

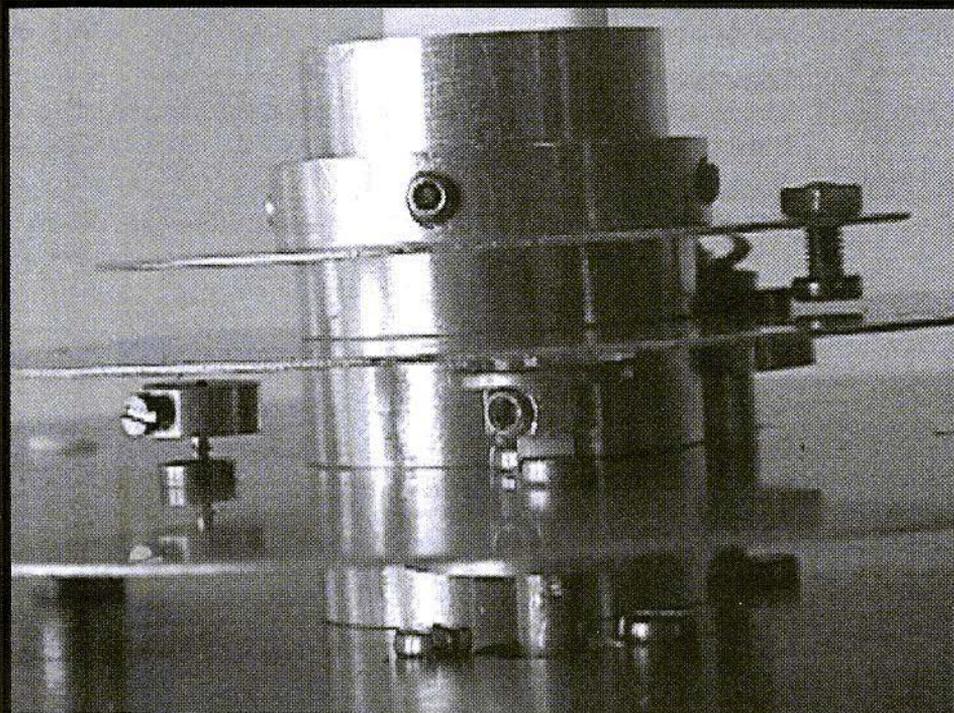


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Microwave Multi Band Feed

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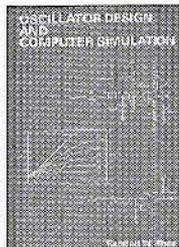
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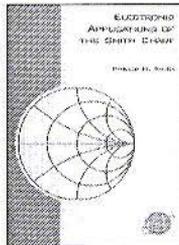


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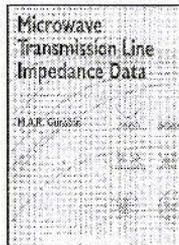
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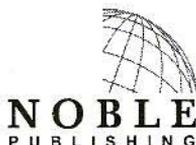
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More good articles this time, I have reduced the line spacing to get more in this time, comments on readability welcome - Andy



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Freddy de Guchteneire ON6UG

Microwave Multi Band Feed Second Generation

The multi-band feed is used primarily for satellite communications and covers the 23 cm band (L), 13 cm band (S), 6 cm band (C) and the 3 cm (X) amateur radio bands. It is possible to extend the range of frequency bands in the dish to 6.

Duplex operation is mandatory for satellite communications. The feed can be used for terrestrial communications by removing the capacitive stubs and changing the feed point or by accepting the less than 3 dB penalty in gain.

The feed is to be used with a primary focus dish having an F/D in the range of 0.4

The antenna is particularly suitable for all the P3D microwave bands.

1.

Introduction

A Multi-band feed was first described in 1996 (1) as part of a lecture during the Amsat-UK Colloquium. A description and practical dimensions can be found in (2).

Use of shorting stubs allowed circular

polarization and 4 bands were covered (L,S,C and X)

However isolation between bands was poor and the mechanical construction was critical.

Further developments made during the following years allowed increased performance and better isolation between each band. Instead of using inductive polarization stubs capacitive stubs were used for better matching and circular polarization.

The resulting feed is more complex and needs careful tuning to achieve optimum matching and polarization. (Fig 1) Matching is better than 20 dB (return loss) and isolation is depending on the bands between 30 and more than 70 dB. For some bands an additional simple filter may be required to isolate and protect receivers and transmitters.

The feed is used with a 1m dish for satellite and terrestrial communications. The dish is homemade on the principle I used for construction of a 2m dish (3). The accuracy of the dish is improved to be better than 1.5 mm and the mesh is 1mm aluminum mesh used for door and window screens. An aluminum ring around the dish increases its stability. The resulting dish is a modified version of the original concept and described in

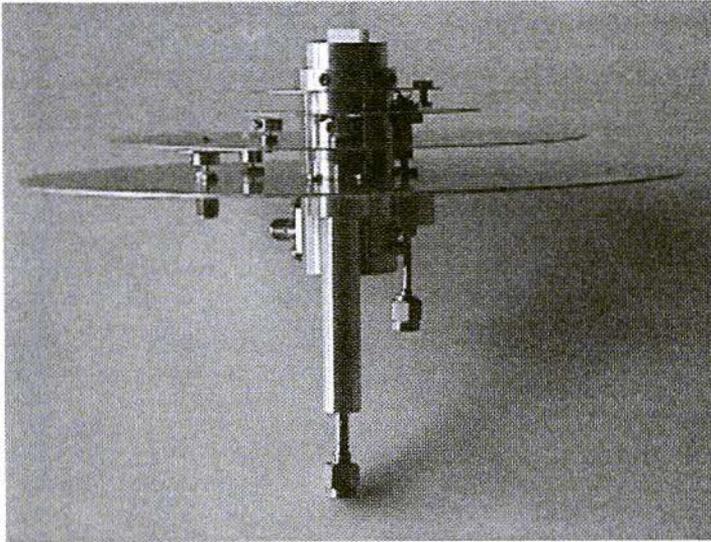


Fig 1 Microwave Multi Band Feed - Second Generation

(4)

A simple single loop extends the range into UHF making the dish antenna capable of using 5 bands. (fig 2) For UHF-band satellite communications the linear polarization does not have serious operating implications. In a normal situation a linear antenna used for circular satellite communication does not have any consequences other than a 3 dB penalty. The effect of the so called

spin modulation is not increased with using a linear antenna. Instead sometimes a linear antenna has advantages when used with circular polarized satellite signals especially for near horizon parts of the orbit.

The dish has been successfully used for portable DX operation on the 23,13,6 and 3 cm bands using the 70 cm capability for talk-back.

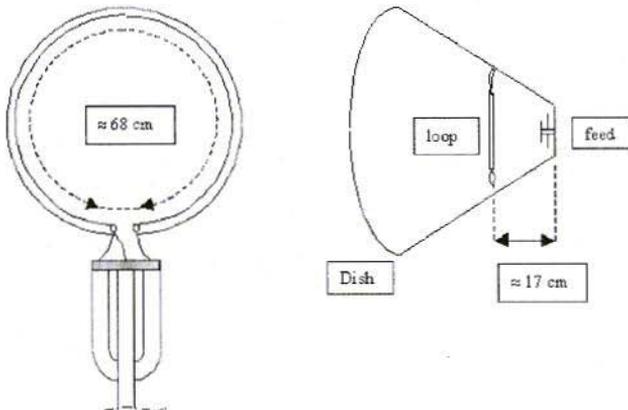


Fig 2

Construction an arrangement of additional loop antenna for 70 cms band

Recently the dish antenna was used on 24 GHz with an add-on feed that simply inserts into the X-band wave guide feed portion. A modification of the design of the X-band part is possible to allow 24 GHz operation in full duplex with other bands. 24 GHz operation is linear polarized.

For L,S and C-band patch antennas with air dielectric are used and it is recommended to read more about patch antennas and the way they work in the literature.

Simplified the patch antenna and polarization is represented in fig 3

Note that the delay section is not only the physical separation between the feed point and the polarization capacitor but includes also the distance to the ground of that capacitor.

The dimension of the patch itself is not critical as long as the feed point is matched and the circularity capacitor is located at the right spot. In essence the patch antenna is considered as a cavity and the radiating part is compared with a double slot antenna.

With the P3D satellite approaching launch the Multi-band feed may provide full operation on all microwave and UHF transponders. A simple VHF antenna is all that is required to be QRV on all transponder bands available in the P3D satellite. The portability of the small antenna system allows operation from almost any place (balcony) with a view to the satellite orbit.

With a 1m dish the gain on S,X and K band is sufficient to allow the use of moderate to high noise figure receive converters while using low power on the up-link.

No expensive high performance converters and power amplifiers will be needed to work with the P3D receivers and transmitters.

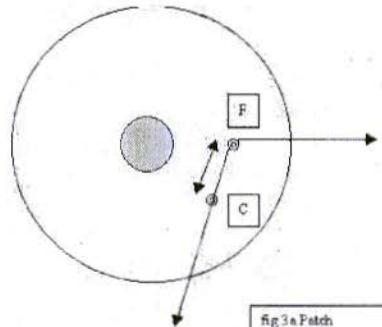


Fig 3a Patch

F = Feed capacitor
C = Polarization capacitor

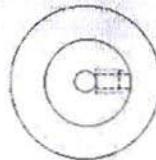
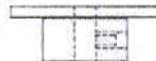


Fig 3b Feed Point and Polarization point Capacitor

Fig 3 Patch Antenna and Polarisation

2.

The X-band feed

This is a standard single mode circular wave guide linear feed. Circular polarization is achieved by inserting a piece of Teflon material (about 4 x 20 x 40 mm) in the right position. Both LHCP and RHCP polarization are possible. For terrestrial operation the Teflon piece is removed.

It is possible to use a circular polarizer with tuning sections instead of using a dielectric to achieve circular polariza-

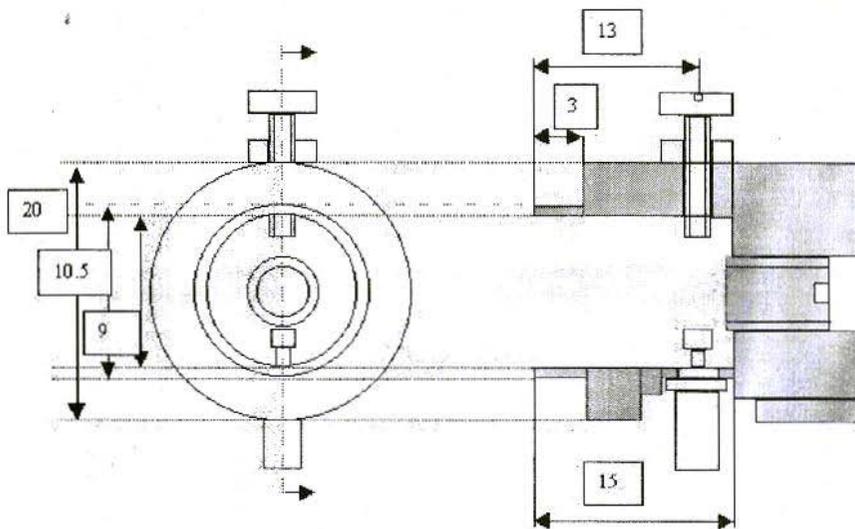


Fig 4 24 GHz Radiator in 10 GHz Feed

tion. With these polarizers better circularity can be achieved at the expense of a more complicated and longer wave guide feed. The use of tuning screws and the long wave guide increase the loss in the feed and so the simpler dielectric was chosen especially as there is no high power used with this feed.

Two matching screws are used to tune the feed for optimum return loss.

No coupling of this feed is to be

expected from other bands. The cut-off frequency of the circular wave guide aids in isolating the radiator from other bands used. For lowest loss a pre amplifier can be coupled directly to the SMA connector and a normal low loss cable or semi rigid cable is used to connect to the receiver. It is possible to use a male SMA connector instead of the normal female to connect directly to a pre amplifier. The SMA connector is the type that has a Teflon insulator extending the connector.

Alternatively a short semi rigid cable can be used to connect the feed directly to a transmit /receive relay and transverter or transceiver. In this case about 1.2 dB loss can be expected from a good cable like Econoflex 143 or equivalent.

The Teflon piece is inserted at 45 degree angle with the feed probe. Do not push the Teflon all the way down to the probe. A better matching is achieved but the loss in the feed due to the Teflon is also increased.

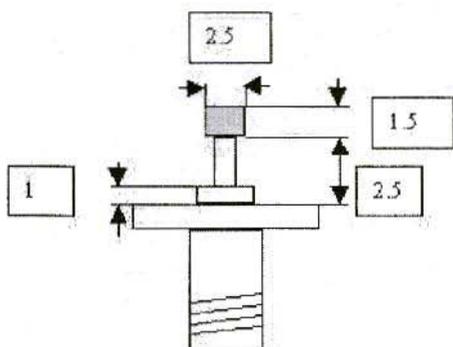


Fig 5 Detail of 24 GHz Radiator

3.

24 GHz operation

If 24 GHz operation is to be included a hole in the back of the X-band wave guide can be drilled to insert the 24 GHz radiator. This principle was first described by DB6NT in his dual band feed. many years ago. With this radiator the 24 GHz radiation pattern is under illuminating the dish. The overall gain of the dish at 24 GHz may be lower but on the other hand pointing is less critical.

The initial 24 GHz feed consist of a small radiator that is inserted in the X-band wave guide feed.

In this case no duplex operation between 24 GHz and 10 GHz is possible. (fig 4 and 5)

Here the dish has some more gain at 24 GHz. For portable operation this last arrangement is preferred.

A 1 m dish has the following specifications :

Frequency	Gain	Beam-width
435 MHz	15 dB	22 degree.
1269 MHz	20 dB	16 degree.
2400 MHz	26 dB	8 degree.
5668 MHz	33 dB	4 degree.
10.45 GHz	38 dB	2 degree.
24.04 GHz	45 dB	1 degree.

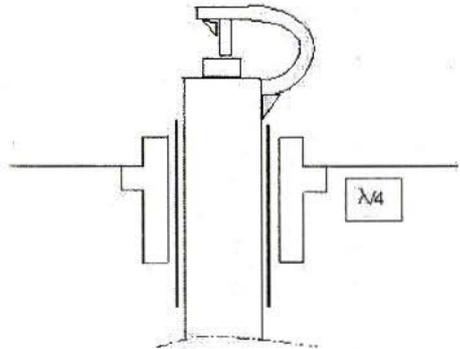


Fig 6 C Band Probe

4.

C-Band feed patch

The C-band feed is somewhat different from the other patch feed arrangements. At this frequency the small dimensions become a problem to mechanically construct the feed point. Instead of using a capacitive coupling an inductive loop is used to fire the patch (fig 6). A small loop at the end of the UT-85 semi rigid probe is located close to the patch. To prevent the loop from touching the patch a small piece of Scotch tape (Scotch

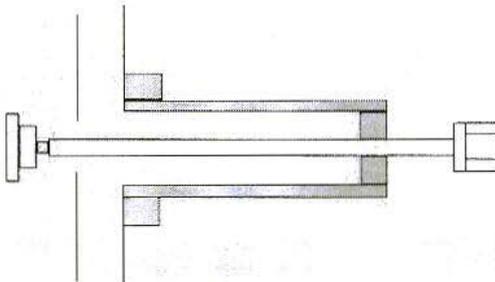


Fig 7
Capacitive
Input
Coupling
Probe

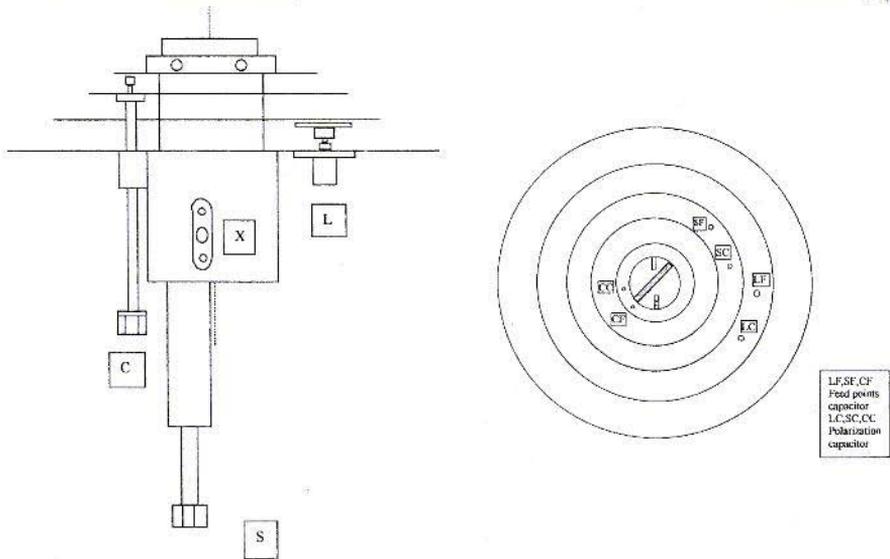


Fig 8 Multi Band Feed

Magic Nr 810) is used. The orientation of the loop is very critical to achieve optimum match and low loss in the feed. This can be easily adjusted with the aid of a signal source and detector. A directional coupler is used to aid in tuning to match the patch to the semi rigid line.

The feed ground is also different. The semi rigid (UT-85) is capacitively grounded at the S-band patch.

To achieve this the semi rigid has 1.5 turns of Scotch tape between the cable and the S-band patch feed through. The so formed $\frac{1}{4}$ wavelength capacitive stub forms a virtual ground at the S-band patch. The matching can be made better then 30 dB and the capacitive coupling increase isolation to more then 26 dB from C to S-band and more then 36 dB from S to C-band

Bandwidth was measured to between 5.5 and 5.7 GHz for 20 dB return loss.

As a consequence the feed can be used for terrestrial C-band operation .

5.

L and S-band Patches

Here a capacitive input coupling probe is used. (fig 7) To achieve a smooth tuning the dielectric of the capacitor is increased with a small piece of plexi-glass or Teflon.

The S-band semi rigid is UT141 and forms a $\frac{1}{4}$ wave stub ground at the L-band patch.

Isolation is 36 dB from L to S-band. The stub increases the isolation from S to L to above 20 dB. This is not enough to isolate the L-band transmitter/receiver from S-band power. A simple low pass filter at L-band can provide for an additional 10-20 dB isolation. (Fig 9) The L-band grounding $\frac{1}{4}$ wave stub further isolates the L-band from the other bands and forms a sleeve balun.

