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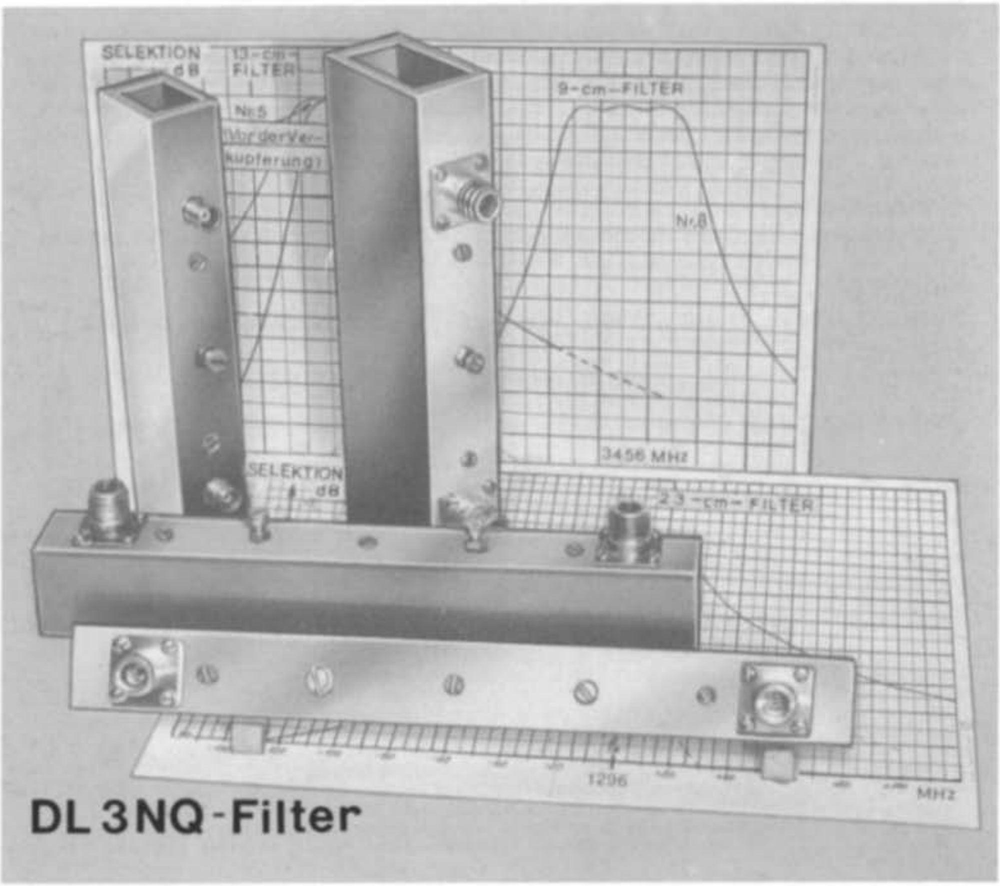

DL 3NQ-Filter

VHF communications

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NARROW-BAND FILTERS for the 23 cm, 13 cm and 9 cm Band

by D. Vollhardt, DL 3 NQ

The following article is to describe three and five stage narrow-band filters and to give their passband curves.

1. INTRODUCTION

Nowadays, more and more semiconductors are being used in the frequency range between 1 and 4 GHz; firstly, this was only the case in the receiver stages, however, they are now often used in the transmit signal path and for the driver stages. Very often, sufficient attention is not paid to providing sufficient selectivity. The disadvantages of this on the receive side were discussed in (1). On the transmit side, similar wideband problems are often encountered. There is a danger that insufficiently suppressed harmonics, unwanted conversion products and SSB sidebands (image frequencies), as well as unwanted spectrums from ATV transmissions could cause difficulties to other services if they are ignored in the future. This is, of course, also valid when using tubes, since the required, or advisable suppression of spurious signals is not always achieved even when amateurs are using manufactured cavity power amplifiers.

2. DEMANDS ON SELECTIVITY AND INSERTION LOSS

In the case of a receive filter, it is sufficient for the image frequency to be suppressed by approximately 20 dB; the insertion loss should be less than 1 dB. In the case of a converter using a first IF of 28 MHz, a filter is required that provides 20 dB attenuation 56 MHz below the required frequency. This demand is not too great.

If, however, not only the lower sideband, but also the local oscillator frequency are to be suppressed by 40 dB, this will require a corresponding suppression already at 28 MHz below the required frequency, in the case of a 28 MHz signal. Such demand can only be fulfilled when using multi-stage filters; one is, however, easily able to accept a somewhat higher insertion loss of 1 to 1.5 dB. Since the signal processing for the 23 cm, 13 cm and 9 cm band is usually based on a 2 m signal, the selectivity at a spacing of 144 MHz should also exhibit sufficient suppression.

3. INTERDIGITAL FILTERS

So-called interdigital filters represent the least amount of mechanical construction since the resonance elements intercouple directly so that intermediate panels are not required, as well as virtually no coupling elements. The input and output coupling can be made with mechanical components which are virtually the same as the actual resonant elements. Such filters allow large, relative bandwidths to be achieved. However, they exhibit excellent characteristics when designed as narrow-band filters, as is shown in the measured values.

By using available aluminium profiles, a construction method has been found that allows such filters to be easily manufactured for the SHF bands. No milling is required, and with the exception of a little skill, only simple tools such as a drill, tap, file etc. are required. If one has access to a lathe, the rest of the filter can be prepared in three to four hours. The coarse frequency alignment is made with the given length of the resonant elements, and the final alignment is made at the hot end with the aid of a 5 mm alignment screw. The input and output coupling are made with the aid of coupling elements, which are very similar to the resonant elements, and are directly connected with the aid of a 3 mm screw to the inner conductor of the coaxial connector **(Figure 1)**.

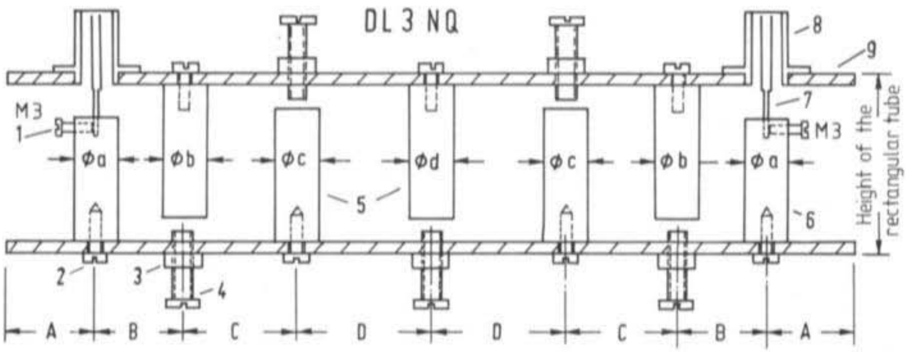

Fig. 1: The principle of the construction of the filter (dimensions given in the table)

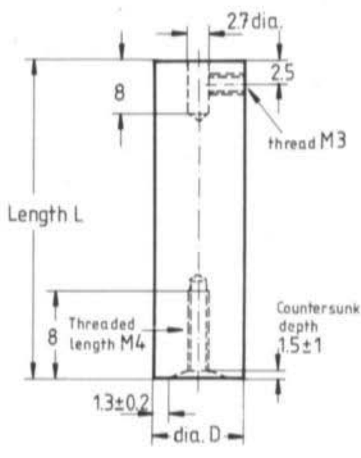

Fig. 2a: Coupling element »a«

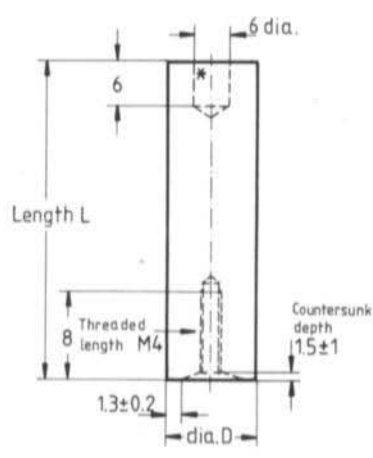

Fig. 2b: Resonant element »b-d«

*) Only drill when the filter is to be aligned to f_0 – 144 MHz
Not required for 9 cm

4. CONSTRUCTION

4.1. Mechanical Construction Details

Hollow aluminium profiles having a rectangular cross section are used as »case« of the filters. These are readily available at hardware stores. The narrow sides are provided with holes of 4 mm dia. and 5 mm threads alternately, or provided with holes for the coaxial sockets. It is more practical for the center line and the center of the filter to be marked and to then mark the position of the other resonant elements from the position of the center resonant element. After this, the holes should be slightly countersunk so that the drill does not slide away. It is advisable for holes of approximately 1.5 to 2 mm dia. to be drilled at first. Afterwards, they can be drilled out to the nominal value of 4 mm. This means that all holes on the center line have now a diameter of 4 mm.

The two outer holes can now be drilled out to the required diameter for the coaxial socket. This is followed by tapping a 5 mm thread into the two 4 mm dia. holes at each side of the center hole. After fitting the coaxial sockets, it is necessary for the burrs to be removed on the inside of the profile. This is not very simple, and care must be taken not to spoil the surface onto which the resonant elements are to be mounted. This process will be very much simplified when a new, well-sharpened drill is used for making the 4 mm dia. holes, in order to ensure that the holes are not pressed in but cut cleanly. The burrs of the outer holes can be removed quite easily, however, a tool must be made for the other holes, for instance, by sharpening a long screw driver and cutting of the burrs carefully.

The hollow profiles can remain electrically open at both ends as long as the filter is not in the direct vicinity of the tank circuit of the power amplifier.

In order to achieve a low insertion loss, it is important that the ohmic losses are kept as low as possible. This means that the resonant elements should be mounted to the inside of the profile at low resistance. This means that a smooth, clean surface connection is required and a good surface pressure. This is achieved by screwing the 4 mm screws tightly, and the former by counter-sinking the resonant elements on the tapped side so that they only possess an edge of approximately 1.3 mm **(Figure 2)**.

Of course, this can be achieved easily when using a small lathe. If one has no access to such a lathe, a certain amount of patience is required when using a relatively fine flat file and several sheets of emery cloth to ensure that the surface is at right angles and is as smooth as possible. A vice should only be used for drilling, for finer work, a small bracket or similar should be used.

Both the conductivity and surface quality of the resonant elements is important with respect to loss. For simplicity, round brass bars can be used for these elements, since it possesses a very uniform surface. The small disadvantage of the relatively poor conductivity can be avoided when the brass parts are copper-plated before mounting, which is not very difficult. Experiments have shown that such copper plating reduces the insertion loss by approximately 0.2 dB.

It is somewhat more elegant for the input and output coupling, as well as the resonant elements to also be made from aluminium. It is, however, necessary in this case for the surface of the elements to be polished, since such aluminium rod is usually damaged due to its softness. This is achieved using a waterproof emery cloth (360 or finer). This can be done by placing a 4 mm rod in an electric drill onto which the elements can be screwed one after another. By adding a little oil, the elements can be polished to give a uniform mat surface. It would be possible to polish them still further, but this would only have an optical effect and will not improve the electrical characteristics. The required 5 mm and 4 mm dia. nuts and 5 mm dia. aluminium screws are available readily in hardware stores, which means that all critical parts of the filter can be made from the same material. This is advantageous with respect to corrision under unfavorable ambient conditions.

Another possibility would be to copper-plate the profiles in addition to the elements. This should cause no problems for a good galvanizing workshop. Experiments on the prototype filters have shown that the improvement in conductivity can reduce the insertion loss by a maximum of 0.4 dB. Another possibility would be in using a little silver between the element and the side of the profile.

Virtually any connectors can be used. The prototype filters were partly equipped with BNC-connectors and partly with N-connectors. The mounting of such and similar connectors is not difficult. A hole of 10 mm diameter is drilled for the BNC-connector type UG-1094/U instead of the 5 mm threaded holes for the tuning screws. The connector is panel-mounted using the supplied nut. In the case of N-connectors, it is necessary for four 3 mm dia. tapped holes to be drilled for mounting. The central hole will probably require a somewhat different diameter according to the manufacturer. It should be only just big enough that the N- (or C) connector is provided with a good ground contact.

After completing the preparations, the resonant elements are mounted one after another commencing at the center and screwed tightly using the 4 mm diameter x 10 mm cylinder-headed or countersunk screws. The input and output coupling elements are provided with a central hole of 2.7 mm dia. at the upper end, as well as with a 3 mm dia. thread at the side into which the inner conductor of the coaxial socket is clamped later. After this, they are mounted into the profile after the resonant elements.

Before mounting the conductors into place, the inner conductors are lengthened either by soldering a metallic tube (hollow rivet of 2.5 mm dia., or brass tube from a ball-point pen), or a piece of copper wire (silver-plated) of approximately 2.6 mm in diameter. A length of approximately 8 to 12 mm is required so that the inner conductor can be clamped together with the 3 mm screw of the coupling element after the connector has been mounted into place.

4.2. Design Details

The table gives all mechanical details for eight prototype filters that should be constructed as shown in Figure 1. These values are the dimensions of the aluminium rectangular profiles, and the lengths and diameters of the elements, as well as their center-to-center spacings within the cavity. The non-variable dimensions of the round elements are given in Figure 2.

The table also gives the 3 dB bandwidths, the insertion losses at the given center frequency f_o (e.g. at 1296 MHz, 2304 MHz, and 3456 MHz), as well as the stop-band attenuation at the frequencies of special interest (e.g. f_o minus 28 MHz, f_o minus 56 MHz, and f_o minus 144 MHz). The stop-band attenuation 288 MHz below f_o is better than 60 dB in the case of the three-stage filters and better than 80 dB in the case of the five-stage filters. Finally what the type of surface the prototype filters have is given: A = brass elements in aluminium profile, B = all parts from aluminium, and C = surface of elements and profiles copper-plated.

No.	Type of filter	Rectangular tube Height x Width x Wall thickness	Spacing between elements (mm) A	B	C	D	Dia. of the elements (mm) a	b	c	d	Length of the elements(mm) a	b-d	Δ f (3 dB) (MHz)	Attenuation values in dB f_o	f_o − 28MHz	f_o − 56MHz	f_o − 144 MHz	Sur-face
1	23 cm 3-stage	60 x 34 x 3	43.0	24.0	43.0	00.0	15	10	11	00	47	48.5	31	0.5	17	35	62	B
2	23 cm 3-stage	60 x 34 x 3	36.0	29.0	50.0	00.0	15	10	12	00	47	48.5	15	1.2	42	61	> 100	A
3	23 cm 5-stage	60 x 34 x 3	36.0	28.0	51.0	56.5	15	11	11	12	47	48.5	7	2.2	80	> 80	> 100	A
4	23 cm 5-stage	60 x 34 x 3	36.0	28.0	51.0	56.5	15	12	12	15	47	48.5	8.5	1.8	76	> 80	> 100	A
5	13 cm 3-stage	40 x 30 x 4	40.0	21.0	39.0	00.0	12	9	9	00	25	26	29	0.95	18.5	37	62	A
6	13 cm 3-stage	40 x 30 x 4	40.0	21.0	39.0	00.0	12	9	9	00	25	26	28	0.5	19	38	63	C
7	13 cm 5-stage	40 x 30 x 4	26.0	20.0	38.0	42.0	12	9	9	10	25	26	17	0.9	44	75	> 90	C
8	9 cm 5-stage	30 x 30 x 3	25.0	23.5	42.5	47.0	12	10	10	11	17	15.3	48	1.0	14	54	> 90	A
	9 cm 5-stage	30 x 30 x 3	27.0	26.5	45.5	51.0	13	10	11	12	17	15.3	Passband curve approx. as for filter No. 7					

Table: Dimensions and Data of 9 different filters

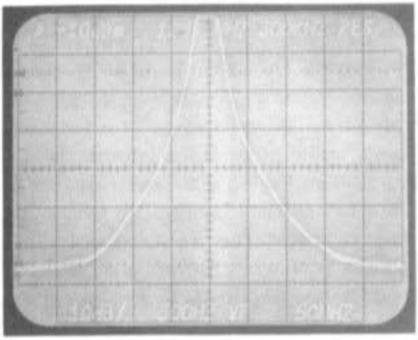

Fig. 3:
Filter no. 1 (23 cm)
hor. 50 MHz/cm, vert. 10 dB/cm

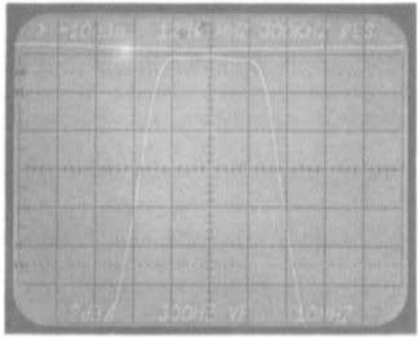

Fig. 4:
Extended display filter no. 1
upper line: 0 dB reference
h. 10 MHz/cm, v. 2 dB/cm

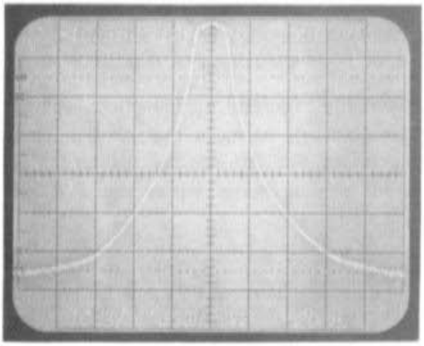

Fig. 5:
Filter no. 2 (23 cm)
h. 20 MHz/cm, v. 10 dB/cm

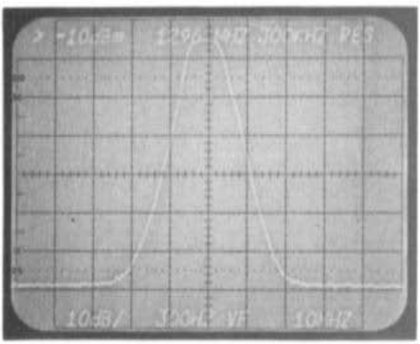

Fig. 6:
Filter no. 4 (23 cm)
h. 10 MHz/cm, v. 10 dB/cm

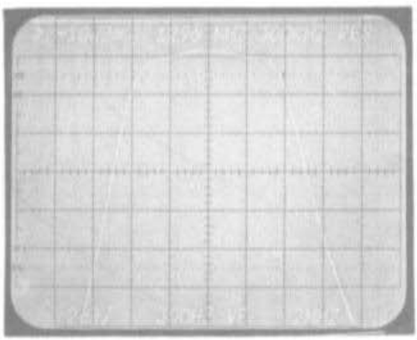

Fig. 7:
Extended display filter no. 4
upper line: 0 dB reference
h. 2 MHz/cm, v. 2 dB/cm

4.2.1. Bandpass Filter for the 23 cm Band (1152 - 1296 MHz)

Unfortunately, the author found no ideal aluminium profiles for the 23 cm band at hardware stores, which meant that the dimensions of the 5-stage filter with 340 mm in length was somewhat unhandy. However, the 3-stage filter with a length of 230 mm was quite acceptable. Aluminium rectangular tubing of 60 x 34 x 3 mm was used, which possess the internal dimensions of 54 x 28 mm. Filters No. 1 and No. 2 are 3-stage filters having a different degree of coupling, No. 3 and No. 4 are 5-stage filters using the same basic design, by which the coupling has been varied slightly using different element diameter. The prototype filters 2, 3 and 4 possess brass elements mounted in an aluminium profile. Prototype filter No. 1 is completely made from aluminium. As has been mentioned previously, the insertion loss could be reduced by 0.2 to 0.4 dB by use of copper-plating.

