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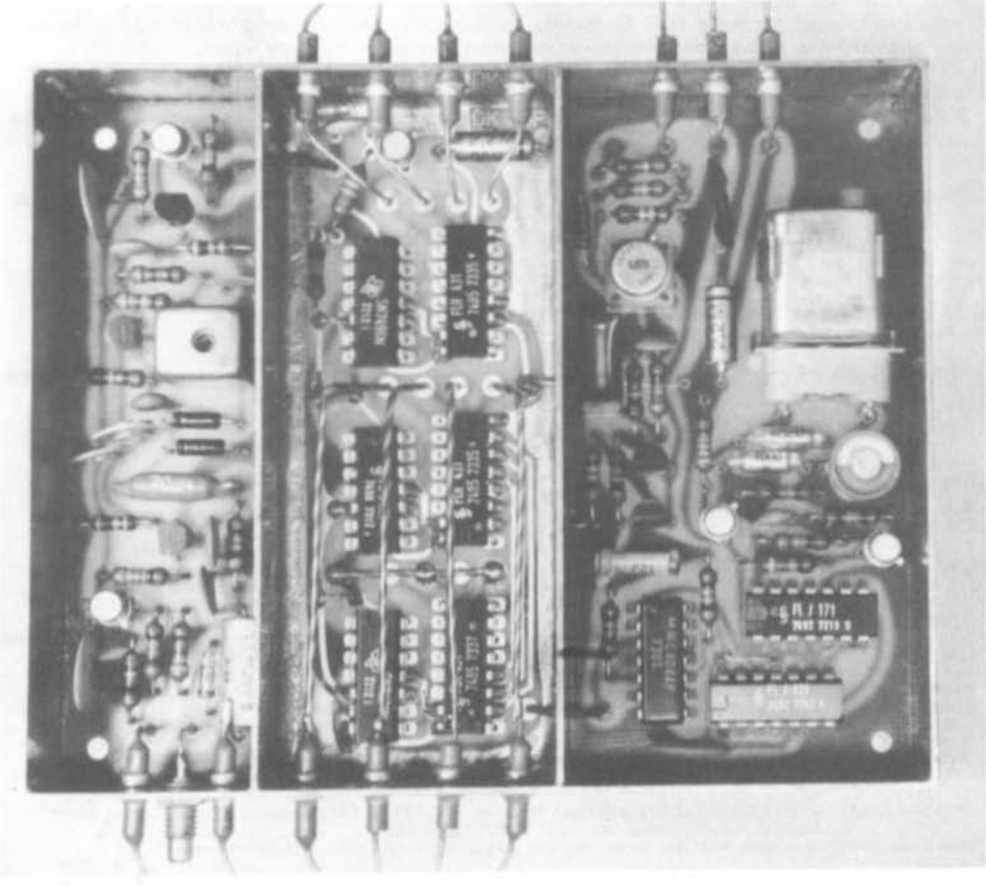

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A 400 CHANNEL SYNTHESIZER FOR 2 m

by J. Kestler, DK 1 OF

A channel spacing of 30 kHz is used for FM transmissions in North America, and in some other countries. A frequency synthesizer is to be described that provides a channel spacing of 10 kHz which is not only suitable for 30 kHz channel spacings but also for other applications. The described synthesizer is suitable for use together with the phase-locked oscillator DK 1 OF 011/014 (1) and will then provide 400 channels in the frequency range of 144 to 148 MHz. Figure 1 indicates which modules can be used in order to construct a complete transceiver.

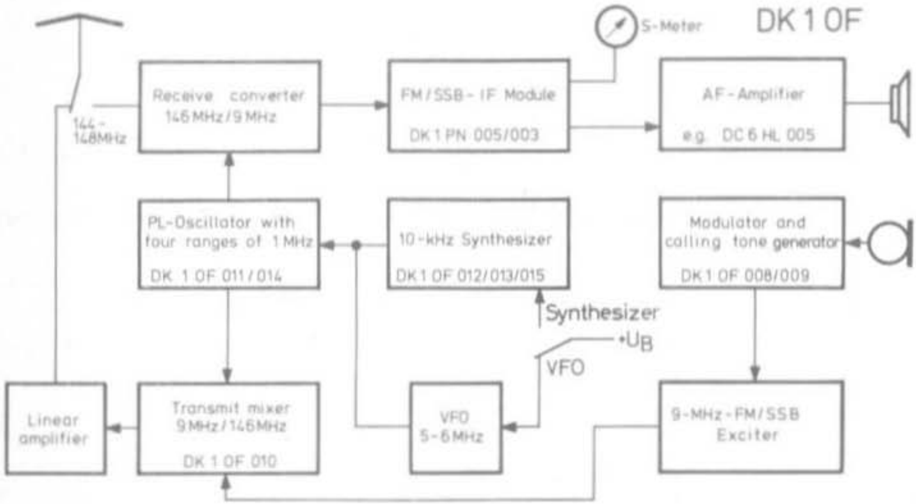

Fig. 1: Block diagram of a FM transceiver for 400 channels with a 10 kHz spacing

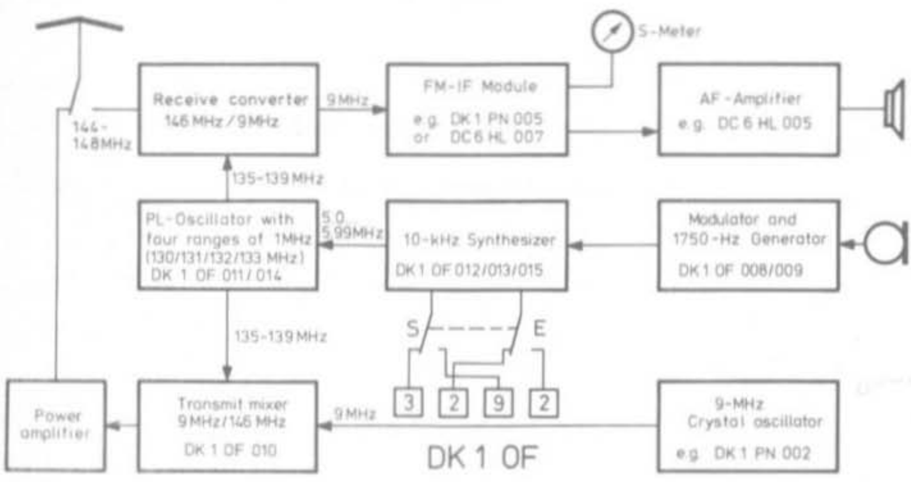

Fig. 2: Block diagram of a SSB/FM transceiver with VFO/Synthesizer

It is possible to use the described synthesizer with slight modifications for the 25 kHz channel spacing used in Europe and elsewhere. Full details are to be given for two types of 80 channel synthesizers having an IF of 9 MHz and 10.7 MHz.