Isolation of the S-band receiver/trans-

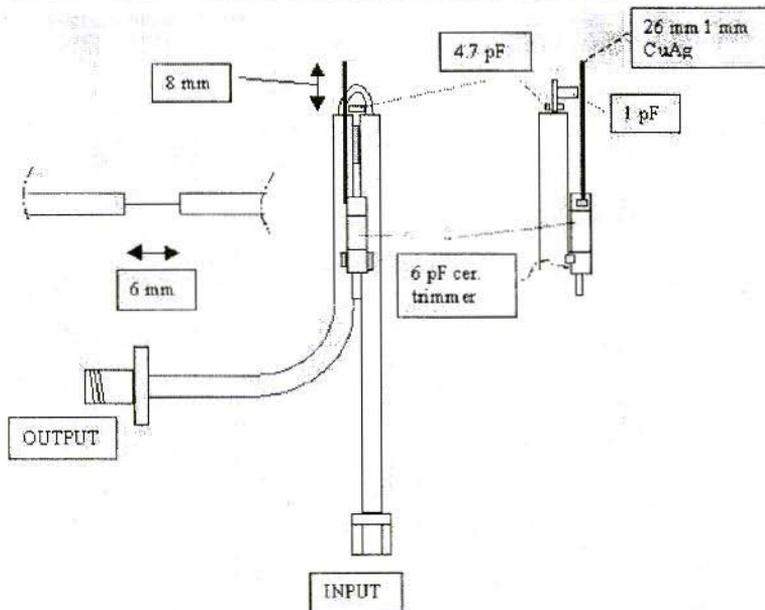


Fig 9 L Band Low Pass Filter UT 141

mitter is in most cases sufficient.

Matching can be tuned to be better than 26 dB.

Isolation from L to C-band is better than 50 dB. No attempt was made to measure the better than 70 dB isolation between L and X-band and C to X-band.

6.

Mechanical construction

The mechanical construction of the feed is complicated and beyond the normal construction facilities of most radio amateurs. The use of a lathe and mill is mandatory.

Also tuning is a matter of experience as both the input and the polarization capacitor are inter-dependent. An antenna range and suitable measuring equipment are mandatory. A network

analyzer is highly recommended.

In order to help radio amateurs in constructing this feed I hope to make this feed available in the near future provided I find a way to get the feed produced, tuned and distributed.

The dimensions given in the drawings are as a rough indication only and can be further optimized.

7.

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- (1) Proceedings of the 11th Amsat-UK Colloquium 1996
- (2) Amsat-DL Journal nr 1 1997
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Matjaz Vidmar, S53MV

A simple TNC for Megabit Packet Radio Links

1.

Computer interfaces for packet-radio

Computers were essential parts of packet-radio equipment right from its beginning more than two decades ago. Since at that time computers were not easily available and were much less capable than today, most amateurs started their activity on packet-radio with an old ASCII terminal. The ASCII terminal required an interface called TNC (Terminal Node Controller).

The TNC interface led to a standardization of the protocol used and to a worldwide acceptance of the AX.25 standard.

Today there are many different interfaces called TNC. The most popular is the TNC2, originally developed by TAPR (Tucson Area Packet Radio) and afterwards cloned elsewhere. Lots of software was written for the TNC2 too, ranging from simple terminal interfaces to complex computer interfaces and even network nodes.

As more powerful computers became available, some functions of the TNC were no longer required. In fact, some

early TNC software, designed to work with dumb ASCII terminals, represented a bottleneck for efficient computer file transfer or multiconnect operation. Most functions of the TNC were therefore transferred to the host computer using the simple KISS protocol, originally developed for TCP/IP operation only. Unfortunately, the KISS protocol adds additional delays in any packet-radio connection.

Today most computers allow a direct steering of a radio modem up to about 10kbit/s, making the TNC completely unnecessary. For higher speeds, different interface cards were developed. These cards are plugged directly into the ISA bus of IBM PC clones to avoid the delays and other problems caused by external interfaces.

The development of new TNCs and related software almost stopped. Although there is lots of software available for the popular TNC2 or its clones, most of the software is of very poor quality. For medium-speed operation at 38400 or 76800bps, even the simple KISS software is unreliable. The popular TNC2 will not go much beyond 76800bps, while more recent TNCs lack software support.

The first megabit PSK transceivers [3], [5], [6] were intended for packet-radio

node interconnects [1]. Their price and complexity was relatively unimportant compared to the reliability requirements and radio-range performance.

For megabit operation, the packet-radio nodes had to be equipped with Direct-Memory Access (DMA) controllers and efficient software. Besides our Slovenian "SuperVozelj" nodes and the similar Italian "Itanet" project, most packet-radio node hardware and software was unable to operate at megabit speeds. In most parts of Europe, the FlexNet network is effectively blocked by the inefficient 9600bps FSK modems.

The Zero-IF technology made simple and efficient megabit PSK radios very affordable even in the form of NO-TUNE projects [2], [4], [7], [8], [9]. The next step was to develop a simple user interface. Marko Kovacevic, S57MMK, modified and improved the popular canadian "PI" packet interface card [10]. The S57MMK card includes a Z8530 SCC serial interface chip, plugs into an ISA slot in a PC-computer bus and uses the PC DMA for fast data transfers. Strictly technically speaking, a DMA interface card represents the best technical solution for megabit packet-radio.

Unfortunately, an ISA card can not be plugged into any computer bus. Some computers even do not have ISA slots, like laptops. The operation of some computers is unreliable on its own even without additional hardware. Poorly-written software and some other peripheral devices may not tolerate a packet-radio interface card on the same bus. An additional interface card, regardless of how well it is designed and built, may cause problems with incompatible hardware and software.

Since new computers are equipped with faster, incompatible and difficult-to-use buses, it is increasingly more difficult to design packet-radio interface cards that plug directly into the computer bus.

TNCs in different forms are therefore coming back, using one of the standard ports available on any computer like RS232, printer port, ethernet connector or even USB. Since any such interface adds additional delays, using a TNC is not the best technical solution.

Unfortunately, a TNC has to be used in all cases when a plug-in interface card on the computer bus can not be used. Therefore suitable TNCs have to be developed. In this article I am going to describe a very simple TNC that connects to the RS232 port on the computer. The TNC allows the operation with the described 1.2Mbit/s PSK radios [2], [4], [7], [8], [9], but on most computers the real bottleneck is the 115.2kbit/s RS232.

2.

Megabit TNC design

Megabit serial interfaces usually require DMA hardware for fast and efficient data transfers. DMA controllers are therefore used in the "SuperVozelj" [1] and in other high-speed packet-radio node projects. Besides adding complexity, a DMA controller also requires a critical handshake with the CPU, since the CPU and the DMA must be synchronized to access the same bus.

The development of high-speed serial interfaces is slowly moving away from dedicated hardware like DMA controllers. For example, the MC68360 communications processor includes four fast serial ports with DMA data transfers that are internally built with a single RISC processor. The actual protocol as well as data transfer to or from the CPU bus is selected by simply changing the software running on the RISC processor.

In the case of amateur packet-radio, most AX.25 frames are relatively short. AX.25 therefore requires lots of



processing for moderate data transfers. The required CPU processing power usually exceeds that of a DMA controller. A detailed analysis shows that a DMA controller may be completely unnecessary even for megabit AX.25 terminals.

A simple solution was therefore sought for a megabit TNC, including the selection of easy-to-get components. A careful analysis of the instruction set and required clock cycles of the popular 68000 microprocessor family has shown that megabit transfers from or to the well-known Z8530 serial-communications controller can be handled by interrupts, provided that a few registers are dedicated permanently to the interrupt-service routines.

For the first experiments I modified an old "SuperVozelj" CPU card with a MC68010 processor. The experiments have shown that for simplex 1.2288Mbit/s operation with PSK radios [2], [4], [7], [8], [9], a minimum clock speed of about 15MHz is required, in good agreement with the estimates of the interrupt-routine clock cycles. A completely new TNC CPU board was then designed around up-to-date CMOS parts like the MC68HC000 CPU and Z85C30 SCC.

The wiring diagram of the CPU, memory and serial port is shown on Fig.1. Just like the "SuperVozelj" node, the 16-bit CPU in the TNC boots from a single, cheap and relatively slow 8-bit 27C256 EPROM. The 16-bit instructions are read from the 8-bit EPROM in two steps, while the upper 8-bits of the instruction are temporarily stored in the 74HC374 latch. The content of the EPROM is copied to the 16-bit-wide RAM immediately after reset to allow fast instruction fetching without any wait states.

The Z85C30 serial-communications controller is connected to the lower 8 bits of the data bus. Simple and fast interrupt routines are triggered by the two out-

puts: /REQA (radio channel to INT3) and /REQB (RS232 interface to INT1). The latter are usually used for DMA transfers. The /INT output of the Z83C30 is not used, since the latter includes many different interrupt sources, unnecessarily slowing down the interrupt-service routine or in the worst case delaying an important high-speed data transfer because of an unimportant status interrupt.

Since the /INT output of the Z85C30 is not used, the interrupt vector for the MC68HC000 has to be provided externally. The MC68HC000 is able to generate interrupt "autovectors" on its own by asserting /VPA during an interrupt acknowledge, however this may add up to 15 additional wait states for every interrupt! The interrupt vector is therefore provided by the 74HC244 in exactly the same way as the /VPA, but without additional wait states.

The MC68HC000 microprocessor, different memory chips and Z85C30 SCC require some bus interface logic as shown on Fig.2.

To avoid metastable problems, all clocks are derived from a single crystal oscillator. The crystal oscillator supplies the MC68HC000 clock and wait-state generator directly. The same clock is divided by 2 by the first half of the 74HC74 to obtain the Z85C30 PCLK. The 74HC138 is used to decode the upper address lines. The 2N2369 transistor is used to protect the CMOS-RAM content while the +5V power supply is turned off and during reset. Of course, the CMOS RAM receives a continuous supply voltage +CMOS at all times from a backup battery.

The /RESET signal is first buffered (1/4 74HC08) and is supplied to the microprocessor (pins /RESET and /HALT) and Z85C30 (both /SCCWR and /SCCRD low at the same time).

Unlike the more powerful members of the MC68k family, the MC68HC000 does not have a vector-base register

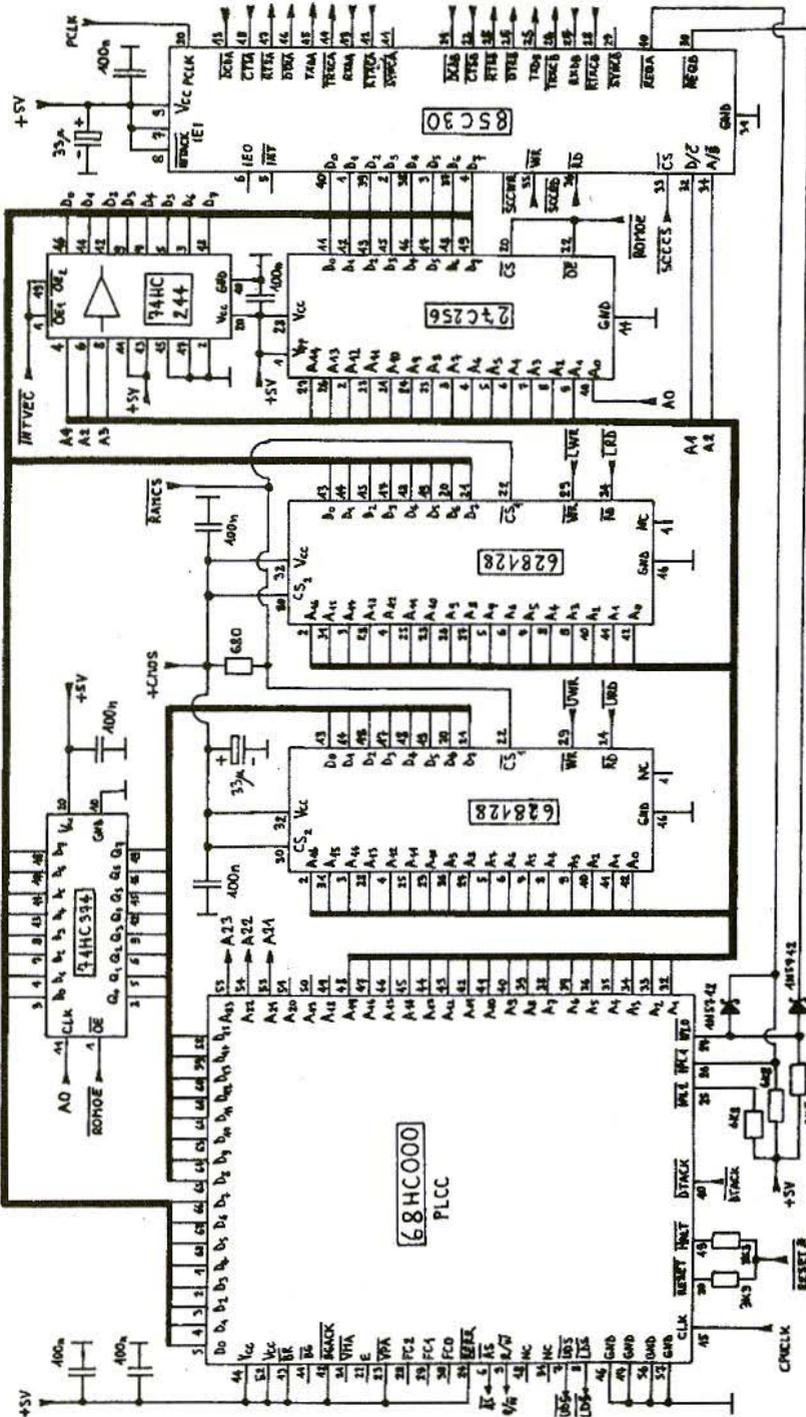


Fig 1 CPU Memory and Serial Port

/UWR, /LWR, /UWR and /LRD control the access to the RAM, the signals /SCCWR and /SCCRD steer the Z85C30 and the signal /ROMOE enables the EPROM.

3.

Practical TNC construction

In the practical construction I decided to build the TNC as a series of modules that can be connected together as required. The power supply, radio modem and RS232 interfaces are therefore built as separate modules on their own printed-circuit boards. The TNC printed-circuit board therefore only includes the CPU, memory and serial port. In this way the same TNC board can also be used for other purposes, for example with different modems or as a simple two-channel packet-radio node.

The CPU, memory, serial port and bus interface logic are located on a double-sided printed circuit board with the dimensions of 120mmX120mm. Both sides of the printed-circuit board are shown on Fig.3. The corresponding component location is shown on Fig.4. The module only has two connectors.

A four-pin connector is used for the power supply: main +5V supply, battery-backed +CMOS, /RESET signal and ground.

Both channels of the Z85C30 serial port are available on a 20-pin connector together with +5V, +CMOS and ground. Besides serial data inputs and outputs, the control lines /DCD, /CTS, /RTS and /DTR as well as both clocks RTXC and TRXC are available for both channels. These allow to connect the serial port to different modems and/or other interfaces.

The MC68HC000 microprocessor is usually available in a 68-pin PLCC

package and is generally used in GPS receivers, fast hard-disk controllers and other applications where the the processing power of a simple 8-bit microcontroller is not sufficient. Since the soldering of a PLCC package is not a simple operation, the printed-circuit board is designed for a standard PLCC socket that spreads out the PLCC pins to a comfortable 2.54mm (0.1") grid. The pins of the 68-pin PLCC socket are arranged in two rows on each of the four sides of the square socket. The socket pinout is therefore somewhat different from the bare PLCC package as shown on Fig.5. The MC68HC000 microprocessor has two unused pins labelled "NC" in the 68-pin PLCC version.

The microprocessor should be carefully pushed in the PLCC socket. On the other hand, a special tool is required to extract the microprocessor from the socket. Special tools can be avoided if the PLCC socket is further inserted in a 68-pin PGA socket (large version 11X11) that is finally soldered into the printed-circuit board.

While building the megabit TNC, it makes sense to install all integrated circuits on good-quality sockets. The TNC will work with 128kbyte RAM chips (628128) or with 32kbyte RAM chips (62256). In the latter case, the 28-pin 32kbyte RAM chips should be inserted in the bottom part of the 32-pin sockets so that pin 1 of the RAM corresponds to pin 3 of the socket. The TNC will also work with 32kbyte, 64kbyte or 128kbyte "cache" RAM chips from old "486" motherboards. Since these RAMs are packaged in narrower packages, their pins should be spread out or a suitable adapter should be built to fit in the wider standard RAM socket.

The relatively short bus connections allow high clock speeds for the TNC CPU. Experiments have shown that the maximum clock speed does not depend much on the CPU used, but rather on

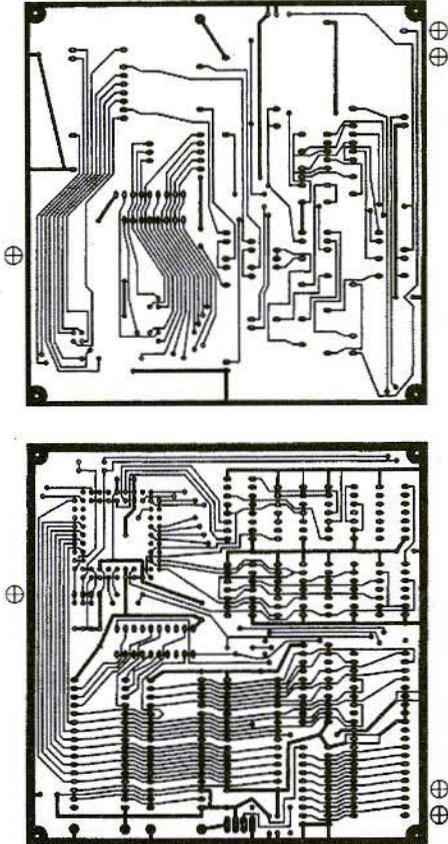


Fig 3 TNC Double Sided Printed Circuit Board

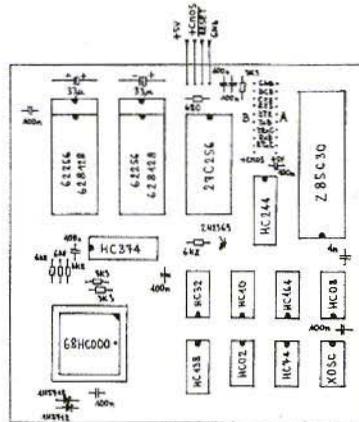


Fig 4 TNC Component Location

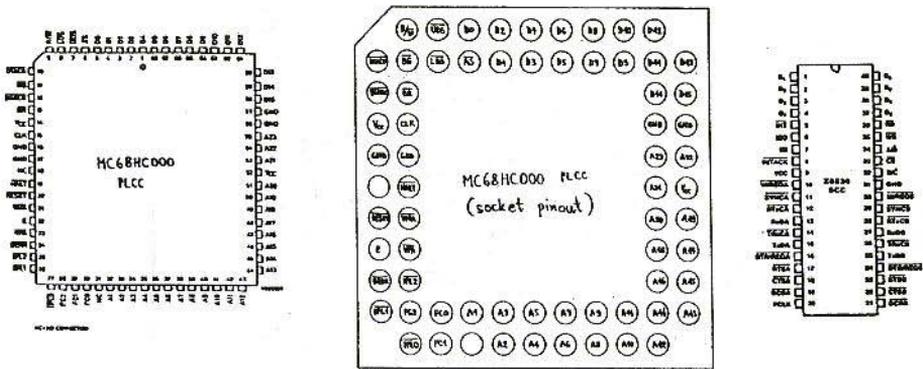


Fig 5 CPU and Serial Port Pinouts

the RAM access time. A 10MHz version of the MC68HC000 will work with 70ns RAMs up to 33MHz, while a 16MHz version of the MC68HC000 will reach 40MHz with 20ns "cache" RAMs. The CMOS version of the SCC chip Z85C30 also allows much higher clock frequencies than specified: a 8MHz version usually works perfectly at 15MHz PCLK (CPU at 30MHz).