Filter No. 1 is an excellent receive filter, e.g. for use in front of a mixer; filter No. 2 is especially designed for use as a transmit filter, and is just as suitable for filtering the output signal of a local oscillator chain, since the filter can also be tuned to 1268 MHz or 1152 MHz. **Figures 3 to 5** show the selectivity curves of filters No. 1 and No. 2, as well as the passband range of filter No. 1 (enlarged). **Figures 6 and 7** show the selectivity curves as well as the passband range of the 5-stage filter No. 4 in a similar manner. This filter possesses a very steep slope, and is therefore suitable for suppressing spurious signals from ATV transmitters which are operating in the frequency range of 1250 - 1280 MHz according to the band plan.

The highest selectivity is offered by filter No. 3, however, it possesses an insertion loss of 2.3 dB. At a frequency of $f_0 \pm 20$ MHz, the attenuation amounts to more than 60 dB. This means that this filter can be used with success for suppressing RADAR interference which is becoming more and more prevalent. However, this is only possible when the RADAR frequency is not directly within the amateur band.

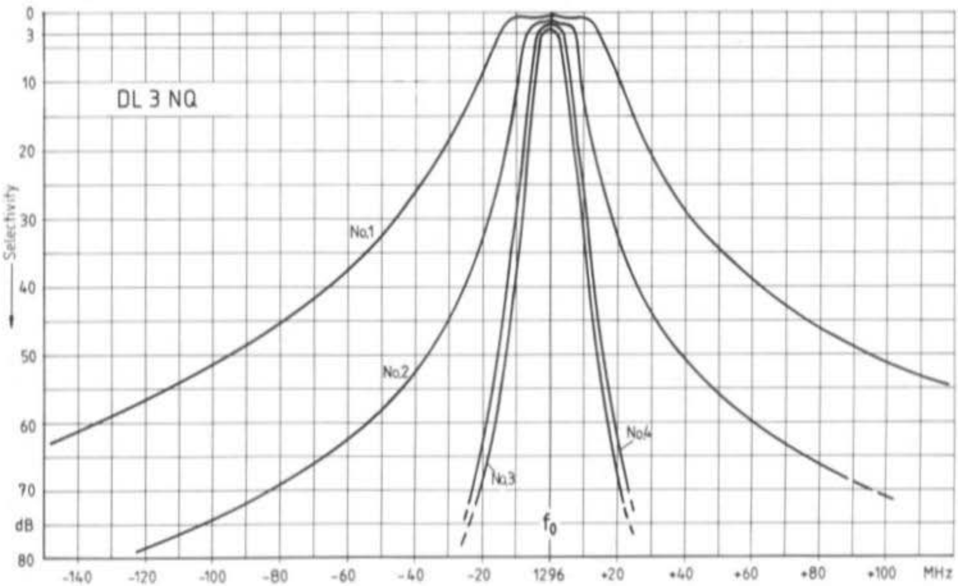

Fig. 8: Selectivity curves of the 23 cm bandpass filter

Since the different scales used in the oscilliscope traces make it difficult to make direct comparisons, the swept frequency curves are given again in the same scale in **Figure 8**.

Some of these curves were made using a swept frequency system, whereas others were made point-for-point using an automatic measuring system and printed out numericaly. The specific attenuation values that are given in the table were not taken from the swept frequency curves, but measured with the aid of an accurate signal generator and thermal-coupler powermeter.

4.2.2. Bandpass Filter for the 13 cm Band (2160 - 2304 MHz)

Aluminium rectangular tubing of 40 x 30 x 4 mm can be used for the 2304 MHz filter; the inner dimensions are 32 x 22 mm. This results in a length of approximately 250 mm even for a 5-stage filter. The resonant elements are 26 mm long, and the input and output coupling elements are 25 mm. The other dimensions can be taken from the table.

Experiments were made during measurements on these prototype filters in order to establish what effect a copperplating of the inner and outer conductors had on the insertion loss. The values have already been given. The main differences are to be found in the insertion loss in the passband range, however, no noticeable difference in the stop-band range was noticed.

As has already been mentioned, no really high selectivity values are required in the receive mode in order to obtain the required image rejection. More important is a low insertion loss. For this reason, a closely coupled 3-stage filter is recommended. This filter possesses a 3 dB bandwidth of approximately 29 MHz when not copper-plated, and already provides virtually 20 dB attenuation at the SSB spacing of 28 MHz. The insertion loss is less than 1 dB. The

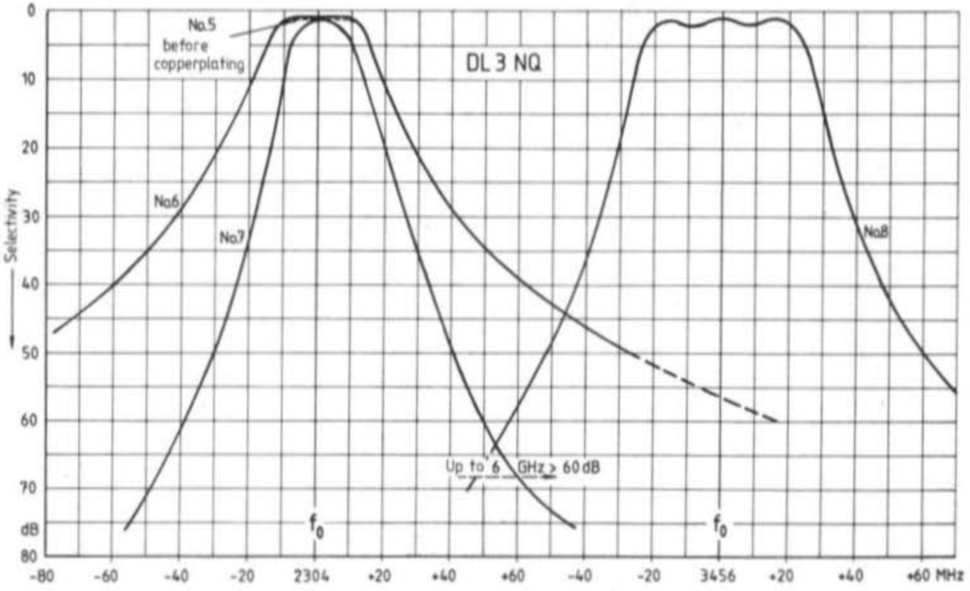

Fig. 9: Selectivity curves of the filters for 13 cm and 9 cm

5-stage filter is more loosely coupled and provides a high selectivity even when not copper-plated that is satisfactory for virtually all applications. Since no swept frequency system was available for this frequency, the given values were taken from the print-out of another professional measuring system. The resulting curves are given in **Figure 9**. When one compares the design of filters No. 1 to No. 4 with those of filters No. 6 and 7, it will be seen that it would not be too difficult to extrapolate element spacings and diameter for a filter having even looser coupling. This will result in a filter having an insertion loss of less than 2 dB, and a 3 dB bandwidth of approximately 12 MHz. Otherwise, the information given in section 4.2.1. is valid.

The ultimate selectivity at very high frequencies was also examined in the case of filter No. 7. It was found that the attenuation remained better than 60 dB up to 6000 MHz; on increasing the frequency still further, the rectangular tube gradually becomes a waveguide and more and more drops in attenuation are observed.

4.2.3. Bandpass Filter for the 9 cm Band (3312 - 3456 MHz)

The aluminium profile 30 x 30 x 3 mm can be used for the 9 cm band with inner dimensions of 24 x 24 mm. Unfortunately, these dimensions are also not quite ideal, and the large internal width means that the length of a 5-stage filter amounts to 276 to 300 mm. However, the advantage is a readily available, strong casing for the filter.

The resonant elements are only 15.3 mm in length and the input and output coupling elements 17 mm. The spacings and diameters are given in the table for filter No. 8. These are for a filter that already shows an over-critical coupling. This has led to a certain amount of ripple and to a widening of the 3 dB bandwidth, which is given on the right-hand side of Figure 9. If the center frequency of such a wide filter is aligned to the operating frequency, it will not be possible for the selectivity to be high 28 MHz below the center frequency. This is tolerable as long as a 28 MHz SSB system is not used, which is hardly likely in the case of a 9 cm system.

However, in order to give some basic design details for a narrow-band filter for 9 cm, spacings and diameters are given in the last line of the table which have been extrapolated from the values of filter No. 4 (23 cm) and No. 7 (13 cm). Unfortunately, it was not possible for the author to confirm these design values in practice before publication of this article.

5. INFORMATION REGARDING THE ALIGNMENT OF THE FILTERS

The alignment of multi-stage filters is still simple if the coupling of the circuits is under-critical, and when the dimensions guarantee that the frequencies of each stage are in the vicinity of the required frequency. The latter is the case due to the given element length of all filters, and the former is the case with filters No. 3 and No. 7. Such filters only need to be terminated with 50 Ω and aligned in several runs for maximum output power.

If less attention is to be paid to the best possible selectivity within the passband range, or if a lower insertion loss, or a wider bandwidth is required, it is necessary to try and obtain a critical coupling (flat passband) or a slightly over-critical coupling. This is the case with filters No. 2, No. 3, and No. 6. For the most optimum alignment of these filters, a signal generator with exact frequency adjustment will be required in addition to a correct termination, and a certain amount of patience.

Of course, the alignment can be made far more simply and quickly when using a swept-frequency system. This is virtually inavoidable for alignment of over-critically coupled filters with more than three stages. Fortunately this is not very often necessary in practice.

For amateur radio applications, one does not usually require a precision filter, but more the highest possible attenuation at a certain spacing from the required frequency, and the least possible insertion loss. An ideal flat top, or perfectly symmetrical slopes are not usually necessary (unless you are designing filters for publication, hi !).

Anyway, it is hardly ever the case that reactive components are not present at the input and output and are compensated for in the overall system. The main importance is to obtain a clean signal by alternately trimming the alignment screws; this can best be made with the filter connected to the system via coaxial cables that are then shortened or lengthened by $\lambda/4$. If one also has the possibility of checking the suppression of an unwanted signal, the values given in the table should be reached or bettered.

The author would like to thank H. Gunkel, DJ 5 ZE, W. Sauerwein, DL 8 JT and especially G. Schwarzbeck, DL 1 BU, for their great assistance especially with respect to the measurements on the filters under professional conditions.

SHF TRANSMIT CONVERTER WITH A VARACTOR DIODE WITH HIGH EFFICIENCY AND LOW INTERMODULATION

PART 1: THEORY OF FREQUENCY CONVERTION WITH NON-LINEAR REACTANCES

by H. Fleckner, DC 8 UG

Until recently, linear transmit converters (SSB, ATV) for the 23 cm and 13 cm band have virtually all used tubes such as the 8255, 2 C 39. Usually, a power mixer is used so that a reasonable power output is provided directly after the mixer. Diode mixers that are used in receive converters could also be used in transmit converters, but it is somewhat difficult for the radio amateur to amplify an output of approximately 0.1 mW up to several tens of watts. This two-part article is now to describe how output power levels of approximately 1 W can be achieved in a mixer stage at very low intermodulation levels without using tubes. This is possible using a power up-converter. Such circuits have been used in professional UHF-SHF technology for quite some time, but are still virtually unknown for amateur radio applications. A similar circuit was given in (7).

Part 1 of this article is to describe the principle of frequency conversion using non-linear reactances, and part 2 will describe a transmit converter 144 MHz / 1296 MHz comprising three modules. The local oscillator of this transverter operates from 12 V and provides at least 0.5 W at 1152 MHz. This means that an output power of approximately 0.3 W is available at 1296 MHz for an input drive level of approximately 0.2 W at 144 MHz. The intermodulation rejection will be better than 20 dB and spurious signals will be suppressed by more than 46 dB in the built-in filter.

1. INTRODUCTION

In contrast to ohmic resistances, reactances, both capacitive and inductive, store energy. Non-linear reactances are thus non-linear stores that are formed, for instance, by the junction capacitance of a semiconductor diode, or by saturation of an inductance having a ferromagnetic core.

Whereas non-linear resistances convert a portion of the generated or converted AC-voltage into heat, non-linear reactances convert this energy at virtually no loss, which means that a frequency conversion is, in principle, possible with an efficiency of 100 %. When converting the frequency using non-linear energy storage, it is even possible, under certain conditions, for a power gain to be achieved. It can be defined as the ratio of the input signal power to the output frequency power, or of the carrier frequency power to the output frequency power. The energy required is supplied by the carrier frequency, or signal frequency voltage.

Varactors are mainly used as reactances in such systems, and virtually always Schottky or PN-varactors in the form of junction or storage varactors. In the case of Schottky and junction varactor diodes, the voltage-dependent capacitance of a biased Schottky or PN-junction is used. In the case of a storage varactor, the diode is controlled between the flow area and the blocking direction of a PN-junction at such a speed that the minority carriers injected into the neighbouring conductor areas cannot recombine (1). The conductive area is the area outside of the junction; minority carriers are electrons in a P-semiconductor and vice versa. Recombination is the process by which free electrons are replaced in the holes.

Storage varactors are fed so that their characteristic curves (capacitance as a function of voltage) is more abrupt and thus more non-linear than would be the case with junction varactors. Such a characteristic curve is given in **Figure 1**. It will be seen that the charge storage due to the non-recombination of the minority carriers provides, theoretically, an infinitely large capacitance when driving it into the flow area. This causes current pulses with a high harmonic content that are responsible for the high efficiency when used for frequency multiplication and conversion. Furthermore, the relationship between the diode voltage and the diode charge caused by the current is cubed in the case of an abrupt PN-junction. This means that, theoretically speaking, no intermodulation products can occur. In practice, such mixers offer intermodulation rejections in the order of 40 dB.

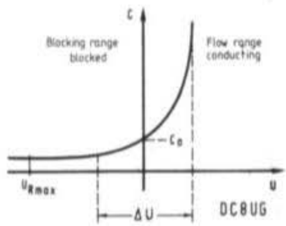

Fig. 1: Capacitance-voltage characteristic of a storage diode

2. PRINCIPLE OF FREQUENCY CONVERSION

One must differentiate between two different types of frequency converters or mixers: up-converters and down-converters. Up-converters convert a relatively low frequency f_1 (e.g. 144 MHz) with the aid of a higher frequency f_2 (1152 MHz) to a new frequency (1296 MHz). In the case of a non-inverting up-converter, the new frequency represents the sum of both these frequencies: $f_3 = f_1 + f_2$. In the case of inverted up-converters, the output frequency is the difference between these two frequencies: $f_4 = f_2 - f_1$. The circuit of the mixer is designed so that power is only converted to the required frequency.

It has been found that the inverted mixer circuit has a tendency to instability whereas, the non-inverting converter operates in an absolutely stable manner, (2). If the inverted mixer is therefore no longer considered, our attention can be concentrated on non-inverting up- and down-converters. The following designations are valid in subsequent considerations:

Input signal frequency f_1 (e.g. 144 MHz)
P_1: power at f_1

Oscillator frequency f_2 (1152 MHz)
P_2: power at f_2

Output frequency f_3 (1296 MHz)
P_3: power at f_3

The power conversion using non-linear reactance mixers can be calculated according to the power distribution laws according to Manley and Rowe. This was described in a general manner in (2); it gives the distribution ratio of an infinite number of powers in the case of just as many independent frequencies. No differentiation should be made here between input and output power, but only the amount of such. The following relationships are valid for non-inverting up-converters:

$$P_3/P_1 = f_3/f_1 \qquad (1)$$

$$P_3/P_2 = f_3/f_2 \qquad (2)$$

$$\text{with } f_3 = f_1 + f_2$$

The output power P_3 is greater than the power of the input signal to the same ratio as the frequency. The output power is also greater than the oscillator power P_2.

In the case of down-converters, it will be seen that a loss will be exhibited at the same ratio as the power gain in the case of up-converters. For this reason, conversion using non-linear reactances is only used with up-converters, and non-linear resistors are used for down-converters.

It is therefore advisable to concentrate on non-inverting up-converters. If equations (1) and (2) are used together with a 23 cm up-converter, the following relationships will result:

$$P_3/P_1 = 1296/144 = 9$$

In the case of an input frequency of 144 MHz, a conversion gain of 9 is obtained when referred to the output frequency. The following conversion gain factor results between the oscillator power and output power:

$$P_3/P_2 = 1296/1152 = 1.125$$

These ideal values are valid under two conditions:

The reactance must have no loss (cut-off frequency of the diode f_C virtually infinite);

The circuit should only convert power at the three frequencies.

Without discussing whether this can be realized in practice or not, the following values result as example:

An oscillator power of $P_2 = 1.6$ W is available. This means that the theoretical output power of the frequency converter is:

$$P_3 = P_2 \times 1.125 = 1.8 \text{ W}$$

The required input signal power is then:

$$P_1 = P_3/9 = 0.2 \text{ W}$$

The overall efficiency of the circuit, e.g. the ratio of output power to the sum of the input powers is, of course, 1.

In a second example, an input signal frequency of 28 MHz is to be considered. In this case, the conversion gain of the oscillator power to output power is lower ($1296/1268 = 1.02$), whereas, the signal power P_1 can be far less, ($P_3/P_1 = 1296/28 = 46.3$). This means that one requires less drive power for the same output power, but more oscillator power. Since the local oscillator power can only be obtained with considerbale effort and since a sideband spacing of only 28 MHz requires a considerable amount of selectivity, and since such selectivity will reduce the efficiency, the use of an input signal of 144 MHz is far more favorable.

In a third example let us study the relationships for the 13 cm band: when using an input signal of 144 MHz, a local oscillator frequency of 2160 MHz is required. The ratio $P_3/P_2 = f_3/f_2 = 2304/2160 = 1.07$, which we have designated RF efficiency, is less favorable than in the case of the 23 cm band. The required input signal power is thus one sixteenth of the output power. In the case of a local oscillator power of 1 W, the following will be valid:

$$P_1 = 0.06 \text{ W} \quad \text{and} \quad P_3 = 1.07 \text{ W}.$$

3. LOSSES ENCOUNTERED WHEN USING STORAGE DIODES FOR FREQUENCY CONVERSION

After describing the theory of a lossless, non-inverting up-converter using non-linear reactances, this section is to discuss the losses encountered when using real varactor mixers and to discuss their effect on the efficiency. These losses can be split into two areas:

Losses in the external circuit

Losses in the varactor itself

The losses occuring in the external circuit are caused entirely by the Q of the filters and matching circuits. These can be brought to a minimum using good UHF practice, selecting the right material, and by using large, smooth surfaces. The losses within the varactor diode can also be split into two groups:

Losses in the blocking range

Losses in the pass range

When operating in the blocked area, the diode can be assumed to operate with its junction capacitance in parallel with an ohmic resistance (3). A portion of the RF energy is converted into heat in this resistance. The value of this component is dependent on the current.

In the forward, or current flow direction various types of losses occur:

Recombination losses:

These are caused by the finite life of the minority carriers in the flow area: This means that a portion of the minority carriers recombine and cause a DC-current which causes these losses together with the connected DC-voltage (diode !). A correct selection of the storage time t_S with respect to the period duration of the RF-voltage can bring an improvement here, namely:

$$t_S > t_{HF}$$

Hysteresis losses:

These were examined in (3) in detail. They are caused by the after-effects of the residual charge on switching from forward to blocking direction and can be minimized by correct selection of storage time and switching time.

Losses due to conductivity modulation:

Overdriven diodes cause a modulation of the conductivity, which causes additional heat to be generated by the current.

The losses occuring in the flow range cannot be avoided with storage diodes. The improvement of the efficiency obtained by using the abrupt capacity-voltage characteristic in the flow range always provides better results than when using Schottky or junction varactors only biased in their blocking range (4).

When considering unavoidable losses, it is necessary for the individual parameters to be selected so that maximum efficiency is obtained. In the case of power up-converters, a good RF efficiency $\eta = P_3/P_2$ is sought in contrast to parametric low-signal mixers (e.g. varactor up-converters in receivers). This means that frequency f_1 (144 MHz), which has no effect on this efficiency, must provide a considerably higher drive to the varactor than voltages f_2 and f_3. Such operation is described as parametric operation using the modulated signal source as pump. Its power is used to gradually compensate for the losses. The reason for this was discussed in (1) and is as follows:

Less effort is required at low frequencies to produce high pump powers than at higher frequencies.

