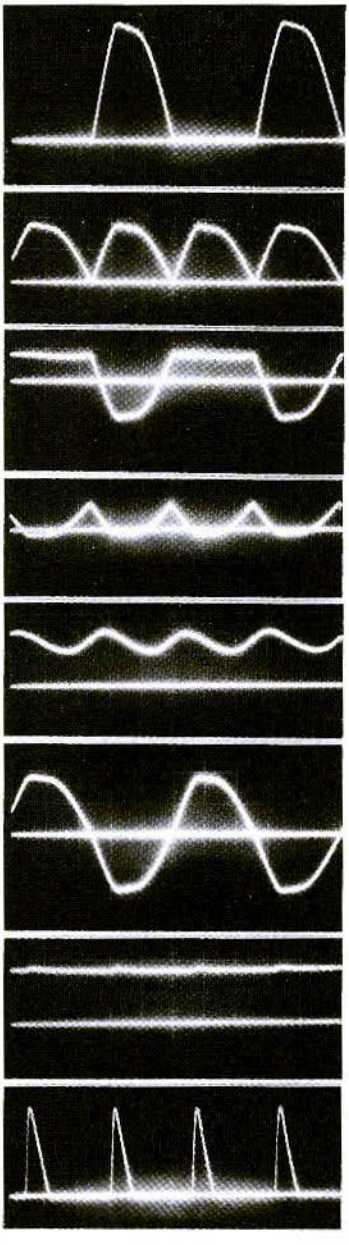


Fig.5: Nominal Load, Full-Wave Rectifier

Expanation of Fig's.3 to 5; Waveforms in Model Converter

a: Collector Voltage (10V/div)

b: Collector Centre-tap (10V/div)

c: Base Voltage (1V/div)

d: Base Centre-tap (1V/div)

e: Primary Current (10mA/div)

f: Secondary Voltage (10V/div)

g: Rectified Secondary Voltage (10V/div)

h: Secondary Current (10mA/div) (Load Voltage 8.2V/div)

i: Secondary Current (20mA/div)

The true zero line is in each case mixed in with double illumination. X is always 20µs/div.

in the first case, or $R_p = 2 \cdot R_L$ in the second. If we disregard the losses, then this comes out as equally $R_p \approx 4\ k\Omega$.

The illustration of all AC voltages and currents arising should make things easier to understand. In Fig.3, the waveforms are shown for no load. Fig.4 is then valid for the drainage coil circuit under nominal load, and Fig.5 for peak rectification. We shall omit any discussion of the curves.

The drainage coil circuit has little significance in practice, although it can improve the efficiency. It has been shown here because certain waveforms repeat themselves primarily and secondarily in it. The secondary side RFI suppressionis also simplified with it, if such filtering is needed at all, because lines lead out of the apparatus from there.

The determining components were measured on this model, as well as the total data:

Balance coils:
20mH + 2.9 Ω; Q ≈ 500 (20 kHz)

Transformer:
Primary 12.5mH + 4.2Ω, Q ≈ 200 (10 kHz)
Secondary 12.5mH + 7.3Ω, Q ≈ 120 (10 kHz)
Stray inductance 4.8µH
Coupling capacity 350pF

Capacitor: 20.5nF + 0.7Ω, Q ≈ 1,150 (10 kHz)

No-load current:
1.3mA, inclusive of 0.3mA base current

Drainage coil rectification:
Primary 12V / 15.1mA / 181mW
Secondary 11.1V / 13.5mA / 150mW
$\eta = 0.83$

Charging capacitor rectification:
Primary 12V / 13.3mA / 160mW
Secondary 16.2V / 8.1mA / 131mW
$\eta = 0.82$

An analysis of the losses reveals that they are essentially brought about through the saturation voltage of the transistors (0.2V) and the forward drop voltage of the rectifiers (0.7 to 1V).

The chokes could actually be wound to have even higher resistance and inductance. That would benefit the suppression of ripple in the input and output currents. 1.4 times the number of turns would double R and L, the additional voltage drop would be only 45mV, and the ripple would be halved.

4. TRANSFORMER CALCULATIONS

The calculations for a transformer involve, among other things, weighing up the iron and copper prices, the volumes and the heating. Generations of engineers have calculated and tested, until the mains transformer as per DIN 41300 [3] finally came about. There are still a few more degrees of freedom here (frequency, material, AL value), which makes a single precise solution for a given case very improbable. For this reason, a type of calculation is used here which accomodates an individual project and is based on a well-filled components store. Briefly, we've already got a few cores and we'd like to know what to wind round them.

The model converter above, for example, took this particular form because of what was available - a number of dia 18 x 14 AL 400 pot-cores (N.B. Don't forget the 10^{-9} H multiplier!) and styroflex capacitors up to 20nF.

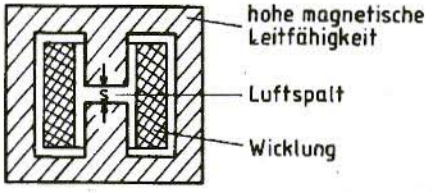


Fig.6: The Magnetic Circuit; the Reactive Power is located in the Air Gap;
***Hohe magnetische Leitfähigkeit* = Highly magnetically conductive, *Luftspalt* = Air gap, *Wicklung* = Winding**

Then, for example, the frequency 10 kHz and the resonant circuit grade under load $Q_L = 5$ were pre-set. But it is also possible to pre-set the power and the peak flux density, which are certainly more normal start conditions. The core size referred to here is sufficient for power levels from 5mW to 200W if the frequency is varied between 0.5 and 100 kHz and the A_L value between 630 and 25. There will be difficulties with cooling at the top end, as the heating has not been taken into account here. But 10-20W would certainly be worth trying. The figures show how flexible the calculation must be.