Since all tested combinations of CPUs and memories always achieved a maximum clock frequency of at least 25MHz, a CPU clock of at least 20MHz is recommended for the megabit TNC. The exact clock frequency also depends on the useable division modulus inside the Z85C30 to obtain standard data rates. If the 1.2288MHz clock on the radio side comes from the scrambler module [2], [4], [8], [9], then the Z85C30 dividers only supply the RS232 baudrate.

A RS232 baudrate of 115.2kbit/s can be obtained from a CPU clock frequency of 14.7456MHz, 22.1184MHz or 29.4912MHz. Although the TNC may work at 14.7MHz, this clock is too low for reliable operation at 1.2288Mbit/s on

the radio side. Therefore 22.1MHz or 29.4MHz are recommended. Of course, any clock frequency can be used for the CPU if an external clock of 1.8432MHz is supplied to RTXCB. While using the TNC for lower kilobit data rates on the radio port, there is much more freedom in choosing the clock frequency.

4.

Radio and RS232 interfaces

The megabit TNC is designed to work with different modems. The most common combination is a scrambler/bit-synchronizer on channel A and a RS232 interface on channel B of the serial-communications controller Z85C30.

The scrambler/bit-synchronizer [2], [4], [8], [9] includes many modem functions like clock recovery and descrambling for the received data and clock generation and data scrambling for the transmitted data. The scrambler is connected to the Z85C30 channel A with five signal wires plus ground. The signals



include: received data (RXD), clock (connected to RTXC), carrier detect (/DCD), PTT command (/RTS) and transmitted data (/TXD). In this case the input /CTS, output /DTR and clock TRXC are not used on channel A. An unused CMOS input should be connected to ground or to any other signal.

Some other modems for lower kilobit rates may require internal clock recovery inside the Z85C30. In this case, the modem is connected with just four wires besides ground: received data (RXD), carrier detect (/DCD), PTT command (/RTS) and transmitted data (TXD). The TNC software programs the Z85C30 to supply the regenerated clock from the internal DPLL on the TRXC output. An external jumper is required to bring this clock to RTXC. Of course, the correct divider modulo should be set by the software.

If more than one modem is installed in the TNC, then the channel A inputs RXD, /DCD and RTXC have to be switched among the different modems and/or other clock sources. All of the modems should receive their supply voltage at all times so that it is not necessary to switch the channel A outputs /RTS and TXD.

A TNC is usually connected to a host computer through a RS232 interface. Unfortunately, the latter represents a real bottleneck for a megabit TNC, since the highest RS232 baudrate is 115.2kbit/s on most computers. The RS232 interface also includes a polarity inversion on all signals and a logic-level shift to higher positive and negative voltages.

A very simple RS232 interface can be built with the integrated circuit MAX232 as shown on Fig.6. The MAX232 includes two RS232 transmitters, two RS232 receivers and two DC/DC converters to obtain +10V and -10V from a single +5V supply. Unfortunately, the MAX232 also includes lowpass filters on both transmitters to limit radio interference. These lowpass

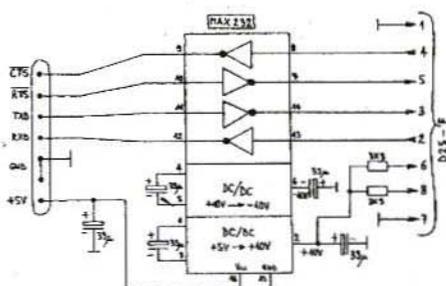


Fig 6 RS232 Interface

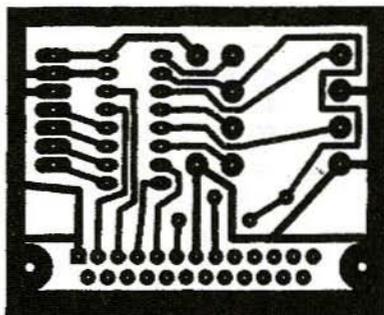


Fig 7 Rs232 Printed Cicuit Board

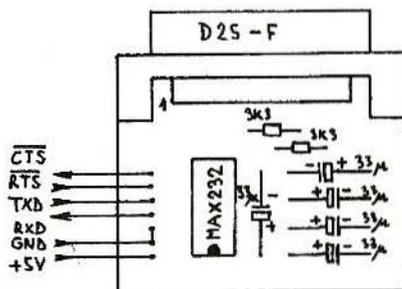


Fig 8 RS232 Component Location

filters also limit the available data rate to about 150kbit/s.

The RS232 interface is built on a small, single-sided printed-circuit board with the dimensions of 54mmX44mm, as shown on Fig.7. The printed-circuit board is supported by an angled female D52 connector so that additional mounting screws are not required. The corresponding component location is shown on Fig.8. The female D25 connector is wired to fit the male RS232 D25 connector on PC COM ports.

While using the RS232 interface on channel B of the Z85C30, the /DCD input and /DTR output remain unused. An unused CMOS input should be connected to ground or to any other signal. If the internal clock source in the Z85C30 is used, then a jumper from TRXC to RTXC is required. An external clock source can also be connected to the RTXC input. In the latter case, the external clock frequency should equal 16-times the desired baudrate or 1.8432MHz for 115.2kbit/s.

5.

Power supply and reset circuit

The described TNC is intended to be operated from a 12V power supply with the negative grounded like most radio-amateur equipment. Of course, the digital circuits of the TNC require a +5V supply. If any program parameters are stored in the CMOS RAM, then the latter requires a battery-backed supply +CMOS. Finally, the TNC requires a reset circuit that operates reliably regardless of the power-up or power-down sequences.

The circuit diagram of the power supply and reset circuit is shown on Fig.9. A simple switching regulator is used to obtain +5V from the available +12V external supply. The switching regulator

achieves an efficiency of about 80%, thus halving the power consumption and heat generation when compared to a conventional linear regulator. The +CMOS supply is backed by a 3.6V 60mAh NiCd battery. An additional protection of the CMOS RAM content is provided by the /RESET signal by disabling the RAM chip select.

The reset circuit senses the input voltage to the power supply. The /RESET signal is only released after the input voltage exceeds 9.5V. An additional delay is generated by the 1.2kohmX470uF RC time constant. At power down, the 470uF capacitor is discharged immediately through the 1N4001 diode, so that the /RESET signal is applied before the switching regulator stops operating. Finally, the reset circuit includes a small hysteresis (47kohm resistor) to generate a clean /RESET signal on its output.

The TNC power supply is built on a single-sided printed-circuit board with the dimensions of 75mmX45mm, as shown on Fig.10. The corresponding component location is shown on Fig.11. Several mounting pads are provided for different-style NiCd batteries.

The total current drain of the described TNC equipped with a scrambler/bit-synchronizer and RS232 interface amounts to about 200mA at +5V or about 100mA at +12V with the described power supply. At these current levels, the BD138 switching transistor does not require a heatsink. On the other hand, the 100uH chokes must be able to handle these currents too: chokes of the size of 1/2W resistors or even larger should be used in the power supply.

6.

Megabit TNC software

The TNC software should allow a fast and reliable data transfer between the



are those from NordLink. Unfortunately, no one removed some important bugs from the original WA8DED code that corrupt the data and crash the TNC software or the host computer. Even worse, the same bugs were also transferred to the software that emulates a HOSTMODE TNC on a PC computer (TFPCX, TFKISS etc). For this reason, the HOSTMODE protocol is now almost forgotten.

The only RS232 protocol used by most TNCs seems to be the KISS (Keep It Simple Stupid) protocol. The KISS protocol simply transfers the same, unprocessed AX.25 frames over the RS232 interface. The AX.25 protocol is therefore handled by the host computer except for a few timing functions tightly connected to the radio modem, like receive/transmit changeover.

The KISS protocol transmits the data as 8-bit bytes on an asynchronous serial interface. The beginning and the end of AX.25 frames are marked with FEND (\$C0) characters. When a FEND character appears inside an AX.25 frame, it is replaced by the sequence FESC, TFEND (\$DB, \$DC). The special character FESC also requires a replacement if appearing inside an AX.25 frame. FESC is replaced by FESC, TFESC (\$DB, \$DD). By definition, the KISS protocol does not make any use of the RS232 control lines (CTS, RTS etc).

The KISS frames are equipped with an additional control byte in front of the AX.25 frame. The control byte allows the selection of up to 16 radio ports on a multiport TNC (upper four bits) and setting a few parameters of each radio channel (lower four bits). An AX.25 frame is identified by the lower four bits set to zero, while a value different from zero sets some TNC parameters. An exception is represented by the control byte \$FF that is usually used to switch the TNC (back) to another mode of operation.

Most KISS frames have the first (con-

trol) byte set to zero (\$00), since we only have single-channel TNCs and most KISS frames carry AX.25 frames. Setting the TNC parameters TX delay, TX tail, pperistance and slottime and selecting simplex or duplex operation as well as quitting the KISS mode of operation with special control bytes proved to be a poor solution in practice. Any errors on the RS232 interface may program the TNC in some useless operating mode and crash the AX.25 link.

Although the RS232 link includes just a short length of cable between the TNC and the host computer, transmission data errors occur frequently on the RS232 interface. The main cause is the poor design of PC interrupts. A lower-priority interrupt may stop the interrupt-driven RS232 data transfer leading to loss of data. Serial ports with FIFO registers may save some data, but they are still unable to guarantee an error-free data transmission.

RS232 errors can be tolerated in TCP/IP operation, since all TCPIP frames include their own, internal CRC checksum. Of course, RS232 errors corrupt the data in conventional AX.25 contacts. Special KISS protocols including a two-byte CRC at the end of the frame, just like in the HDLC AX.25 frames, have been developed to detect and automatically reject corrupted frames.

Of course, both the TNC software and the host-computer software should detect automatically what kind of KISS protocol is being used: the last two bytes in a frame may be a 16-bit CRC computed in different ways or simply two valid data bytes. SMACK (Stuttgarts Modifiziertes Amaturfunk-CRC-KISS) is a KISS protocol with two CRC bytes at the end of each frame. The SMACK frames are marked with a \$80 control byte at the beginning of each frame, since there are no TNCs available with more than 8 radio ports.

SMACK is using the same CRC poly-



mial-divison algorithm as conventional AX.25 HDLC frames, except that the polynomial generator is started from an all-zero condition. The latter is a poor choice and may not detect some types of errors on the RS232 interface.

The FlexNet group corrected this problem by developing their own CRC algorithm. KISS frames with a FlexNet-CRC are marked with a \$20 control byte at the beginning of each frame, since multiport TNCs are no longer being used.

The requirements for the TNC software are exactly defined: the TNC should communicate in the conventional-KISS protocol without CRC as well as in both CRC versions, SMACK and FlexNet-CRC. All three KISS protocols are being used by almost all available packet-radio software running on host computers. The critical timing and operational parameters of the TNC are best burned once forever in the TNC EPROM rather than being modified by RS232 errors in an uncontrolled way.

The megabit TNC has some additional requirements. Since the radio interface is an order of magnitude faster than the maximum RS232 speed, the TNC should check the callsigns of all frames and reject useless frames addressed to other stations. An even better solution is a small packet-network node to handle the retries on both the radio and RS232 sides.

The described TNC was first tested with simple KISS software including both CRC variants. Besides the currently-available KISS/SMACK/FlexNet software, many other software upgrades are planned for the described TNC, including a small packet-radio node similar to the SuperVozelj nodes.

All software is available as documented 68k ASM source as well as compiled EPROM files on our packet-radio mailbox S50BOX and the VHF Communications web site.

The current KISS software supports a simple interface to adjust the important program parameters like filter callsign, TX delay, TX tail, pperistence and KISS protocol (no-CRC, SMACK or FlexNet). These parameters are simply typed in the "unproto" mode and transmitted as UI (beacon) frames to the callsign "TNC". The TNC will answer with UI frames too, showing the current parameter settings. This simple protocol resulted very reliable and at the same time accessible at any time from any packet-radio terminal program.

The current software version copies the default parameters from the EPROM after every reset and does not make any use of the CMOS battery. However, future versions will probably use the nonvolatile CMOS RAM at least to store the program parameters and callsign.

There are also a few restrictions imposed by the hardware simplicity of the described megabit TNC. Three address registers A4, A5, A6 and two data registers D6 and D7 are used at all times by the high-speed interrupt routines, so they can not be used by other TNC software. The radio-port interrupts are designed for simplex operation only.

7.

Megabit TNC applications

The described megabit TNC was initially intended to prove that a simple circuit can perform much better than complex and expensive hardware. In particular, complicated DMA circuits and other interfaces are probably not required for high-speed packet-radio. The described megabit TNC probably shows the evolution of future packet-radio hardware: most problems will be solved with standard, inexpensive parts programmed for our purposes. Practical experiments with the described TNC

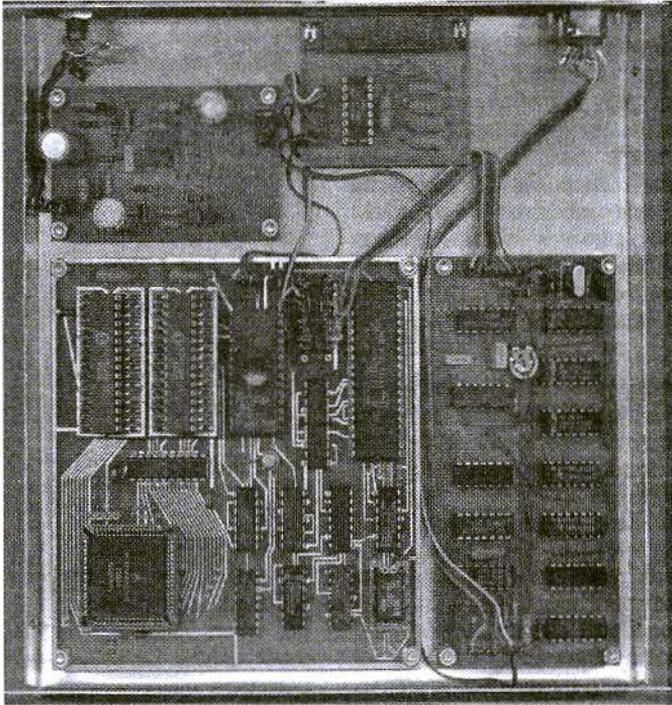


Fig 12 The Completed TNC

equipped with simple KISS software and FlexNet driver running on the host computer have shown an average data throughput of about 25kbit/s or 3kbyte/s in a real network with many other users active on the same channel at the same time. A small-node software in the TNC to handle separately the retries on the radio side as well as on the RS232 side could probably double this figure before reaching the 115.2kbit/s RS232 bottleneck.

Some improvement could also be obtained with a better driver than FlexNet, since the timing constants of the latter are programmed for 9600bit/s modems and can not be changed.

For the packet-radio user, a megabit TNC on the RS232 interface is certainly a slower solution than a DMA card in the computer bus, mainly thanks to the

115.2kbit/s RS232 bottleneck. The megabit TNC is therefore an interesting solution only when a DMA card can not be used, like laptop computers, or due to conflicts with other hardware and/or software in the same computer. Of course, a megabit TNC equipped with an ethernet, USB or even parallel printer-port interface could perform much better.

For the network developer or sysop, the megabit TNC may have advantages too. Most important, one single user can never get all of the capacity of a megabit channel thanks to his/her RS232 bottleneck, thus blocking the access to other users. A megabit TNC also allows testing new protocols in the network without changing the actual user terminal software running on host computers.

A megabit TNC also represents an ideal



interface to older node hardware, like the RMNC/FlexNet nodes used elsewhere in Europe. The efficient megabit PSK transceivers, although described in detail in several places: [2], [3], [4], [5], [6], [7], [8], are not widely used outside Slovenia and Italy, probably just because popular packet-radio node hardware can not operate at megabit speeds directly, without a megabit TNC interface. Finally, the described megabit TNC equipped with CMOS parts can also work as a simple, very low-power digipeater already with the currently-available KISS software. The overall current consumption of the TNC is about 100mA at 12V. Adding about 200mA for the 23cm PSK transceiver, the whole digipeater can be powered with a 50W solar panel and a large "diesel" car battery.

8.

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- [5] Matjaz Vidmar: "13cm PSK Transceiver for 1.2Mbit/s Packet Radio, Part-1", VHF-Communications 3/1996, pages 130-147.
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Harald Fleckner, DC 8 UG

Teflon, Epoxy or RO 4000 Base Material ?

A supplement to the article in issue 2/2000

Three low-noise pre-amplifiers with the same structure in a comparison test, with source degeneration for 1,200 to 1,400 MHz.

1. Introduction

RO 4000 substrates belong to the group of Teflon-free carrier materials (duroplastics), which are made use of for mass-produced applications in the low cost sector (telecommunications / mobile radio-telephones).

This type of polymer substrate is fibreglass-reinforced (type 1080) and ceramic-filled.

The new substrate combines good mechanical properties (comparable with epoxy fibreglass laminate FR-4) with very good high-frequency properties and is also available at a considerably more favourable price than RT/duroid 5870.