1. OPERATION

As has been previously mentioned, the described synthesizer is suitable for feeding the phase-locked oscillator (1) instead of the VFO. Since the phase-locked oscillator possesses four crystal oscillators, it was decided to divide the total range of 135 to 139 MHz (operating frequency 144 - 148 MHz, IF = 9 MHz) into four bands of 1 MHz each. A synthesizer is required which is able to cover the range of 5.00 to 5.99 MHz in steps of 10 kHz.

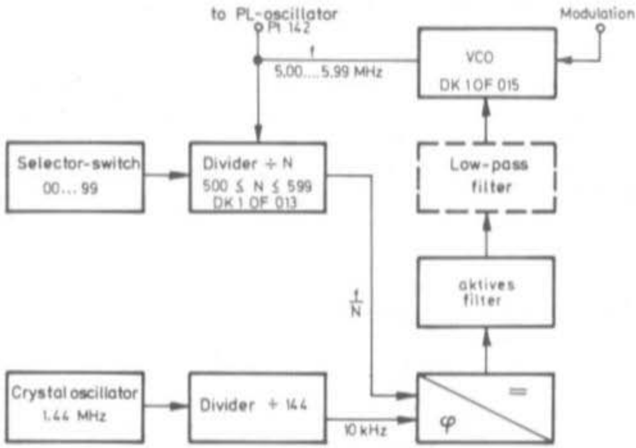

Fig. 3: Block diagram of the 10 kHz synthesizer

The block diagram of the synthesizer is given in Figure 3. The VCO operates in the above frequency range. The VCO signal is passed to the output, and to the variable frequency divider DK 1 OF 013 where its frequency is divided by a factor n of between 500 and 599 depending on the position of the channel selector. The output frequency f/n of this module is then passed to the phase comparator. It would also have been possible to obtain the 10 kHz reference frequency from a 1 MHz crystal oscillator and dividing this by 100. However, it is possible that the steep 1 MHz pulses would be injected into the intermediate frequency of 9 MHz.

The output voltage of the phase comparator drives the varactor diode of the VCO via the active filter, and possibly an additional lowpass filter so that the two input frequencies of the phase comparator coincide. When locked in, the following is valid:

$$\frac{f}{n} = 10 \text{ kHz} \qquad \text{or} \qquad f = n \times 10 \text{ kHz}$$

In practice, the output voltage of the phase detector is not a pure DC-voltage but always possesses a certain component of the phase comparator frequency (10 kHz in our case) and harmonics thereof in the form of a superimposed

AC-voltage. The active filter is not able to suppress these residual voltages completely, since its cut-off frequency may not be too low in order to ensure that the control time and short-term stability of the signal is satisfactory. However, if these residual AC-voltages are fed to the VCO, unwanted sidebands (frequency modulation) will be generated. These spectral lines will be present with a spacing of 10 kHz from the required signal and will therefore be present in the adjacent channels. This effect is very characteristic for all frequency synthesizers, and the lower the phase comparator frequency, the greater will be this interference. It can be avoided if the additional lowpass filter is used that is indicated within dashed lines in Figure 3. It is important that it should not possess any noticeable phase shift between the input and output voltage in the frequency range up to approximately 3 kHz, in order to ensure that the control circuit remains stable. In spite of this, it must provide a high attenuation at 10 kHz. A practical circuit will be described at the end of this description.

It should be noted that the subsequent phase-locked loop greatly attenuates the previously described sidebands. It is only when large demands are to be placed on the adjacent channel selectivity where such a lowpass filter will be required in the control line of the VCO. Since it is not often that a 10 kHz spacing will be used, this will probably not be necessary for practical operation.

2. CIRCUIT DESCRIPTION

The circuit diagram of the 10 kHz synthesizer is given in Figure 4. It comprises three modules, DK 1 OF 015 (VCO with matching buffer and output stages), DK 1 OF 013 (variable divider) and DK 1 OF 012 (reference oscillator, phase comparator, active filter). The field effect transistor T 151 operates as oscillator stage for the VCO. The frequency is determined by inductance L 151 and the two varactor diodes D 151 and D 152. The feedback is made inductively via the source of the transistor which operates in a common drain circuit. The tuning voltage (from connection Pt 153) is fed via the ferrite choke L 153 to the two tuning diodes. The AF-input Pt 155 is provided for direct frequency modulation of the synthesizer. It can be directly connected to the output of an AF-preamplifier such as DK 1 OF 008. Capacitor C 151 provides a galvanic isolation of AF and tuning voltage. It should be a plastic foil type. In order not to interfere with the dynamic behaviour of the phase-control circuit, the AF-voltage is fed via the decoupling resistor R 151. An AF voltage of approx. 0.4 V (RMS) is required at Pt 155 for a frequency deviation of $\pm$ 5 kHz.

The output signal of the frequency synthesizer is taken from the source of transistor T 151 via a capacitive voltage divider, and passed via the low-reactive buffer stage T 152. The emitter follower (T 153) which is used as output stage provides an additional isolation and low-impedance output at Pt 151. The use of diode D 153 allows a VFO to be connected to the synthesizer at the same time, without having to use a relay or switch. The changeover switching is then made simply by connecting the operating voltage to the synthesizer or VFO.

The RF output voltage can be taken from the output resistor R 152 of the VCO transistor. It is then passed via a further buffer (T 154) and the emitter follower T 155 to module DK 1 OF 013 (programable frequency divider). Good bypassing is important here so that no 10 kHz impulses are able to enter the VCO.