Core constants which make the calculation easier can be found in the relevant catalogues for ferrite cores. [2] is an example of such a catalogue, which also contains useful formulae for calculation. There we can find (among other things):

AF - magnetically active cross-section
AL - inductance factor
AR - resistance factor
AN - winding cross-section

It may be questioned from the catalogue whether the ferrite material is suitable above app. 20 kHz. Moreover, the permissible peak flux density should be looked up - it is always between 0.1 and 0.3 T. With an unknown material, we should always tend to the lower side. You don't go up to the peak value unless you want to "squeeze out the last drop".

The reactive power in the resonant circuit becomes effective in the air gap of the magnetic field (Fig.6). Air doesn't weigh much, costs nothing, creates no losses and can not be saturated. For practical reasons, the ferrite core must be there, with its specified operation limits (overload level). It limits the air volume to such a small part of the total volume that the reactive power becomes so low that it can still be fed in by the coil.

The more power has to be converted, the bigger the air gap must become! If the AL value of a core is known, there is no need to reckon with s at any point. Only when there is a core present without an air gap, or when one with a gap too small for relatively high power has to be "bored out", is it necessary to calculate the gap insert thickness.

The induction law is the basis of all calculations:

$$u = -L \cdot \frac{di}{dt} = -w \cdot \frac{d\varnothing}{dt} = -w \cdot AF \frac{dB}{dt} =$$

$$= -w \cdot AF \cdot \mu \cdot \frac{dH}{dt} = -w \cdot AF \cdot \mu_0 \cdot \frac{w}{s} \cdot \frac{di}{dt} \quad (1)$$

u and i are the instantaneous values for the voltage and the current, w is the number of turns.

ϕ is the flux in Webers (Wb = V · s),

B is the flux density in Teslas
$T = (V \cdot s)/m^2$,

H is the field strength in A/m,

μ_0 is the magnetic field constant:

$$\mu_0 = 1.257 \cdot 10^{-6} \cdot (V/A) \cdot (s/m)$$

(previously 1(G/Oe), but with the legislation on dimensional units in 1970, the Gauss, the Oerstedt and the Maxwell have fallen out of favour).

For the sine converter, all waveforms for u, i, B and H are sinusoidal. Using equation system (1), we can, since:

$$(d/dt)(\sin \omega t) = \omega \cdot \cos \omega t$$

then transcribe, for the amplitudes (peak values) and also for the effective values:

$$\hat{U} = \omega \cdot L \cdot \hat{I} = \omega \cdot w \cdot \hat{\Phi} = \omega \cdot w \cdot AF \cdot \hat{B} =$$
$$= \omega \cdot w \cdot AF \cdot \mu \cdot \hat{H} = \omega \cdot \frac{w^2}{s} \cdot AF \cdot \mu_0 \cdot \hat{I} \quad (2)$$

From the second and last terms, using comparison of coefficients, we derive

$$L = (w^2/s) \cdot AF \cdot \mu_0 \quad (3)$$

and since L is also $AL \cdot w^2$,

$$AL = (\mu_0 \cdot AF)/s \quad (4)$$

This equation makes it possible to calculate an air gap for a desired AL value.

No further grounds need be given to show that for a high degree of efficiency the L and C quality factor must be high. The initial Q, Q_o, can be estimated from the coil resistance and the inductance, if the core losses are disregarded:

$$Q_0 = \omega L/R = \omega \cdot (AL/AR) \quad (5)$$

The AR value specified in the catalogues relates to the total coil cross-section AN. Since the resonant circuit coil takes in only a part, t, of AN (e.g. 2/3), the AR value must be corrected:

$$AR_{korr} = AR/t \quad (6)$$

For a high degree of efficiency, the resonant circuit quality for a load, Q_L, should not be too high. A value of 5 has proved itself. If importance is placed on a particularly good curve form and frequency constancy, Q_L should be selected still higher, but a sufficient interval with Q_o must be maintained:

$$Q_L \leq Q_0/10 \quad (7)$$

With the voltage values from Fig.1, the parallel resistance, Rp, to the resonant circuit can be determined. It is independent from the type of equation if only the converted power is taken into account:

$$R_p = (\pi^2/2) \cdot (U_0^2/N) \quad (8)$$

If Q_o is disregarded, Q_L is then:

$$Q_L = \frac{R_P}{\omega L} = \frac{\pi^2}{2} \cdot \frac{U_0^2}{N \cdot \omega \cdot AL \cdot w^2} \quad (9)$$

and since, in accordance with the first and fourth terms of equation (2)

$$\hat{u} = \pi \cdot U_o = w \cdot AF \cdot \omega \cdot B$$

we finally obtain:

$$w = \frac{\pi}{2} \cdot \frac{U_0 \cdot \hat{B} \cdot AF}{Q_L \cdot N \cdot AL} \quad (10)$$

This determines the number of turns for the resonant circuit coil. Selecting a suitable transfer ratio to the secondary side is a trivial matter. Further algebraic conversion then gives us:

$$f = \frac{Q_L \cdot N \cdot AL}{\pi \cdot \hat{B}^2 \cdot AF^2} \quad (11)$$

and

$$C = \frac{\hat{B}^2 \cdot AF^2}{AL \cdot \pi^2 \cdot U_0^2} \quad (12)$$

If we put the cart before the horse, as happened, for example, in the model converter, we first obtain B from equation (12).

With the value obtained in this way (here 75mT) and the pre-set frequency 10 kHz, N from equation (11) is calculated at 180 mW, and finally w from equation (10) at 178.

It should always be borne in mind that the legal units for f, N, C and L are derived units. Here it is always better to use the basic units, i.e.:

1/s instead of Hz

V · A instead of W

(A · s)/V instead of F

(V · s)/A instead of H

and to check for each calculation whether all the units, together with multiplicators, form an equation.