The Mauritz company kindly made some specimens of Rogers RO 4003 substrate available to the author, so that he could make some comparison tests.

Some details of the mechanical and electrical values of this new microwave-

base material:

Dielectric constant ϵ_r : $3.38 \pm 0,05$

Substrate thicknesses : 0.2 3,15 mm

Copper layer : 35 or 17,5 μm

Dielectric loss factor : 0.0027

(10 GHz / 23°C)

(RT/duroid 5870: 0.0009,FR-4 : 0.035)

Comparison values for insertion loss for the same microstrip circuits on FR-4, PTFE and RO 4003 with a substrate thickness of 0.81 mm.:

Specifications from ROGERS at $f = 10$ GHz [1][2]:

FR-4	:-0,8 dB /inch
PTFE (Teflon)	:-0,1 dB /inch
RO 4003	:-0,2 dB /inch

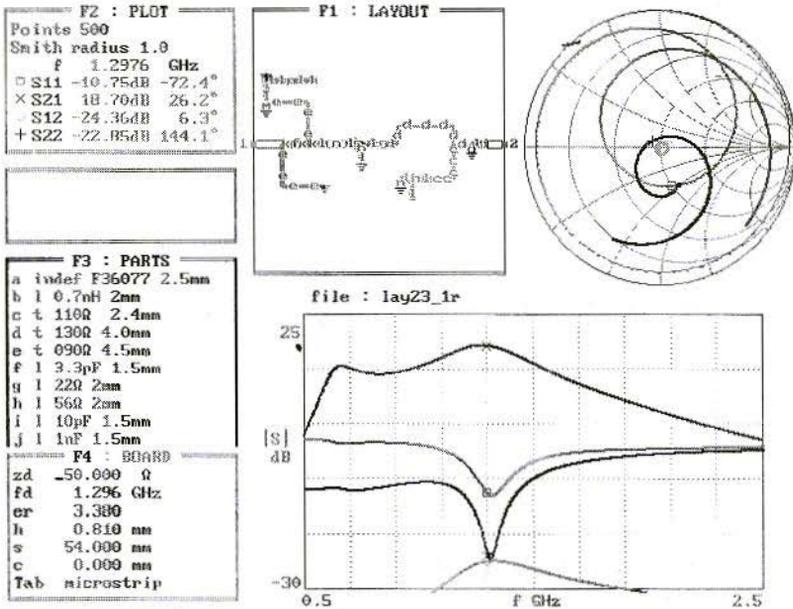


Fig 1 Puff Simulation of the Test Amplifier

2. Readings

The third comparison amplifier was constructed on RO 4003 with a substrate thickness of 0.81 mm. and simulated using PUFF. Fig. 1 shows the results of the simulation and Fig. 2 the associated layout.

The noise factor, amplification and return loss were then determined in the circuit input and output, under the same measurement conditions.

All the readings are illustrated for comparison in Figs. 3, 4, 5 and 6.

As expected, the RO 4003 base material is absolutely evenly matched with the Teflon material RT/duroid in this frequency range. (Fig. 3)

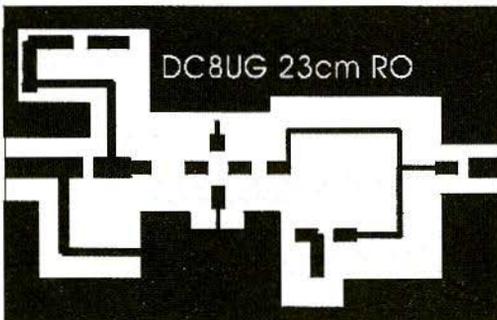
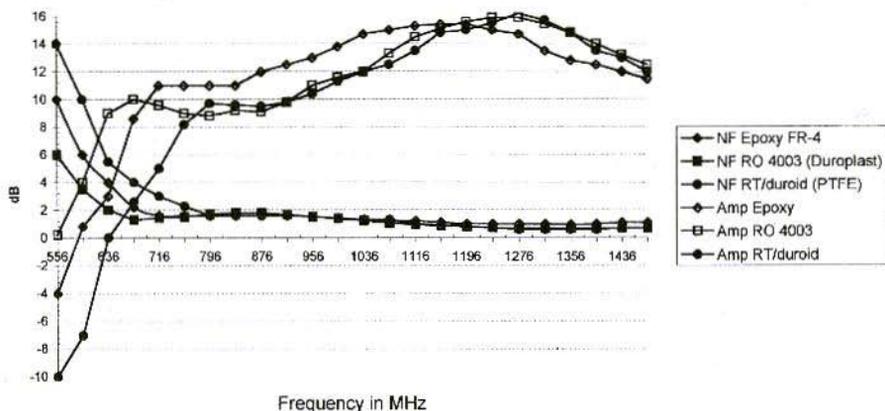


Fig 2 Printed Circuit Board Layout for the Test Amplifier



Fig 3 Comparison of Readings for Amplification and Noise Factor



The noise factor obtained for two assemblies tested, using the GaAs-FET ATF 36077, was 0.7 dB (see Fig. 4) and the amplification was exactly 16 dB.

The return loss values in the output and input essentially corresponded to the readings from the Teflon assembly. (Fig. 5 and Fig. 6)

3. Summary

The RO4003 base material is a splendid alternative to the more expensive Teflon base material. The electrical properties below 18 GHz are not significantly different from those for the RT/duroid. Because of the rigidity of the material, the mechanical properties are as good as

Fig 4 Detail View of Noise Figure Comparison

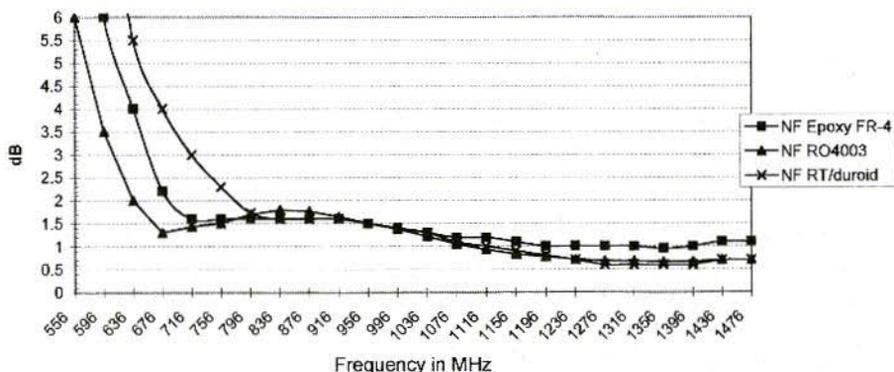
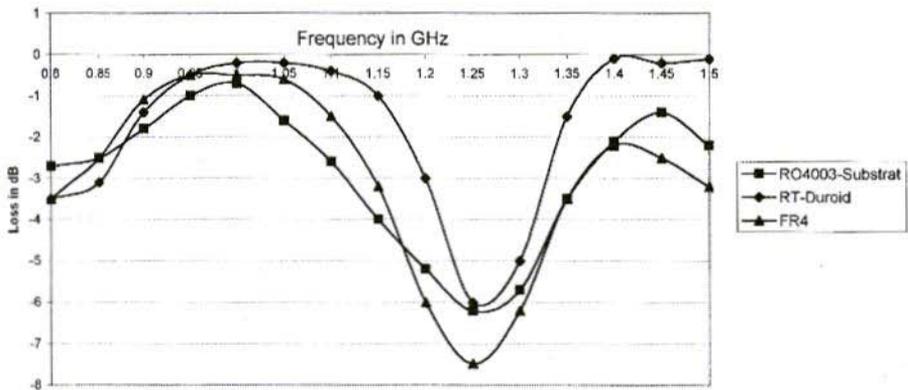




Fig 5 Comparison of Input Return Loss Values



those for epoxy fibreglass laminates, even with thin substrates, i.e. it is very easy to drill and throughplate and, in contrast to Teflon fibreglass material, it keeps its shape and is temperature-stable to over 280 °C .

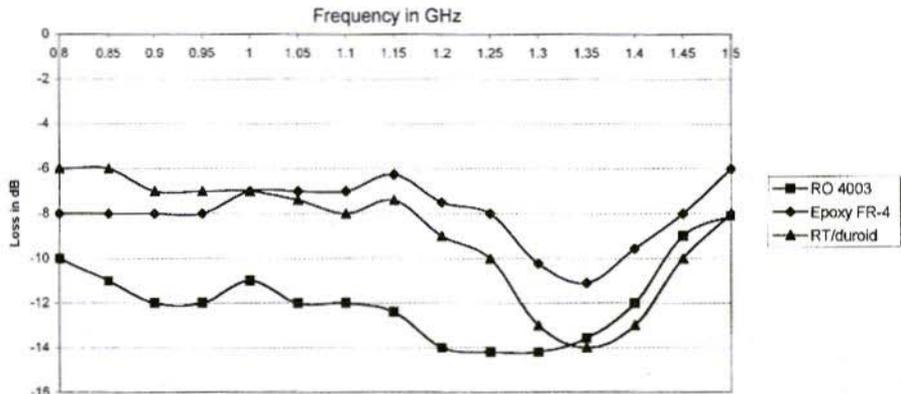
The price for this material is about 40% lower than the price for comparable material made from RT/duroid, and some of the companies involved have recently also been offering it in small board sizes.

4. Literature

[1] Rogers Corporation, DATA RO 1.4000, RO4000 Series, High Frequency Circuit Materials, www.rogers-corp.com/mwu

[2] Mauritz GmbH + CO, ROGERS RO 4000 Base material system for high-frequency multi-layer, www.mauritz-hamburg.de

Fig 6 Comparison of Output Return Loss Values





Gerhard Schmitt, DJ 5 AP

Low Pass Filter for 2 m and 70 cm

Low-pass filters are described which are easy to assemble and fulfil both the demands of the licensing authority and the self-imposed requirements to reduce harmonics coming from 2 m and 70 cm transmitters. This subject appears particularly topical in an age in which the problems of electromagnetic compatibility are consciously being addressed.

1. Introduction

The authors motivation for building his own equipment or aials springs mainly from his activities in connection with contests. A few years ago, for example, in the telegraphy competition (Marconi contest), during the 70 cm contest (which was still in existence at that time), several stations were heard in the CW range of the 70 cm band, which did not respond to targetted calls. It soon became clear that these stations were not really transmitting in the 70 cm band, but in the 2 m band, and we were receiving only their harmonics. As some of the stations in question were over 100 km away, it is easy to imagine what the harmonic suppression arrangements were like on the 2 m transmitter being

used there.

Following this experience, the 2 m high-level stage at our club station (DK0PX on the Kochelsberg/Schwäbische Alb) was topped by a low-pass filter as a precaution.

A little later, problems again occurred when simultaneous contests were taking place in the 70 cm and 23 cm bands. When transmissions were taking place at 70 cm, the receiver of the 23 cm station was completely jammed. It became clear that the 70 cm high-level stage, which had 2 x 2C39B tubes, was producing too many harmonics at 23 cm, which could be eliminated by the introduction of a low-pass filter.

2. Requirements for harmonic filter

- Adequate harmonic suppression better than 40 dB up to at least the 5th harmonic. The remaining 20 dB (to arrive at the 60 dB required by the licensing authority) could certainly be assumed to be present in the operating transmitter.

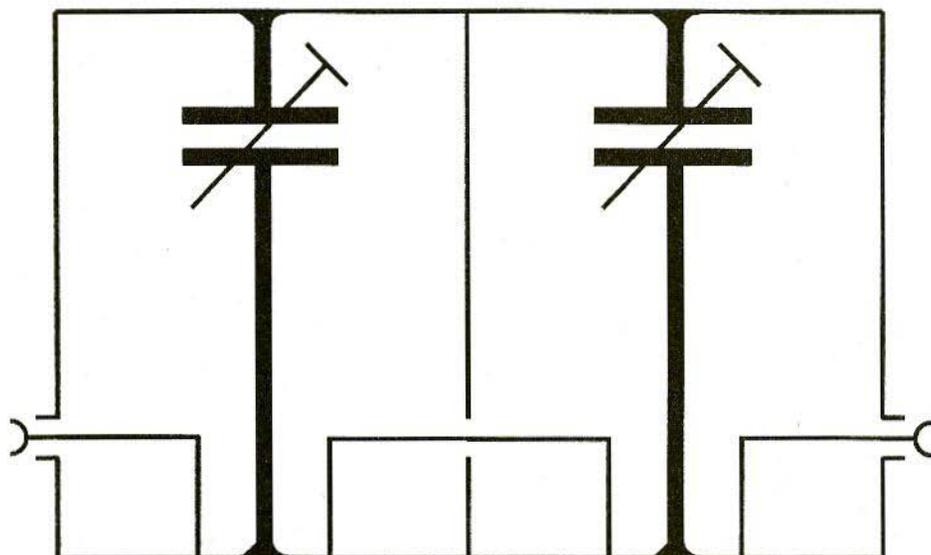


Fig 1 The Basic Band Pass Filter

- No attenuation troughs, at least not up to the 5th harmonic wave.
- Good SWR, below 1:1,2;
- Low insertion attenuation of under 0,5 dB on the operating frequency;
- Transmittable output matched to individual transmitter output; optimal up to 750 W;
- Simple assembly without expensive special components;
- Smoothing should be possible using means available to amateurs;
- No excessive dimensions.

3.

Band-pass or low-pass filter?

Both-band pass and low-pass filters can

be used for harmonic suppression.

3.1. Band-pass filter

These consist of single-circuit and multi-circuit resonant circuits, tuned to the individual operating frequency ($\lambda/4$ circuits) (Fig. 1). Only the operating frequency can pass through. All other frequencies below and above it are suppressed. Unfortunately, the filtering of undesirable frequencies above the operating frequency is usually inadequate as a result of undesirable resonances.

Advantages:

Band pass filters can be made really small for high-frequency outputs of up to approximately 50 Watts.

The use of band-pass filters can also be effective for reception when it is necessary to keep signals from nearby powerful transmitters away from the receiver input.

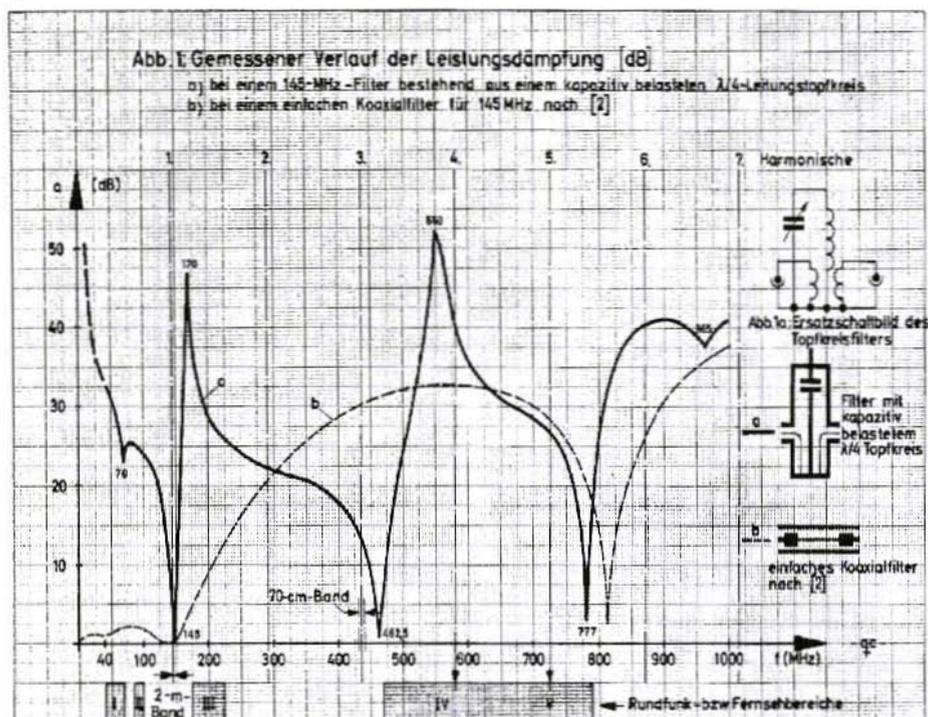


Fig 2 Attenuation of Band Pass Filters

Disadvantages:

Due to their assembly using $\lambda/4$ technology, band-pass filters encounter resonance problems again at $3/4 \lambda$. When a two-circuit band-pass filter assembled by the author was put to work for 435 MHz, the 1296 MHz frequency was suppressed by a mere 6 dB.

These resonance values can certainly be pushed upwards using ingenious constructions; if you're unlucky, you end up with an attenuation trough in the vicinity of the fourth harmonic, instead of on the third, and there, for example at 2150 MHz, another radio service creates interference.

If the maximum permitted power is used, or even at power levels of 200 Watts, voltages in the KV range arise in

the resonant circuit capacitors of a band-pass filter. So, to avoid flashovers, the tuning capacitors, and thus the entire filter, have to be suitably bulky. A two-circuit band-pass filter for the 2-m band can easily exceed the dimensions of a transistor amplifier with an output power of 200 Watts.

For the attenuation gradients of band pass filters, see Fig. 2 from [1]. The band-pass filters from DJ1ZB described in [2] and [3] can be used pretty successfully at high-frequency power levels of up to app. 50 Watts.

3.2. Low-pass filter

These are dimensioned in such a way that all frequencies below their operating frequency are allowed through with scarcely any attenuation. However,

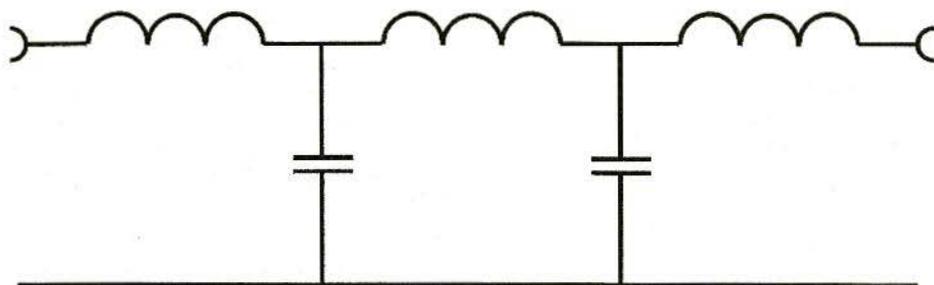


Fig 3 The Basic Low Pass Filter

above the operating frequency the more or less steep rise in attenuation begins. Most low-pass filters are multi-section pi-filters (Chebyshev filters); see Fig.3 for a basic wiring diagram.