The impedance of the varactor is far higher at lower frequencies than at higher frequencies, which means that less power is required for the same voltage drive.

A mixer of this type can be considered to be a non-inverting up-converter pumped at IF level. If the varactor is driven at f_1 and f_2 at the same high level, one will leave the range of parametric operation, which must also be considered during the calculation (5). Such mixers are called power up-converters.

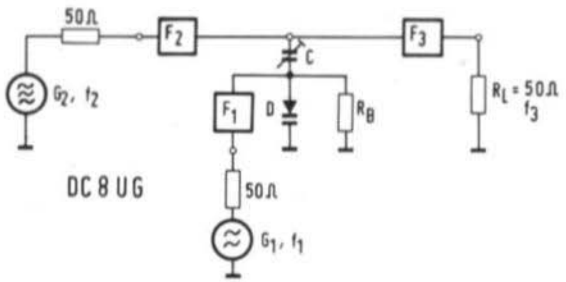

Fig. 2: Block diagram of the up-converter

4. BLOCK DIAGRAM OF THE MIXER

The block diagram of a mixer equipped with varactor diode is given in **Figure 2**. The diode is fed with oscillator power at frequency f_2 via a two-stage filter F_2 which also matches the generator impedance of 50 Ω to the impedance of the varactor. At the same time, the load resistor R_L is matched to the diode via a two-stage filter F_3 which is aligned to the output frequency f_3. Well-proven circuits used by frequency multipliers were used here. The capacitor C in series with the diode blocks the intermediate frequency f_1 (signal frequency) from the RF circuits. The IF signal is fed to the diode via a transformation circuit F_1. This can consist, for instance, of two resonant circuits, or an air-spaced coil as choke and using a toroid transformer for impedance conversion.

Such a mixer was examined in (1). The following favorable values were obtained using a storage diode type MA 4597 E (cut-off frequency 153 MHz, break-down voltage 63 V, mean capacitance 1.2 pF) at a signal frequency of f_1 = 70 MHz and an oscillator frequency f_2 = 2300 MHz:

RF efficiency $\eta = P_3/P_2 = 50\,\%$ with $P_3 = 28\ \text{dBm} \triangleq 0.6\ \text{W}$	
3 dB bandwidth:	12.5 MHz
Lower sideband:	– 53 dB
Second upper sideband:	– 39 dB
Harmonics of the IF:	– 60 dB
Intermodulation rejection (two-tone) :	40 dB

Due to the low IF, an auxiliary circuit was provided for $f_2 + f_1/2$.

These measured values were used as orientation values during the development of a mixer for the 23 cm band. It was reproducible in virtually all points. One can therefore assume that at least 50 % of the oscillator power is converted into output power. The signal frequency power and rest of the oscillator power are converted into heat.

In principle, such mixers can be designed for all amateur bands in the UHF and SHF range. They are especially interesting for applications in the 23 cm and 13 cm bands. Part 2 is to describe a mixer complete with local oscillator for the 23 cm band, and development and a description of such a converter for 13 cm is also planned for the future.

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A LOCAL OSCILLATOR MODULE FOR 200 mW AT 1152 MHz

by J. Dahms, DC 0 DA

The local oscillator module to be described here can be easily constructed. With its output level of approximately 200 mW, it is suitable for a large number of applications. It is, for instance, suitable for driving the 8255 tube in the DC 8 NR transmit converter described in (1); tube V 1, which is used in this transverter to increase the output level of the oscillator module DC 8 NR 006 to the required level for mixing, will no longer be required if the described local oscillator is used. Inspite of its higher output power, the new oscillator chain is more simple and easier to construct: only one power transistor is used, no parts are to be found on the lower side of the board, and the narrow $\lambda/4$ coaxial circuit has been replaced by a $\lambda/2$ stripline circuit. Of course, the local oscillator module shown in **Figure 1** is provided with screening panels made from tin plate or PC-board material. The screening is completed in a RF-tight manner by mounting the base plate with the aid of the eight screws shown in the photograph, and providing a top cover. A subsequent bandpass filter comprising two $\lambda/4$ circuits, as was described in (2), is used for selectivity. When constructed in this manner, the unwanted second harmonic of the varactor tripler (768 MHz) will be suppressed by more than 60 dB.

Another application will be for the 3 cm band: The module could be followed by a frequency multiplier (nine times) to end up with a frequency of 10 368 MHz. A further application would be to provide a low power transmitter for 1296 MHz.

If a 92 MHz crystal is used instead of the 96 MHz crystal, an output frequency of 1104 MHz will be obtained which is suitable for driving a frequency multiplier to 3312 MHz. This frequency will be required for 9 cm / 2 m transverters. Finally, if a crystal of 105.66 MHz is used, one will obtain a frequency of 1268 MHz which is suitable for 28 MHz/1296 MHz applications.

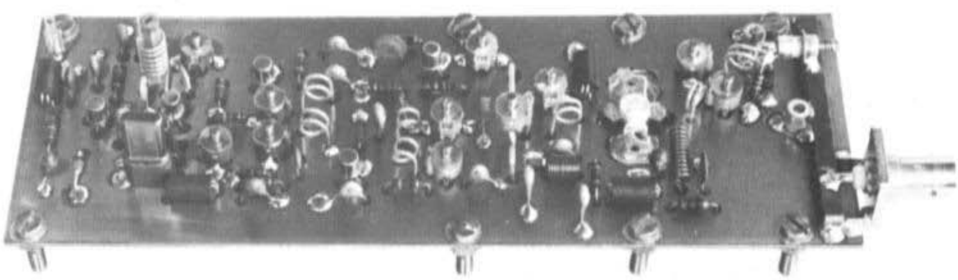

Fig. 1: 1152 MHz oscillator chain without casing

1. CIRCUIT, CONSTRUCTION AND ALIGNMENT

The circuit diagram of the module is given in **Figure 2**. The first few stages of the oscillator chain need not be described in detail since they are very similar to the local oscillator module described in (3). The crystal oscillator frequency of 96 MHz is doubled twice and the power level at 384 MHz amplified in three stages. The small power transistor C 1 - 12, which drives the varactor tripler via a matching network, is to be found in the last stage. A varactor diode type BB 105 is used which is used extensively in UHF television tuners. The output circuit for 1152 MHz is a capacitively shortened $\lambda/2$ air-spaced stripline.

If bridge Br 1, which usually grounds the crystal, is disconnected, it is possible for an additional circuit to be provided for narrow band frequency modulation. A suitable circuit for this was described in (1).

Figure 3 shows the component locations and the conductor lanes on PC-board DC 0 DA 005. The dimensions of this board are 195 mm x 70 mm; it is double-coated but does not possess through contacts. The ground surface on the component side of the board remains virtually intact so that a large ground surface is available. The copper coating is only removed with the aid of a 3 mm dia. drill where holes are made through the PC-board to the conductor lanes. This type of construction has been described a number of times in this magazine.

Sufficient room is provided around the edges so that tin plate or PC-board material can be soldered into place and used as screening panels.

2. COMPONENT DETAILS

T 1:	BC 108 or other AF transistor
T 2:	BF 173, BF 199
T 3, T 4:	2 N 5179, BFX 89
T 5:	BFW 30, BFY 90
T 6:	C 1 - 12
D 1:	BB 105 or BA 149
D 2:	C 9 V 1 zener diode

With the exception of L 6 and L 11, all inductances are wound from 1 mm dia. (18 AWG) silver-plated copper wire.

L 1 :	6 turns wound on a 5 mm dia. coil former with VHF core. Spacing between turns: 1 mm; hot end facing the board.
L 2, L 3 :	1.5 turns wound on 5 mm former; pulled out to fit the hole spacing on the board, spaced approximately 5 mm from the board.
L 4, L 5 :	1.5 turns on 4 mm former; pulled out to fit the hole spacing on the board; spaced approx. 2 mm from the board surface. Tap on L 5 approx. 3/4 turn.
L 6 :	Resonant line 1.5 mm dia. (15 AWG) silver-plated copper wire; straight length: 30 mm, ends bent back and placed through holes on the PC-board so that a spacing of 4 mm results. Tap: 11 mm straight length from the cold end
L 7 :	1.5 turns wound on a 5 mm former, pulled out to fit the spacing on the board, spaced approx. 2 mm from the board.
L 8:	1.5 turns, 5 mm former, close wound, spaced approx. 2 mm from the board.
L 9 :	1.25 turns on 6 mm former, pull out slightly
L 10 :	0.5 turns, 5 mm former, U-shaped bent towards the surface of the board; bent approx. 1 mm from the ground surface.
L 11 :	4 mm wide copper strip; straight length 60 mm, ends bent back at 90° and soldered to ground so that a spacing of 4 mm results. The trimmer capacitor is soldered in the middle of this line at the side, the tap for the trimmer capacitor to the varactor is approximately 13 mm of straight length to one end of the stripline; the tap for the BNC-connector approx. 10 mm straight length from the other end.

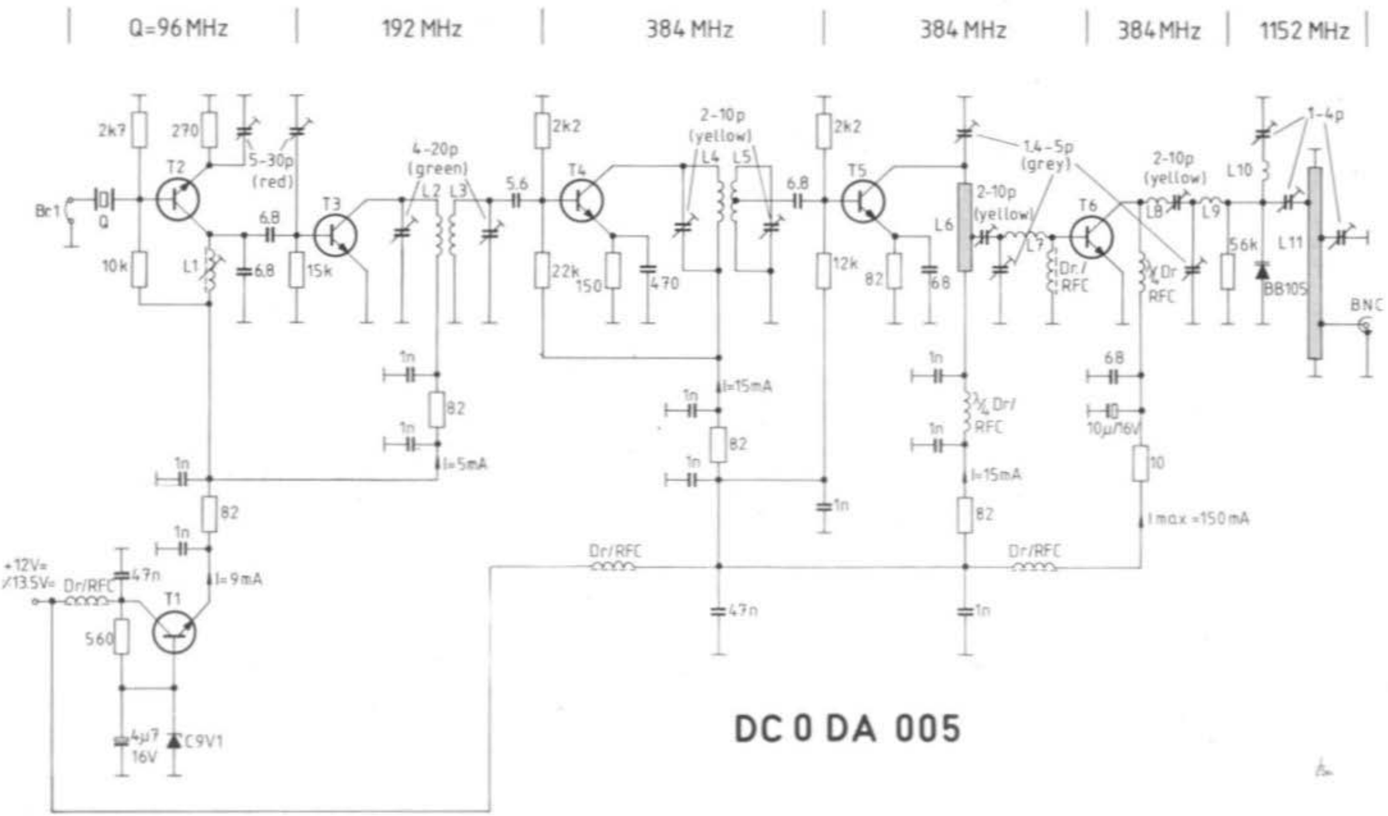

Fig. 2: Oscillator chain with 6 stages for 200 mW at 1152 MHz

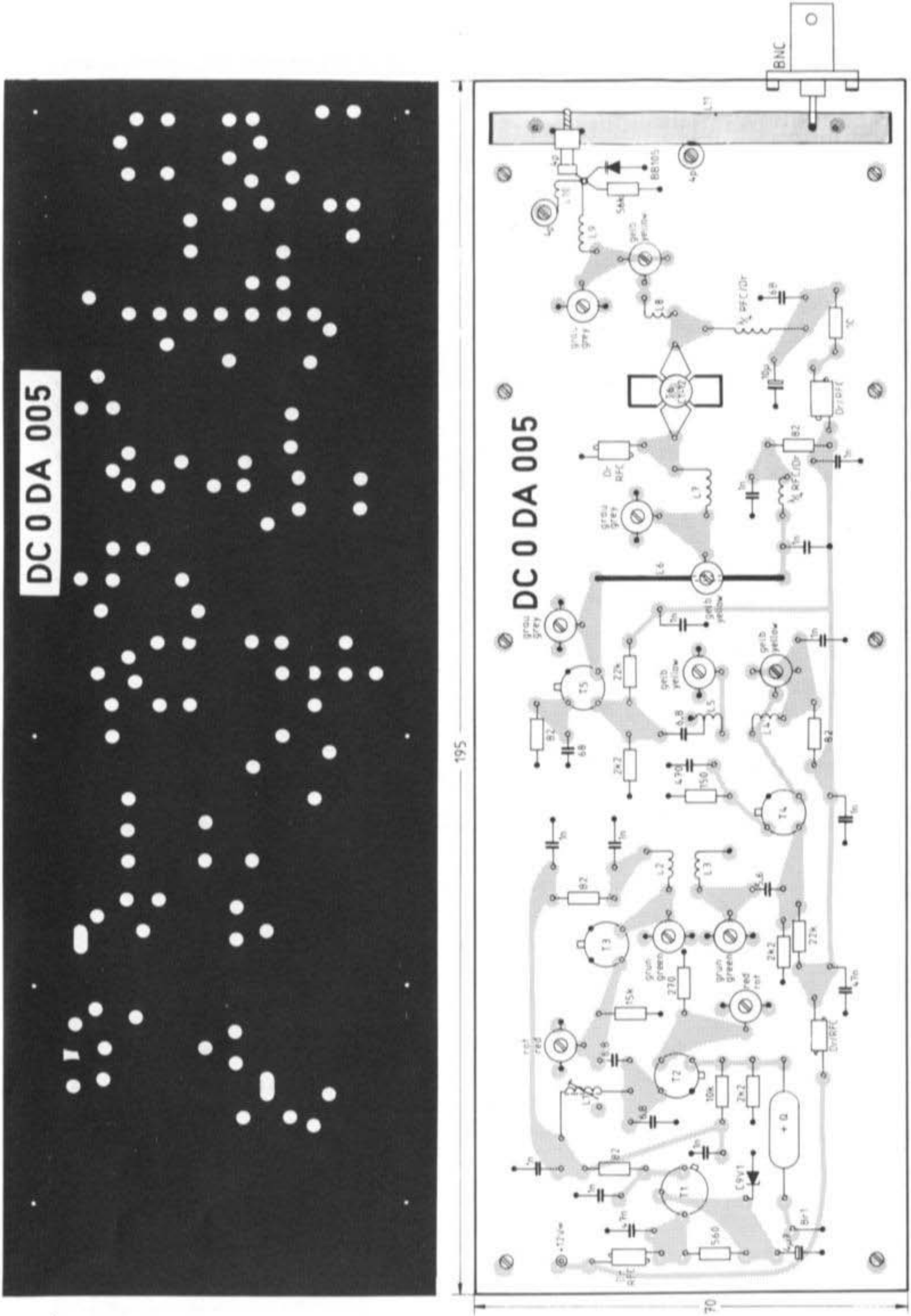

Fig. 3: PC-board DC 0 DA 005 for the oscillator chain

The trimmers for the idler circuit, coupling and tuning of the stripline L 11: small ceramic tubular trimmers, approx. 4 pF max. capacitance (Philips).

All other trimmers: plastic foil or ceramic disk trimmers of 7 mm dia. (Philips, Stettner).

All resistors: if possible capless carbon resistors for 10 mm spacing. All bypass and coupling capacitors: ceramic disk types for 5 mm spacing.

Required are further: 4 pieces, 6 hole core (Philips 4312 020 36700); 2 pieces λ/4 chokes constructed from 0.5 mm dia. (24 AWG) enamelled copper wire wound on a 3 mm former, self-supporting, wire length approx. 195 mm.

It is necessary during the alignment for all trimmers, commencing with the collector trimmer of T 5 to the tuning trimmer of the output stripline circuit, to be aligned several times alternately for maximum output power. Further details regarding the alignment of such stages have been given in the various references. In the preliminary alignment process, it is advisable for a bandpass filter for the output frequency to be used; the filter described in (2) is satisfactory. A sensitive power indicator simplifies the alignment of the module, even if it is not calibrated. The absorption frequency meter described in (4) is of great assistance since all frequencies present in this local oscillator module can be measured.

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LOOP-YAGI ANTENNAS

by R. Lentz, DL 3 WR

1. INTRODUCTION AND GENERAL DESIGN DETAILS

The elements of Yagi antennas can be in various forms: the most common is the linear type using straight elements. There are, however, also the quad antennas (square) with its derivatives ring loop (round) and delta loop (triangular). This article is to discuss the ring loop Yagi antenna. Such loop Yagi antennas possess elements having a circumference of one wavelength.

Controversy has existed for many years between the normal, linear Yagi fraternity and that of the loop Yagi and other quad antennas. Publications giving practical data (2, 3) and technical comparisons between both types of Yagi antennas have only tended to increase the controversy. Normally, the selection of an antenna is, however, made according to such practice as mechanical construction possibilities, appearance of the antenna, or cost. This article is to bring attention to a new paper (4) which gives detailed design details for loop Yagis with optimum gain.

The authors describe a very simple design procedure. Constant element spacings and element diameters are used. The spacing of the reflector could be increased somewhat (5), in order to improve the front-to-back ratio.

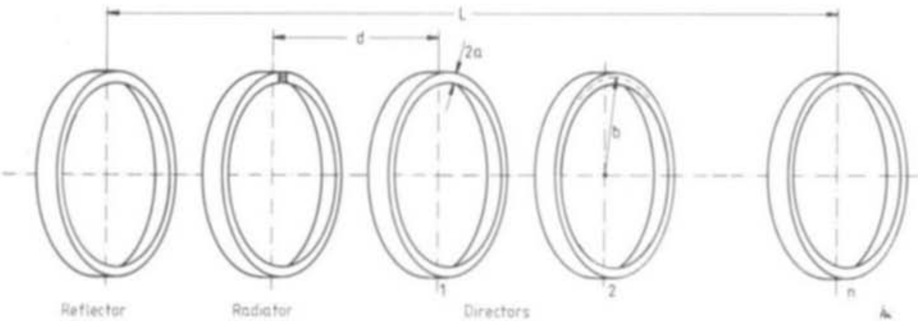


Fig. 1: A Loop-Yagi antenna according to (4), with loop thickness 2a, loop diameter 2b, loop spacing d, and antenna length L

Figure 1 shows the general form of a loop Yagi antenna. This article is now to discuss design details based on this:

Usually the basic parameter is antenna gain or antenna length. The curves given in **Figure 2** (from 4) show the bandwidth and gain (in dB) as a function of antenna length.