Q_0 plays no part in this calculation. So a separate check must be carried out to see whether equation (7) is fulfilled. For the quality of the capacitor, C, to be sufficiently high:

$Q_c = 1/d = 1/\tan\delta$

types are used utilising styroflex, mica or polypropylene dielectrics, which have a Q_c value exceeding 1,000 in the audio-frequency range. Q_0 is then predominantly determined by the coil, and values of a few hundred can be attained throughout.

The base winding is not critical. It can be dimensioned for a peak voltage of between 1 V and $0.5.U_{BEmax}$ of the transistors used. It ensures that the base current flows into the correct transistor, which is guaranteed, even at a few 10's of mV's of potential difference.

5. CALCULATIONS FOR CHOKE

The choke, Dr, has to fulfil three conditions:

1. The inductive resistance must be high, as against the primary inductance of the transformer. The transistors switch it alternately in parallel to each part winding of Tr (primary side).

If we want $L_{Dr} \geq q.L_{Tr}$, with $q > 10$ being desirable, then if the same core size is used as for Tr it means:

$$w \geq \frac{\pi}{2} \cdot U_0 \cdot \hat{B} \cdot \sqrt{\frac{q \cdot AF}{AL \cdot Q_L \cdot N}} \qquad (13)$$

2. The pre-magnetising by the DC current, $I_o = N/U_o$ must not lead to any exceeding of B, thus:

$$w \leq \frac{\hat{B} \cdot AF \cdot U_0}{N \cdot AL} \qquad (14)$$

3. The ohmic resistance must not be too high. It is calculated at

$$R_{Dr} = AR \cdot w^2 \qquad (15)$$

The DC drop voltage is allowed to become only a few per cent of U_o.

So if $\rho = (I_o \cdot R_{Dr})/U_o$, then

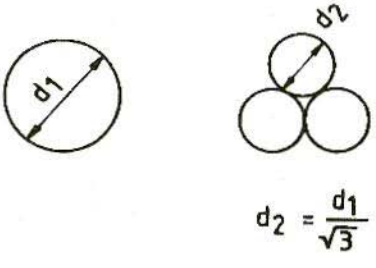


Fig.7: Home-Made Radio-Frequency Litz wire with same Cu Cross-Section - Easier Coiling, Reduced Skin Effect

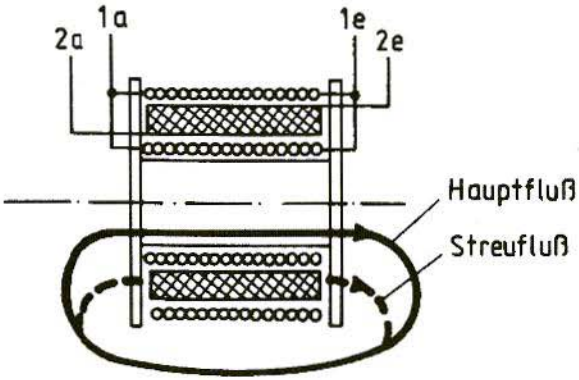


Fig.8:
Forcing Coil Prevents Stray Flux;
***Hauptfluß* = Main flux,**
***Streufluß* = Stray Flux**

$$w \leq U_0 \cdot \sqrt{\frac{P}{N \cdot AR}} \qquad (16)$$

is valid.

In general, it will be established that a wide margin for w exists. Depending on the priority, we can optimise in the direction of a low voltage drop or a low current ripple.

6. CALCULATION OF WIRE DIAMETER

The coil cross-section available, AN, is given in the catalogue, or is determined through the measurement of the window height and window width on the bobbin. As regards the transformer, the winding will usually be assigned the biggest part of the cross-section which gives the highest apparent power (reactive power and effective power geometrically added). That should always be the resonant circuit winding.

The wire diameters, d, are calculated, on the assumption of a Cu filling factor of 0.5 and a part, t, of the coil cross-section, at:

$$d = \sqrt{[(2/\pi) \cdot t \cdot (AN/w)]} \qquad (17)$$

The sum of all t values in one coil must not exceed 1! For the choke, t is naturally 1.

7. A FEW USEFUL TIPS

Bobbins for ferrite cores often have a very delicate structure. Coils with several parallel-wired thin wires create less stress than a thick one which has to be "forged in". A home-made litz in accordance with Fig.7 can even counter skin effect, which becomes noticeable at higher frequencies.

At low AL values, in particular, the stray inductance can become considerable. The known disc winding for reducing this requires special bobbins which may possibly be difficult to get. The less well-known push forcing is just as effective and easier to wind (Fig.8). The coil with the highest apparent

power - i.e. here the resonant circuit coil - is divided into a top winding and a bottom winding. All other windings are placed in between. No stray flux can exist in the volume between the top winding and the bottom winding. It would result in different voltages being induced in the two part-windings. But this can not occur, due to the parallel connection.

This forcing coil is therefore generally recommended. It helps the model converter transformer to achieve a stray inductance of only 0.038%.

The only disadvantage is that the primary and secondary windings lie opposite each other twice. The coupling capacity becomes greater, and two good insulations are necessary if high potential differences arise.

The resonant circuit capacitor may also be mounted secondarily. That can be advantageous if this means its value or its electric rating can be adapted in the desired manner. Such an example is shown later (Fig.10).

With many high-voltage converters, a capacitor is not needed at all if the internal capacitance is sufficiently high. But a high Q_0 value can not then be attained.

Instead of a single capacitor, it is also possible to provide several, on all coils, which are then to be dimensioned for values corresponding to the power fraction of the coil in question.