Advantages:

If suitably dimensioned, low-pass filters have scarcely any attenuation troughs above their operating frequency.

As indicated here, they can be made really small, even for outputs of up to 750 Watts! The filters for the 2-m and the 70-cm band developed by the author have housing dimensions of 111 mm x 55 mm x 30 mm; for power levels of up to app. 50 Watts, they can be even smaller.

Disadvantages:

As mentioned, all frequencies below the operating frequency are allowed through with scarcely any attenuation. Thus interfering mixed products stemming, for example, from the frequency processing of a transmitter (transceiver, transverter), and lying below the operat-

ing frequency, are also scarcely attenuated by a low-pass filter. On the other hand, with the transceivers / transverters which are commercially available today it can be assumed that interfering spurious have already been largely suppressed during the signal processing, and thus just do not reach the transmitter high-level stages in the first place.

When the advantages and disadvantages of band-pass and low-pass filters had been weighed up, preference was given to the simpler low-pass filters and to their smaller formats.

4.

Low-pass filter for the 2 m band the first experiments

Literature on low-pass filters (LP filters) for power levels up to 750 Watts, or higher, can be found in [4], [6] and [7]. The filter was originally assembled in accordance with [6].

The circuit bears some resemblance to Fig. 3. Due to the high level of HF power to be transmitted - up to 2,8 kW - the capacitors used were not "normal" ceramic wafer types, but short pieces of Teflon-insulated coax cable (RG 142), with about 1 pF/cm.

Thus, for the two capacitors, each providing 18 pF, two pieces of coax cable are required, each being 18 cm length. The insulation is stripped from each of the coax cable pieces 1 cm. from their ends. The internal conductors are soldered together and the braiding is soldered to earth.

The assembled filter was put through its paces by an OM friend in the range from 145 MHz to 500 MHz. The values obtained for the band-pass attenuation and attenuation gradient appeared initially to give cause for euphoria. But this quickly evaporated when the filter was measured in a higher range on another occasion! Above 500 MHz there were sometimes tremendous attenuation troughs which make the filter almost unusable for harmonic suppression depending on the format and the tolerances.

One might object that it makes no difference if the attenuation troughs are outside the harmonics of the operating frequency. However, it is doubtless better and more sensible to construct the low-pass filter in such a way that it registers no attenuation troughs until far into the GHz range!

An attempt was made to push the attenuation troughs up into a higher frequency level by dividing the coax cable sections into several small pieces. However, this did not improve matters sufficiently. Nor did it help to divide the capacitors into two RG213 units, each registering 18 pF and each being 9 cm long, as described in [7]. The values for harmonic suppression were already insignificant at 350 MHz.

The filter from [4] must behave in a similar way above 500 MHz with regard

to attenuation troughs.

Considerably better values could be obtained for harmonic suppression at high frequencies if the coax cable sections functioning as capacitors were replaced by ceramic wafer capacitors providing 18 pF each. From app.230 MHz to 800 MHz, the attenuation gradient remained almost uniformly on 45 dB, then fell to app.30 dB at 1100 MHz and fell further to a miserable 10 dB between 1400 MHz and 1800 MHz, the limit of measurement.

The reasons for the poor values obtained from the assembly using the coax cable sections as capacitors lie in the fact that at higher frequencies these coax cable sections do not just act as capacitances. They mainly behave as inductances, with undesirable resonances. Moreover with this type of filter, the relatively large coils also contribute to resonance phenomena at high frequencies, due to the undesirable capacitance. Fig. 4 shows a test assembly of the LP filter from [6] and Fig. 5 shows the attenuation gradient.

5. The development of low-pass filters with excellent attenuation gradients

During further research, the author hit on the idea of making capacitors from printed circuit board material with a copper coating on both sides. At the same time, additional capacitors were added at the input and at the output of the filter. This makes the coils smaller and they have fewer windings. The entire system has a dual positive effect: the printed circuit board material capacitors have practically no natural resonances up to very high frequencies; the coils have less undesirable capacitance because their dimensions are smaller.

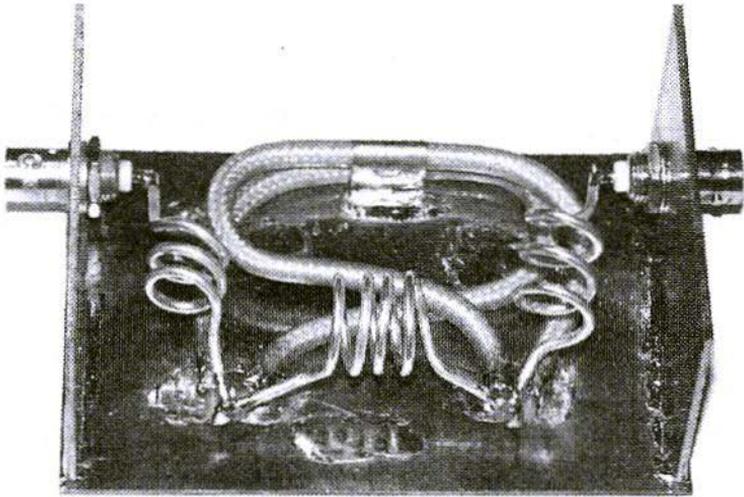


Fig 4 Test Assembly of the Low Pass Filter

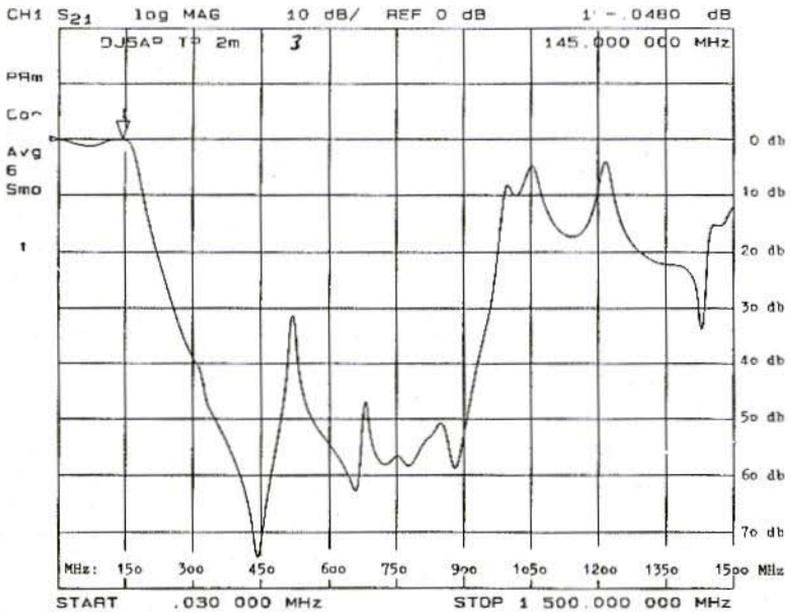


Fig 5 Attenuation of the Low Pass Filter

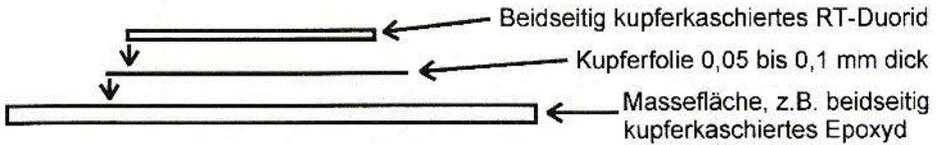


Fig 6 Assembly of Foil Which Projects Around the Outside of the Laminate Capacitor

For price and procurement reasons, the first tests were carried out using epoxy material of varying thickness, with copper coatings on both sides. The values from C and L from (5) served as the basis for the calculations for the filter.

The first preliminary tests were carried out at 145 MHz. First the capacitance per cm² was determined for epoxy material with a thickness of 0,79 mm. A square piece was then cut out (with a wide pair of shears) with each edge being 10 cm long, and then the capacitance was determined for this 100 cm², using a capacitance meter. The reading was 640 pF, which meant that for 1 pF an area of 0,15625 cm² was required. For 47 pF the total was 7,34 cm², for 27 pF an area of 4,22 cm² was needed. The 47 pF capacitor took the form of what was still a very small square, with each edge being 27,1 mm long, while the edge length for the 27 pF model was 20,5 mm. The problem of linking this (almost) inductance-free capacitor with the earth surface was solved in a simple manner. One side of the copper coating is tin-plated at certain points, especially on the edge. A piece of thin copper or brass foil 0.05 to 0.1 mm thick is cut in

such a way that the edge length is app. 3 to 4 mm longer than the length of the piece of epoxy material. The foil is tin-plated on both sides and then soldered to the epoxy laminate. The next step is that the earth surface is likewise tin-plated at the points where the capacitors are to be positioned, and then the foil, which projects under the epoxy laminate, is soldered to the earth surface all around. Fig 6 makes this clearer.

For an LP filter, four such capacitors are needed. With a little practise, these can be created in less than 30 minutes. The coils can also be rapidly wound. Due to the better trimming (compression or extension of the coils), silver-plated copper wire with a thickness of 1 mm was used for the first tests. 2-mm. wire was used in the final versions, because the power levels set could be high.

Following successful trimming, which will be described in more detail later, splendid values were obtained for the LP filter function: a steep rise in the attenuation values from approximately 200 MHz, which diminished only slowly down to 45 dB at over 2 GHz without any attenuation troughs.



Unfortunately the epoxy material initially used is already very lossy, even at 145 MHz. When an LP filter with capacitors made from epoxy material was inserted between a 2-m transmitter with an output power of 50 Watts and an EME Wattmeter, a power dissipation of app. 15 % was observed, i.e. app. 0.7 dB at 145 MHz. While the coils remained almost cold, the epoxy capacitors had already warmed up very noticeably a few seconds after being switched on. The author wanted to obtain some more precise information and ran the LP filter at 145 MHz with a power level of 600 Watts HF. The result was that little clouds of smoke were rising from the capacitors after about 10 seconds. This is easy to explain: when 600 Watts of power are fed in, with a 15 % loss, about 90 Watts are lost in the LP filter.

LP filters with experimental constructions, which had capacitors made from epoxy material 0.5 mm or 1.5 mm thick, brought no improvement; the losses at 145 MHz were of the same order of magnitude. Later designs for 70 cm gave even worse attenuation values at 435 MHz, so that the experiments with the capacitors made from epoxy material were broken off. This pre-determined the direction for subsequent developments: the circuit used plus the use of large-surface capacitors made from printed circuit board material gave excellent readings for the harmonic suppression we just needed to find a printed circuit board material with low losses.

6.

The final format for 145 MHz

Some pieces of RT-Duroid 5870 with copper coatings on both sides - 0,5 mm and 0,79 mm were picked up very cheaply at a jumble sale. This printed circuit board material, which is made from Teflon substrate, is used by com-

mercial operators and amateurs alike for assemblies at up to well over 10 GHz and is also obtainable at thoroughly affordable prices from distributors through "normal" outlets. The losses at 145 and 435 MHz are minimal. Initially, the capacitance per cm² was determined here too. For RT-Duroid with a thickness of 0,5 mm, a value of 0.20833 cm² was obtained for 1 pF; for RT-Duroid with a thickness of 0,79 mm, the value was 0.32 cm² for 1 pF. For a capacitance of 47 pF, this would mean an area of 9.79 cm² was needed with a thickness of 0.5 mm. For a thickness of 0.79 mm, the area required would be 15.04 cm². In order to fulfil the requirement for spatial compactness, the 0.5 mm thick substrate was used for the 2-m format. For 47 pF, the length of the squares edges is only 31.3 mm, and for 27 pF only 23.7 mm. There are no disadvantages if the required area is obtained using rectangles. 37.5 mm x 26.1 mm edge lengths once again give 9.79 cm². However, those carrying out their own experiments should bear in mind that the ratio of length to width should not exceed 3 : 1, so that no undesirable inductances "creep into" the large-surface capacitor.

For the assembly of the first 2-m filter with capacitors made from RT-Duroid, a purchased tinplate housing was used, in the interests of a mechanically uncomplicated construction, with the dimensions 111 x 55.5 x 30 mm. A suitable housing can naturally also be soldered together from printed circuit board material or tinplate strips this has no basic influence on functionality.

The use of the easily solderable tinplate housing makes it very easy to solder on the capacitors made from RT-Duroid. The RT-Duroid laminate is well tin-plated all round and then positioned in the tinplate housing with the tin-plated surface at the marked place on the base. The base is heated below the capacitor with a 100 to 150 Watt soldering iron after a few seconds the solder is

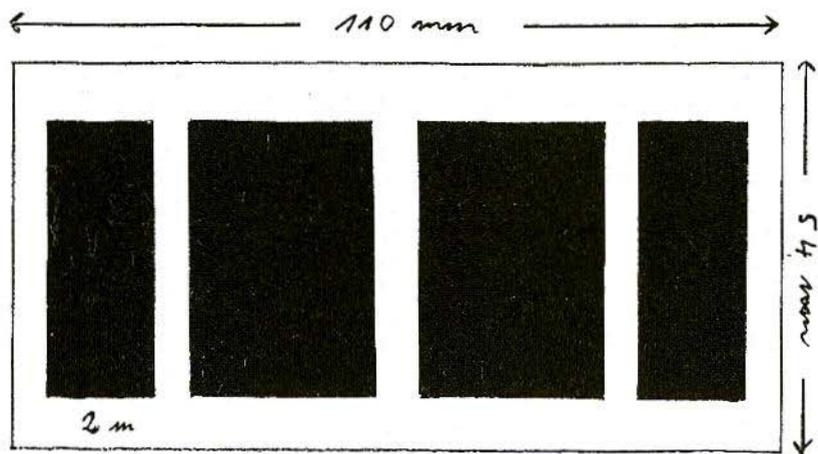


Fig 7 Board Layout for 2 m Low Pass Filter

flowing, and a large area of the capacitor is connected to earth. If you are using a home-made housing made of printed circuit board material, this simple method of mounting the capacitors is naturally not available. The RT-Duroid laminate, as described in Section 4, are soldered to the earth surface, with a thin intermediate layer of copper or brass foil.

If the filter is to be operated at full power - 750 Watts it is recommended that 1 mm of copper coating should be removed around the rim at the upper edges of the laminate capacitors. This can be done with the help of a file, which is held very flat against the surface. This considerably increases security against voltage flashovers.

In the interests of a simple assembly, and to make the design easier to copy, the author developed boards for 2 m and 70 cm, which held capacitors C 1 to C 4. For 2 m, RT-Duroid with a thickness

of 0.5 mm was used, and 0.79 mm thick material was used for 70 cm. The boards fit into a purchased tinplate housing with the dimensions 111 mm x 55,5 mm and a height of 30 mm. A home-made housing constructed from tinplate, brass plate or printed circuit board material can naturally also be used. If printed circuit board material is used which has a copper coating on one side only, make sure that the copper-coated sides face inwards. Of course, the coax sockets at the filter input and output are then to be positioned in such a way that the flanges of the sockets face inwards, so that the high-frequency earthing paths remain as short as possible.

Both N-sockets and BNC-sockets with flanges can be used. If the filter is being used at full power - 750 Watts then naturally only the N-type can be considered. In this case, you are recommended to use N-sockets without flanges, as otherwise, if the usual flanged sockets are used, there may well be problems of

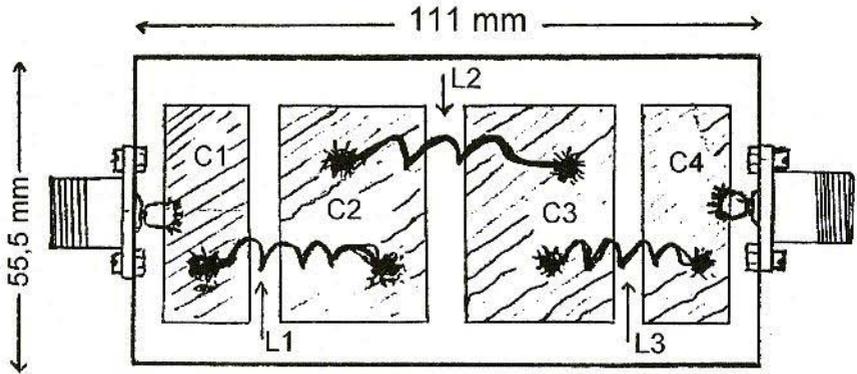


Fig 8 Component Mounting for 2 m Low Pass Filter (Plan View)

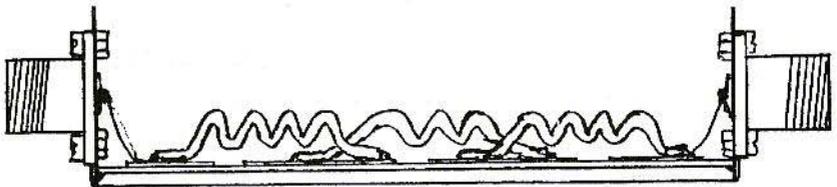


Fig 9 Component Mounting for 2 m Low Pass Filter (Side View)

space when the sockets are fitted to the surfaces of the tinfoil housing, which is only 30 mm high.

PL connectors and / or SO sockets should in no case be used at 435 MHz; even at 145 MHz, the use of these "telephone sockets with screening" can no longer be recommended.

Fig. 7 shows the board layout for 2 m using 0.5-mm thick RT-Duroid.

Coil data for 145 MHz LP filter:

L 1, L 3 = 2 mm dia. silver-plated copper wire, wound on 6 mm mandrel (auger shank), 25 mm long, winding interval app. 3 mm

L 2 = as L 1 and L 3, 30 mm long, winding interval app. 6 mm

Irrespective of what board material is used, the values for C 1 and C 4 should each be 27 pF, and those for C 2 and C 3 should each be 47 pF.

Figs. 8 and 9 show how the components should be mounted on the board.

When the boards are fitted into a housing, the earth surface (underside of board) is soldered to the housing walls all around. A distance of app. 2 mm should be maintained from the underside of the board to the bottom edge of the housing.

The internal conductors of the sockets should be connected to the capacitors via the shortest low-inductance paths. To this end, the internal conductors are pinched all the way to the Teflon insulation (use stable side cutting pliers) and are soldered to the capacitors by means of 5 mm wide thin copper foils.

If no sockets are used, but the filter is, for example, inserted directly into the high-level stage of a transmitter, the coax cable is fed through suitable bores in the surfaces of the housing and the cable screening is soldered to the housing on the inside. A cable with Teflon insulation (RG 142) is suitable for higher power levels.