According to the required bandwidth, one of the three d/b curves is selected, which is normally $d/b = 1$ for amateur radio applications. After this, **Table 1** which was also given in (4), is used to obtain the ratios L/λ and b/λ for the required d/b ratio.

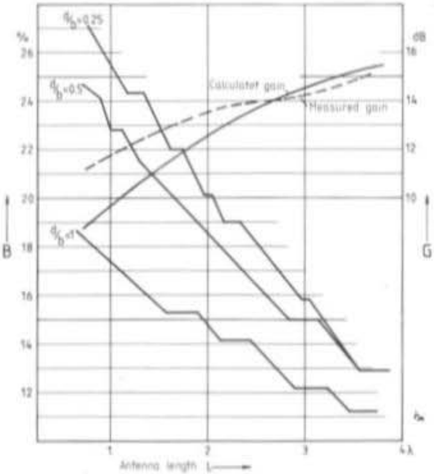


Fig. 2:
Design curves for Loop-Yagi antennas according to (4). The standardized antenna length L/λ results from the required gain. According to the required bandwidth, the required curve d/b is selected. All other values required can be taken from table 1

Ed. The bandwidth mentioned above and used later on in this paper should not be confused with the conventional bandwidth of an antenna and matching system (– 3 dB bandwidth). Meant is the transmission range of the loop Yagi structure for a specific wave type. The – 3 dB bandwidth of the antenna is considerably narrower. However, d/b = 1 is usually sufficient for amateur applications.

This is followed by calculating the wavelength λ for the center of the required band:

$$\lambda = \frac{300\,000}{f}$$

With f in MHz, λ will be in mm.

It is possible after calculating the λ-value to calculate value b, followed by the thickness of the loop material 2 a, and finally the spacing between the loops d. The number of elements including reflector is obtained by dividing the approximate antenna length by spacing b.

d/b = 1		d/b = 0,5		d/b = 0,25	
L/λ	b/λ	L/λ	b/λ	L/λ	b/λ
0,73 - 0,87	0,146	0,78 - 0,98	0,142	0,81 - 1,00	0,140
0,88 - 1,44	0,145	0,99 - 1,45	0,140	1,01 - 1,40	0,138
1,45 - 2,55	0,143	1,46 - 1,99	0,138	1,41 - 1,80	0,137
2,56 - 3,36	0,142	2,00 - 2,51	0,137	1,81 - 2,18	0,135
3,37 - 4,03	0,140	2,52 - 3,28	0,135	2,19 - 2,55	0,135
		3,29 - 3,92	0,134	2,56 - 3,17	0,132
				3,18 - 3,65	0,131
				3,66 - 3,84	0,129

Table 1: Design data for Loop-Yagi antennas according to (4)
a/b = 0.01; L is the antenna length

Example:

The wavelength corresponds to 235 mm in the case of a loop Yagi with a center frequency of 1275 MHz. At an antenna length of 3.5 λ the lowest bandwidth is in the order of 10 %, or 130 MHz, and the gain is approximately 15 dB (Ed: presumed to be dB_i).

The ratio d/b = 1 is taken from Table 1 at L = 3.5 λ : b/λ = 0.140

It is now possible to calculate the loop radius: b = 0.140 x λ = 32.9 mm

Loop diameter 2 b = 65.8 mm, and loop circumference C = 2 π b = 206.7 mm

With a = 0.01 x b, the following will result:
material thickness 2 a = 0.02 x 32.9 mm = 0.66 mm
Spacing between loops: at d/b = 1, it follows that d = b = 32.9 mm

Number of elements including reflector: $N = \frac{L}{d} = \frac{3.5\,\lambda}{d} = \frac{822.5\text{ mm}}{32.9\text{ mm}} = 25$

To summarize, the following values are obtained:

Frequency:	1275 MHz	Loop diameter:	65.8 mm
Bandwidth:	130 MHz	Loop circumference:	206.7 mm
Gain:	15 dB_i	Material thickness:	0.7 mm
Antenna length:	approx. 85 cm	Loop spacing:	32.9 mm
Number of elements:	25		

1.1. Construction Details

Table 2 contains data for a short and a long loop Yagi antenna for both the 2 m and the 70 cm amateur band. The elements can be made from metal tape, tube, or solid rod. It is only necessary for the diameter and the thickness 2a to be maintained. The metallic boom can pass through the center of the loops and the loops be mounted using insulated supports, or the loops can be directly mounted to the metallic boom at the point of minimum voltage maximum current). Such an antenna is shown in **Figure 3**.

A sophisticated method of making such an antenna would be to use printed loops on PC-board material. However, it would then be necessary for the effect of the board material on the resonant frequency to be determined, and taken into consideration. If an insulated boom is used, it will be possible for the elements to be mounted simply using a dual-component glue. Further details regarding loop Yagi antennas were given in (5) and (6).

Parameter	2 m band		70 cm band	
Gain (dB_i)	> 11	15	> 11	14
Loop radius (mm)	288	279	96	95
Loop thickness (mm)	6	6	2	2
Element spacing (mm)	288	279	96	95
Loop circumference (cm)	181	179	60.4	60
Antenna length (λ)	1.7	4	1.7	3
Antenna length (cm)	320	780	105.6	201.3
Number of elements	12	29	12	21
Bandwidth (MHz)	22	16	65	56

Table 2: Design data for Loop Yagi antennas as a basis for experiments

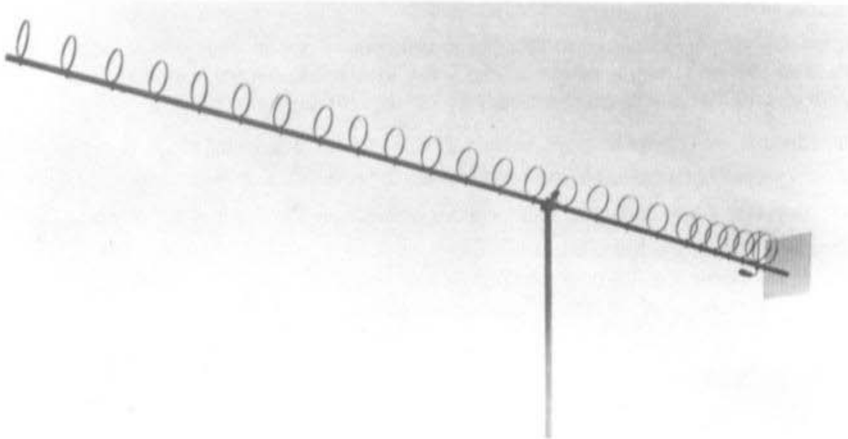


Fig. 3: Photograph of a Loop-Yagi antenna for the 23 cm band with a published gain of 20 dB

2. A LOOP YAGI ANTENNA FOR THE 23 cm BAND ACCORDING TO (7) AND (8)

Such a loop Yagi antenna for the 23 cm band has been marketed in England for several years and has proved to be popular. A photograph of this antenna is shown in **Figure 3**. The manufacturers give a gain of 20 dB for such an individual antenna. The construction details given in **Figure 4** show that constant loop spacings are not used in this case but differing spacings, as is often the case with conventional long Yagi antennas. In addition to the loop-reflector, a rectangular reflector comprising a metal grid is used. The element mounting is very simple: they are directly screwed to the boom. The balun system used is very clever: it consists of semi-rigid coaxial cable (9) having an impedance of 50 Ω and is fed via the boom after a length of approximately $\lambda/4$. It is then soldered through the mounting screw of the radiator loop so that it possesses ground potential at this position.

2.1. Construction Details

The construction of this antenna is simple, but the dimensions must be exactly maintained. The thickness of the loop material is critical. It is possible for it to be thinner than given, but the gain will drop noticeably if thicker material is used.

Since an error of 2.5 mm corresponds to approximately 1 % of the wavelength, or approximately 13 MHz, it is recommended when marking the holes on the boom to always measure from the same position. It is possible, for instance, for all positions to be marked from the radiator element, so that measurement errors are not added.

The loop strips (copper for the radiator, and aluminium for the other elements) overlap. The two holes are drilled with a spacing corresponding to the given circumference. The strips are not formed into rings until after the drilling process.

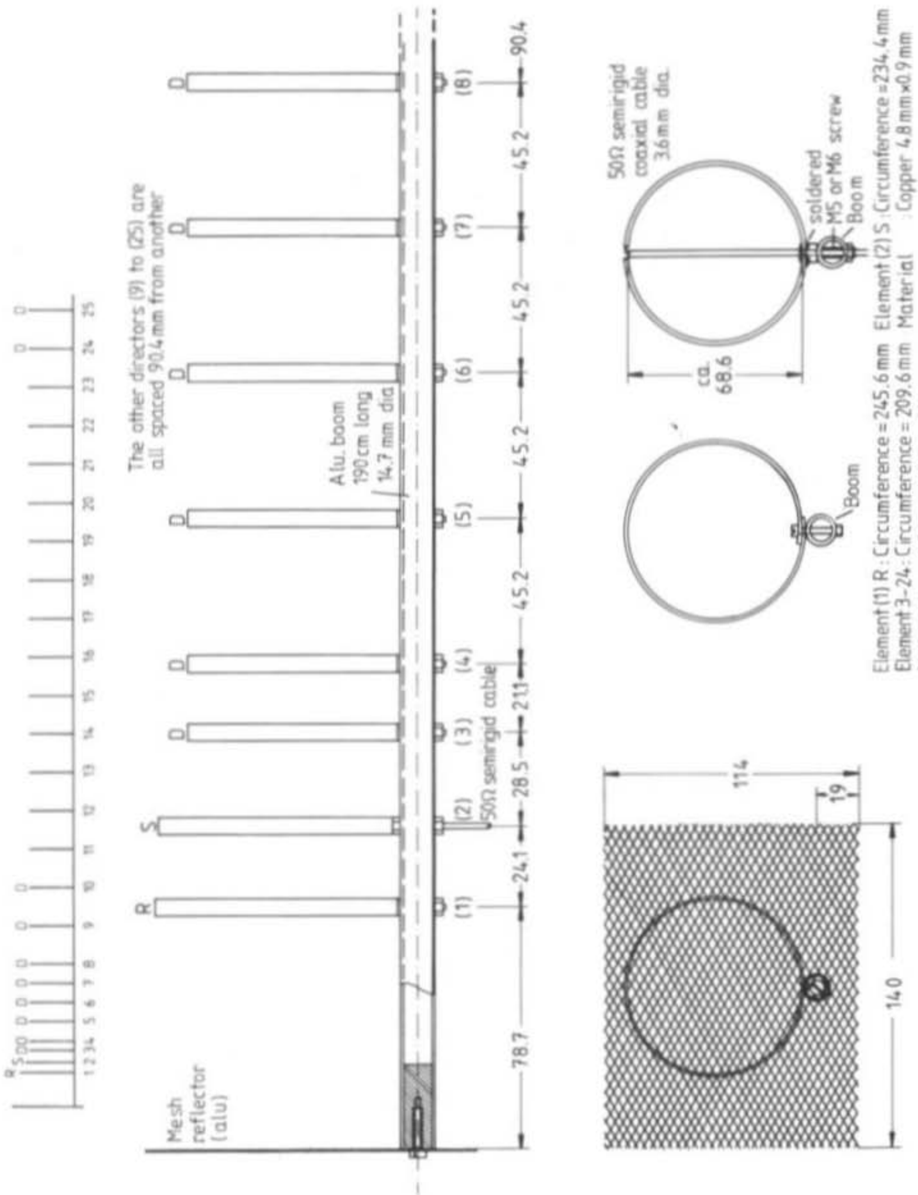


Fig. 4:
Construction details of the antenna shown in Figure 3. Dimensions for 432 MHz and for 2304 MHz are obtained by exact recalculation of all dimensions in the same relationship as the wavelengths

The radiator loop, which is manufactured from copper strip, possesses a 3.6 mm hole through which the semirigid coaxial cable is passed. The radiator loop remains open at the top and is then depressed so that it is approximately 6 mm wider than high. The matching and gain of the antenna are dependent on this measure, in other words on the length of the balun. One end of the radiator loop is soldered to the outer conductor, and the other to the inner conductor of the coaxial cable. In order to obtain the given gain, it is necessary to feed the semirigid cable through the boom. In order to do this, a screw is drilled through lengthwise and the free end of the coaxial cable is passed through it. Loop and cable are then soldered to the head of the screw after the most favorable matching and/or highest gain has been determined by adjusting the length of the balun. After placing this part through the boom, the height of the depressed radiator loop will be somewhat higher due to the screw head and will then coincide approximately with the axis of the other loops.

Finally, the copper radiator, all screws, and all soldered joints should be protected with a suitable polyurethan-lacquer after which the whole antenna can be protected with a conventional lacquer.

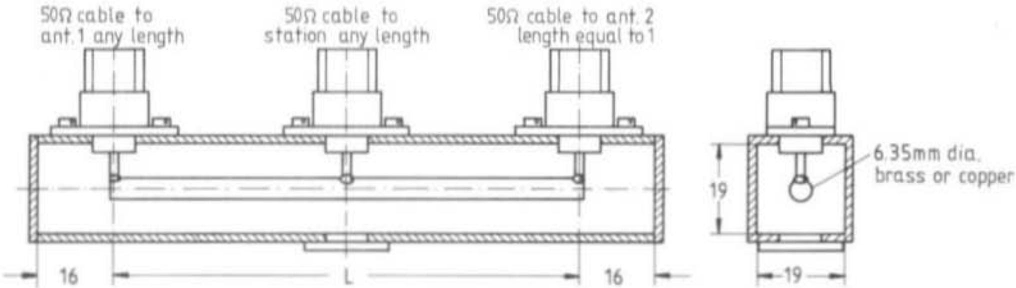
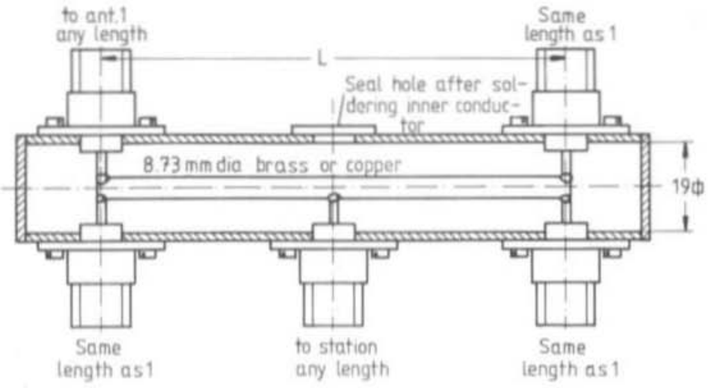


Fig. 5: A 50 Ω power divider/combiner for interconnecting two antennas



Connectors: N or BNC

Frequency MHz	Length L mm
2305	65
1297	116
433	347

Fig. 6: A 50 Ω power divider/combiner for interconnecting four antennas

2.2. Stacking the 25 Element Loop Yagi

The described antenna can be stacked in the vertical or horizontal plane, or even four used in an array. Many experiments made by G 8 DIC and G 3 JVL (8) have shown that 3 λ is the most favorable stacking distance. This corresponds to approximately 70 cm in the case of the 23 cm band. It is known that mast tubes interfere less with the radiated field when they are at right angles to the polarization plane. For this reason, the mounting structure for a four-antenna array should not be square but in the form of an H.

The power divider/combiner shown in **Figure 5** is suitable for matching and combining two antennas. This unit is designed for an impedance of 50 Ω. The two 50 Ω feeder cables to the antennas can be of any length, but their electrical length must be equal. The feeder cable to the station should also be of 50 Ω impedance, its length is not critical.

A power divider/combiner for four antennas operating according to the same principle is shown in **Figure 6**. In this case, four identical antennas can be combined onto a single feeder cable, or four identical power output stages to a single output.

2.3. Applications at Other Frequencies

Both the loop Yagi antenna shown in Figure 4, and the power divider/combiner can also be constructed for the 70 cm, or the 13 cm band. It is only necessary, in the case of the antenna, for all dimensions to be recalculated in relationship to wavelength. In the case of the power dividers/combiners it is only necessary for the length L to be recalculated. As is given in (7), such re-dimensioning has been successful.

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A NEW TYPE OF PREAMPLIFIER for 145 MHz and 435 MHz RECEIVERS

by M. Martin, DJ 7 VY

Very rapid advances have been made in the last few years in the development of semiconductors. It seemed that the field effect transistor (FET) was getting more and more popular over bipolar transistors, especially with respect to the cut-off frequency, and noise at UHF and SHF (e.g. GA-FET of NEC or Plessey). However, the bipolar transistor is still of importance in receive technology, especially since a new type of feedback allows them to be made more linear and exhibit less noise than a FET.

1. INTRODUCTION

The most important criterions for judging the quality of receivers are the sensitivity and the intermodulation rejection. Sufficient selectivity can be easily obtained using crystal filters. A high intermodulation rejection can be best achieved using high-level Schottky mixers and matching amplifiers between mixer and crystal filter (1). Input circuits can be built using frequency separation filters for separating the output components of the mixer: oscillator frequency f_0 minus receive frequency $f_{in} = IF_1$ to the amplifier and $f_0 + f_{in} = IF_2$ to a 50 Ω resistor.

Such circuits possess an SSB noise figure F = 10 ≙ 10 dB and a third order intercept point (IP) of approx. 30 dBm (the term IP was explained in detail in (1)). There are mixers available that possess an IP of 40 dBm ≙ 10 W ! (450 $), but their applications are limited due to the IP list of the crystal filterm which is in the order of 25 to 28 dBm. Since the mixer noise figure of 10 dB can be improved easily using a preamplifier, and since the overall IP of the system is reduced to the value of the preamplification minus selectivity losses, only as little preamplification as necessary should be used. If the gain is fixed at approximately 20 dB, the mixer noise figure component will be approximately one hundredth of the overall noise figure, thus 0.1 in the case of F = 10, which is sufficiently low.

The overall IP would be 10 dBm with the required high IP of the preamplifier, which represents an excellent value in comparision to many of the commercially available units that operate in the order of approximately – 30 to – 20 dBm.