Although initially it seems like an academic game, this can be the only possibility of making the power transfer good enough if you have a transformer with considerable leakage (e.g. with an air core). Sine converters of, for example, 27 MHz, are built for use as industrial generators!

If it's a question of frequency constancy, those parts of the ferrite core should be glued with ttwo-part epoxy adhesive at the point where they are in contact with one another (but not at the air gap).

Since cores with a considerable air gap are always used here, the change in the AL value caused by the adhesive layer can usually be neglected. The adhesion also reduces "whistling". Due to magnetostriction, the core changes its volume in accordance with the instantaneous flux. It is an (almost) ideal isotropic zero-order radiator.

This sound radiation can not be impeded, and another frequency must be selected if the noise can not be tolerated. But microscopically small particles can hammer against one another at the interface, which generates even more noise. The two-part epoxy adhesive fastens them reliably.

Dipping the finished coil into clear polyurethane lacquer provides adequate protection against a hostile environment and decreasing adhesive force in the insulation film and gives an attractive appearance.

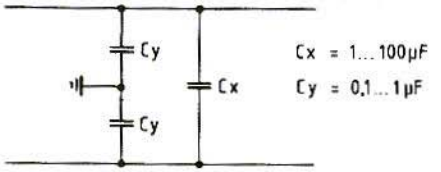


Fig.9: XY RFI Suppression Filter

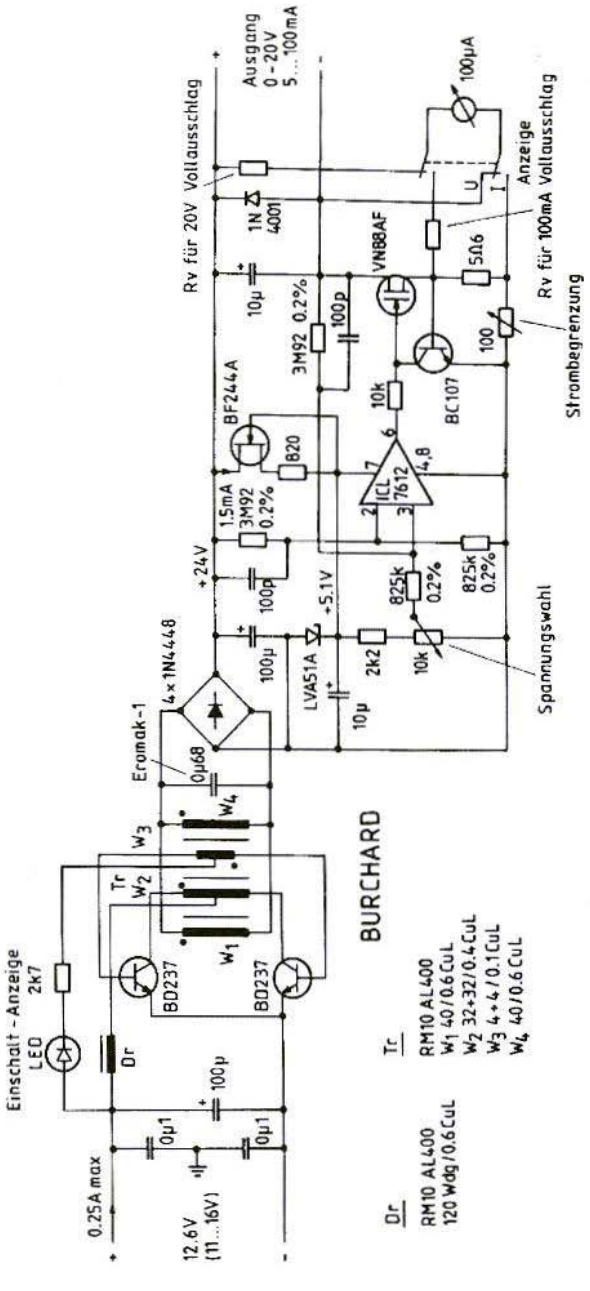


Fig.10: Field Power Pack Apparatus with Floating Inputs and Outputs;
Einschalt-Anzeige = On indicator, Spannungswahl = Voltage set
Strombegrenzung = Current limit set, Fuhr = For'
Vollausschlag = End-scale deflection, Ausgang = Output, Anzeige = Display

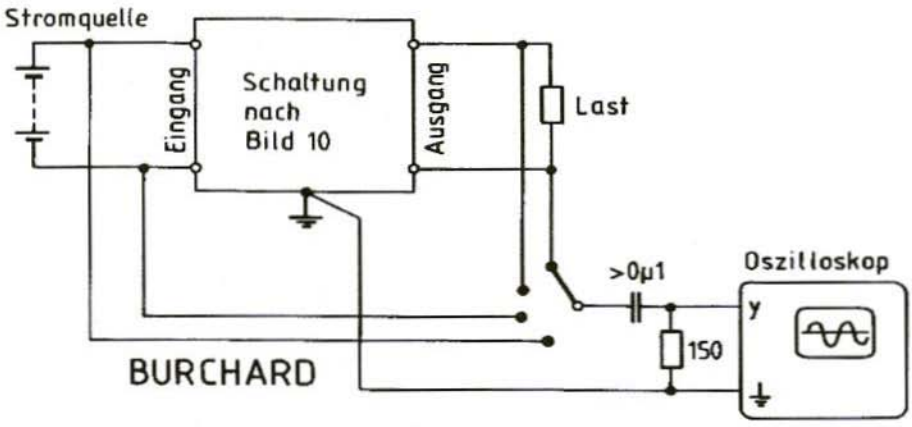


Fig.11: RFI Voltage Measuring Set-Up;
***Stromquelle* = Current source, *Eingang* = Input, *Schaltung nach Bild 10* = Circuit as per Fig.10, *Ausgang* = Output, *Last* = Load, *Oszilloskop* = Oscilloscope**

8. SELECTING TRANSISTORS

Lower requirements are imposed here on the switching times and SOA (safe operation area) than for other converters of the same frequency. The voltage rating can be read off directly from Fig.1 and the current switched is fairly precisely Io. But, as for any other converter, requirements are also imposed on the collector saturation voltage and the base current demand. Thus the B maximum should be approximately at Io, and B there should exceed 100.