The filters were tested at 750 Watts HF. No voltage flashovers took place; the coils became warm to the touch after operating for several minutes.

Attenuation levels of 40 dB are already being achieved for the second harmonic, 290 MHz. By the third harmonic in the amateur radio band (435 MHz), the figure is an impressive 70 dB. The attenuation gradient gradually dips to app. 45 dB at 1800 MHz, and then rises again to 60 dB at 2800 MHz. Only above app. 3300 MHz do attenuation troughs down to app. 20 dB appear. The attenuation in the 145 MHz transmission band is 0.13 dB!

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Gunthard Kraus, DG 8 GB

Earthing in HF and Microwave Circuits ... A Case for Puff

In any circuit with active components, we are confronted with the problem that certain lines or parts of circuits do not belong to the purely AC voltage circuit. Using appropriately adjusted earthing points, we can ensure that these components fulfil only DC tasks. And these earthing points are the subject of this article. In practise, we see every possible structure on printed circuit boards, and ask ourselves what advantages this structure is intended to convey. Which is precisely where this article is intended to be of use.

1.

Definition of Term Correct Earthing

We should first agree on the definition below for the term correct earthing.

A circuit point or a connection line can be said to be correctly earthed if, within the frequency range under consideration, the impedance (Z) between this point and earth is lower than 5 Ohms.

Let us now put the miscellaneous standard measures under the microscope, taking this point of view, originating

from developers practise.

2.

The Earthing Capacitor

This measure is certainly the oldest in electronics. In theory this method is first class, since it is indeed well-known that the AC resistance of a capacitor decreases as the frequency increases, and thus the connection to earth can only become even better.

2.1. Aluminium electrolytic capacitors

In practice, though, things look rather different. At very low frequencies, polarised aluminium electrolytic capacitors are used, but these display a series resistance which rises with the frequency and a reactive component of the impedance which also rises, due to inherent inductance. Moreover, it is not officially mentioned that the quality of electrolytic capacitors has deteriorated. The mechanical dimensions have been continuously reduced, but any radio or television technician will be happy to confirm that, for example, modern electrolytic

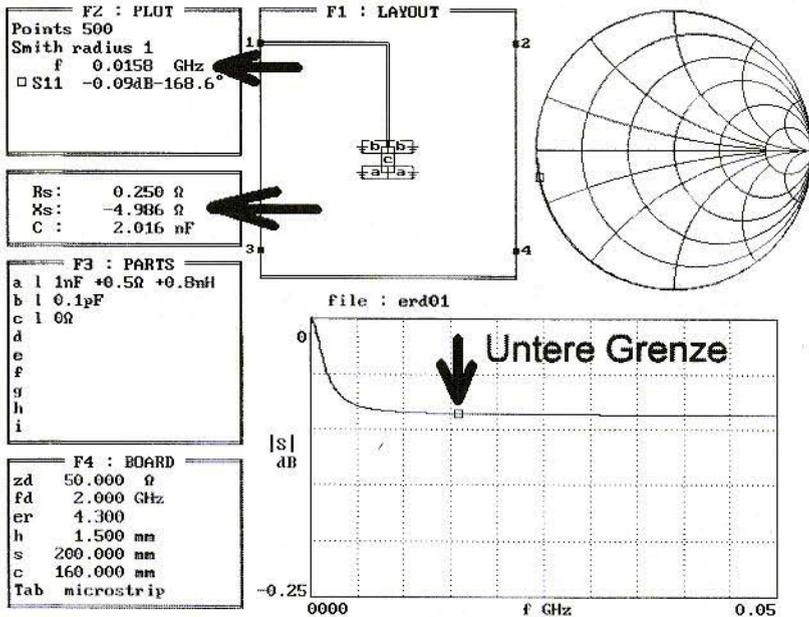


Fig. 1: Determination of lower limit for use of 2 parallel-wired 1-nF SMD capacitors as earthing (Untere Grenze = Lower limit)

capacitors are the cause of the problem in 50% of repairs of TV sets. They no longer break down, but usually slowly dry out, because the seals become porous. The unfortunate result of this is that, in addition to the capacity drop, which usually would not be so awful, there is very rapidly a sharp rise in the series resistance and that's bad!

2.2. Tantalum electrolytic capacitors

Tantalum electrolytic capacitors behave considerably better here. Sintered from powder, they basically represent semi-conductors (i.e. diodes) which are operated in the filter attenuation band. Their advantages lie in their small dimensions, their low series resistance (approximately 1-2 Ohms) and their low inherent inductance of a few nH. This gives excellent coupling capacitors and / or

wide-band earthings. In particular, in parallel circuits with several capacitors, they cover a range from low frequency to far above 100 MHz.

Let us not conceal their negative characteristics. Like any semi-conductor, they react extremely sensitively to overvoltage and / or pole reversal. In the most favourable case, disruptive short-circuits then occur. In the most unfavourable case, however, there can be scarcely predictable reductions, great or small, in the insulation resistance, which are to some extent dependent on temperature.

One factor here is often misunderstood. It will certainly say somewhere in the data sheet that, in the interests of service life, standard tantalum electrolytic capacitors should not be subjected to a switching current pulse exceeding approximately 0.3 A.

The reason for this is very simple. If this value is exceeded, it can quickly be

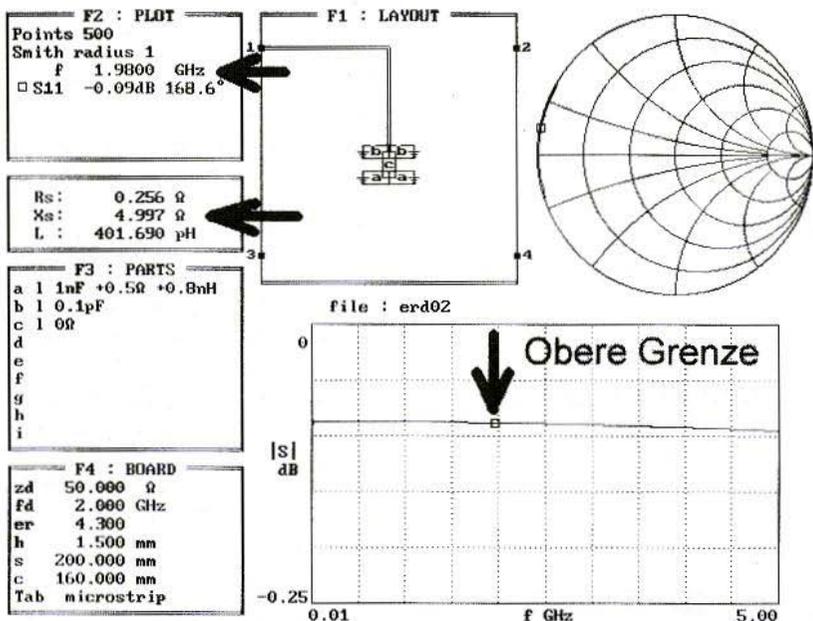


Fig. 2: Upper limit for earthing from two parallel-wired 1-nF SMD capacitors (Obere Grenze = Upper limit)

come so hot at individual points within the sintered material (in which, indeed small grains are in contact with each other) that a melt-on occurs, leading to a short circuit.

It makes no difference whether the current limitation required is brought about through the power supply itself or using a pre-resistance.

2.3. Ceramic capacitors

From 100 MHz upwards, the only capacitors still used are essentially ceramic models. Their quality is adequate for all requirements up to 2 GHz, even with the cheapest standard formats.

However, one should, as far as possible, attempt to use only SMD solutions, for only their low inherent inductance (< 1 nH) eliminates the tiresome inherent

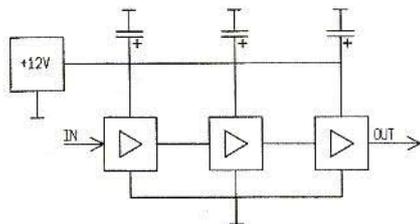
resonances of the wired-up copies. Here it is particularly advantageously to have parallel circuits of at least two capacitors, which means that the Ohmic resistance and the inherent inductance are smoothly halved.

We can use Puff to show how such a layout behaves at various frequencies. To do this, four components are parallel-wired in a simulation:

- Two SMD capacitors, each with 1 nF + 0.5 Ohms series resistance + 0.8 nH inherent inductance, and
- Two capacitors, each with 0.1 pF, each simulating the size 0805 SMD pad.

It is recommended that analyses should be distributed into an upper and a lower range for such very broad-band layouts

Falsch!



Richtig!

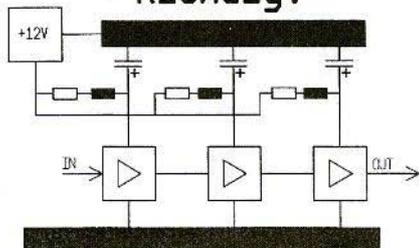


Fig. 3: Wrong and right layout for a multi-stage broad-band amplifier (Falsch = Wrong, Richtig = Right)

and for the Puff representation, restricted to 500 dots. Fig. 1 therefore shows the behaviour from 0 to 50 MHz, in the form of the reflection factor S_{11} , for this parallel circuit, which will supply the values for the impedance curve of the layout.

For this purpose, use Page up or Page Down to go to a specific frequency, place the cursor in field F2 on S 11 and key in the equals sign, =.

The active and reactive components of the input resistance immediately appear in the dialogue window, and the associated Substitute dummy component is displayed as an inductance or capacity value. Now we look for the frequency at which the capacitive reactive component of this layout undershoots 5 Ohms. In Fig. 1, this happens at 15.8 MHz.

In a second pass, we take the range from 10 MHz to 5 GHz and there we find, at

precisely 2 GHz, 5 Ohms for the (now inductive) reactive component (Fig. 2). It specifies the upper limit of the usable frequency band.

Anyone who wants to know why the actual value zero was not taken as the lower frequency limit for Puff can repeat the simulation using this value. The blemish arising in the representation in the Smith chart, due to the fact that the step widths at very low frequencies are now far too great, is avoided using the setting proposed.

The parallel-wired layout of ceramic capacitors is thus effective and correspondingly popular. To set the bottom frequency limit still lower, we simply replace the ceramic-SMD capacitors with a parallel circuit of several tantalum electrolytic capacitors, in which the loss resistances and inherent inductances are approximately twice or three times as great.

2.4. Notes on sources of errors

Unfortunately, we can find some circuit technology earthing errors in power supply systems not only in DIY projects, but even in industrially produced circuits.

Fig. 3 shows a wrong and a right format for a multi-stage broad-band amplifier as an example:

a) There are no protective resistors for the tantalum electrolytic capacitors.

b) The power supply lines of the individual stages must have star connections to the power supply and must be de-coupled from one another. With the help of a choke coil and / or an Ohmic resistor, we can thus then create an LC or RC low pass in each feeder with the tantalum electrolytic capacitor. Only in

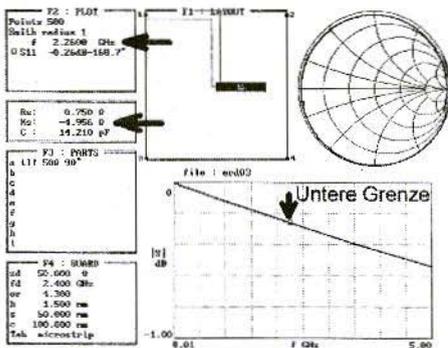


Fig. 4: Determination of lower limiting frequency using a 50-Ohm microstrip line on no-load as earthing (Untere Grenze = Lower limit)

this way can we reliably prevent part of the high level of the last stage from going back to the input section and causing the layout to oscillate.

The HF choke used, however, must not be of too high a quality (to avoid resonance effects). Here it is thoroughly normal to wire Ohmic resistors into the circuit. This not only attenuates but improves the low-pass effect, especially at lower frequencies, where the inductive resistance is known to be still low. This simultaneously provides the protective resistors for the tantalum electrolytic capacitors.

c) The points connected directly to earth (= without capacitor) within an amplifier stage must not just be looped on to the next stage, or an unnecessary oscillation risk will be created.

A version mounted on printed circuit boards in accordance with the laws of microwave engineering is ideal, and almost oscillation-proof, even at low frequencies. Coated on both sides, this

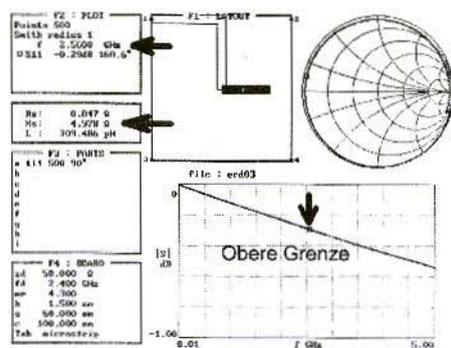


Fig. 5: Upper limit using 50-Ohm microstrip line on no-load as earthing (Obere Grenze = Upper limit)

version thus has an underside which is an integrated earthing and earth level. In the illustration it is shown as a blocked-in rectangle. The direct earthings are brought about through suitable through-platings, which can be implemented in the form of full tubular rivets (silvered, diameter 0.8 mm.).

The screening action of such choke / capacitor layouts at various frequencies can also be determined by means of Puff. But you can find out from [1] how expensive this can be, for example, for a low-noise oscillator.

3.

Earthing Through Microstrip Lines

At higher frequencies, if discrete components slowly fail, we resort to sections of line working at no-load as earth connections. It is well known that such a line represents a short circuit at $l = \lambda/4$.

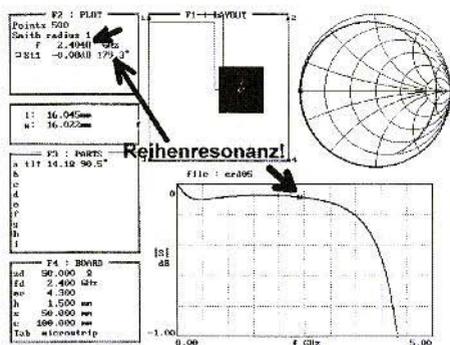


Fig. 6: Using a square pad with an edge length $l = \lambda/4$ as earthing (Reihenresonanz = Series resonance)

Thus, for example, we can take such a lossy $\lambda/4$ line with 50 Ohms for 2.4 GHz and simulate its behaviour for a printed circuit board made from epoxy material FR4, using the data:

Thickness 1.5 mm, $r = 4.3$ and

Loss factor $l_t = 0.02$

We are less interested in behaviour under resonance here, for it is well known that only the small loss resistance remains behind there (approximately 0.8 Ohms). Use Page Up or Page Down to move the cursor until keying in the equals sign at S11 gives you an inductive or a capacitive reactive component of 5 Ohms (Figs. 4 and 5). Frequencies associated with this are 2,260 MHz and 2,560 MHz earthing would thus be usable only in a narrow range 300 MHz wide. This method is therefore also deliberately used only for narrow-band applications or oscillators.

Things work considerably better, with a broader band, if the section of line working at no-load is altered to such an extent that the length and width are the same and form a square. If we also make the area of this square sufficiently big, the line, because of the low impedance level, acts like a big capacitor with a correspondingly low reactive impedance, even at really low frequencies (i.e. at $l < \lambda/4$).

If the frequency is then increased, the $\lambda/4$ resonance follows, with the short circuit at the input. Not until this resonance is exceeded does the reactive component increase again, due to the inductive behaviour, but with a low-Ohm line the inductance is indeed also very low. This leads us to expect a decidedly wide usable frequency range, in which the impedance is sufficiently low.

This can be tested for 2,400 MHz using Puff. With the same printed circuit board data as before, the length and width at this frequency are to be given values which correspond to a quarter of the wavelengths. These include lengths and widths of approximately 16 mm. and an impedance level of 14.1 Ohms (Fig. 6).

If we now simulate the behaviour of this line and look again for the points with a reactive component of 5 Ohms, we find the frequency values 2,867 MHz and 1,973 MHz are suitable. This layout would also fulfil the requirements laid down in a range of approximately 900 MHz. For a direct comparison, the impedance curves ($Z = f(\text{frequency})$) for the two sections of line have been plotted in a joint diagram (Fig. 7).

The advantage of the square section of line in broad-band applications is once again easy to see here. Unfortunately, it is very cumbersome for practical application at low frequencies.

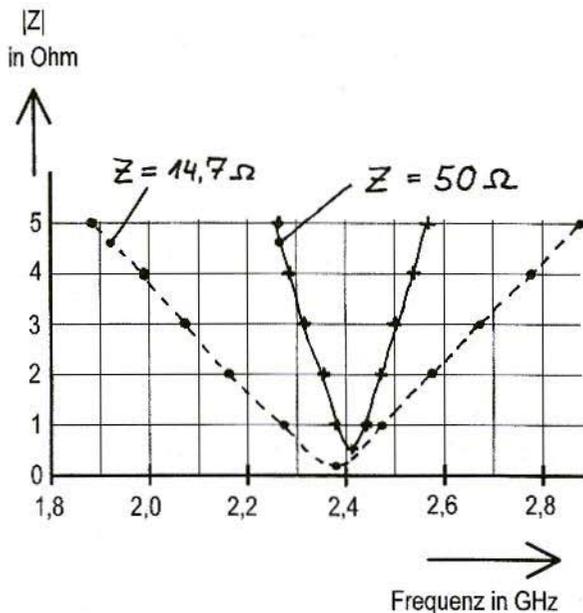


Fig. 7: Impedance curve $Z = f(\text{frequency})$ for the two line sections investigated (Frequenz = Frequency)

4. Earthing Through Radial Line Stubs

Now let us look more closely at the mysterious circuit segments which are found in nearly all microwave circuits from approximately 2 GHz upwards. Fig. 8 shows such a layout, from type application note 1091. How do such layouts work, and what advantages do they offer?

Interestingly, they are used by almost all microwave circuit developers, although there is scarcely any literature about it. Only an application note from HP [2] taken from the Internet gives them their correct name (radial line stubs) and names some literature sources [3], [4].

It does, though, become apparent very quickly that the calculation formulae

given there are very lavish and are tailored more for use on mainframe computers. The HP note also describes the application in a very precise and enlightening manner:

"...problems of location and parasitics of low impedance shunt stubs were solved by using fan-shaped open stubs with the narrow end connected to the main transmission line..."