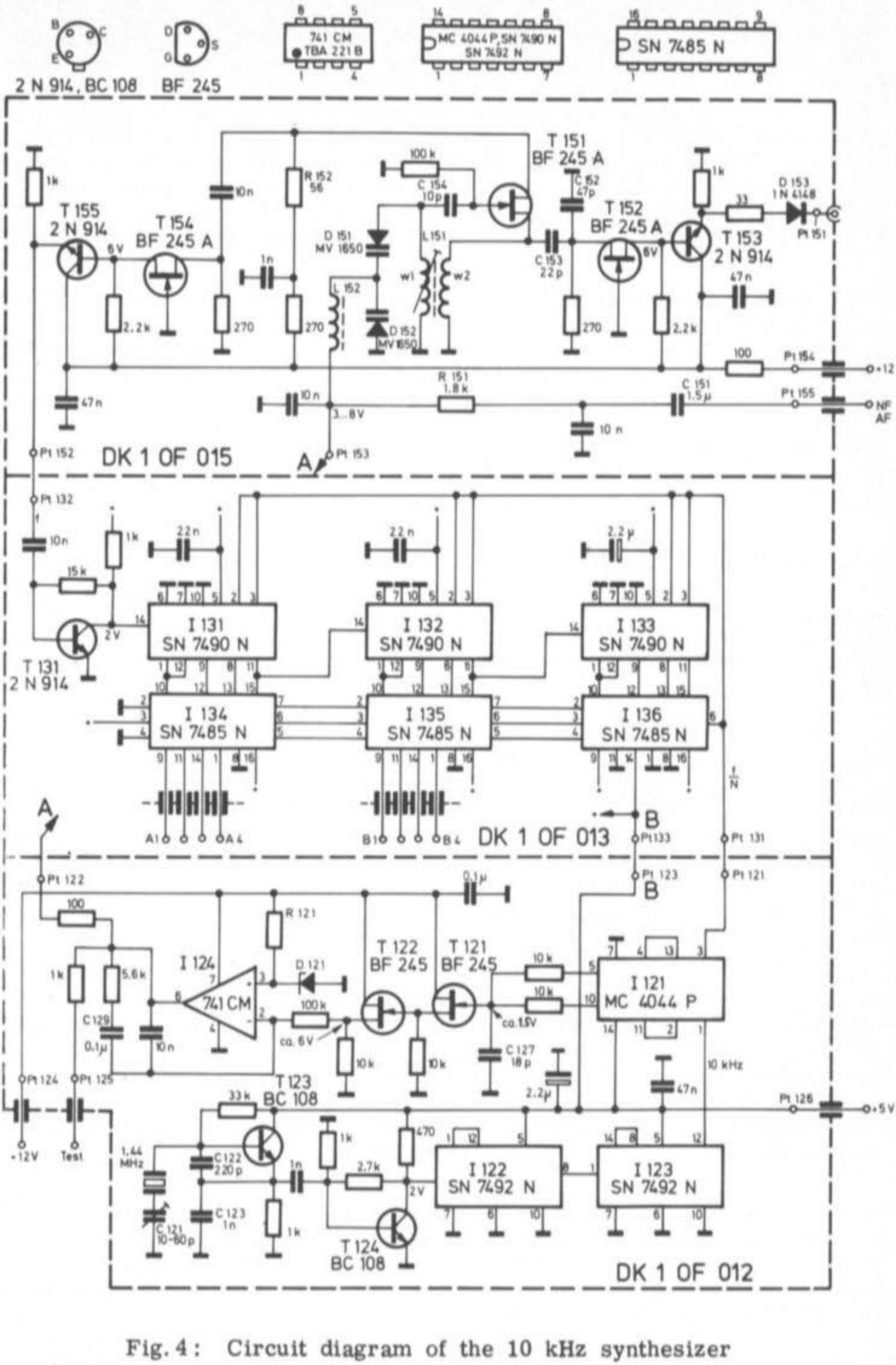

Fig. 4: Circuit diagram of the 10 kHz synthesizer

The VCO frequency available at connection Pt 132 then drives transistor T 131 which matches it to the required TTL level and passes it to the input of the variable divider. The operation of this divider has already been described in (2). The same circuit is used with the exception of the number of stages. The programming inputs A 1 (LSD = least significant digit) to B 4 (MSD = most significant digit) are connected to the appropriate outputs of the two BCD - coded selector switches, whereas the inputs of the highest decimal position (I 136) are bridged (C = 5). These connections are not provided as conductor lanes on the PC-board so that the module remains as universal as possible.

The output pulses of the variable divider are then passed via connection Pt 131/ Pt 121 to the input of the phase comparator I 121. The following circuit comprising T 121, T 122 and I 124 is the same as was used for the phase-locked oscillator (1). The tuning voltage for the VCO is fed from Pt 122 to Pt 153. Module 013 is provided with an operating voltage of + 5 V via Pt 123/Pt 133. A meter can be connected to connection Pt 125 for alignment and monitoring purposes (range 10 V). The voltage of the zener diode D 121 is dependent on the pinch-off voltage of transistors T 121 and T 122 as was the case with module DK 1 OF 014. Further details regarding this are given in Section 4.

In the lower part of Figure 4, a crystal oscillator using a clapp-circuit (T 123) and a matching stage (T 124) will be seen. The circuit corresponds mainly to that of module DK 1 OF 004. The 1.44 MHz crystal-controlled signal is divided by a total of 144 (12 each) in I 122 and I 123 and is then fed to the reference frequency input of the phase detector.

3. CONSTRUCTION

The PC-boards DK 1 OF 012, 013 and 015 have been developed for accommodating the frequency synthesizer. DK 1 OF 013 is double-coated and possesses through-contacts. The two other boards are single-coated. Figure 5, Figure 6 and Figure 7 give these PC-boards with component locations. Screening panels are provided in a similar manner to that used in the phase-locked oscillator and are shown in Figure 8. They should be provided with the required holes for the interconnection leads. Bridge "a" from Pt 122 to Pt 153 is made below the PC-board.

In order to avoid interference caused by harmonics of the fast TTL pulses, it is necessary for the whole module to be carefully screened. With the exception of connection Pt 151, which is provided with a miniature coaxial socket, all connections should be made via feedthrough capacitors as can be seen in the photograph (Fig. 8). After completing the alignment, a cover made from double-coated PC-board material should be soldered to the upper and lower side so that a completely screened case results.

Since the additional lowpass filter described in section 1 will not be necessary in most cases, no room has been provided for it within the synthesizer. If required, it can be mounted outside of the actual synthesizer on a small board. The interconnection leads to connections Pt 122 or Pt 153 should then be made via two further feedthrough capacitors.