2. EXAMINED CIRCUITS

Up to now, virtually only preamplifiers equipped with FETs were used for high-performance receivers. When using dual-gate MOSFETs, intercept point (IP) values of 4 dBm with noise figure F = 1.8 ≙ 2.5 dB could be achieved. Enhancement FETs such as SD 201 provided somewhat better values with IP = 6.5 dBm and F = 1.6. However, the best results were obtained using the preamplifier given in **Figure 1**. This preamplifier possesses the following specifications:

Gain: G = 18.5 dB (144 MHz), 19.5 dB (440 MHz)
Noise figure F_{144} = 1.35 ≙ 1.3 dB, F_{440} = 1.6 ≙ 2.0 dB
Third order intercept point: IP = 14 dBm (14 + 19 = 33 dBm at the output)
1 dB compression: 18 dBm ≙ 63 mW output power
Isolation: 22 dB
Bandwidth: 70 to 570 MHz, see **Figure 2**
Input SWR: 144 MHz: 1.4; 440 MHz: 2.8
Intermodulation-free, dynamic range: 102 dB (see appendix)
Operating voltage: 12 V / 21.5 mA

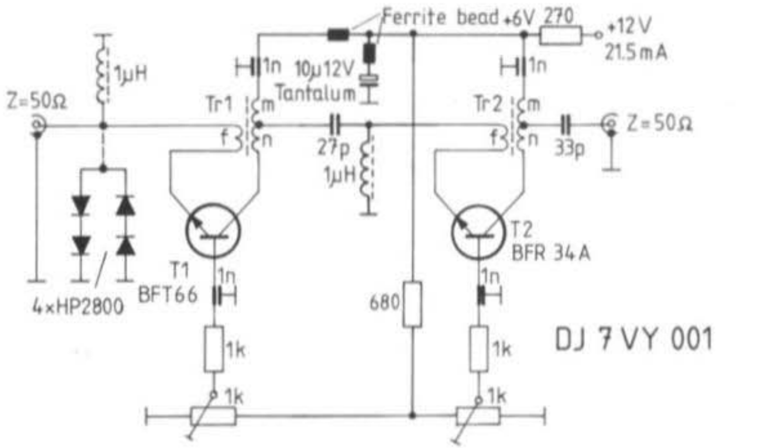

Fig. 1: Low-noise 145 MHz and 440 MHz preamplifier with high linearity

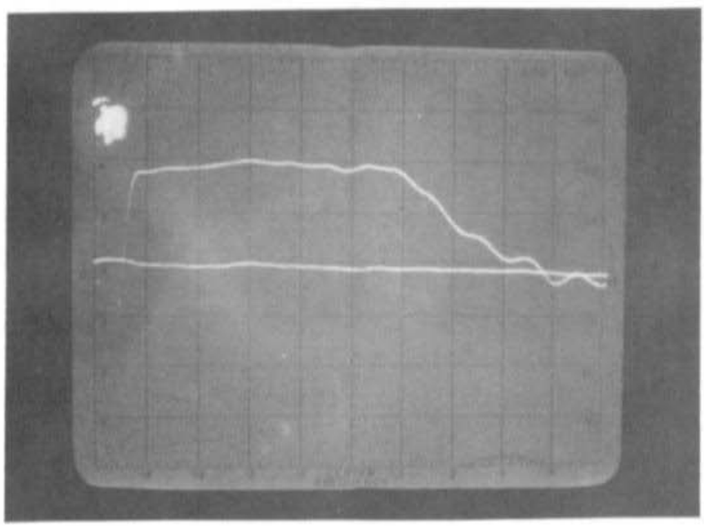

Fig. 2: Frequency response of the amplifier shown in Figure 1
Horizontal: 100 MHz/division – Vertical: 10 dB/division

These excellent values result from the use of a new type of feedback principle (2), by which energy is fed back to the input of the transistor in antiphase using an exchange feedback winding in a similar manner as used in a Meißner oscillator. This is obtained without deteriorating the original noise figure of the transistor which would usually be the case when using conventional feedback circuits. The linearity increases in this manner as a function of the feedback f, m, and n are feedback windings of the transformers Tr 1 and Tr 2 (see Figure 2).

The following relationships are valid for the design of the circuit:

Gain $G = m^2$

$n = m^2 - 1 - m$ for $Z_{in} = Z_{out}$

$R_{\text{load of the transistor}} = (n + m)\, Z_{out}$ for $f = 1$

In the case of m = 2, 3, 4, gains of G = 6, 9.5, 12 dB result, and load impedances of 3, 8, and 15 Z_{out}. By suitable design of the circuit and cascading various different stages, virtually any IP values and noise figures can be obtained. Since the input impedance corresponds to the impedance connected to the output, the low input SWR values represent a further advantage with respect to FET preamplifiers whose noise matching and the resulting high input SWR cause an increase in cable attenuation between the antenna and preamplifier. **Figure 3** shows the frequency response of a BFR 34 A amplifier between 0 and 1 GHz using various types of feedback transformation. The 0 dB line is to be found at the center. Above this, a line will be seen with 4.5 dB gain showing the frequency response of a stage with m = 4, n = 2 and f = 2; the upper frequency limit of approximately 1 GHz was reduced to 600 MHz after noticing a tendency to oscillation after connecting a bandpass filter. This was neutralized using a 1.5 pF capacitor from collector to ground. The next curve shows a version with m = 3, n = 5 and f = 1 with 9.5 dB gain, which is for a two-stage circuit as shown in Figure 1. The upper curve shows a version with m = 4, n = 11, and f = 1 having 12 dB gain and using 0.12 mm enamelled copper wire.

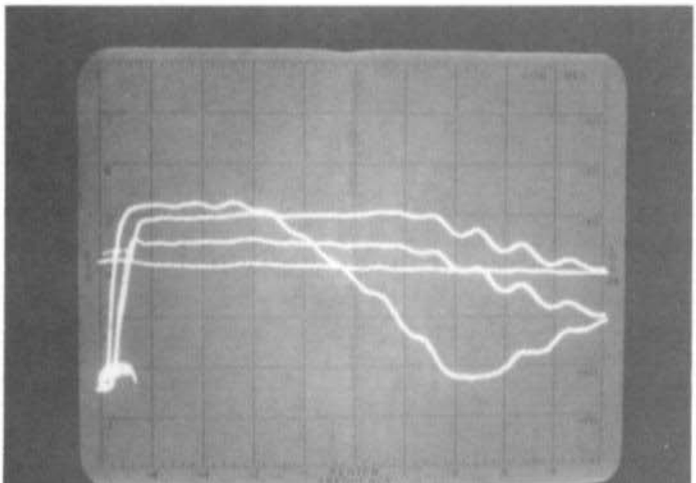

Fig. 3: Wideband amplifiers with different degrees of feedback
Horizontal: 100 MHz/division – Vertical: 10 dB/division

During experiments to find the best method of feedback transformation, it was found that the best results were provided using two-hole ferrite cores (Siemens B 62152 A 8 - X 17, material U 17). Transformers manufactured using these cores exhibit very low loss, a close coupling and very low stray inductivity, so that the gain values determined in practice differ very little from those calculated theoretically. The cores should be woundas shown in **Figure 4** with the input winding f on the inside, winding m above it, and winding n on the outside with the largest diameter.

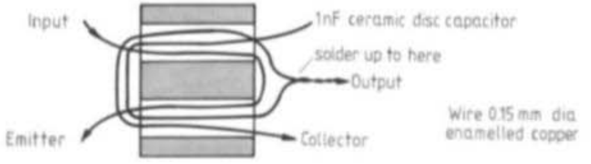

Fig. 4: Feedback transformer for wideband amplifiers

The lower frequency limit of this type of transformer feedback can be reduced to less than 1 MHz by using suitable cores and larger coupling and by-pass capacitors, as well as using higher inductivity for the emitter chokes. The upper cut-off frequency is in the order of 700 MHz, however, this method can still be used above this frequency using directional couplers as long as the delay of the signal through the active element is not too great.

Figure 5 shows the selectivity curves of various types of filter that can be used in front of and behind wideband amplifiers. In order to obtain the required image rejection for the receiver, a filter as shown in **Figure 6** should be used between mixer and amplifier. If the linearity in the direct vicinity of strong broadcast transmitters should not be sufficient inspite of the high IP value, a bandpass filter as shown in **Figure 7** or a high-pass filter as shown in **Figure 8** can be used, however, the noise figure will be reduced by the value of the insertion loss of the filter.

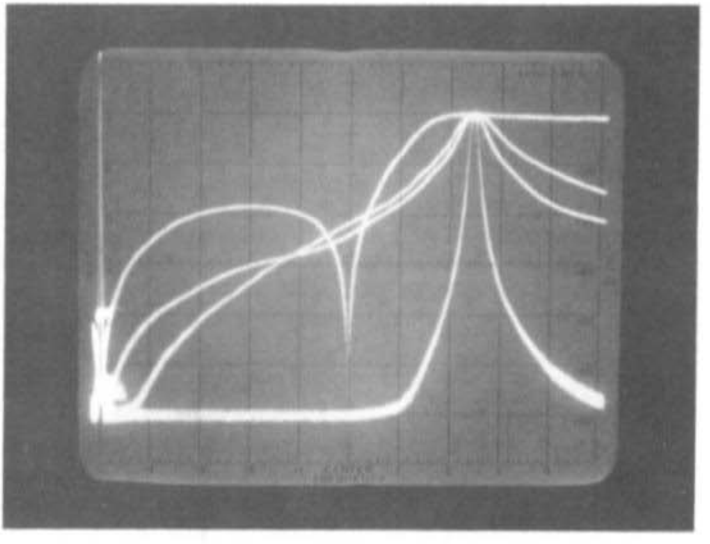

Fig. 5: Selectivity curves of the filters shown in Figures 6, 7, and 8, as well as a – 0.1 dB helical filter
Horizontal: 20 MHz/division – Vertical: 10 dB/division

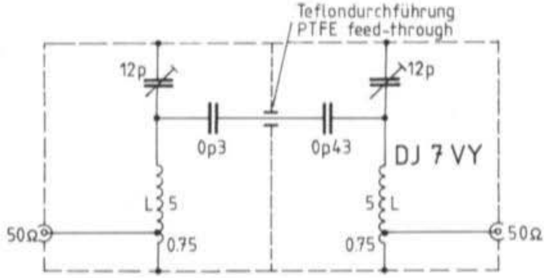

Fig. 6:
145 MHz bandpass filter
L: 5 turns of 2 mm dia. silver-plated copper wire, length 25 mm, 17 mm dia.
case: 60 x 50 x 100 mm
insertion loss = 1.34 dB
selectivity at 127 MHz: 50 dB
selectivity at 100 MHz: 76 dB
1 dB bandwidth: 2 MHz
3 dB bandwidth: 2.7 MHz

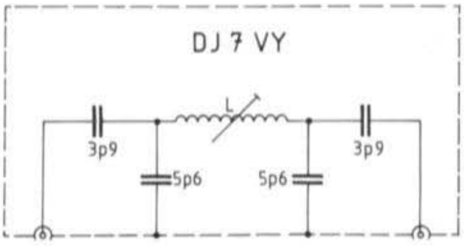

Fig. 7:
145 MHz input circuit
L: 4 turns of 2 mm dia. silver-plated copper wire, length 12 mm, 20 mm dia.
aligned by pulling
case: 60 x 50 x 100 mm
insertion loss = 0.25 dB
selectivity at 127 MHz: 14 dB
selectivity at 100 MHz: 24 dB
1 dB bandwidth: 4 MHz
3 dB bandwidth: 8 MHz

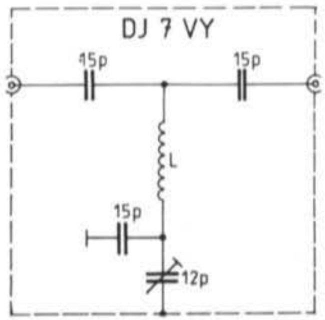

Fig. 8:
130 MHz high-pass filter
L: 3 turns of 2 mm dia. silver-plated copper wire, length 8 mm, 14 mm dia.
align trimmer to 100 MHz
case: 60 x 50 x 100 mm
attenuation: 145 MHz: – 0.06 dB
134 MHz: – 1 dB
130 MHz: – 3 dB
100 MHz: – 57 dB
Stop band range: – 30 dB: 97 - 103 MHz
Stop band range: – 40 dB: 99 - 101 MHz

If the amplifier is to be used at 435 MHz, it will be necessary for filters to be provided at the input and output, e.g. a $\lambda/2$ filter at the input, and a bandpass filter comprising two $\lambda/4$ circuits in front of the mixer.

3. CONSTRUCTION

The amplifier can be constructed using PC-board DJ 7 VY 001 whose dimensions are 45 mm x 35 mm **(Figure 9)**. The BFT 66 is placed through a 4.9 mm hole in the board at point »x«, and the transistor is directly soldered to the ground surface (as quickly as possible). When mounting the transformer, the connection wires to emitter, collector and to the disk capacitor should be kept as short as possible (less than 3 mm). The connection leads of the BFT 66 should be shortened down to 2 mm, which is also valid for the emitter and collector connections of the BFR 34 A which are bent up. The base connection of the BFR 34 A is directly soldered to the 1 nF disk capacitor and is also used for mounting. In the case of the transformers, attention should be paid that the twisted tap is tinned right up to the winding, preferably after winding f and m + 1 are mounted. The operating points are adjusted with a mA-meter in the associated collector lead: T 1 = 6 V / 5 mA, T 2 = 6 V / 13 mA. After this, the preamplifier is ready for operation. Caution: Due to the DC-feedback via the 270 Ω resistor, the adjustments will interact.

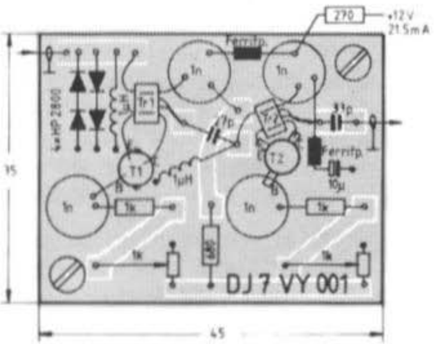

Fig. 9: PC-board DJ 7 VY 001

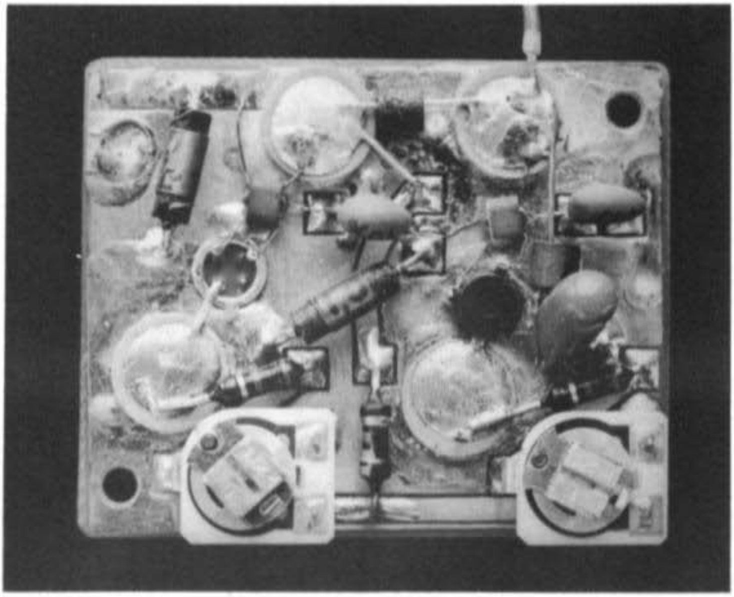

Fig. 10: Photograph of the author's prototype

The filter connection can be made with the aid of a short piece of coaxial cable, remembering that any additional tenth of a dB on the input side will reduce the noise figure by the same value. If the amplifier is to be mounted at the antenna together with an antenna relay one can expect an increase in sensitivity in the order of 3 dB even when a low-loss cable of 21 mm diameter has previously been used. Example: 30 m RG 17 with 1 dB at 145 MHz, additional attenuation due to SWR = 3: = 0.5 dB, noise figure of the FET preamplifier F ≙ 2.5 dB, = 4 dB (and far more when using inferior cable). In comparison, the amplifier would provide 1.3 dB plus insertion loss of the filter.

When the preamplifier is mounted at the antenna, the input should be protected using four antiphase Schottky diodes. If the output coupling capacitor is deleted, the supply voltage can be fed via the coaxial cable, which can be normal TV cable or similar. In this case, the coupling capacitor of 27 pF between the two stages should be reduced to 18 pF and a 270 Ω resistor connected in series with a 2.7 µH choke at the end of the cable. The coupling capacitor in the shack should amount to 1 nF. If a very cheap, longer cable having an attenuation of 6 to 8 dB is used, it is advisable for a BFR 34 A or BFT 66 amplifier with winding m = 3 to be used in front of the image rejection filter.

Finally, it should be mentioned that this amplifier is also excellent for use as a VFO amplifier for high-level Schottky mixers requiring 17 dBm ≙ 60 mW oscillator level. Due to its very low noise figure, this amplifier will not deteriorate the sideband noise of the local oscillator.

4. APPENDIX

Dynamic range of an amplifier (see (1))
Natural noise threshold: − 174 dBm/Hz bandwidth
Receiver bandwidth 2.4 kHz: 2.4 kHz / 1 Hz ≙ 33.8 dB
Preamplifier with F = 1.3 dB and input IP = 14 dBm
0 dBm = 1 mW at 50 Ω ≙ 223.6 mV
Sensitivity threshold S = − 174 + 33.8 + 1.3 = − 138.8 dBm ≙ 25.4 nV

Input power of two signals whose third order intermodulation products correspond to the sensitivity threshold:

P_{in} = 1/3 (2 x IP + S) = > 1/3 (28 - 138.9) = − 36.96 dBm ≙ 3.17 mV

Intermodulation-free dynamic range: 138.9 − 36.96 = 101.94 dB !

This means: Two signals of 3.17 mV at the input of the preamplifier result in two intermodulation products of 25.4 nV and thus corresponding to the noise floor of the preamplifier with F = 1.3 dB and a bandwidth of 2.4 kHz.

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ANTENNA SPLITTING FILTER FOR BROADCAST AND 144 MHz

by J. Kestler, DK 1 OF

Usually two antennas are required when a car radio and a 2 m mobile station are installed in a vehicle. Experiments have shown, however, that the car radio will work just as well from a 5/8 λ 2 m antenna. This 2 m antenna was mounted on the luggage compartment of the automobile, and the conventional car antenna (980 mm long when extended) used for comparison was mounted on the front wing. It was found that a λ/4 rod antenna for 145 MHz mounted on the roof of the vehicle was considerably better for VHF-FM broadcast reception (87.5 to 104 MHz), but slightly worse than the car radio antenna in the long and medium wave range. The home-built receiver (1) was used for VHF-FM broadcast reception, and commercially available car radio for MW and LW.

This article is to describe an antenna splitting filter that allows a 2 m mobile antenna (λ/4 or 5/8 λ) to be used simultaneously for broadcast reception. Construction and alignment are relatively simple, and no special measuring equipment is required.

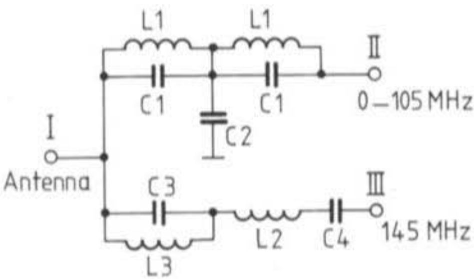


Fig. 1:
An antenna splitting filter for broadcast/2 m

1. CIRCUIT OF THE FILTER

The circuit diagram of the splitting filter is shown in **Figure 1**. It will be seen that each of the signal paths commences with a parallel resonant circuit. The circuit L 1 / C 1 of the upper path is tuned to block the 2 m frequency, whereas L 3 / C 3 is aligned to approximately 95 MHz which is the center frequency of the VHF-FM broadcast band. The upper signal path represents a complete low-pass filter using a T-link; it possesses deeper slopes, and an attenuation pole at 145 MHz. The equivalent circuit diagram for 95 MHz is given in **Figure 2**. The parallel circuit comprising L 1 / C 1 has an inductive effect here, since this frequency is below the resonant frequency of 145 MHz. This results in a low-pass filter in conjunction with C 2, which possesses a passband in the order of 95 MHz. The resonant circuit L 3 / C 3 isolates the 2 m station connected to III from the VHF-FM broadcast signals from the antenna, and C 4 provides the isolation at the lower frequencies for the medium and long-wave bands.

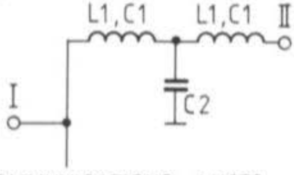


Fig. 2:
Equivalent circuit diagram for 95 MHz

Fig. 3: Equivalent circuit diagram for 145 MHz

The equivalent circuit diagram for 145 MHz is given in **Figure 3**. The resonant circuit comprising L 3 / C 3 is tuned to 95 MHz and has a capacitive effect at 145 MHz. Inductance L 2 extends the lower signal path to form a series resonant circuit which forms a low-loss path for 2 m frequencies to the antenna. **Figure 4** shows the attenuation curves for the signal path I-II, and **Figure 5** for path I-III. The lower passband at approximately 60 MHz is of no practical use, but results from the circuit used. A two-pole comprising four reactive impedances (L 3, C 3, L 2, C 4) always has three resonance positions (in our case two series and one parallel resonance).