If we now set the base current to $0.5.B_{max}$, this automatically results in a low saturation voltage. This rule always leads to over-dimensioned transistors, which require no special protection during the inrush phase. The power loss for the two transistors is essentially no higher than $U_{sat}.I_o$. A cooling surface can usually be dispensed with. Only at high converted power levels (I_o >2 to 5A) will more precise consideration and calculation be required.

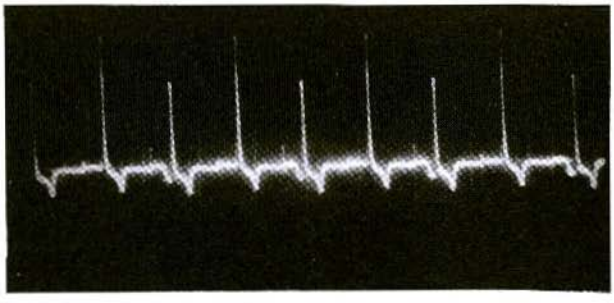


Fig.12:
Radio Interference Voltage at Output with 10V across 200Ω Load

Y: 20 mV / div
X: 50 µs / div

Cheap low-frequency transistors such as BC 107 or the like are suitable for low power levels. For extremely low input voltages (solar cell or a single storage battery cell), AC153's can also be used. For medium and higher power levels, transistors can be found of the types, for example, which are used for audio output stages. And naturally all types recommended for sold as square wave converters can be used here too.

9. SELECTION OF RECTIFIERS

There are a large number of semi-fast rectifiers for voltage converters which can be used here, although their price is markedly higher than that of universal diodes (1N4148, 1N4007, etc.).

The best technical data are obtained using Schottky rectifiers, which can process both high current levels and high frequencies. The sinusoidal form imposes less stress on the rectifier. Try an experiment using any diodes available before making a special purchase. Ge gold-bonded diodes from the old days are better for low power levels than anything produced since!

10. RFI SUPPRESSION

A clean electro-magnetic environment is just as important as other technical data. Now the sine converter can be filtered more easily than all other types of converter. On the primary side, an X-Y filter as per Fig.9 is sufficient in most cases, as also used in mains-operated pieces of apparatus with divergent capacitor values. The power supply can actually be unearthed, but can also be earthed on one side. Operating several devices using a converter of this kind from one power source is as problem-free as doing it from the mains. Earth loops and other inter-actions through the power source no longer arise.

The secondary side needs no filtering at all if no lines lead out of the equipment from there. However, if any do, as in my field laboratory power pack (Fig.10), the radio interference emitted is directly proportional to the coupling capacity. It can be detected using a measurement instrument set-up adapted from the VDE 0875 radio interference measuring method (Fig.11). For equipment as per Fig.10, this gives an oscillogram as per Fig.12.

It can be observed that small transients arise during commutating and rectifying which have a comb spectrum going right up to high frequencies. What can be seen here may still just be tolerable. If one side of the output is earthed, these interference effects disappear. Experience shows that, with a floating output, the radio interference voltage shown can affect radio reception in the close vicinity. Mains operated series-controlled laboratory power units can emit similar interference signals with a floating output. Of course, the resulting frequency there is 100 Hz.

RFI supression is thus important, and should be included in every project.

11. LITERATURE

[1] P.F.v.Eldik & P.Cornelius (1964): Transformers, Chokes, Transducers and Leakage Field Transformers; 2nd edition Philips Technical Library, Eindhoven

[2] Ferrite; Siemens Data Book Siemens AG, Munich

[3] Small Transformers, Transformers, Chokes; Characteristic Data DIN 41300, Beuth-Vertrieb, Berlin

[4] Laboratory Book, Vol. 5; 1st edition (1971) pp. 340 - 341 AEG-Telefunken Franzis-Verlag, Munich

The Parabolic 24cm Preamplifier

A fully weather-proofed very high-quality preamplifier covering 1200 MHz to 1360 MHz. Mounted in a sealed diecast enclosure with weatherproof N-type sockets for input and output. DC powered via the output socket for remote mast-head mounting.

GAIN >40dB across the band
NOISE FIGURE <1.6dB

£135 + £7.50 post and packing

KM Publications, 5 Ware Orchard, Barby, Nr.Rugby, CV23 8UF, U.K.
Tel: 01788 890 365; Int: +44 1788 890 365
Fax: 01788 891 883; Int: +44 1788 891 883
ALL MAJOR CREDIT CARDS ACCEPTED - above prices +5%

Matjaz Vidmar, S53MV

GPS/GLONASS Receiver Hardware and Software Update #1

Since the publication of the series of articles about the GPS/GLONASS receiver in UKW-Berichte/VHF-Communications there have been a few modifications of the hardware and software of

these receivers. The original articles describe the operation of the GPS software V122 and GLONASS software V39. The current update describes the new GPS software V125 and GLONASS software

V42. The new GPS V125 and GLONASS V42 include the following modifications:

(1) Improved internal operation of the software. The new software is able to handle the overflows of the hardware counters correctly thus almost eliminating the occurrence of the "T" error.

(2) Additions to the command set:

(2.1) In menu #5, key #0 will shift the privileged RX channel.

(2.2) In menu #7, key #0 has a new function: in AUT mode it operates as before while in MAN mode this command sets the carrier frequencies of the currently selected satellites.