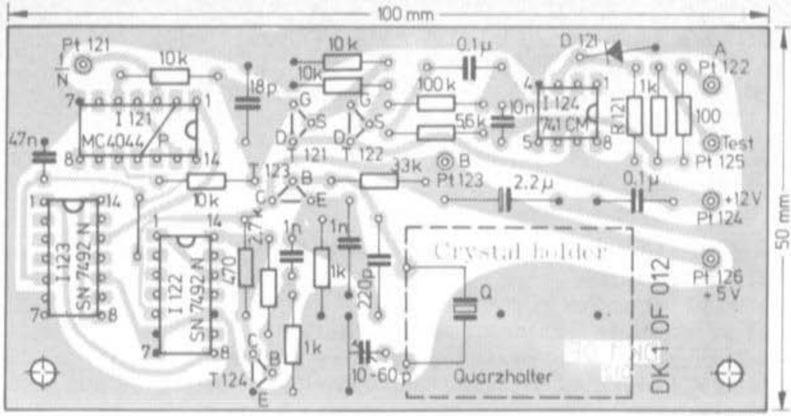

Fig. 5 : PC-board DK 1 OF 012 with component locations

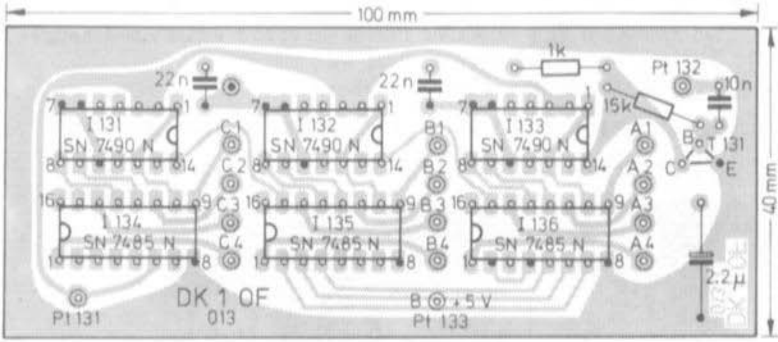

Fig. 6 : PC-board DK 1 OF 013 with component locations

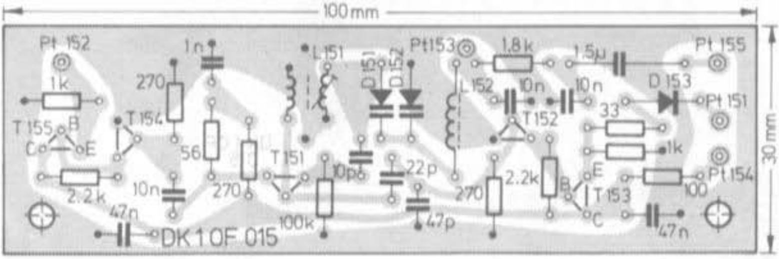

Fig. 7 : PC-board DK 1 OF 015 with component locations

3.1. COMPONENT DETAILS

T 121, T 122: BF 245 A, B or C (TI) or similar FET
T 123, T 124: BC 108, BC 413 or similar silicon transistor
T 131: 2 N 914, 2 N 706 or similar switching transistor
T 151, T 152, T 154: BF 245 A (TI) or similar FET
T 153, T 155: 2 N 914, 2 N 706 or similar switching transistor

D 121: Zener diode of series BZy 85, BZX 55 or BZX 83 or similar; value see section 4.
D 151, D 152: MV 1650 (Motorola) or two pieces BA 124/65 (AEG-Tfk)
D 153: 1 N 4148, 1 N 914 or similar.

I 121: MC 4044 P (Motorola)
I 122, I 123: SN 7492 N
I 124: 741 CM (various manufacturers) or TBA 221 B (Siemens)
I 131, I 132, I 133: SN 7490 N
I 134, I 135, I 136: SN 7485 N

Crystal 1.440 MHz, HC 6/U
Crystal holder for horizontal mounting

C 121: Approx. 10-60 pF ceramic or plastic foil trimmer 10 or 7 mm dia.
C 122: 220 pF styroflex
C 123: 1000 pF styroflex
C 127: 18 pF ceramic tubular or disc capacitor
C 129: 0.1 μF plastic-foil capacitor, spacing 10 mm
C 151: 1.5 μF (uncritical) plastic-foil capacitor, spacing 15 mm
C 152: 47 pF ceramic disc capacitor, spacing 5 mm
C 153: 22 pF as C 152
C 154: 10 pF as C 152

L 151: w 1: 45 turns; w 2: 5 turns of cotton-covered enamelled copper wire in special coil set
L 152: approx. 70 μH miniature ferrite choke

13 feedthrough capacitors of 2.2 nF or more for solder mounting

All resistors for 10 mm spacing.

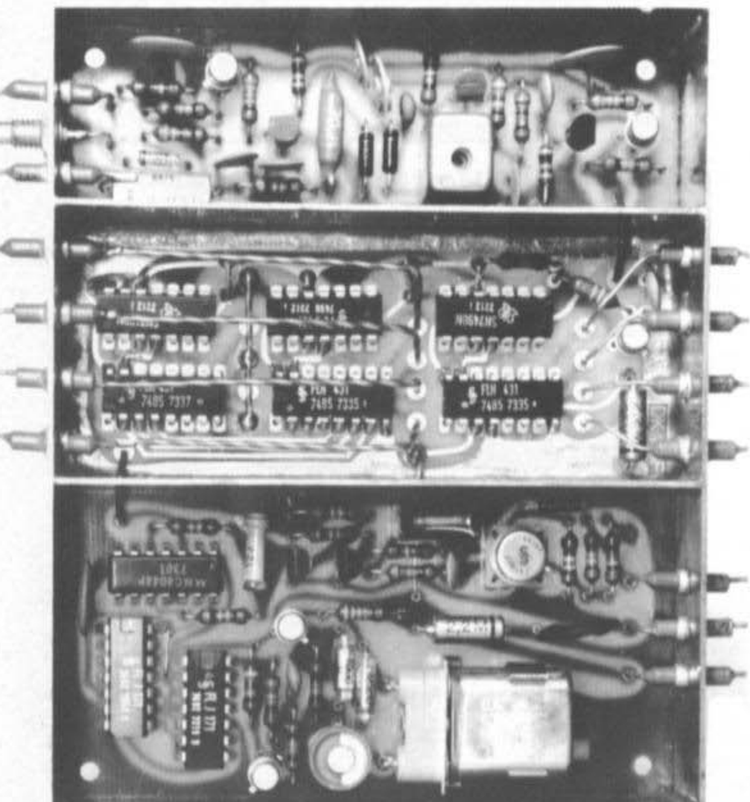

Fig. 8 : Author's prototype with modules DK 1 OF 012, 013, 015

4. ALIGNMENT

Disconnect the interconnection between connections Pt 153 and Pt 122. Connection Pt 153 is fed with a variable bias voltage of 0 to + 12 V. The output of the synthesizer (Pt 151) is terminated with a 1 kΩ to ground and connected via 1 nF to a frequency counter or shortwave receiver (5 to 6 MHz, via an attenuator if required). The operating voltage of + 12 V is now connected to Pt 154 and Pt 124. Align inductance L 151 so that the frequency range of 5 to 6 MHz is obtained with a bias voltage of 3 V to 8 V. The given voltage range is not critical and deviations of ± 1 V are permissible at the lower limit and ± 2 V at the upper limit.

The operating voltage of + 5 V (approx. 300 mA) is now connected to connection Pt 126, after which the operation of the crystal oscillator and subsequent frequency divider are checked. The most favourable means of doing this is with the aid of an oscilloscope which is connected to the output of I 123 (pin 12). If not available, a high impedance head-set can be used to monitor the output frequency of 10 kHz.