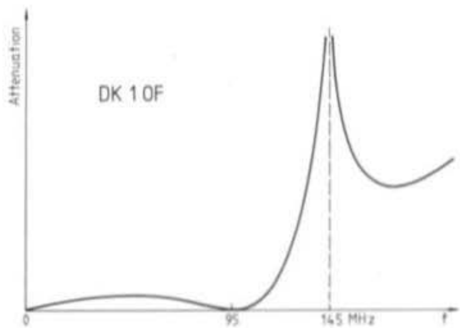


Fig. 4: Attenuation curve path I - II

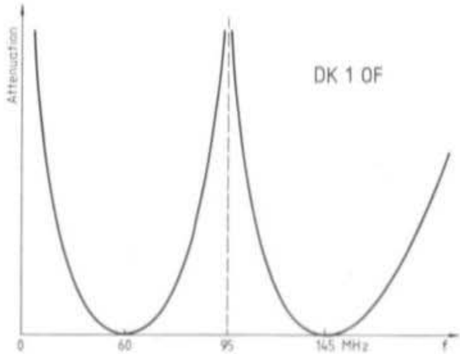


Fig. 5: Attenuation curve path I - III

2. MECHANICAL CONSTRUCTION AND COMPONENTS

Figure 6 shows a photograph of the antenna splitter. It is enclosed in a TEKO box type 372 which has been provided with four independent chambers. All coils are wound on 6 mm dia. Trolitul coil formers using 1 mm dia. silver-plated copper wire. Each of these coil formers is provided with a brown core. The component values are as follows:

	Calculated	Installed
L 1	47.8 nH	2 turns
L 2	54.9 nH	2 turns
L 3	36.5 nH	1.5 turns
C 1	25.2 pF	27 pF
C 2	33.5 pF	33 pF
C 3	76.9 pF	39 + 39 pF
C 4	43.9 pF	47 pF

Disk or tubular ceramic capacitors are suitable. For power levels in excess of approximately 25 W, it is necessary for capacitors to be used for C 3 and C 4 that have a sufficient breakdown voltage (100 V), and NPO-ceramic should be used.

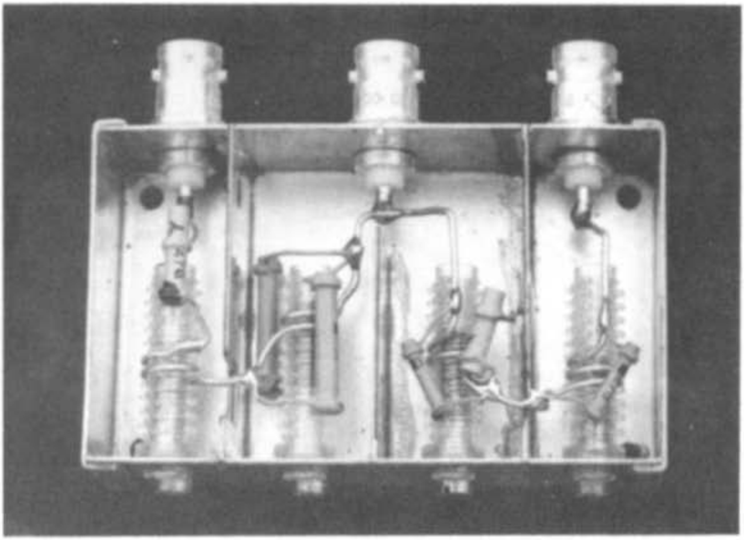


Fig. 6: Photograph of the author's prototype

3. ALIGNMENT

The alignment can commence after the filter has been wired and the cover mounted into place. It is not necessary for this to be made in the vehicle in conjunction with the mobile antenna, but may be made in the radio shack.

Firstly, connection I should be connected to a VHF-FM broadcast antenna, and connection III to the broadcast receiver. A strong signal in the order of 95 MHz is now selected and inductance L 3 is aligned for minimum field strength. After this, position I is terminated with a 50 Ω resistor and connection III via a reflectometer to a 2 m transceiver. Inductance L 2 is now aligned for best standing wave ratio at a frequency of 145 MHz. This completes the alignment of the lower path of the filter.

This is followed by connecting position I to a 2 m antenna and connecting the 2 m transceiver to II. A strong signal should be found in the vicinity of 145 MHz and the alignment cores of both inductances L 1 tuned alternately for minimum signal strength. Since the stop-band attenuation is in the order of 80 dB, it may be necessary to make a sked with a local station in order to obtain a signal of sufficient field strength.

If one should find that the core of any of the coils must be completely inserted or extracted, it will then be necessary to increase or reduce the number of turns of the coil in question by approximately a quarter or half a turn so that resonance is obtained reliably.

4. MEASURED VALUES

The described splitting filter was measured in a professional laboratory using a signal generator and level receiver. The following values were determined:

Signal path I-III :

Insertion loss	144 to 146 MHz	0.1 dB
Standing wave ratio	144 to 146 MHz	< 1.1
Stop-band attenuation	95 MHz	46 dB
3 dB bandwidth		15 MHz

Signal path I-II :

Insertion loss	90 to 105 MHz	0.1 dB
Insertion loss	80 to 110 MHz	< 0.3 dB
Insertion loss	0 to 30 MHz	< 0.1 dB
Stop-band attenuation	144 to 146 MHz	> 85 dB

In order to demonstrate the efficiency of the described splitting filter the following experiment was made: a relatively weak signal was tuned in on the VHF-FM broadcast receiver using a 4-element 2 m Yagi antenna. Even though a VHF output power of more than 400 W was fed to connection III at 145 MHz, no deterioration of the VHF-FM broadcast reception was observed. However, the coil formers of L 2 and L 3 melted after approximately two minutes of continuous carrier, since the splitting filter is not designed for such a high power level. After exchanging the inductances, the experiment was repeated using a power level of 100 W. After approximately ten minutes of continuous carrier, the case of the filter was only handwarm and the inductances showed no adverse affects. This power output can therefore be classed as the maximum power rating using this type of construction. The insertion loss of 0.1 dB corresponds to a power loss of 2 %, which means that 2 W will be converted to heat in this filter. Higher power levels are very seldom used for mobile operation.

5. APPENDIX

Of course, it is possible for the described splitting filter to be designed for other frequency ranges. It should, however, be noted that path I - III is relatively narrow-band (max. ± 5 % of the center frequency) and its frequency must be higher than that of path I - II. The equations required for recalculation are as follows:

$$L_1 = \frac{R}{2\pi f_1} \cdot \left[1 - \left(\frac{f_1}{f_2} \right)^2 \right] \qquad C_1 = \frac{1}{4\pi^2 f_2^2 L_1} \qquad C_2 = \frac{1}{2\pi f_1 R}$$

$$L_2 = \frac{R}{2\pi f_2} \qquad L_3 = \frac{R}{4\pi f_2} \cdot \left[\left(\frac{f_2}{f_1} \right)^2 - 1 \right] \qquad C_3 = \frac{1}{4\pi^2 f_2^2 L_3}$$

$$C_4 = \frac{1}{\pi f_2 R}$$

where: R = antenna impedance (50, 75 Ω)
f_1 = cut-off frequency of path I - II,
f_2 = center frequency of path I - III

Finally, an example for checking the calculation:

An 11 m CB mobile antenna (27.1 MHz) is also to be used for a car radio for long, medium and short-wave reception. The passband range for the car radio should go up to 10 MHz so that the 31 m short-wave band is included. The impedance is 50 Ω. When using the previously mentioned equations, the following values will result:

$L\,1 = 0.688\ \mu H$	$C\,1 = 50.3\ pF$
$L\,2 = 0.294\ \mu H$	$C\,2 = 318\ pF$
$L\,3 = 0.933\ \mu H$	$C\,3 = 272\ pF$
	$C\,4 = 235\ pF$

6. REFERENCES

(1) J. Kestler: A Stereo VHF/FM Receiver with Frequency Synthesizer
VHF COMMUNICATIONS (7), Edition 2/1975, pages 66 - 77

THE 70 cm FM TRANSCEIVER ULM 70
Part 4: Mechanical Construction and Wiring

by I. Sangmeister, DJ 7 OH / H. Bentivoglio, DJ 0 FW / H.J. Franke, DK 1 PN

This part of the description assumes that the reader requires the cabinet and individual pieces that are to be described. If, however, the reader wishes to combine the modules in part 2 and 3 in a different manner, he will only need the wiring diagram.

9. PARTS REQUIRED FOR COMPLETING THE TRANSCEIVER

1 Nickel-cadmium accumulator 12 V

Connectors:

Cn 1, Cn 2:	Miniature coaxial connectors
Cn 3:	BNC-connector for single-hole mounting
Cn 4:	Battery connector with switching contact (e.g. as DIN 45323)
Cn 5:	AF connector for microphone (e.g. DIN connector)

1 Silicon rectifier for polarity protection, e.g. 1 N 4001
1 silicon diode, e.g. BAY 94 or 1 N 4148
1 cabinet, e.g. type 9516.15 (Ettinger, 8 München 70, Florian-Geyer-Str. 1)
1 250-µA-meter for rectangular cutout 23.5 x 7.5 mm
1 relay, e.g. type 2 KH 3950 (HI-G d'Italia) or RH-12 (National)

1 32-Ω loudspeaker, 0.7 W, 40 mm x 18 mm

La 2:	pilot lamps 12 V / 30 mA for illumination of the S-meter
P 1 / S 1:	10 kΩ potentiometer, log., with switch, 4 mm dia. shaft (volume)
P 2, P 3:	10 kΩ potentiometer, log., 4 mm shaft (tuning)
S 2, S 4:	multi-position switch, single wafer, 2 x 5 contacts
S 3:	Sub-miniature toggle switch
S 5:	Toggle switch (ON-OFF)
S 6:	Miniature sliding switch

1 fuse 0.5 A, slow-blow, 5 x 20 mm, mounted on the sliding switch.

10. PREPARATION OF THE CABINET, AND OTHER PIECES

The front panel is firstly prepared as shown in **Figure 22**, the rear panel as in **Figure 23**, and the supplied mounting plate as shown in **Figure 24**. Of course, it is possible for the cutouts for the loudspeaker to be made in a more simple manner. After this, it is necessary for a few small pieces to be made:

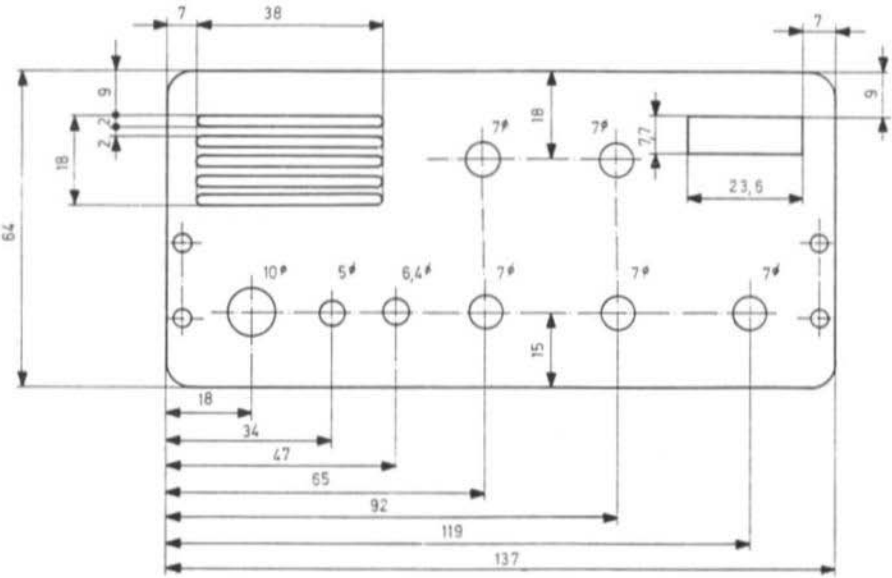


Fig. 22: Holes and cutouts in the front panel

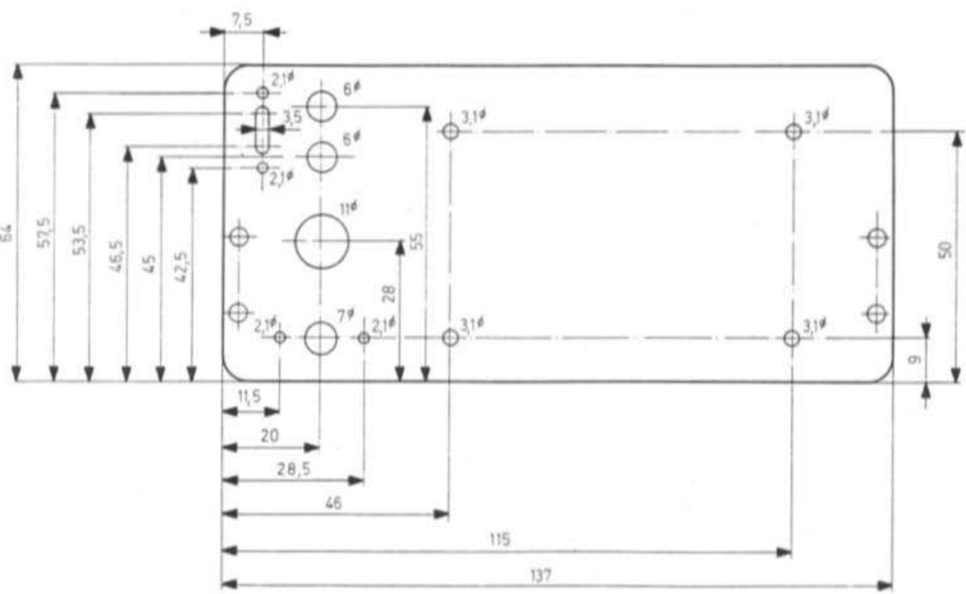


Fig. 23: Holes and cutout in the rear panel

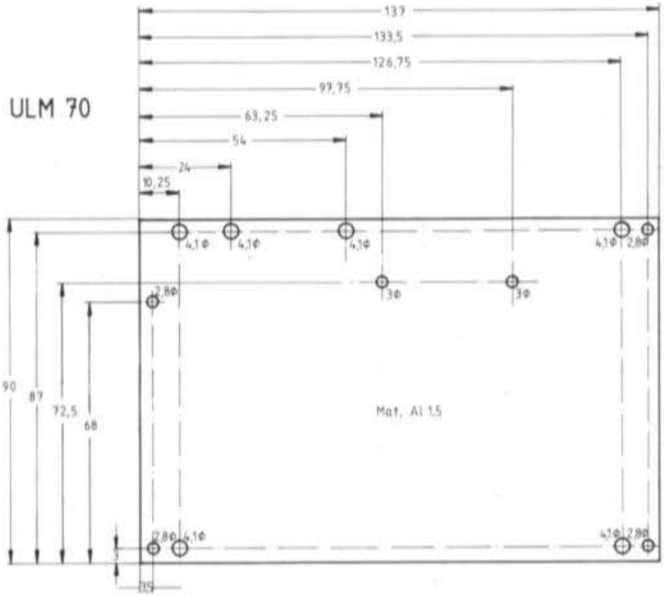


Fig. 24: The mounting plate

Two heat sinks constructed from aluminium or copper for transistors T 6 and T 7, which are shown in **Figure 25a** assembled. Four spacers as shown in **Figure 25b** manufactured from aluminium or brass which are used for supporting the receiver board. Six spacers as shown in **Figure 25c** manufactured from soft aluminium or brass that are later rivetted into the mounting plate, and support the transmit board. Finally, two brackets are required with foam rubber insert, as shown in **Figure 25d**, for mounting the accumulator.

11. MOUNTING AND WIRING

Figure 26 gives a cross section showing how the fully equipped and aligned PC-boards should be mounted. The six spacers should firstly be rivetted to the mounting plate and the two heat sinks screwed into place. Transistors T 6 and T 7 are now provided with heat-conductive paste, after which the transmit board is mounted. This is followed by fixing the receiver board temporarily to the mounting plate and placing the nuts into the guide slots on the side panels of the case. It is possible, after this, for everything to be screwed tightly into place.

The front and rear panels can now be completed and screwed to the side panels. After this, it is only necessary for the internal wiring of the transceiver to be wired as shown in **Figure 27**. However, this need not be described in detail. The following four photographs **(Figure 28 to 31)** should provide sufficient information.

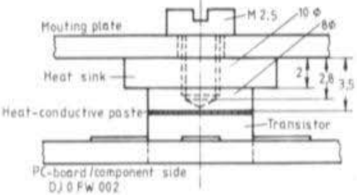


Fig. 25a:
Heat sink for T 6
and T 7 (assembled)