(2.3) In menu #7, key #1 has an additional function: the receiver will display the Keplerian elements of the GPS/GLONASS satellites as decoded from the almanac data before entering the total RESET sequence.

(3) A simple bi-directional RS-232 interface is included, requiring only a few additional hardware components to be installed in the receiver.

(3.1) The RS-232 interface output circuit is shown on Fig.1. Because of hardware limitations, the bit rate can only be set to 1000bps. The output data format is a serial asynchronous transmission including a start bit, 8 data bits, no parity and one stop bit. The output signal level ranges from 0V to +5V only, although these levels are usually accepted by most RS-232 receivers. The signal polarity is inverted as usual in RS-232: 0V represents a logical "1" while +5V represents a logical "0". The data output matches the LCD display

content, the display clear command being replaced by a CR/LF combination.

(3.2) The RS-232 interface input circuit is shown on Fig.2. Because of hardware limitations, the bit rate can only be set to 100bps or 10 times slower than the output rate. The RS-232 input can be used to issue commands identical to those coming from the 8-key keyboard. Only ASCII characters "0", "1", "2", "3", "4", "5", "6" and "7" are therefore accepted as valid commands. All other codes are simply ignored. The data format is 8 bits, no parity, one or more stop bits. The signal polarity is inverted as usual in RS-232.

Since the PC6 input is now used for the RS-232 interface input, it can no longer be used to select the display type, CRT or LCD. The latter selection is thus pre-set in the software.

The new receiver software is supplied in three files. The .SRC file is intended to be compiled and executed on a DSP computer, equipped with a GPS/GLONASS hardware board. The .BIN file is a compressed version of the program to be burned in a 27C256 EPROM for a stand-alone GPS/GLONASS receiver.

Finally, the .ASM file is only supplied as an information about the internal operation of the main interrupt routine that includes all of the digital signal processing. GTERM is a 1000bps RX, 100bps TX RS-232 terminal program for the DSP computer.

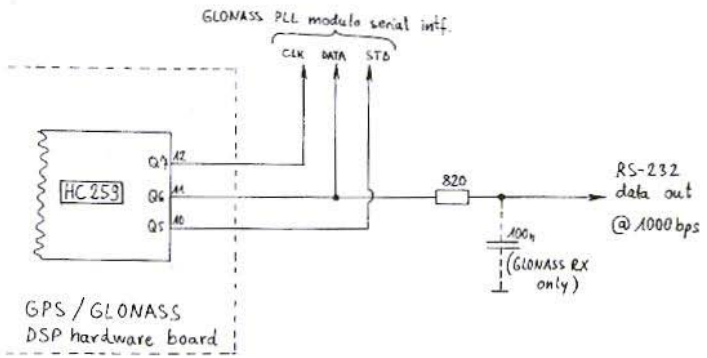


Fig.1: RS-232 Interface Output Circuit

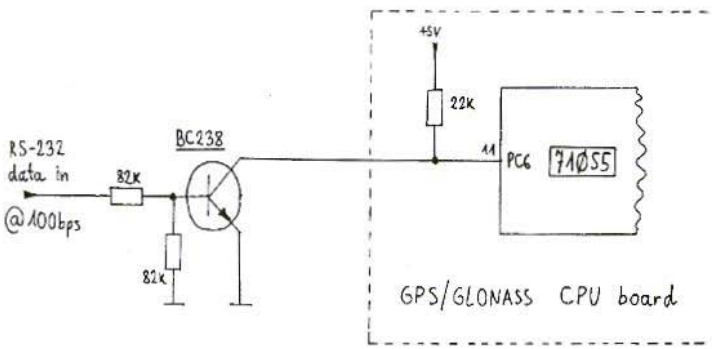


Fig.2: RS-232 Interface Input Circuit

COMPLETE KITS & PCB's

KIT	DESCRIPTION	ISSUE	No.	PRICE
DB6NT-001	Measuring Amp up to 2.5 GHz	4/93	06382	£ 60.00
DB6NT-002	Frequency Divider to 5.5. GHz	4/93	06383	£100.00
DC8UG-007	5W PA for 13cm	3/94	06938	£286.00
DJ8ES-001	23cm FM ATV Converter	1/91	06347	£ 93.00
DJ8ES-002	Digital Frequency Indicator	1/91	06352	£ 53.00
DJ8ES-003	IF Amplifier	1/91	06355	£ 55.00
DJ8ES-004	Demodulator	1/91	06362	£ 72.00
DJ8ES-019	Transverter 144/28 MHz	4/93	06385	£143.00
DJ8ES-019mod	Transverter 50/28 MHz	2/95	06392	£143.00
DJ8ES-020	Hybrid Amplifier for 144 MHz	1/94	06387	£179.00
DJ8ES-021	13cm FM ATV Exciter	2/94	06388	£ 67.00
DJ8ES-022	28/432 MHz Transverter Oscillator	2/95	06395	£ 59.00
DJ8ES-023	28/432 MHz Transverter Converter	2/95	06396	£114.00
HF-MESS	PC Software for the DJ8ES projects	TBA	--------	£ 32.50
OE2TLZ 001	PreMixer for 23 and 13cm	1/97	--------	£115.00
YT3MV	Low Noise L-Band Aerial Amplifier	2/92	06358	£ 29.80