The channel selector-switches are now connected temporarily to the data inputs A 1 to A 4 and B 1 to B 4. A short impulse (of approx. 50 ns, frequency approx. 10 kHz) should be present at point Pt 131 at all positions of the preselector switches. Where no suitable oscilloscope with a bandwidth of approx. 50 MHz is available, the following procedure can be used: Connection Pt 131 to the clock input of a TTL-flip-flop (such as SN 7472, or half SN 7473). A mean DC voltage of approximately 2 V should be indicated at the output of this module. If a headset is connected, a 5 kHz tone should be audible.

The bias voltage (Pt 153) is now adjusted to 5 V and a high-impedance voltmeter (min. 10 kΩ/V, range 10 V) connected to the source of T 122. The preselector switch is now switched through the various positions from channel 00 to 99 and the measured voltage must jump to a lower value. The channel number by which the jump occurs can be altered by varying the tuning voltage. The magnitude of the voltage jump should be between 1 V and 2 V and the mean value between 4 V and 8 V. This mean value is mainly dependent on the pinch-off voltage of the field-effect transistors T 121 and T 122, which unfortunately possess large fluctuations. For this reason, it is not possible for any details to be given as to the most favourable type (BF 245 A, B or C). In the author's prototype, a mean value of 6.2 V was obtained with a transistor type BF 245 A and BF 245 B.

The zener diode D 121 should stabilize a voltage that is equivalent to this mean value. Any deviations from this will mean that the tuning speeds will not be identical to higher and lower frequencies. A fine alignment of the zener voltage is possible with resistor R 121.

Finally, connection Pt 122 is re-connected to Pt 153. The voltage at testpoint T 125 should correspond to approximately 3 V at channel 00 and to approximately 8 V at channel 99. If required, this can be corrected by aligning L 151.

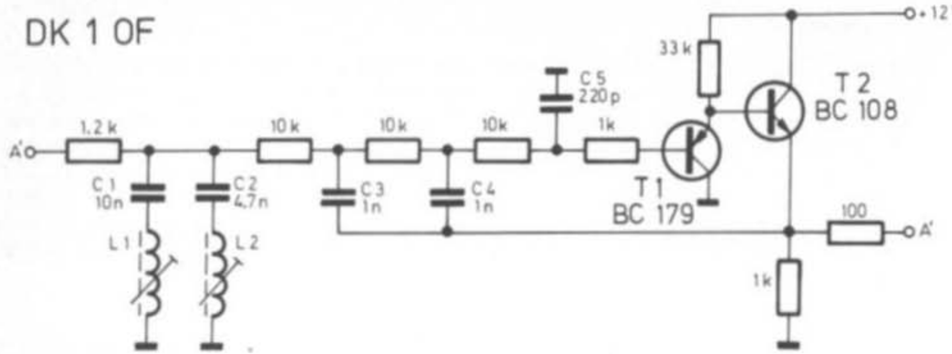

Fig. 9: Active low-pass filter with absorption circuits for 10 kHz and 20 kHz

5. ADDITIONAL LOW-PASS FILTER

As can be seen in Figure 9, the circuit is based on an active low-pass filter as described by DJ 4 BG in (3). In order to ensure that the phase-comparator frequency and its second harmonic are effectively suppressed, two additional absorption circuits are provided. They are aligned to 10 kHz and 20 kHz. Low-tolerance plastic-foil capacitors (± 2%) should be used for C 1 and C 2 since the alignment range of the potted cores is relatively small.

The stopband range of the active filter does not commence until approximately 15 kHz. This high cut-off frequency is necessary in order to ensure that the phase distortions in the range below 3 kHz are low. Otherwise, the phase-control circuit would not be stable and it would mean that the generated frequency would fluctuate periodically around the selected value.

As previously mentioned, the subsequent phase-locked loop provides a considerable extra selectivity due to its relatively low bandwidth (approx. 3 kHz). Third-order harmonics (spaced ± 3 kHz from the required signal) will therefore no longer interfere. This means that the active filter will not normally be necessary for most applications and the two series-resonant circuits will be sufficient for attenuation of first and second order sidebands (± 10 and ± 20 kHz).

5.1. COMPONENTS FOR THE FILTER

T 1: BC 179, BC 213, BC 415 or similar silicon PNP AF transistor
T 2: BC 108, BC 413 or similar silicon NPN AF transistor

L 1: 25.3 mH; 400 turns in potted core 14 x 8; A_L = 160
L 2: 13.5 mH; 290 turns in a potted core as for L 1

C 1: 10 nF styroflex
C 2: 4.7 nF styroflex
C 3, C 4: 1 nF styroflex
C 5: 220 pF styroflex

6. CHARACTERISTICS OF THE SYNTHESIZER

The short-term stability of an oscillator can be judged simply by monitoring the signal in the SSB or CW modes of a receiver. On testing the described synthesizer, even the sixth harmonic (30 MHz) provided a clean superheterodyne tone. This means that the oscillator is not only sufficiently stable for FM and AM, but also for SSB and CW. The long-term drift is only dependent on the ageing and temperature response of the crystal; it amounts to approximately 10^{-6}. The crystal oscillator can be pulled by approximately $\pm$ 300 Hz with the aid of the series trimmer.

The adjacent channel suppression amounts to approximately 30 dB without phase-locked loop and the filter given in Figure 8. With the additional low-pass filter, the first-order sidebands amounted to -60 dB and the third-order by -80 dB.

The tuning speed amounts to 100 ms/MHz which should be sufficiently fast for practical operation.

7. OPERATION OF THE SYNTHESIZER WITH 25 kHz SPACING

Since the described synthesizer covers a frequency range of 1 MHz, only two crystals will be required in the phase-locked oscillator when used in the European 2 m band from 144 MHz to 146 MHz. The modifications required are limited to the VCO-inductance, variable divider, reference oscillator and active filter.

7.1. A 9 MHz INTERMEDIATE FREQUENCY

As shown in the table in (1), the output frequency range of the phase-locked oscillator is from 135 MHz to 137 MHz. If crystal frequencies of 130 MHz and 131 MHz are used, the frequency synthesizer should operate in the range of 5.0 MHz to 5.975 MHz in steps of 25 kHz. This means that no modifications are necessary to the VCO. The following is valid for the variable divider:

$$n_{min} = \frac{5000 \text{ kHz}}{25 \text{ kHz}} = 200 \quad \text{and} \quad n_{max} = \frac{5975 \text{ kHz}}{25 \text{ kHz}} = 239;$$

It is now necessary for C = 2 to be wired instead of C = 5; since the number 2 corresponds to 0010 in BCD-code, the following is valid: $B_1 = B_3 = B_4$ = logic 0 and B_2 = logic 1. This is obtained by grounding pins 1, 9 and 14 of I 136 and connecting pin 11 to + 5 V. The channel selection is made via A_1 to A_4 and B_1 to B_4 between 00 and 39.