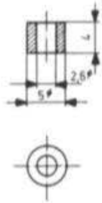


Fig. 25b:
Spacers (4 pieces)
for board 001

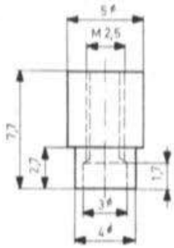


Fig. 25c:
Spacers (6 pieces)
for board 002

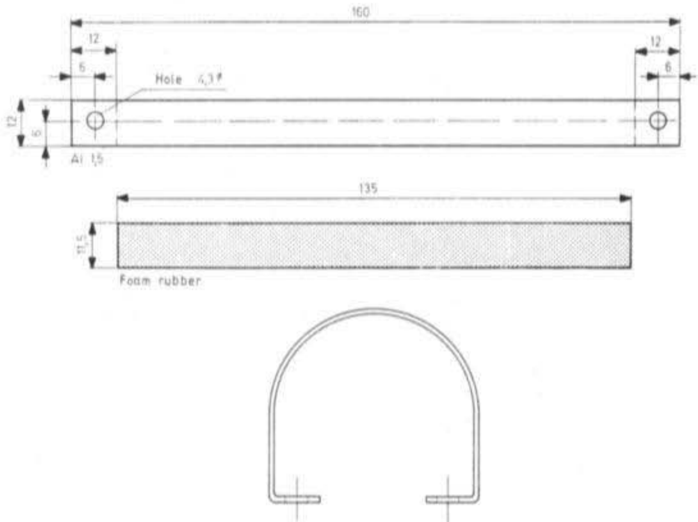


Fig. 25d: Mounting bracket (2 pieces) for the accumulator

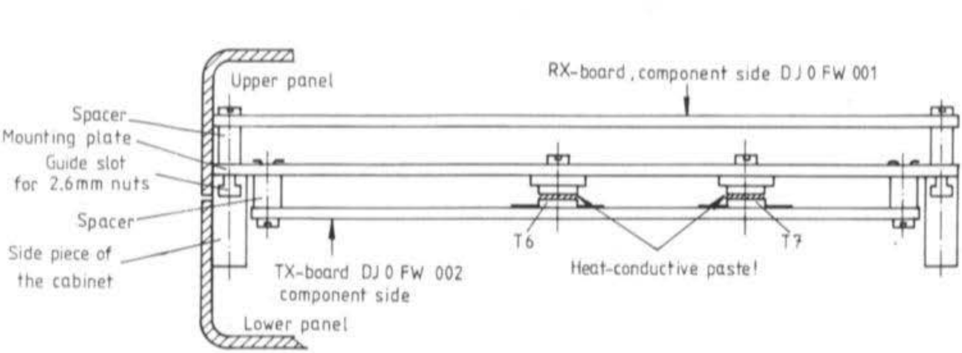


Fig. 26: Cross section of the transceiver

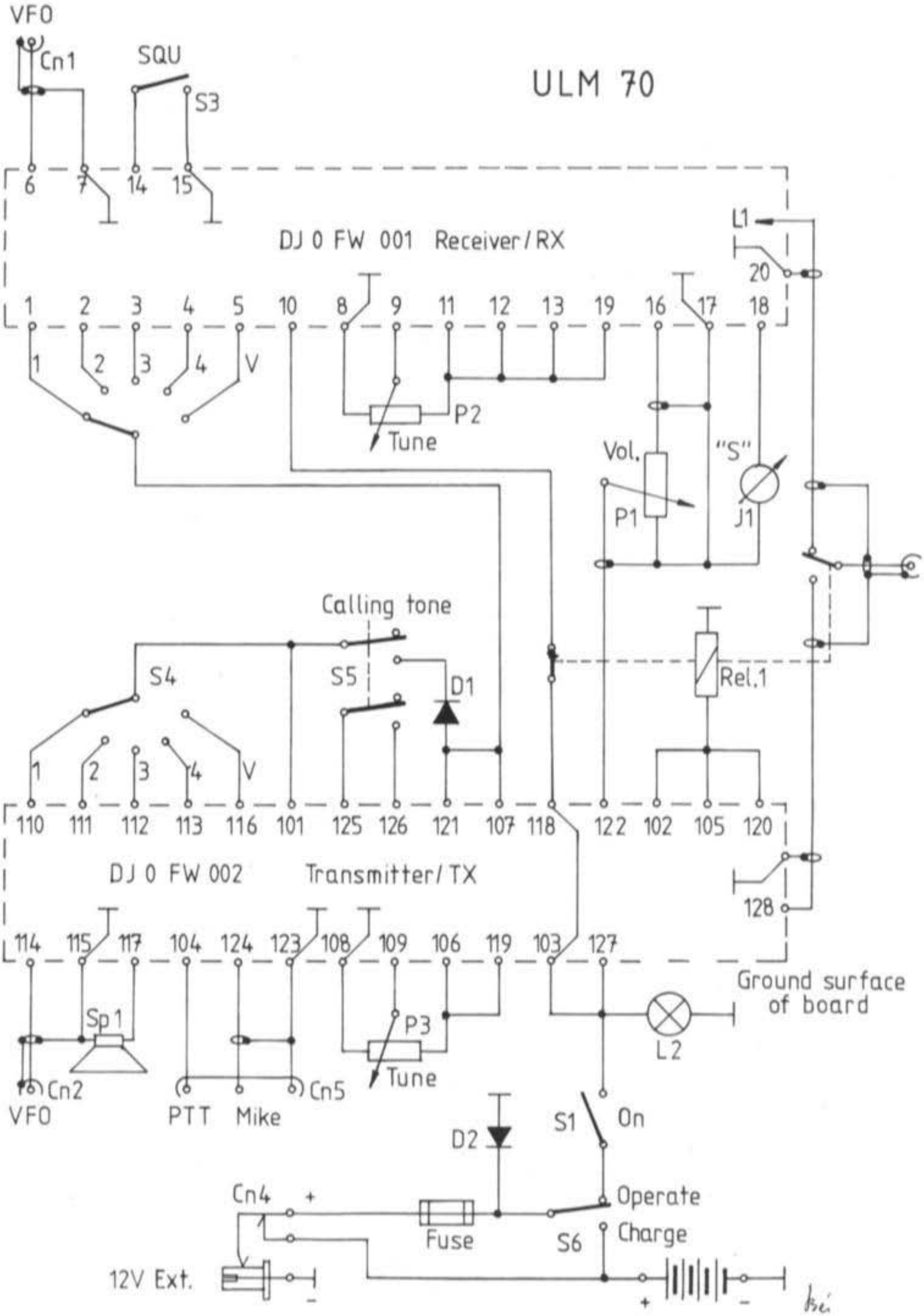


Fig. 27: Internal wiring diagram of the transceiver

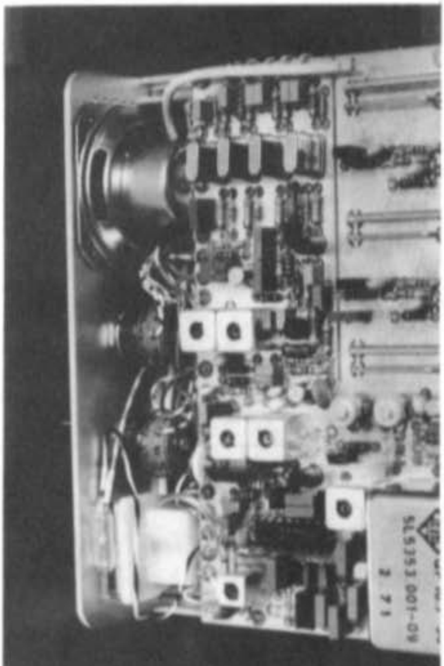


Fig. 28: Front panel and receiver module

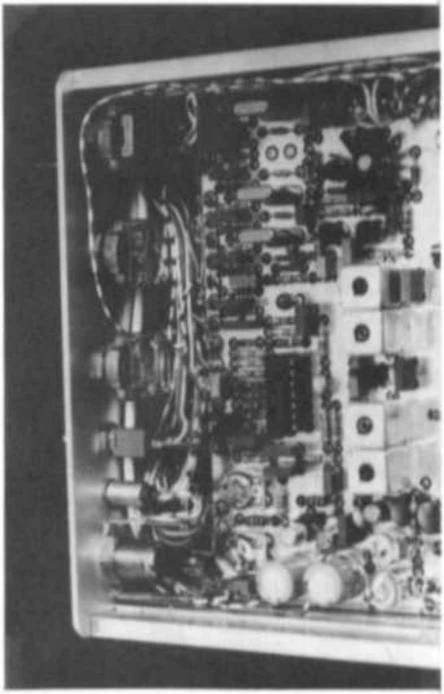


Fig. 29: Front panel and transmit module

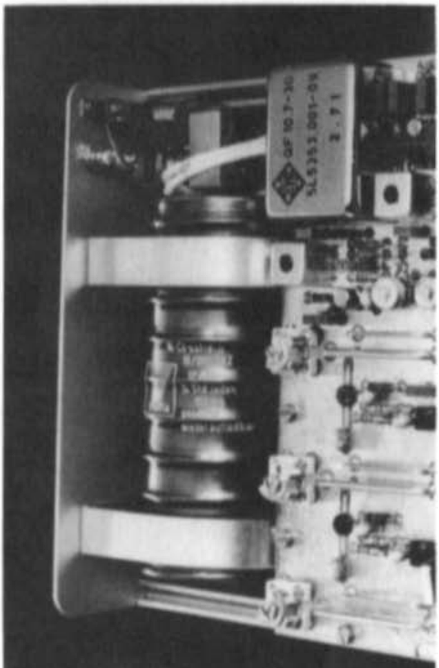


Fig. 30: Rear panel with accumulator, antenna relay and fuse

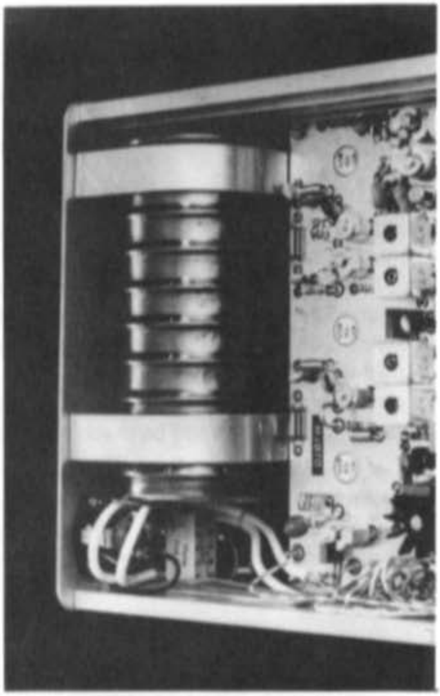


Fig. 31: Rear panel with antenna relay and charging connector

CALCULATION OF DISTANCE AND ANTENNA DIRECTION FROM TWO QTH-LOCATORS

O. Schmidt, DL 3 OV

Methods of calculating the distance from QTH-locators were described by W. Kraus (1), H. Jirikowski (2) and F. Landstorfer (3). The first mainly used logarithmic tables because electronic calculators were not readily available at that time. The second concentrated his calculation instructions on the TI SR-50 calculator. However, since here are now many other calculator types on the market, the author wished to describe a method of calculation that is not designed around a specific calculator but can be used with any calculator that can process trigonometric functions. In contrast to (3), no deriations are to be given but briefly how these calculations can be made in practice.

The fundamental is, of course, the calculation of the great circle distance between two points on a sphere according to spherical trigonometry. However, before the well-known navigational formula can be utilized, it is necessary to decode the longitude and latitude coordinates from the QTH-Locator, so that the calculation of the great circle distance can be made.

1. DECODING THE QTH-LOCATOR

The individual parts of the QTH-locator are as follows:

Longitude signs (meridians)	M			m	λ
	↑			↑	↑
QTH-Locator	**F**	**J**	**3**	**6**	**g**
		↓	↓		↓
Latitude signs		L	l		β

It is now necessary for the letters to be converted to numbers. This is done using the following system:

Capital letter:	...	R	S	T	U	V	W	X	Y	Z	A	B	C	D	E	F	G	H	I	J	K	L	M	...
Number:	...	-8	-7	-6	-5	-4	-3	-2	-1	0	1	2	3	4	5	6	7	8	9	10	11	12	13	...

For »l«, it is possible to directly extract the first number of the QTH-locator (3 in our example), as long as the second number is not a 0 (1-9). If it is a 0, it is necessary to deduct 1 from the numeral »l«. In our example l = 3, but would l = 4, for instance, with EK 50 d.

The second numeral of the QTH-locator can be taken for »m«. This would be 6 in our example. However, when it is a 0, it should be changed to m = 10.

The values for λ and β can be taken from the following table:

Value for λ :	1	2	2	2	1	0	0	0	1
Last letter of QTH-locator :	**a**	**b**	**c**	**d**	**e**	**f**	**g**	**h**	**j**
Value for β :	0	0	1	2	2	2	1	0	1

Example 1: FJ 36 g

$M = 6,\ m = 6,\ \lambda = 0 \qquad L = 10,\ l = 3,\ \beta = 1$

Example 2: EK 50 d :

$M = 5,\ m = 10,\ \lambda = 2 \qquad L = 11,\ l = 4,\ \beta = 2$

The next step is to convert these numerals to location data. One must differentiate between ones own location index »o« and the second location index »s«.

Own location:

$$u = 40 + L_o - \frac{L_o}{8} - \frac{\beta_o}{24} \tag{1}$$

$$v = 2 \times M_o + \frac{m_o}{5} + \frac{\lambda_o}{15} \tag{2}$$

Second location:

$$x = 40 + L_s - \frac{L_s}{8} - \frac{\beta_s}{24} \tag{3}$$

$$y = 2 \times M_s + \frac{m_s}{5} - \frac{\lambda_s}{15} \tag{4}$$

The following is calculated from the results of equations (4) and (2):

$$z = y - v \tag{5}$$

2. CALCULATION OF THE GREAT CIRCLE DISTANCE

The last step is to insert the values u, x and z into the general formula for calculation of the great circle distance. The calculator is now switched to the degree mode »D« or »DEG«. In order to ensure that those readers that are not used to such calculations do not make errors, the calculation is to be made in two parts, and given in the same way as it would usually be inserted into a calculator.

$$\cos z \times \cos u \times \cos x + (\sin u \times \sin x) = \cos d \tag{6}$$

and finally:

$$\text{inv} \cos d \times 111.2 = \text{distance in km} \tag{7}$$

Many calculators have a button marked »$\cos^{-1}$« instead of »inv cos«, also one will sometimes see »arc cos«. If statute miles or nautical miles are required the numeral 111.2 (km/deg.) should be replaced by 69.1 (St.miles) or 60 (Naut.miles).

If the calculator is not able to solve parenthetical problems, or if it does not possess a memory, it will be necessary to note down the sine or cosine value with an accuracy of six decimal positions.

The calculation procedure appears to be complicated, but can be carried out rapidly after a little practice. If several calculations are to be made from a single location, the calculation will be simpler since the values u and v need only be calculated once for ones own location, and can be inserted again and again into equations (5) and (6). The handbooks of expensive scientific calculators often give special programs for calculating the great circle distance. There are also a number of calculators for marine and aeronautical navigation that have fixed programs for great circle calculations. In fact there are so many systems that they are not to be mentioned in any detail here.

Just one more tip:

The geographical latitude of the NW-corner of a QTH-locator field has the value u (or x); the geographic longitude results from v (or y) – 2.2 degrees.

3. A COMPLETE EXAMPLE

Required is the distance from Erlangen FJ 36 f to Manchester, England, ZN 69 f.

$$M_o = 6 \qquad M_S = 0$$
$$m_o = 6 \qquad m_S = 9$$
$$\lambda_o = 0 \qquad \lambda_S = 0$$
$$L_o = 10 \qquad L_S = 14$$
$$l_o = 3 \qquad l_S = 6$$
$$\beta_o = 1 \qquad \beta_S = 2$$

$$u = 40 + 10 - 3/8 - 1/24 = 49.5833...$$
$$v = 2 \times 6 + 6/5 + 0/15 = 13.2000$$
$$x = 40 + 14 - 6/8 - 2/24 = 53.1666...$$
$$y = 2 \times 0 + 9/5 + 0/15 = 1.8000$$
$$z = 1.80 - 13.20 = -11.40$$

$$\cos d = \cos(-11.40) \times \cos 49.5833 \times \cos 53.1666 + \sin 49.5833 \times \sin 53.1666$$
$$\cos d = 0.381006 + 0.609371 = 0.990377$$
$$d = \cos^{-1}(0.990377)$$
$$d = 7.9550288$$

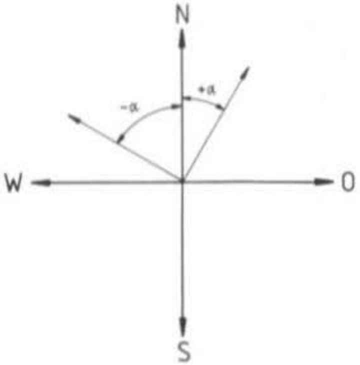
Distance D = d x 111.2 = 884.6 km

4. CALCULATION OF THE ANTENNA DIRECTION

The following equations can be used if the exact direction to a certain QTH-locator field is required, for instance, for long distance communication tests with narrow beam antennas:

$$\sin \alpha = \frac{\sin z \times \cos x}{\sin \alpha} \quad (8) \qquad \alpha = \text{inv} \sin \alpha \quad (9)$$

The values of z, x, and α can be taken from equations (5), (3) and (6).



Direction of the angle from North

The angle α is referred to North, as can be seen in the illustration, with positive sign in a clockwise, and minus sign in an anticlockwise direction.

There is, however, one difficulty: The sine values are not inambiguous in that each value between 0° and 90° repeats itself between 90° and 180°. This means that one is not sure, whether the angle is α or $(180 - \alpha)$. In most cases, this can be determined by a quick look at a map, but one can be mistaken at long distances.

One can, however, clearly define whether the calculation is correct using the following method: After calculating the angle α for the required location select another location slightly further North. In other words: $x' = x + 0.01$. If the second angle α' is less than α, the first calculation will be correct. However, if the second value α' is greater than the first (α), the first value must be recalculated according to the following formula:

For positive values α: $\alpha = 180 - \alpha_{\text{indicated}}$ (10)

For negative values α: $\alpha = (\alpha_{\text{indicated}} + 180) \times (-1)$ (11)

5. EXAMPLE OF AN ANTENNA DIRECTION CALCULATION

Own location: FJ 36 g – Second location: AJ 45 h

According to Section 1, the following are calculated:

$u = 49.58333...$ $\quad x = 49.50$

$v = 13.20$ $\quad y = 3.0$

$z = -10.2$

This results in: $d = 6.61418$ and $D = 735.5$ km.

Calculation of the angle is as follows:

$$\sin\alpha = \frac{\sin(-10.2) \times \cos 49.50}{\sin 6.61418} = -0.998475$$

$$\alpha = -86.8353 \approx -87^\circ$$

Now the second calculation as a check:

$$\sin \alpha' = \frac{\sin (-10.2) \times \cos 49.51}{\sin 6.61418} = -0.998271$$

$$\alpha' = -86.6302^\circ$$

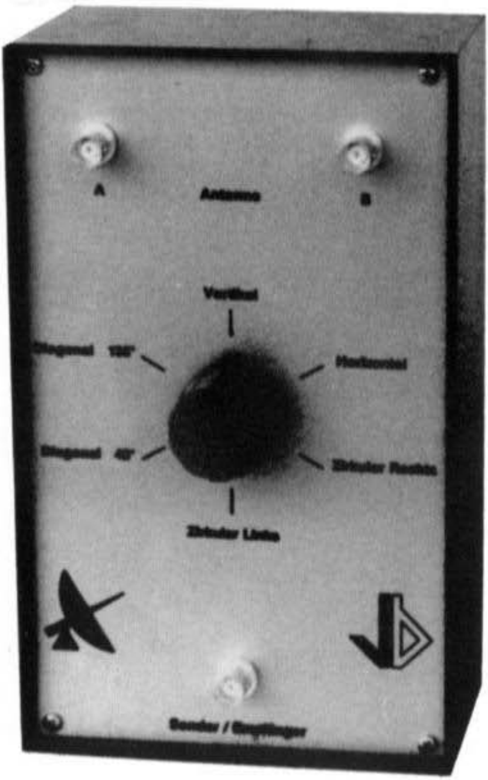
The second value is less so that the first value for α is correct. Although the QTH-locator shows the second location to be further south than ones own location, the antenna direction is 3 degrees North of West !

6. REFERENCES

(1) Dr. W. Kraus: Möglichkeiten der Entfernungsbestimmung von UKW-Funkverbindungen Amateurfunk-Magazin, Edition 1/1973, Pages 20 - 24

(2) H. Jirikowski: Berechnung von Entfernungen aus QTH-Kennern mit dem Elektronen-Rechner CQ-DL 46 (1975), Edition 12, Pages 707 - 709

(3) F. Landstorfer: Bezeichnung des Standorts und Entfernungsberechnungen CQ-DL 48 (1977), Edition 8, Pages 293 - 301



APPLICATIONS OF C-MOS CIRCUITS

by G. Heeke, DC 1 QW

Complementary C-MOS integrated circuits can be used for numerous applications in digital electronics. They offer a multitude of advantages and some disadvantages over TTL-integrated circuits. However, it is necessary to take the specific characteristics of MOS-ICs into consideration when designing circuits around them. This article is to discuss operating and other criterion of these circuits without going too deeply into MOS-technology.

1. DESIGNATIONS AND INTERCHANGEABILITY

C-MOS circuits are offered by numerous manufacturers. There are a large number of standard circuits available for basic logic, and also highly-integrated systems which allow complex circuits to be realized.

The standard circuits of the various manufacturers are nearly always interchangeable as is often indicated in the similar designations. For instance, an integrated circuit containing four NAND-gates is designated CD 4011 by RCA, HD 4011 by Harris, HEF 4011 by Philips, and MC 14011 by Motorola. Further letters are added to give information on the temperature range, case, and operating voltage range. The inexpensive versions in plastic dual-in-line cases are suitable for amateur applications. The operating temperature range of these is between 233 and 358 Kelvin, corresponding to – 40°C to + 85°C.

The integrated circuits designated 74 C... represent circuits having the same function and connections of the TTL circuits with the same designations. However, it is seldom possible for them to be directly exchanged, for instance, to reduce current drain. As will be mentioned later; it is also necessary to take the cutoff frequency and the different input and output load factors into consideration.

2. OPERATING VOLTAGE AND LOGIC LEVEL

MOS integrated circuits operate over a wide operating range; voltages from + 3 V to + 15 V, and even up to + 18 V in the case of the ceramic version with its better heat dissipation. This usually allows use of an existing voltage source in the equipment, in contrast to TTL circuits that require a stabilized input voltage of + 5 V. Since it is not usually necessary to stabilize the operating voltage of C-MOS ICs, no dissipation heat and power loss is consumed in a voltage stabilizer circuit.