PCB	DESCRIPTION	ISSUE	No.	PRICE
DC8UG-PA	5W PA for 13cm	3/94	06936	£ 19.25
DC8UG-NT	Power Supply for the PA	3/94	06937	£ 7.75
DJ8ES-001	23cm FM ATV Converter	1/91	06347	£ 10.75
DJ8ES-019	Transverter 144/28 MHz	4/93	06384	£ 17.75
DJ8ES-020	Hybrid Amplifier 144 MHz	1/94	06386	£ 17.25
DJ8ES-021	13cm FM ATV Exciter	2/94	--------	£ 18.50
DJ8ES-022	28/432 MHz Transverter Oscillator	2/95	--------	£ 16.50
DJ8ES-023	28/432 MHz Transverter Converter	2/95	--------	£ 16.50
DJ8ES-027	Milliwatt Meter for the PC	4/96	--------	£ 13.50
DJ8ES-028	0.9 - 1.5 GHz Synthesiser for the PC	1/97	--------	£ 13.50
DJ8ES-029	Interface Card	3/97	--------	£ 15.00
DJ8ES-030	VCO with PLL	3/97	--------	£ 15.00
DJ8ES-031	Mixer / LO	3/97	--------	£ 15.00
DJ8ES-032	Pin Diode Controller	3/97	--------	£ 15.00
DJ8ES-033	A/D Converter with Detector	4/97	--------	£ 15.00
DJ8ES-034	Direct Mixer with Logarithmic Display	4/97	--------	£ 15.00
HF-MESS	Software for the DJ8ES PC projects	TBA	--------	£ 32.50
DF9PL-001	30 Volt PSU	1/93	06378	£ 9.80
DF9PL-002	Pre-Stabiliser	1/93	06376	£ 10.20
DF9PL-003	Precision Stabiliser	1/93	06377	£ 11.20
OE2TZL-001	PreMixer for 23 and 13cm	1/97	--------	£ 15.00

Minimum Shipping charge £6.75 * Credit Card orders +5% * Address data at bottom of Contents page

VHF Communications

Back Issues

Back issues of VHF Communications Magazine are available at £2.50 each

Issues available: 1, 2, 4/1972 * 2, 4/1973 1/1974 * 1, 2, 3, 4/1975 * 3/1976 * 1, 4/1977 * 1, 2/1978 * 1, 3/1979 1980 to 1996

less 2/80, 1/81, 4/81, all 82, 3/87, 1/88, 2/89, 3/89, 1/90 all 91

Price includes surface shipping, for Air Mail add £0.75 per issue. Credit card orders +5%

A complete set of available issues to 4/1996
£120.00 plus shipping at cost

KM Publications, 5 Ware Orchard, Barby, Nr.Rugby, CV23 8UF, UK

Tel: (0)1788 890365 Fax: (0)1788 891883

Email: vhfsales@vhfcomm.co.uk Secure WWW Form

https://www.vhfcomm.co.uk/saleform.htm

VHF *COMMUNICATIONS*

A Publication for the Radio Amateur Worldwide

Especially Covering VHF, UHF and Microwaves

Volume No. 29 . Winter . Edition 4/1997

Publishers

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

Tel: +44 1788 890365
Fax: +44 1788 891883

Email:
michael.j@vhfcomm.co.uk

Editor

Mike Wooding G6IQM

VHF *COMMUNICATIONS*

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

The 1997 subscription price is £18.00, or national equivalent. Individual copies are available at £4.75, or national equivalent each. Subscriptions should be addressed to the national representative shown in the next column. Orders for individual copies of the magazine , back issues, kits, binders, or any other enquiries should be addressed directly to the publishers.

NOTICE: No guarantee is given that the circuits, plans and PCB designs published are free of intellectual property rights. Commercial supply of these designs without the agreement of the Author and Publisher is not allowed. Users should also take notice of all relevant laws and regulations when designing, constructing and operating radio devices.

© KM PUBLICATIONS

All rights reserved. Reprints, translations, or extracts only with the written approval of the publishers

Translated by: Inter-Ling Services, 62 Caldecott Street, Rugby, CV21 3TH, UK

Printed in the United Kingdom by: Cramphorn Colour Printers Ltd., 15c Paynes Lane, Rugby.

AUSTRALIA - W.I.A. South Australia Division, GPO Box 1234, Adelaide, SA 5001, Australia. Tel/Fax: 8 8522 4859

BELGIUM - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133-77980 Fax: 09133-779833 Postgiro Nbg. 30445-858

DENMARK - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

FRANCE - Christianne Michel F5SM, SM ELECTRONIC, 20bis Avenue des Clairions, F-89000 AUXERRE, France Tel: (86) 46 96 59

FINLAND - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

GERMANY - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133 7798-0. Fax: 09133 779833

GREECE - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

HOLLAND - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

ITALY - ADB ELETTRONICA di Luchesi Fabrizio IW5ADB, Via Vecchia Romana 266, 55100 ANTRACCOLI (LUCCA), Italy Tel: 0583 952612. Fax: 0583 91109

NEW ZEALAND - Judith Holland, AUCKLAND VHF GROUP Inc., P.O. Box 10 138, AUCKLAND 1030, New Zealand Tel: 0 9 412 7659

NORWAY - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: +46 40 16 42 66. Fax: +46 40 15 05 07. GSM: 0705 16 42 66
e-mail: waveline@algonet.se http://www.algonet.se/~waveline

SOUTH AFRICA - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

SPAIN & PORTUGAL - JULIO A. PRIETO ALONSO EA4CJ, Donoso Cortes 58 5° -B, MADRID 15, Spain. Tel: 543 83 84

SWEDEN - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: 040 16 42 66. Fax: 040 15 05 07. GSM: 0705 16 42 66 e-mail: waveline@algonet.se http://www.algonet.se/~waveline

SWITZERLAND - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

UNITED KINGDOM - KM PUBLICATIONS , 5 Ware Orchard, Barby, Nr.RUGBY, CV23 8UF, UK. Tel: +44 788 890365. Fax: +44 788 891883. Email: vhfsubs@vhfcomm.co.uk