The reference frequency must now correspond to 25 kHz since it always corresponds to the channel spacing. The simplest manner of achieving this, is to replace the 1.44 MHz crystal with one of 3.6 MHz. However, other crystal frequencies are possible as will be seen in the following table:

Crystal Frequency	Division Factor	I 122	I 123
2.5 MHz	100	SN 7490	SN 7490
3.0 MHz	120	SN 7490	SN 7492
4.0 MHz	160	SN 7490	SN 7493
4.8 MHz	192	SN 7492	SN 7493
6.4 MHz	256	SN 7493	SN 7493

However, it is then necessary to change the PC-board DK 1 OF 012 since the integrated circuits SN 7490 and SN 7493 possess different pin connections.

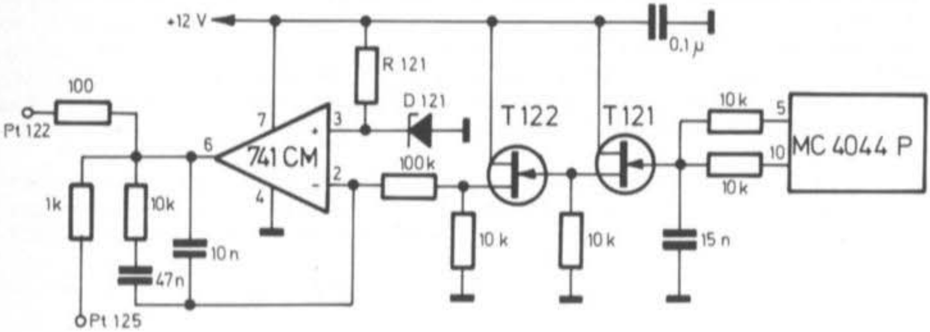

Fig. 10: Low-pass filter modified for a phase-comparator frequency of 25 kHz

The active filter must also be changed due to the alteration of the phase comparator frequency (25 kHz instead of 10 kHz). The corresponding circuit diagram is given in Figure 10. The components not given in this diagram are identical to those given in Figure 4.

7.2. A 10.7 MHz INTERMEDIATE FREQUENCY

The required local oscillator range is from 133.3 MHz to 135.3 MHz and crystal frequencies of 129.3 MHz and 130.3 MHz are used. This means that the required frequency range of the synthesizer is from 4.000 to 4.975 MHz in 25 kHz steps. The number of turns on the VCO inductance L 151 is therefore increased from 45 to 50 turns. The coupling winding need not be altered. The variable divider must be now programmable from:

$$n_{min} = \frac{4000 \text{ kHz}}{25 \text{ kHz}} = 160 \quad \text{to} \quad n_{max} = \frac{4975 \text{ kHz}}{25 \text{ kHz}} = 199;$$

This means that the following is valid: C = 1, e.g. C_1 = logic 1, C_2 = C_3 = C_4 = logic 0. Pins 1, 11 and 14 of I 136 should therefore be grounded and pin 9 connected to + 5 V. The channel selection (inputs A and B) is made between 60 and 99. The same is valid here for the reference oscillator and the active filter as was given in section 7.1.

7.3. PRACTICAL EXPERIENCE

The 9 MHz version of the frequency synthesizer described in section 7.1. is used in the 2 m station of the author. Extensive tests have proved the suitability of this oscillator for SSB operation. Although no additional, low-pass filter is used, the adjacent channels (± 25 kHz) are suppressed by more than 80 dB when referred to the required signal. Higher order sidebands could not be determined. The tuning speed of the 25 kHz synthesizer amounted to approximately 50 ms/MHz so that no delay is noticeable even during duplex operation (via repeaters).

8. REFERENCES

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LOSSES ENCOUNTERED WHEN INTERCONNECTING CABLES HAVING THE INCORRECT IMPEDANCE

by Dr. P. Brumm, DL 7 HG

A radio-frequency signal will be transferred at low loss when the impedance Z_{out} of the generator and the terminating impedance Z_t of the consumer have the same impedance as the interconnecting cable, and when the impedance of this cable is constant over its whole length. These conditions cannot be fulfilled completely in the VHF, and especially not in the UHF or SHF range. Extremely low-loss cable is often only available from surplus sources and the correct connectors are often not obtainable. This means that the amateur is usually forced to make compromises. However, the effects of these compromises on the transfer of radio-frequency energy is virtually unknown. There are probably hundreds of radio amateurs that lose half an S-point or more during the transfer between transmitter and antenna without realizing the reason why they do not radiate so well. Others, on the other hand, become "matching perfectionists" and believe, for instance, that a standing-wave ratio (SWR) of, for example, 1.5 is the reason for considerable losses. The actual losses involved were described in (1).

The following information is to be given in order to be able to establish the actual losses involved and in order to find the most favourable construction. It is assumed that the cables and coupling pieces used are in perfect condition and that no other impedance jumps are exhibited. Corroded or unsuitable coupling pieces cause uncertain impedance jumps of the inner or outer conductor and therefore cause interferring fields outside of the cable. It is not possible for these effects to be calculated. Furthermore, the following considerations only possess a good accuracy for low-loss cables. However, this is no limitation of the usability in practice since the losses caused by impedance jumps are not very important in the case of high attenuation levels.

The following table gives several values for the standing-wave ratio S, the reflected power P_r and the corresponding transfer losses A_{tr} in dB. A_{tr} is the ratio between the actual power transferred at a given S and that power that would have been transferred if S = 1 were present (ideal matching):

S	1	1.1	1.2	1.5	2	2.5	3
P_r (%)	0	0.25	1	4	11	18	25
A_{tr} (dB)	0	-0.01	-0.04	-0.18	-0.5	-0.9	-1.3

It will be seen that the loss A_{tr} only amounts to 0.5 dB with an SWR of 2. This low loss in signal strength will not be audible even with the weakest DX-signals (an S-point is approx. 5 to 6 dB). Noticeable losses will not be present until the SWR is greater than 2.

For this reason, any system that possesses an SWR of 2 or less can be classed of operating perfectly since it is completely immaterial whether the actual SWR amounts to 2 or 1 or any intermediate value.

The following information is to give several simplified formulas and rules which will allow one to estimate whether a system is able to fulfill the criterion $S \leq 2$ or whether it must be improved.