Consequently, we can introduce a low-Ohmic resistance at a three-dimensionally limited point in the circuit and that is exactly what we want for earthing. I am reminded here of the concept of tapering, which in the past was an important method for broad-band transformations and matchings. We are referring here to lines or cavity conductors which continuously alter their dimensions and thus their wave resistance,

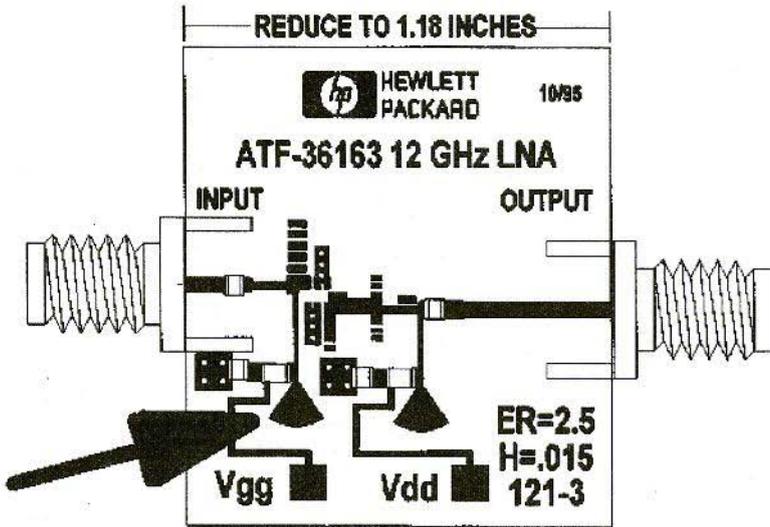


Fig. 8: Application example for radial line stubs (12 GHz LNA)

with the usable frequency range being markedly increased through this measure and the line length required for this reduced.

First a few basic observations:

a) Because the line end is under no-load here too, the associated series resonance must consequently be introduced somewhere (recognisable at the 180° phase angle at S11) with the short circuit at the input. The knife-sharp tip of the structure is very advantageous here, so that the earthing can be positioned precisely at the desired spot on the printed circuit board.

b) If we make the central angle very small, the construction becomes more and more like a normal but narrow and also high-Ohmic microstrip line, with its narrow-band resonance. So increasing the central angle must produce the opposite effect, namely broad-band low impedance.

c) And, assuming the function of tapering has been described correctly, the layout would even have to turn out markedly shorter than a standard $\lambda/4$ line!

We can now use a little trick in our investigation with Puff:

One of these radial line stubs is chopped up into many short line sections, but all with the same length, which are wired in series. The line width of each partial section is then selected to be small enough to obtain the best approximation to the original structure.

For a frequency $f = 2,400$ MHz, we select the radius of the associated full circle to be precisely the same as the $\lambda/4$ line length in the previous example, i.e. approximately 16 mm.. Let a piece be cut out of the full circle with a central angle of 45 degrees.

It is helpful if the entire structure can be

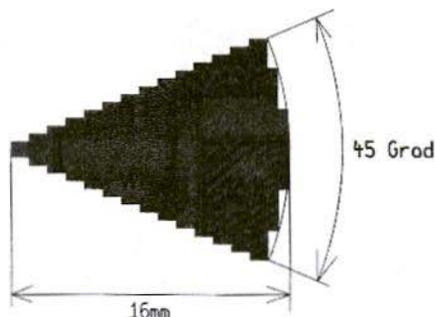


Fig. 9: Radial line stub assembled from 16 elements (Grad = Degrees)

drawn out with a scale of 10:1 on millimetre paper or squared paper, and the dimensions determined for the strip

line conductor widths required or they can simply be worked out on a pocket calculator.

We are working here with 16 elements in all (N.B.: you can list a maximum of only 18 components in field F3 of Puff!). Each element has a length of 1 mm, and the widths are graduated in such a way that the best approximation to the ideal stub is obtained (Fig. 9). The following individual data for the simulation can be measured out from the drawing of the line sections or calculated with the help of the geometrical formulae. The numbering for the line sections starts on the left, i.e. at the tip of the circle sector (Table 1).

Now the hard work begins. You can certainly key in the lengths of the microstrip line sections in the Puff F3 field directly in mm, but not their mechanical widths. To do that, you first have to assign any impedance to each of the 16 sections and vary it until the desired conductor width is set as a

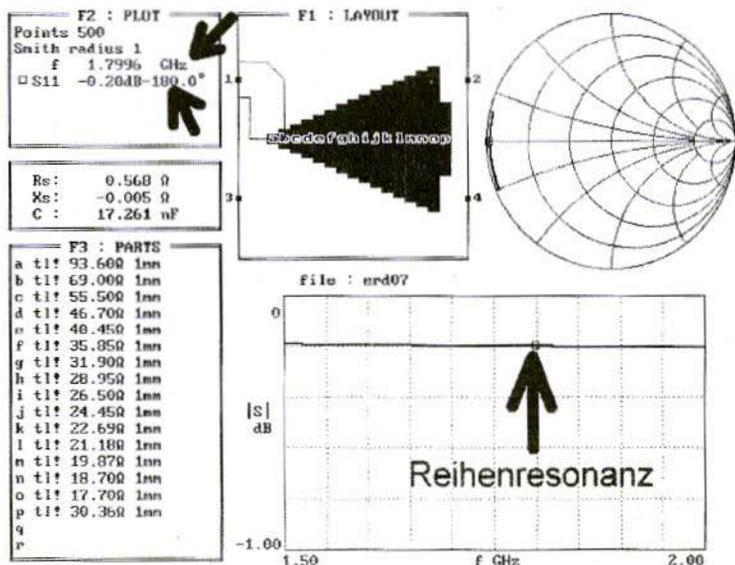


Fig. 10: Radial line stub with a central angle of 45°: S 11 curve between 1.5 and 2.5 GHz (Reihenresonanz = Series resonance)

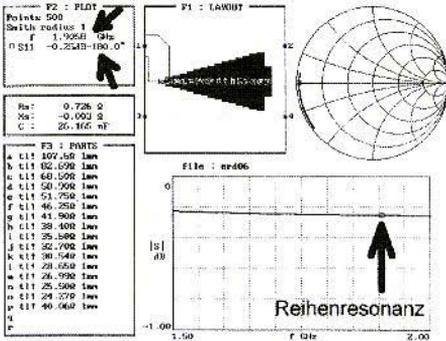


Fig. 11: S 11 curve with a 30° central angle (Reihenresonanz = Series)

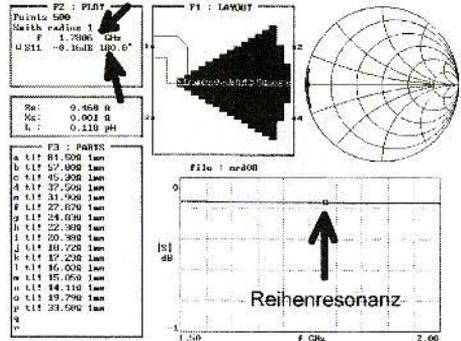


Fig. 12: S 11 curve with a 60° central angle (Reihenresonanz = Series resonance)

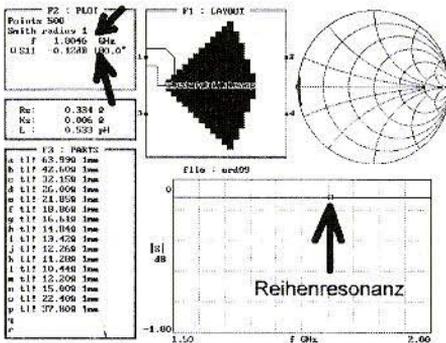


Fig. 13: Radial line stub with a 60° central angle (Reihenresonanz =

reaction when the equals sign is keyed in. Then the stub structure is assembled in field F1, and the simulations are begun for the range 1.5 2.5 GHz.

Not only does Fig. 10 show the complete list of stripline sections, together with their associated impedances, in field F3, but also the resonance frequency of the earthing (= reactive component precisely 0 Ohms, and thus phase angle of 180° at S11) can be read off at approximately 1.8 GHz. Thus, for a central stub angle of 45°, we have a mechanical reduction of the layout of at least approximately 25%, as against the standard $\lambda/4$ line.

Just for interest, the calculations and

Leitungstück Nr.:	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
Breite in mm	0,816	1,63	2,45	3,265	4,08	4,90	5,71	6,53	7,35	8,16	8,98	9,80	10,61	11,43	12,24	6,12
Länge in mm	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0	1,0

Table 1: (Leitungstück = Line section, Breite = Width, Länge = Length)

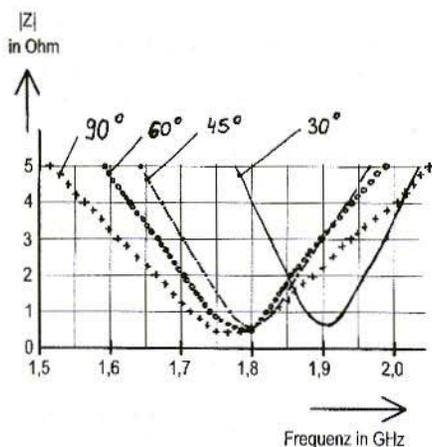


Fig. 14: $Z = f(\text{frequency})$ for radial line stubs with 30/45/60 and 90° central angles (Frequenz = Frequency)

simulations were repeated again for the central angles 30°, 60° and 90°, in order to obtain a feel for optimal dimensioning. Figs. 11, 12, 13 now show both the required impedances of the line sections for the three central angles investigated and also the associated simulations with the resonance frequencies, which now differ only insignificantly.

To be able to carry out a direct comparison between the characteristics of the 4 stub versions investigated, a summary diagram was drawn up for the impedance curve of the four different stub versions, using Puff, in the frequency range from 1.5 to 2 GHz (Fig. 14). If we look closely at this picture, we can obtain the following knowledge from it:

a) With a central angle of 30°, the taper effect is already weakening markedly, and the behaviour corresponds more and more to the normal $\lambda/4$ line. This can be seen, above all, from the fact that the reduction is no longer so effective and

the band width is diminishing.

b) Between 45° and 90°, the reductions are almost identical in the order of 25% - but the broad-bandness naturally increases with the central angle, as was conjectured.

Practical tips:

As a developer, you can lay in a stock of such radial stubs for the main frequency ranges in which you work.

The procedure required for this can be demonstrated using the example of the design frequency of 2,400 MHz already used previously.

Step 1:

Decide on a central angle between 45° and 90° - for example, 60°.

Step 2:

The design frequency can now be selected to be higher by a factor of 1.25. Thus we can enter 3.0 GHz in field F4.

Step 3:

Decide on the mechanical length required for a standard $\lambda/4$ line for the printed circuit board material used and for the frequency of 3 GHz. This gives the radius of the full circle from which the stub will be cut out.

Step 4:

Now sketch the stub, replace it by a series of max. 18 series-wired microstrip line sections see above and enter these sections in field F3. Then comes the wearisome task of determining the matching impedance for each line section.

Step 5:

Now assemble the stub in field F1 from the individual line sections and then

start the simulation for S 11.

N.B.:

Don't be afraid of carrying out two separate simulations, one for a narrow band width of 2 to 3 GHz and another for a broad-band range of 0 to 10 GHz. Thus for a subsequent practical application you can have the option of investigating its characteristics even outside the range for which it is intended to be used and, for example, checking for oscillation tendencies. If you have used only the narrow-band version of stub here, Puff immediately reacts with a corresponding error message if the pre-set frequency range is exceeded.

Step 6:

Now switch to a text processing program and print out the *.puf-file just produced, together with any device file existing for an FET or transistor from the Puff directory. If we lay the two print-outs directly next to one another, we can recognise which parts of the stub file must be altered or deleted to make a device file from it.

The device file created in this way can then be stored under a suitable name (ending in *.dev) in the Puff directory. It will then be available as a component for future campaigns.

Step 7:

If, though, you want to turn your attention to another frequency range, you have to repeat the entire procedure using the corresponding data.

Anyone who, like me, tends to make more and more use of the ARRL Radio Designer (in parallel with Puff) to determine the noise figure, the stability factor, k , or an impedance curve, etc., will now have rather more work to do. You can certainly integrate the S-parameter file for the stub generated by Puff into the Radio Designer, but it does need rather more extensive preliminary work:

Not only do all superfluous parts of the Puff file have to be deleted but the unit GHz has to be added to every line of the S-parameter listed following the frequency value. The altered file is then copied into the ARD file currently being worked on to do this, see [5].

5.

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-
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Dipl. Ing. Wolfgang Borschel DK2DO

Circulators and Ring Hybrids

For protecting VHF high-level stages and for decoupling VHF control transmitters

In practice a transceiver is used to drive a power amplifier. If the power amplifier is tuned and a brief self-excitation takes place, which can easily happen in standby mode, the receiver input transistor can be destroyed.

In contests each operator uses his or her own transceiver, but they all use the power amplifier. Matching the transceiver output impedance to the power amplifier input impedance leads to great variations in output, although all transceivers are providing the same driving power.

The solution to these problems usually lies in using circulators and hybrids.

1.

Basics

This article is concerned exclusively with circulators and with hybrid quarter-wave circulators. The way they work is briefly explained, and this is followed by practical assembly suggestions. Circulators are among the oldest microwave ferrite components in technical use. Their transmission behaviour is influenced by a large number of magnetic

and geometric parameters, which makes them difficult for amateurs to manufacture and also makes access to them difficult. Optimisation with regard to band width and transmission losses is also difficult to arrange.

The functioning of the circulator should be described first. Circulators have a pre-magnetised ferrite body as their essential component, which absorbs the electro-magnetic wave in one direction but leaves it uninfluenced in the other.

The way the circulator works should become clear from Fig 1.

The power from the VHF amplifier is passed on to the antenna in the direction of rotation of the circulator. The power reflected from the antenna, e.g. due to a poor SWR, is also passed on to the dummy load in the direction of rotation and is there converted into heat. The VHF amplifier discovers an ideal SWR in every case.

An open (blank) circulator arm can not absorb any power, i.e. it feeds it on to the next arm in the direction of rotation.

In the same way, the circulator protects a transceiver against a poor input SWR from the subsequent amplifier stage. (Fig 2) The transceiver sees an ideal SWR in any tuning position. Because of the special ferrite body, which all

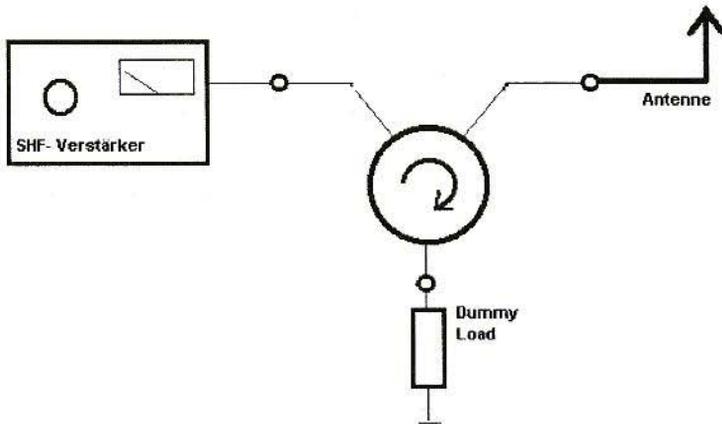


Fig 1 Using The Three Port Circulator to Isolate an Amplifier from an Aerial

circulators have in common, this microwave component is not suitable for production on a DIY basis.

2. Decoupling Between Transceiver and PA Using Hybrid Quarter Wave Circulators.

Quite usable results can be obtained using the hybrid quarter-wave circulator (Fig. 3), if it is correctly dimensioned. It also provides a measuring output for tuning the input circuit. As can be seen in the sketch in Fig 4, this microwave component consists of four $\lambda/4$ circuits with different impedance levels, which are inter-connected to form a ring. However, the power delivered from the input to the usable output is not 100%, as for the circulator, but only

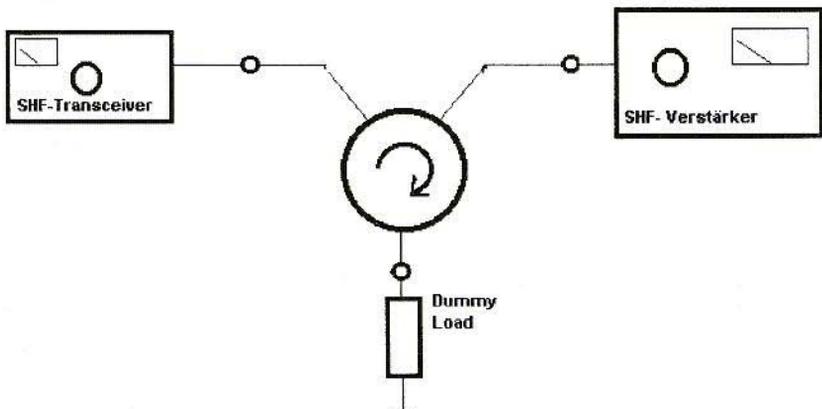


Fig 2 Using the Circulator to Isolate a Transceiver from a Power Amplifier

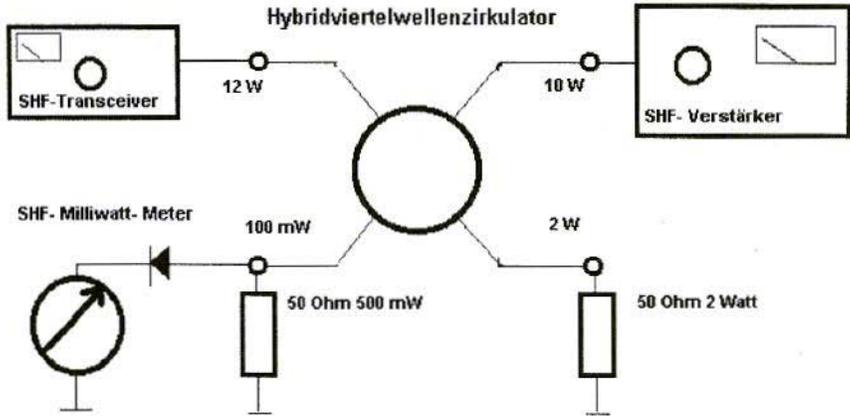


Fig 3 Using the Hybrid Quarter Wave Circulator

a maximum of 86 %. Moreover, the filter attenuation is considerably worse. The four ports of the hybrid in this application are defined according to Fig 5. Power applied at port 4 splits,

rotating clockwise, onto ports 3 and 2. At the measuring output, the signals cancel each other out due to phase opposition. Port 1 is completely de-coupled from port 4. Likewise with this layout ports 2 and 3 are completely de-coupled. However, the input and the usable output of the ring coupler are not completely de-coupled. This is connected with the fact that energy reflected from port 3, rotating anti-clockwise, splits in equal proportions onto ports 4 and 1, whereas power fed in at the input, rotating clockwise, splits onto ports 3 and 2. If port 3 is used as input, then the power supplied, rotating anti-clockwise, splits in equal proportions onto ports 4 and 1. In this case, ports 3 and 2 are completely decoupled. Ports 4 and 3, however, are only de-coupled for half their amplitude, because input and usable output can be transposed in both examples. This means that the power transmitted from the input (port 4) to the usable output (port 3) is reflected from the changing load at port 3 and splits onto ports 4 and 1 in defined proportions. A milliwatt meter connected to port 1 indicates the reflected power at port 3 due to a miss-matched load (connected VHF PA input). The high-level stage input circuit need only be