U.S.A. - WYMAN RESEARCH Inc., 8339S 850W, WALDRON, Indiana 46182-9608, U.S.A. Tel: (765) 525 6452

- Henry Ruh, ATVQ MAGAZINE, 3 North Court Street, Crown Point, Indiana 46307, U.S.A. Tel: (219) 662 6395. Fax: (219) 662 6991

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

WWW: http://www.vhfcomm.co.uk

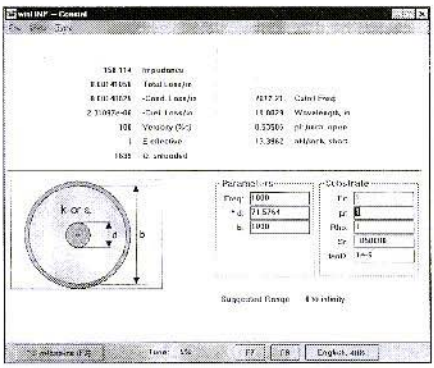
ISSN 0177-7505

AFFORDABLE SOFTWARE FOR PRO AND AMATEUR

*win*LINE

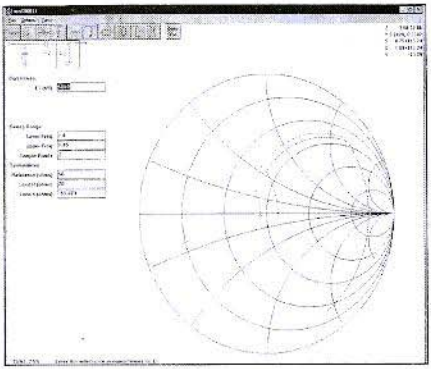
Transmission line impedance calculations for all sorts of lines: coaxial, stripline, microstrip, plus rectangular lines, eccentric lines, trough lines, wire-above-ground plane, and many others. Computes impedance, loss, inductance, capacitance and other parameters. Uses data from *Microwave Transmission Line Impedance Data*. For Windows.

Order NP-11 $195.00



Microwave Transmission Line Impedance Data

by M.A.R. Gunston **Order NP-10 $54.00**



*win*SMITH

The best electronic Smith Chart around! "Build" a ladder network of up to nine elements (L, C, R, transmission line or transformer) and use the chart to tune values until the match is perfect. Written for stand-alone operation, but makes a perfect addition to the book *Electronic Applications of the Smith Chart*. For Windows.

Order NP-5 $79.00

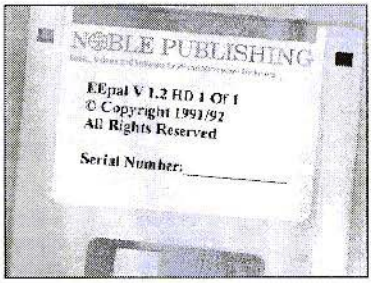
Electronic Applications of the Smith Chart

by Philip Smith **Order NP-4 $59.00**

EEpal

The "Electrical Engineer's Pal" is loaded with useful data and utilities for the active designer. Includes units and conversions, materials and component data, scientific calculator with higher math functions, pads, filters, matching, and more. Has a notepad and calendar, plus a phone list with an autodialer. All this is topped off with a list 1800 EE suppliers! Runs in DOS or a DOS window.

Order NP-20 $195.00



Order by telephone, fax, mail or our Web site!

We accept payment by VISA, MasterCard and American Express or check drawn on a U.S bank. Shipping charges are – U.S.: $5.00 for the first item, $1.00 for each additional item; Canada: $10.00 first item, $2.00 each additional item; Other countries (Air express): $32.00 first item, $10.00 each additional item.

CRESTONE TECHNICAL BOOKS

CRESTONE TECHNICAL BOOKS
Div. of Noble Publishing Corporation
2245 Dillard Street • Tucker, GA 30084 • USA
Tel: (770) 908-2320 • Fax: (770) 939-0157
www.noblepub.com

MAINLINE ELECTRONICS

2 exciting new kits recently added to our catalogue

DIGITAL FREQUENCY DISPLAY

A high-quality microcontroller based unit capable of measuring up to 32 MHz as a basic unit, or by preselecting an IF offset by means of a trimpot it can be used as the frequency readout of a Tx or Rx by adding/subtracting the offset from the LO. Also, by adding a suitable prescaler it can be used up to 8 GHz. 16 character LCD display with 100 Hz resolution - e.g. 14.257 MHz USB - the mode display is set by external resistors and can be programmed to show: AM, FM, CW, USB, LSB, FSK, FAX, or can be left blank.

Kit (no case):	DFD1	£42.46 + VAT
Assembled (no case):	DFD2	£47.20 + VAT

DIGITAL INDUCTANCE/CAPACITANCE METER

A real breakthrough in L/C measuring equipment at the price. A must for home constructors to help identify components and as an aid to winding coils, etc. At 5 samples per second the unit will also track whilst checking adjustable components. Microcomputer based.

Once you own one you will wonder how you managed without it!

Basic specification: 0.001 µH to 100 mH (150 mH typical)
0.1 pF to 1 µF non-polarised (1.5 µF typical)
Auto-ranging and self-calibrating - 16 character intelligent LCD display
4 digit resolution - accuracy typ. 1% of reading - direct display in engineering units, µH, pF, etc. Powered by 9V battery.

Kit inc. box and hardware:	LCMKIT	£79.00 + VAT
Built and tested (no battery):	LCMBLT	£94.00 + VAT

VISA

P&P free in UK, Overseas extra. VAT is 17.5%

MasterCard

MAINLINE ELECTRONICS, P.O. BOX 235, LEICESTER, LE2 9SH, U.K.
TEL: (0)116 277 7648/278 0891 FAX: (0)116 247 7551