1. A SIMPLE IMPEDANCE JUMP

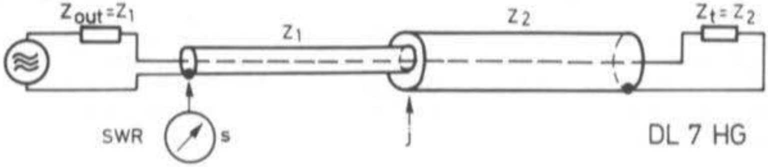


Fig. 1: Simple impedance jump

A jump of impedance from Z_1 to Z_2 is present at position j. Figure 1 shows that $Z_{out} = Z_1$ and $Z_{tr} = Z_2$ exists. The impedance jump j can appear at any position. The resulting SWR is given by:

$$S = \frac{Z_2}{Z_1}, \quad \text{where} \quad Z_2 \geq Z_1.$$

If, however, $Z_1 \geq Z_2$, $S = Z_1/Z_2$ should be formed so that $S \geq 1$ is valid as usual. It would be mathematically more correct to form the complex reflection factor, to determine the amount and then to recalculate the SWR. Since, however, the correct calculation would only complicate this article and since most amateurs are only familiar with SWR, the given definition should be sufficient for our application.

Such a simple impedance jump is relatively harmless with our problems as can be seen in conjunction with two examples:

a) An antenna with a characteristic impedance of 75 Ω is to be connected to a system with a characteristic impedance of 60 Ω (j is transposed to the righthand side of Fig. 1): $S = 75/60 = 1.25$; therefore $A_{tr} = -0.05$ dB.

b) A transmitter with an output impedance of 50 Ω is to be connected to a 60 Ω system (j is transposed fully left in Fig. 1): $S = 60/50 = 1.2$; therefore $A_{tr} = -0.04$ dB.

2. A PIECE OF INTERMEDIATE CABLE WITH INCORRECT IMPEDANCE

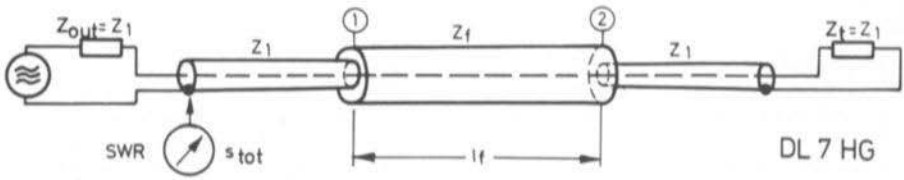


Fig. 2: Impedance jump caused by a piece of cable of incorrect impedance

A piece of cable of the length l_f of the impedance Z_f that differs from Z_1 is to be found in a system which is otherwise correctly matched.

The signal coming from the left in Figure 2 encounters a simple impedance jump $S = Z_f/Z_1$ at position "1". The non-reflected portion of the signal is transferred via l_f and encounters the second impedance jump at position "2" with the same S. The component of the signal reflected at this point is also returned and superimposes itself with the first reflected wave. The phase difference between the two reflected waves is dependent on l_f. The two impedance jumps therefore combine to form a new SWR: Stot, which is now to be calculated.

This calculation is made by using the general transformation formula 1 and by transforming the impedance $Z_1 = Z_{tr}$ at position "1" to position "2", determining the reflection factor and continuing as given above. The resulting formula is relatively complicated. For our applications, several approximations derived from this are sufficient for various lengths of l_f:

2.1. Given: $l_f \leq 0.1\,\lambda_f$; valid is: $S_{tot} = 1 + 2\pi\,(S - \frac{1}{2})\,\frac{l_f}{\lambda_f}$

Note: λ_f is the effective wavelength of the coaxial cable l_f. If its dielectric is not vacuum or air, it is necessary for the wavelength in vacuum λ_0 to be multiplied by the velocity factor $VF = 1/\sqrt{\varepsilon}$ of the actual dielectric (e.g. solid dielectric (PE): VF = 0.66; polyethylene foam (FPE): VF = 0.72 etc.).

2.2. Given: $l_f = \lambda_f/4$ (or $3\,\lambda f/4$, $5\,\lambda f/4$ etc.). Valid is $S_{tot} = S^2$

2.3. Given: $l_f = \lambda f/2$ (or $2\,\lambda f/2$, $3\,\lambda f/2$ etc.). Valid is $S_{tot} = 1$.

This means that if a piece of cable of incorrect impedance is used in an otherwise matched system, and its length is an even multiple of an electrical half-wave, it will have no effect on the impedance! It is therefore possible to compensate the first impedance jump by the second.

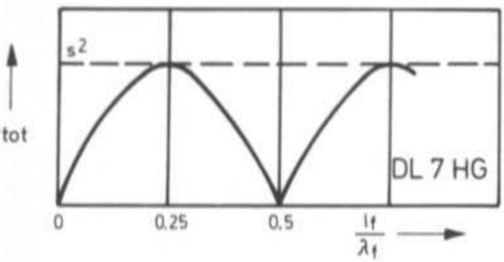


Fig. 3: Run of the standing wave ratio as a function of the electrical length of the cable with incorrect impedance

The whole run of S_{tot} with l_f is given in Figure 3. It can be seen that S_{tot} possesses its maximum value at $l_f = \lambda_f/4$.

This is very useful for estimating the system losses, since if this worst possible value is acceptable no further calculations are needed and a given system will be suitable, immaterial how long the length l_f is.

Example:

c) BNC-interconnection (50 Ω) in a 60 Ω system at 435 MHz:
$S = 60/50 = 1.2$; insulation: teflon (PTFE), $VF \approx 0.7$; $\lambda_0 = 70$ cm, therefore $\lambda_f = 50$ cm; $l_f = 3$ cm therefore $l_f/\lambda f = 0.06$, therefore condition 2.1.: $S_{tot} = 1 + 0.2 = 1.2$. Even if the interconnection were to be $\lambda/4$ it would still be extremely good. At 1300 MHz S_{tot} would be between 1.2 and 1.44 (Fig. 3). e.g. approx. 1.3.

d) Two cables with PL 259 connectors are to be used in a 60 Ω system at 145 MHz with a SO 239/SO 239 coupling piece. These coupling pieces have no defined impedance! Furthermore, longitudinal inductivities appear on the inner conductor and on the transition from the coaxial cable to outer connector of the plug. If we assume that these inductivities are avoided by careful connection, an "impedance" of between 20 and 30 Ω can be estimated from the geometric dimensions. If the calculation is made with $Z_f = 25\ \Omega$, $S = 60/25 = 2.4$ will be valid. With a VF = 0.7 (estimated) and $\lambda_0 = 2$ m, $\lambda_f = 1.4$ m will be valid; actually, $l_f = 4$ cm, which means $l_f/\lambda f = 0.03$, which means that condition 2.1. is valid: $S_{tot} = 1 + 0.38 = 1.38$ and thus $A_{tr} = 0.1$ dB. A value of 1.33 was determined with a precision standing wavemeter, which coincides closely with the calculated values. It will also be seen that this coupling is no longer usable for 435 MHz.

It has been seen that a piece of cable having an incorrect impedance will not cause any great effect on the transfer of the signal as long as it remains in the order of the conventional cable impedances of 50 to 75 Ω. Unfortunately, several coupling pieces are usually to be found between the transmitter and the antenna, which means that the previous considerations are not sufficient for judging such a system.