The wide operating voltage range means no fixed limit values are present for the logic voltage levels of »0« and »L« corresponding to »low« and »high«; these are always referred to the operating voltage. The lower switching level is at approximately 30 % of the operating voltage, and the upper limit at about 70 %. Both values are hardly dependent on temperature. For instance, the following conditions exist in the case of an operating voltage of 10 V: Low level upto approximately + 3 V and high level above about + 7 V. The intermediate range from + 3 V to + 7 V is not defined.

The extremely high impedance of the MOS-ICs results in low current drain. This is one of the most important advantages of this IC family. The input impedance of such circuits is in the order of $10^{12}\Omega$. The quiescent power consumption of the previously mentioned IC with four NAND-gates is 2.5×10^{-9} W at an operating voltage of + 5 V. When switching, the power consumption increases virtually linearly with frequency. At a switching frequency of 100 kHz and an operating voltage of + 15 V, the power dissipation of a gate amounts to 1 mW.

3. EXTERNAL CIRCUITRY

3.1. Input Circuit

Due to the high impedances, the required drive power levels are very low. It is, however, important for correct operation that a sufficient voltage difference is available. If an input is to be connected to low level, it can be grounded to the 0 V line. On the other hand, if a high level is required it can be connected to the positive operating voltage.

The input voltages may be as great as the operating voltage but not exceed it. In other words, when using an operating voltage of + 5 V, the input voltage level should not be greater than this even though an operating voltage of upto + 15 V would be permissible. The lower permissible input voltage limit is – 0.5 V. Negative input voltages can occur in conjunction with excessively long input leads (high inductivity).

Attention should be paid that **all** inputs of the integrated circuit are connected, otherwise undefined conditions can occur that can cause unreliable operation. This is quite in contrast with TTL circuits where unconnected inputs are automatically brought to high level using internal pull-up resistors.

Due to the extremely high input impedances, MOS-ICs are sensitive to high voltages, which can occur as static charges. For this reason, MOS-ICs are supplied packed in such a way that all connections are grounded to another. This is often done using metal rails, or using a black conductive foam material.

MOS-ICs are provided with protective diode structures that usually make them insensitive to such charges; however, it is advisable not to touch the pins of the IC whilst placing it into its socket. The use of sockets is very advisable. Remember, static voltages in the order of 4 kV to 15 kV have been measured on persons working in rooms with synthetic flooring !

If MOS-ICs are to be mounted on plug-in cards and it is possible that the inputs are not terminated when exchanging a card, it is advisable to connect the inputs via shunt resistors to the positive or negative operating voltage.

3.2. Output Circuit

The power dissipation of a MOS-IC should not exceed 200 mW. For this reason, it is not possible for any low-impedance loads to be connected. A short-circuit of the outputs to either the positive or negative operating voltage could case an excessive power consumption that the IC could be destroyed by heat. Capacitive load should not be more than 5 nF in order to keep the charge currents within permissible limits. Inputs and outputs of gates may only be connected in parallel (for instance, to obtain higher output currents) when they are within the same IC. The fan-out of 50 in the case of MOS-ICs means that the output power is sufficient to drive 50 standard gates simultaneously. However, it should be noted that not all MOS-inputs possess a fan-in of 1. Occasionally one will see circuits with a fan-in of 2 or more. In this case, it is possible for only 25 or less circuits to be driven from a MOS-IC. Further details on the input and output load factors are given in the data sheets.

4. DYNAMIC CHARACTERISTICS

The main disadvantage of MOS-ICs over TTL-ICs is that they can only process low frequencies. The upper frequency limits are dependent on the operating voltage since this frequency increases with voltage. Typically, a frequency of 5 MHz is given at an operating voltage of 10 V, which will drop to 2.5 MHz at 5 V. Most manufacturers guarantee 1 MHz or 3 MHz. The maximum frequency can also only be processed when the short rise-times of the input pulses, required in the data sheets, are maintained.

5. LEVEL CONVERTER

In practice, it is very often necessary to combine MOS-ICs with other components in order to use the most favourable component for each application. It is therefore necessary to use matching stages to interface between the various parts.

5.1. TTL to MOS

When MOS-ICs are to be driven by TTL-ICs, and it is required that the MOS-ICs are to operate from the + 5 V operating voltage for the TTL-ICs, it will be necessary to connect a resistor between the TTL-output and the positive line of the operating voltage as shown in **Figure 1**. **Table 1** gives details of suitable resistors.

If the MOS-ICs are to be operated from operating voltages of between 5 and 15 V, resistor R_x should be connected to the higher voltage and the resistance value selected to suit the circuit.

5.2. MOS to TTL

Since conventional MOS-ICs are not able to provide sufficient drive for TTL inputs, special driver stages are available (e.g. CD 4049 or CD 4009). It is then possible to use different operating voltages.

5.3. Industrial Control Logic to MOS

Figure 2 shows an interface circuit for matching industrial logic to MOS.

5.4. MOS to Industrial Control Logic

An interface circuit from MOS to a 24 V system is shown in **Figure 3**.

TTL-Typ	74	74H	74L	74LS	74S
R_x min.	390	270	1K5	820	270 Ω
R_x max.	4K7	4K7	27K	12K	4K7 Ω

Table 1

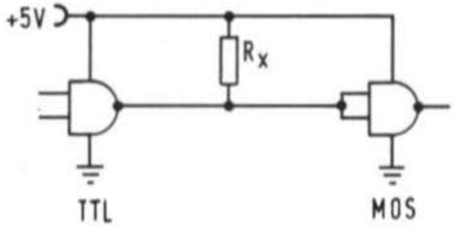


Fig. 1: Level converter TTL to MOS

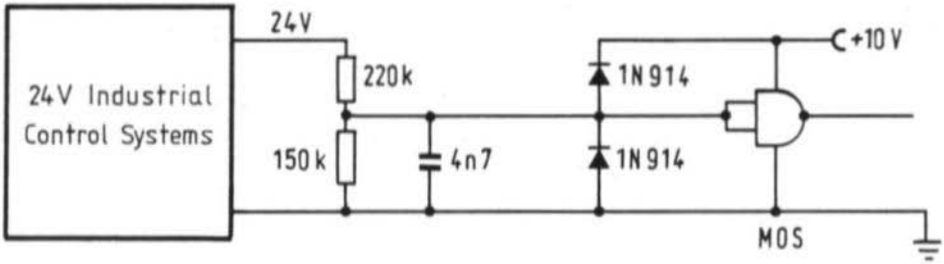


Fig. 2: Level converter 24 V-system to MOS

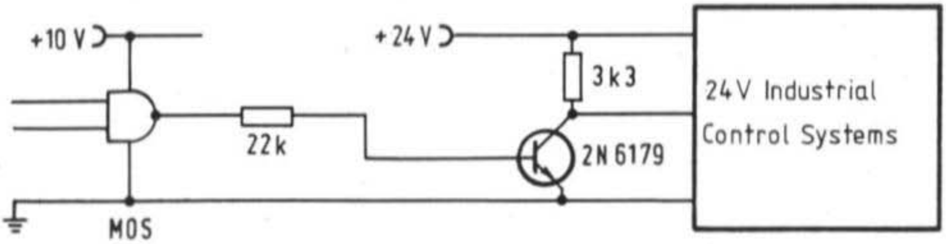


Fig. 3: Level converter MOS to 24 V-system

5.5. MOS (12 V) to ECL (Series 10 000)

Figure 4 shows a suitable interface for MOS to ECL-logic.

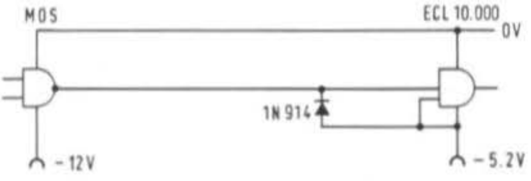


Fig. 4: Level converter MOS to ECL

5.6. ECL to MOS

Complicated level converters are required when MOS-ICs are to be driven by ECL-ICs. The integrated circuit MC 10 125 is suitable for this.

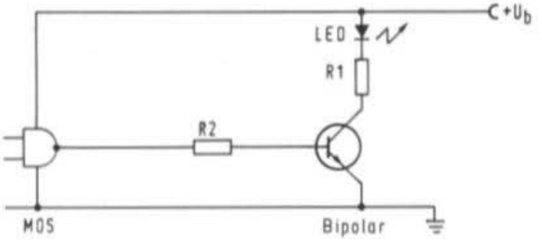


Fig. 5: Driver for an LED from MOS

5.7. MOS to Bipolar Transistor

The MOS-ICs would be loaded too greatly if their outputs were directly connected to the low impedance of bipolar transistors. For this reason, protective series resistors in the order of 4.7 kΩ to 22 kΩ are used for limiting base current. In order to obtain the required collector current, transistors with a high current gain should be used, or even Darlington-circuits.

Figure 5 shows an application for driving an LED. In this case, the following is valid:

$$R\,1 = \frac{U_b - U_{LED} - U_{CE\,sat}}{I_{LED}}; \qquad R\,2 = \frac{U_b - U_{BE\,max.}}{I_{LED\,max} / B_{min}}$$

The various data sheets published by the manufacturers give further examples and calculations for interface circuits.

6. EXPERIENCE GAINED WITH MOS-ICs

During experiments made by the author, it was found the MOS-ICs often provided better results at low frequencies than, for instance, TTL circuits. A special advantage is the high reliability; but also the uncritical operating voltage requirements. Capacitors for reducing the dynamic impedance of the voltage source are seldom required. Also current spikes are far less prevalent during switching processes than with TTL-ICs. The electrical and magnetic coupling to other parts of the circuits are correspondingly low, which is an important consideration in conjunction with RF-circuits.

However, it is disappointing that the values given for the typical cutoff frequency seem often not attainable in practice. This is in contrast to TTL-circuits that often process considerably higher frequencies. For instance, three programmable MOS frequency-dividers from a well-known manufacturer were tested and not one was able to process the listed typical frequency of 5 MHz. The actual limit frequencies were 3.5 MHz, 3.6 MHz and 3.75 MHz respectively.

The integrated resistor-diode network of the MOS-ICs appear to be very effective. The author has not destroyed a single IC due to static charges even though no special precautions were taken.

7. REFERENCES

(1) COS/MOS Today
RCA, 1975

(2) Semiconductor Data Library, Vol. 5
McMOS Integrated Circuits, Motorola

(3) C-MOS Data Book
Harris Semiconductor Corp.

SIMPLIFIED MEASUREMENT OF SPURIOUS SIGNALS OF VHF TRANSMITTERS

by H.J. Brandt, DJ 1 ZB

The author has already described how spurious signals can be determined in VHF transmitters (1). H. Brückner, DL 2 EO, has developed a measuring system for this, which should be advantageous when constructing and aligning VHF transmitters that use frequency conversion principles.

1. MEASURING SYSTEM

The output signal of the transmitter is fed to a dummy load that is connected to a diode rectifier and a low-pass filter with a defined impedance. The output of the low-pass filter is terminated in the measuring receiver (see **Figure 1**). The termination of the low-pass filter must be the same for RF and DC-voltages and should, most favorably, consist of an ohmic resistor. At higher power levels, suitable attenuators or a suitable coupling should be used to ensure that the rectifier is not overdriven.

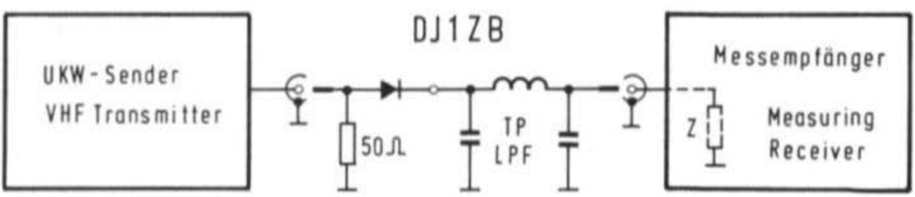


Fig. 1: Principle of measuring system

2. MEASURING PRINCIPLE

A frequency conversion is made in the detector diode between the carrier frequency of the transmitter and any spurious signals. In the case of a low-pass filter with a cut-off frequency of, for instance, 30 MHz, all spurious signals in the range ± 30 MHz from the carrier signal will be present at the output of this filter. However, it will not be possible to see whether these were above or below the original carrier frequency. Of course, this can be determined by studying the frequency plan of the transmitter. For instance, if any residual 136 MHz oscillator signal is still present in the output frequency spectrum of a transmitter using a 9 MHz IF system, the oscillator signal will be indicated at this spacing from the required output signal (144-136 = 9 MHz).

3. DETERMINING THE SPURIOUS SIGNAL REJECTION

After switching on the transmitter, the measuring receiver is firstly tuned through the range up to the cut-off frequency of the low-pass filter. All signals are noted and confirmation that they originate from the transmitter is made by keying the transmitter signal. If the signal is not affected by the keying, this will indicate a spurious signal in the measuring system itself.

This is followed by measuring the DC-voltage at the output of the low-pass filter with the aid of a conventional DC-voltmeter and subsequently measuring the voltage of each individual spurious signal selectively with the aid of the measuring receiver (also in volt). This voltage ratio indicates the spurious rejection with sufficient accuracy and can be converted into dB using the usual formula: 20 log U_{DC}/U_{AC}.

4. MEASURING RECEIVER

Ideally, the measuring receiver should be a professional selective level meter. Of course, only very few radio amateurs will have access to such a unit. However, with some patience a general coverage shortwave receiver with S-meter can be used for this application. In order to obtain a constant input impedance for both DC and AC, an ohmic attenuator of 10 dB or 20 dB is provided in front of the input, and the input of the attenuator is classed as reference point for the measurement.

The indicated S-meter reading is noted when measuring the spurious waves. The input voltage corresponding to this reading is determined subsequently using the substitution method.

4.1. Substitution Method

As is indicated by the name, the spurious signal is substituted by a known signal from a signal generator which is then adjusted until the same S-meter reading is indicated on the measuring receiver. Good signal generators usually have a calibrated output attenuator from which the output voltage can be read off directly.

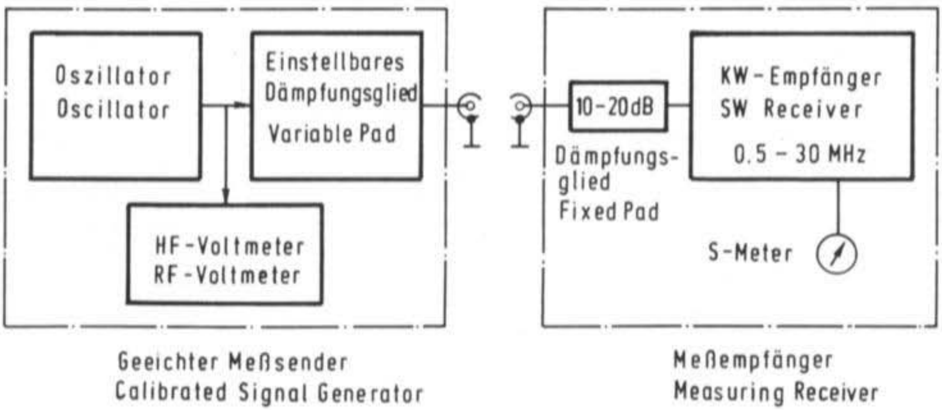


Fig. 2: Amateur selective RF voltage measurement using the substitution method

Figure 2 shows how this calibration can be realized in practice. The output voltage of the oscillator is measured at relatively high level, and subsequently reduced using a continuously-variable or switchable attenuator. It was described in (2) how it is possible to extend a simple signal generator using these circuits. One way of obtaining a defined attenuation value is to use the coaxial attenuator plugs that are available for broadcast and TV applications (community systems).

4.2. Relative Measurements

Even for those amateurs that are not interested in exact measurements, the described method should also be of interest for quick relative measurements. It is thus possible to check the effect of the alignment not only on the required signal but also on the spurious signals. If required, it will be possible to provide additional traps or blocking filters for certain spurious signals and to align these for maximum effect.

5. REFERENCES

(1) H.J. Brandt: Erkennung und Beseitigung von Störschwingungen in Transistorsendern
UKW-BERICHTE 16, Edition 2/1976, Pages 109 - 116

(2) K. Döll: Erweiterung der Anwendungsmöglichkeiten einfacher Prüfsender
Das DL-QTC, Edition 10/1967, Pages 520 - 524

NOTES AND MODIFICATIONS

1. CONVERTER for the 23 cm BAND DJ 5 XA 004 (VHF COMMUNICATIONS 2/1976)

After modifying the layout according to the suggestions given in edition 1/77 on page 59/60, this converter can also be used for ATV signals on that band. The higher IF (38.9 or 50 MHz) however can cause series resonance of capacitor C 7 together with the coupling inductivity L 7. Close coupling with L 8 causes a double-tuned bandfilter having two resonances with a large frequency difference. Thus no single signal maximum can be found. Send an S.A.E. to the publishers please for a circuit diagram with changed matching between the mixer diodes and the IF preamplifier utilizing the modern DG-MOSFET BF 900.

DJ 4 LB

2. ULM 70 – Parts 2 and 3 (VHF COMMUNICATIONS 3 and 4/1977)

When ordering the crystals directly from a manufacturer the fundamental frequency must not be calculated from the center frequency of the required range as given on page 141, but from the frequency of the lowest channel in the required range. This is caused by the fact that the frequency can only be pulled to higher channels. Since the publishers were informed of this only recently, the crystals in the kits oscillate approx. one channel high. So we ask our readers to order crystal R 72 - R 75, not R 75 - R 78 for channel 75, for instance. This is valid for receiver and transmitter crystals until approx. end of 1978.

The capacitance of 2 pF of diodes type 1 N 4148 is high enough to cause mutual detuning of the crystal oscillators. Better suited are diodes type 1 N 4151. Please send an S.A.E. to the publishers for a free-of-charge exchange of 5 diodes per kit. Kits delivered after March 1st, 1978 already contain the better diodes.

Some cases are reported that the crystal oscillator does not deliver enough voltage for the frequency multipliers to operate properly. In such a case one should try several different transistors perhaps with higher gain for T 1. The subsequent frequency triplers can be optimized by finding the best suited value of the emitter resistor.

DJ 7 OH

3. OSCILLATOR MODULE FOR THE LINEAR TRANSVERTER DF 8 QK 001 (VHF COMMUNICATIONS 4/1977)

The required oscillator power of 5 to 10 mW at 1268 MHz can be taken from a slightly modified module DC 0 DA 005 (see this edition of VHF COMMUNICATIONS). Only the stage with transistor T 6 (C 1 - 12) has to be left out.

DF 8 QK himself uses the following oscillator chain: Crystal 70.444 MHz with BF 199 – tripler to 211 MHz (BF 199) – doubler to 422 MHz (BF 199) – amplifier (BF 199) – tripler to 1268 MHz (BFR 34 A or BFR 90/91). This is followed by a bandpass filter of two $\lambda/4$ coaxial resonators. Please send an S.A.E. for a copy of the circuit diagram also containing coil winding data.

DL 3 WR

INDEX TO VOLUME 9 (1977) OF VHF COMMUNICATIONS

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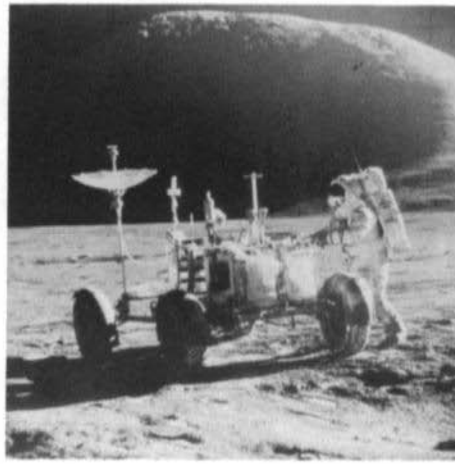

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MN 05	Man on the Moon – Apollo 15
MN 06	Man on the Moon – Apollo 16
MN 07	Man on the Moon – Apollo 17
ST 08	Earth from Space – Europe
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