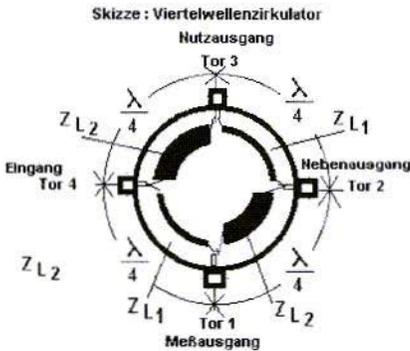


Fig 5 The Four Ports of the Hybrid Quarter Wave Circulator. (Eingang = Input, Nutzausgang = Useable Power Output, Nebenausgang = Dummy Load Output, Messausgang = Measurement Output)

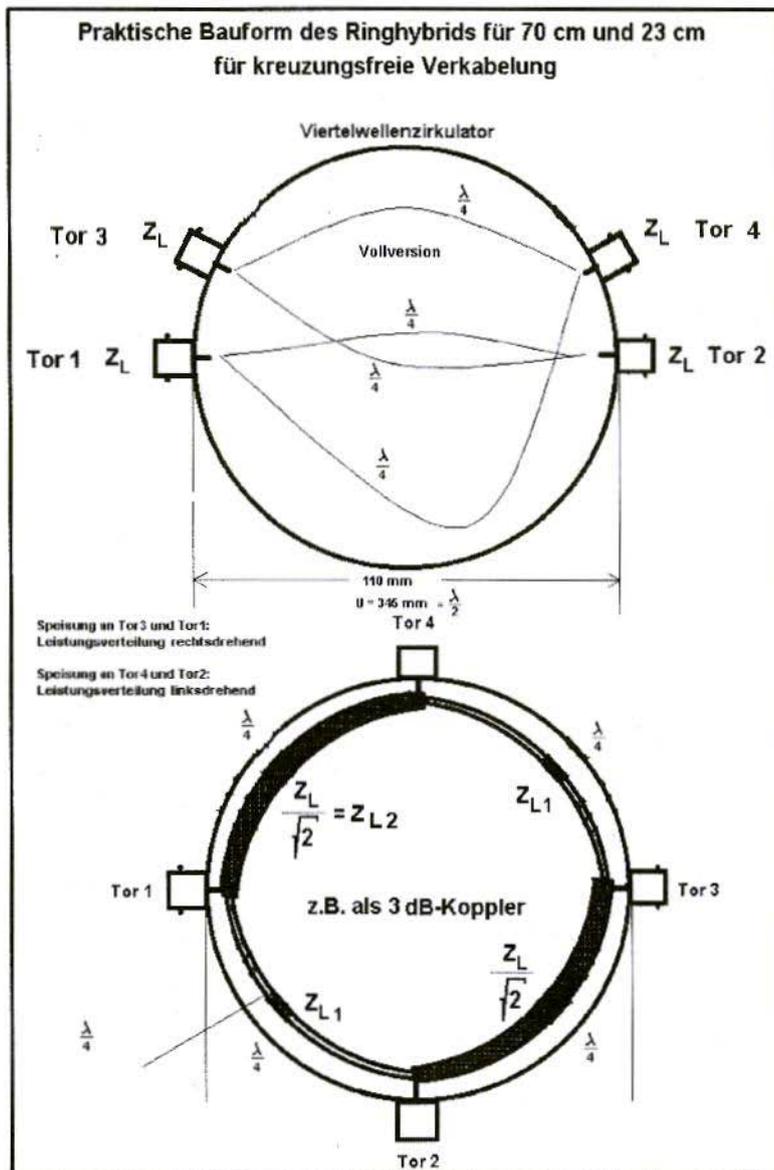


Fig 4 Sketch of Hybrid Quarter Wave Circulator



$$f1(x) = \text{sqr}(50^2 / (1 + (50^2 / x^2)))$$

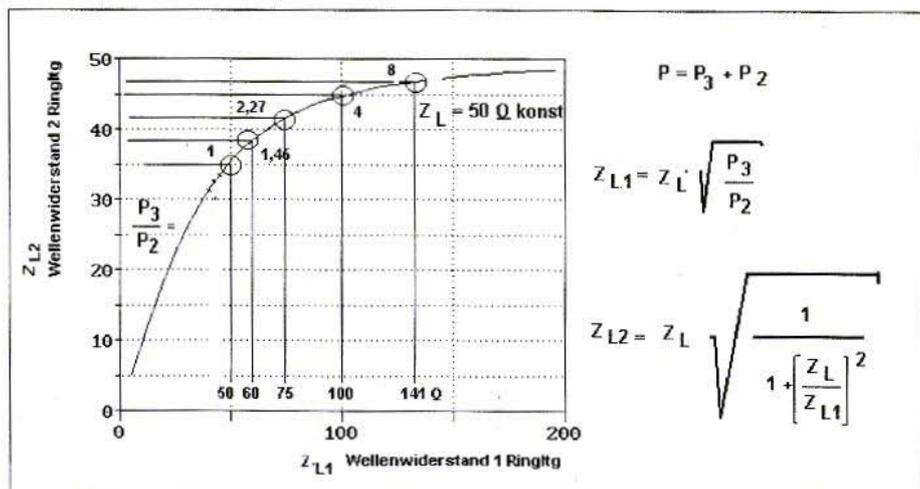


Fig 6 Results of Calculations of Hybrid Quater Wave Circulator

tuned at the minimum amplitude of the milliwatt meter at port 1. The reactive components present in the hybrid are now compensated by the reactive components of the input circuit, i.e.: a conjugate complex tuning has taken place. What can be explained only with

difficulty in theory has an amazing effect in practise. The control transceiver always sees an equally good input SWR almost irrespective of the tuning condition of the VHF high-level stage.

	Z_{L1}	Z_{L2}	Z_L	a_n
$\frac{P_3}{P_2} = 1$	50 Ω	35 Ω	50 Ω	3 dB
1,45	60 Ω	38 Ω	50 Ω	2,3 dB
2,27	75 Ω	41 Ω	50 Ω	1,6 dB
4	100 Ω	44 Ω	50 Ω	0,9 dB
8	141 Ω	47 Ω	50 Ω	0,5 dB

Table 1 Division Ratios Possible

a_n = Durchgangsdämpfung zum Nutzausgang

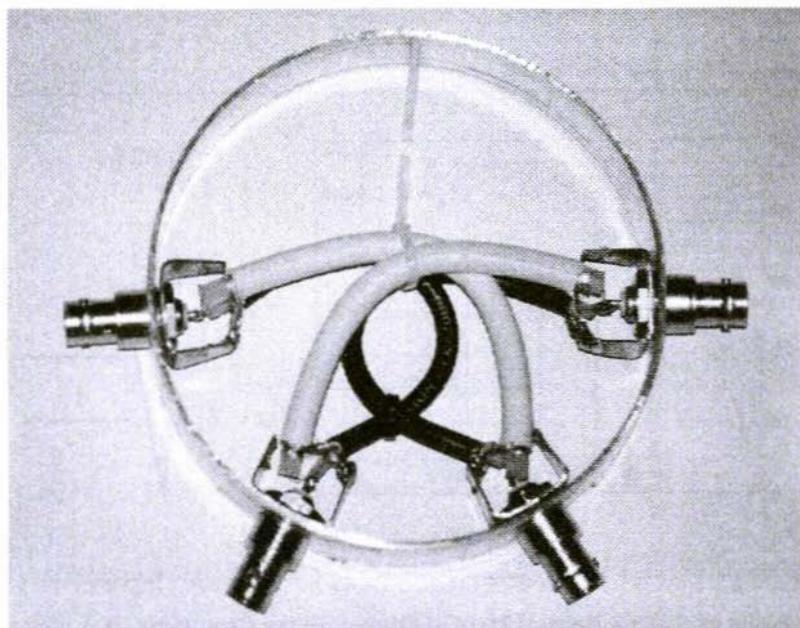


Fig 7 The Completed Hybrid Quarter Wave Circulator

3. Correct Dimensioning of the Hybrid Quarter Wave Circulator

Very few people know that this component can create a power division ratio which departs from 1. The scientific literature contains few descriptions of this characteristic. The disadvantage of this characteristic can be gathered from the scattering matrix using four-pole theory tools. The scattering matrix itself can be taken from [1]; the dimensioning guidelines can be derived from this. The results are shown in the Fig 6.

An analysis of this diagram provides information about how far division ratios can actually be achieved. So that as little as possible of the control power is lost, a division ratio as high as possible is selected. As the division ratio rises,

ZL2 asymptotically approaches the value $ZL = 50 \text{ Ohm}$. The associated value of $ZL1$, however, no longer assumes a value which can be realised. So if practicability is our basic requirement, then any division ratios greater than 8 can be attained. For this ratio, we would need a $\lambda/4$ coax section with an impedance level of 141 Ohms. This can be manufactured if the middle conductor is carefully extracted from a section of flexible 75-Ohm wire and if it is replaced by a piece of enamelled copper wire with 0.2 mm. Z then becomes 125, which approximately corresponds to a division ratio of approximately 6. Other Z values such as 50, 60 and 75 line can be obtained.

So the division ratios listed in the Table 1 can easily be achieved. Division ratios of under 1 would mean that the power at port 2 was higher than at port 3. In this event, we would need only to transpose ports 3 and 2. The table would then have to be determined for the reciprocal



value of P3/P2.

4.

Measuring Results from Assembled Hybrid Circulator

The hybrid ring described was assembled. For this purpose, a large-diameter section of aluminium piping was sawn off and faced. 4 coax sockets should be fitted onto the piping. This structure has proved its worth, because it allows for cabling without intersections. The division ratio was set exactly as determined. The conjugate complex tuning with the input reactive impedances of the VHF high-level stage allows the input reflection factor at port 4 to become infinitesimal (SWR: 1.1).

All lambda/4 circuit sections are cut to 11.5 cm. in the 70-cm band, with consideration being given to the velocity factor ($v = 0.66$). The 141-Ohm circuit

is created as described. In the 23-cm. and 13-cm. bands, it is advantageous to create the hybrid using stripline technology. In still higher bands, it is better to use $\frac{3}{4}$ lambda stripline technology to produce the module. Make sure that the $\frac{3}{4}$ lambda circuit has the same transformation characteristics as the lambda/4 circuit. The circuit lengths for the GHz range are easier to handle.

5.

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Proceedings of 23rd GHz Congress, Dorsten 2000

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Helmut Neidel, DL1IN

Supplement to Article on 5.7 GHz ATV Converter

Issue 1/2000 Pages 24-29

The converter described was designed for an internal operating voltage of 8 V.

This voltage was selected in order to obtain as uniform a distribution as possible of the power loss caused by the MMIC which was used.

Linear Technology have recently made available the type LT1777, a switch controller which eliminates this problem in a simple manner (tnx Ewald, DK2DB, for info!). This is done by replacing the 8 V linear power supply

with the small additional board illustrated in Figs 1 and 2, incorporating the controller IC and a few additional components. The operating voltage of the converter is reduced to 5 V. This changes the following resistance values:

R5:	56 Ohm
R10:	56 Ohm
R11:	47 Ohm
R15:	120 Ohm

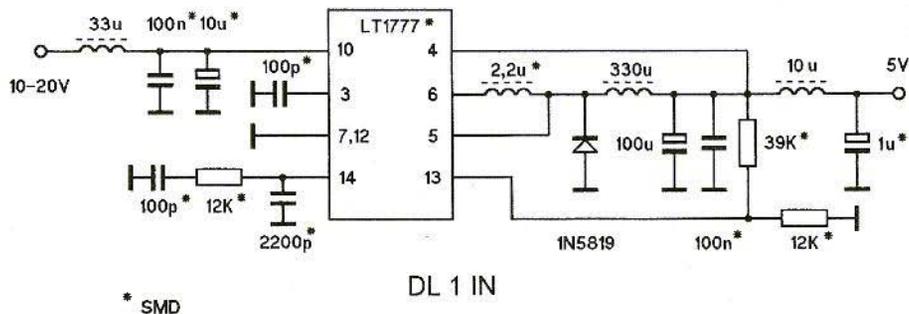


Fig 1 Circuit Diagram of New Switching Controller Power Supply

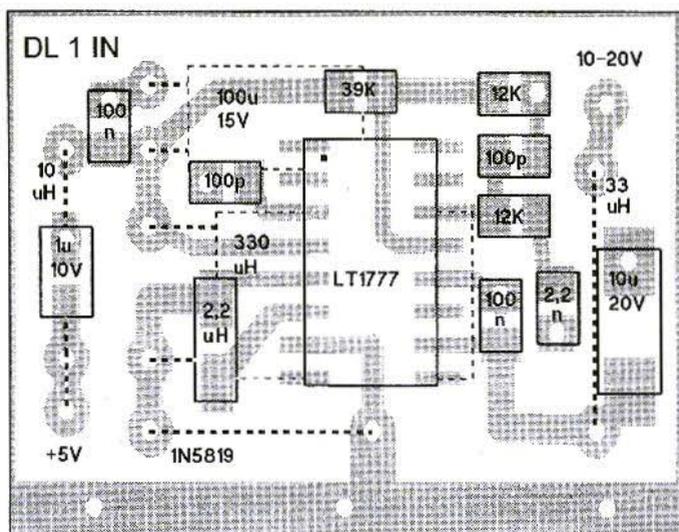


Fig 2 The Printed Circuit Board for the new Switching Controller

R16:	220 Ohm
R17:	47 Ohm
R18:	47 Ohm
R19:	47 Ohm

converter following this change still amounts to only approximately 125 mA for an operating voltage of 12 V.

Literature: Linear Technology Specifications LT1777

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Another friendly gesture is the provision of interesting and well put together Online-Tutorials offering basic introductions to many of the products on offer.

Address:

<http://www.synergymwave.com>

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Focus

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<http://www.dspGuru.com>

<http://www.mrccos.com>

Hamview and Spectran

These names conceal two DSP programs for radio amateurs. With their help (and with the help of the sound card in your PC), you can, for example, filter noisy receiver signals, or even track the signals down in the noise. Both programs were written by the same authors, but they are stored on different homepages.

Addresses:

<http://www.freeyellow.com:8080/members/padan>

<http://www.mx8.xoom.com/spectran>

Narda

Another of the big names in the microwave sector with a correspondingly wide range of components and assemblies on offer. Their homepage is always worth a visit. Among the many attractive things on offer is a professional Freeware-Program (Name: WIN N.R.T. TM) for calculating field strengths in the environment of antennae.

Address:

<http://www.nardamicrowave.com/east/support/software.html>

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Here you can download the usual data sheets, but also Application Notes for Step Recovery Diodes, Tuning Varactor Diodes, PIN-Diodes etc..

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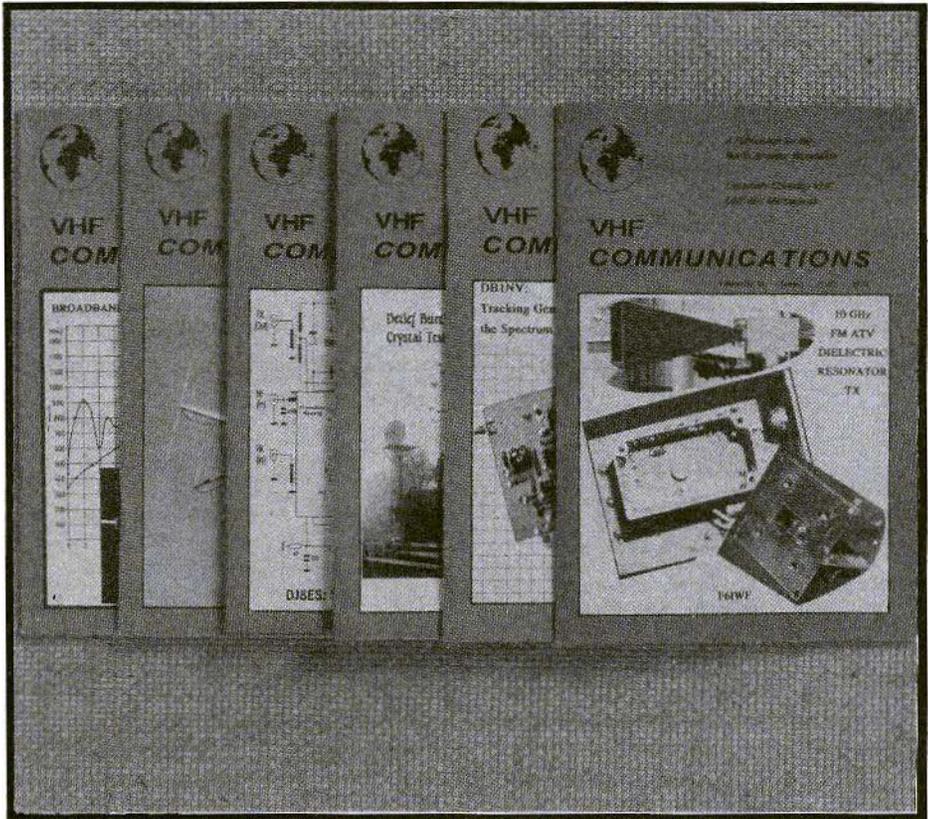
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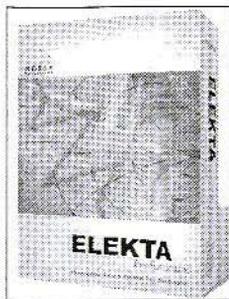
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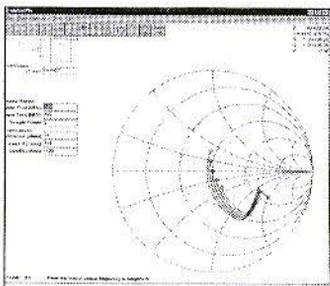
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