3. SEVERAL INTERMEDIATE PIECES HAVING THE INCORRECT IMPEDANCE

A very extensive calculation was carried out for two pieces l_f. The final formula was extremely complicated. However, it is possible to give an approximation for the most unfavourable case for two pieces from which it is possible to estimate the conditions when several pieces of cable are present that have the incorrect impedance (corresponding to 2.2):

If n pieces of incorrect impedance are present in a transmission system and if these pieces possess the individual SWR S1, S2, S3 .... Sn, it is possible for the total SWR S_{tot} to be established for the most unfavourable case as follows:

$$S_{tot} = S_1 \times S_2 \times S_3 \times \ldots S_n$$

Mostly, the actual value will be considerably less than this.

Example:

e) A 435 MHz system is constructed with 60 Ω cables and BNC coupling pieces. Flexible coaxial cable is used in the shack and around the rotator, whereas the actual interconnection cable between the station and the antenna consists of a low-loss, but very inflexible cable. SWR meter and transmit receive relay have an impedance of 50 Ω. The system is as follows:

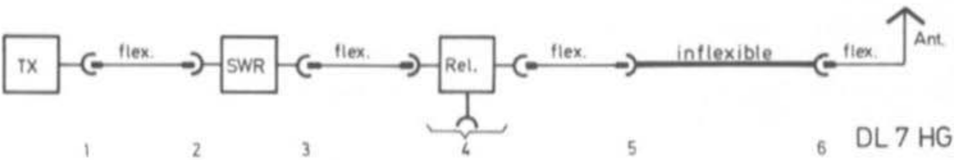


Fig. 4 : Transmit system with several cables and coaxial connections

An S_4 of 1.33 is calculated for the relay. The construction of the SWR meter is unknown (l_f), so that it is necessary for the impedance jump S_2 at the input as well as S_3 at the output to be calculated. We therefore have: $S_1 = S_2 = S_3 = S_5 = S_6 = 1.2$ (see example c) and $S_4 = 1.33$. It is therefore possible for S_{tot} to be $1.2 \times 1.2 \times 1.2 \times 1.33 \times 1.2 \times 1.2 = 3.3$.

It is therefore very easy to obtain the total values of the standing wave ratio by multiplication of the individual values that are harmless when considered individually.

However, this estimation is made considering the worst possible case and the system will most possibly a far better value. This can occur, for instance, by making the cables between 1-2 and 3-4 (Fig. 4) to even multiples of $\lambda/2$ and by matching the transmitter to 50 Ω. In this case, the line from transmitter to point 5 will be seen as 50 Ω impedance which means that only the 60 Ω cables will possess the "incorrect" impedance. Section 2.3. is then valid. If a 60 Ω coupling piece is used for point 6, only one simple impedance jump of 50/60 Ω will remain at point 5 which means that a $S_{tot} = 1.2$ results. Of course, the same effect is obtained when 50 Ω cables is used for the given pieces.

Example:

f) A 23 cm system as shown in Figure 4 will have a higher SWR, since S_1, $S_2 .... = 1.3$ (see example c). It is also possible using suitable cable dimensions to obtain a good standing wave ratio.

Example:

g) A 2 m system similar to Figure 4 is provided with coupling pieces PL 259/ SO 239. An individual PL/SO coupling piece possesses $S = 1.19$ (see example d). This means that the seven coupling pieces can produce $S_{tot} = (1.19)^7 = 3.4$. This means that these coupling pieces cannot even be used at 145 MHz in any great number. However, suitable cable length and matching of the transmit output coupling are also able to reduce the SWR.

All previous considerations are also valid for receive systems, only that the antenna represents the source and the receiver the consumer.

Final note: Most of the "standing wave meters" that are within the price of most radio amateurs are not suitable for <u>measurement</u> of the SWR and are often only suitable for indicating the RF output. One can therefore not expect that the calculated value coincides with a value measured using such a meter. In addition to this, the cable attenuation must be considered, which can be made easily from diagram given in (2).

4. REFERENCES

(1) T. Bittan: Antenna Notebook
VHF COMMUNICATIONS 5 (1973), Edition 4, Pages 220-223

(2) J. Sturm: Standing Wave Ratio and Cable Attenuation
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 85-88

ANTENNA NOTEBOOK

by T. Bittan, DJ Ø BQ / G 3 JVQ

Details on how two or more antennas could be stacked (vertically) or bayed (horizontally) to obtain maximum gain were given (1). Furthermore, a nomogram (2) was given allowing amateurs, to calculate the gain from the vertical and horizontal beamwidth and to estimate whether the manufacturers gain figures can be correct or not.

1. STACKING AND BAYING FOR MAXIMUM DIRECTIVITY

This article is to cover the stacking and baying of antennas to obtain maximum directivity. It has already been mentioned that the beamwidth is reduced approx. by 50% on stacking or baying two identical antennas. Every antenna has a number of null points in its polar diagram which allow unwanted signals to be rejected. It is possible by adjustment of the spacing between the antennas to shift these nulls to the required angle from the main beam. The angle of the null from the main beam as a function of the spacing between the antennas is given in the following table:

Degrees from main beam	Spacing in λ	Degrees from main beam	Spacing in λ
5/175°	5.700	50/130°	0.654
10/170°	2.880	55/125°	0.610
15/165°	1.970	60/120°	0.578
20/160°	1.540	65/115°	0.553
25/155°	1.180	70/110°	0.533
30/150°	1.000	75/105°	0.518
35/145°	0.874	80/100°	0.508
40/140°	0.780	85/ 95°	0.501
45/135°	0.710	90/ 90°	0.500

2. INFINITE FRONT-TO-BACK RATIO

Most well-designed Yagi antennas provide a front-to-back ratio of 20 - 25 dB. This may not be sufficient for some applications (The author has a high power 2 m station with an ERP of 30 000 W within line-of sight.).

It is possible to increase the front-to-back ratio to an infinite value by altering the phase relationships between the two antennas. This is achieved in the following manner: The two antennas are mounted as shown in Figure 1 so that one antenna leads the other mechanically by λ/4. Assuming that the upper antenna leads, the operation can be described as follows: A signal from the required direction will reach the upper antenna λ/4 (90°) before the lower antenna. If a delay line of λ/4 (λ/4 x VF) is provided between the upper antenna and the stacking cable, the signal from both antennas will have the same phase relationship and full gain will be provided.

A signal arriving from the rear, however, will reach the lower antenna λ/4 before the upper antenna, resulting in a phase lag of the upper antenna of 90°. The delay line of λ/4 will provide a further phase lag of 90° which results in a total phase shift of 180° causing complete cancellation of the unwanted signal.

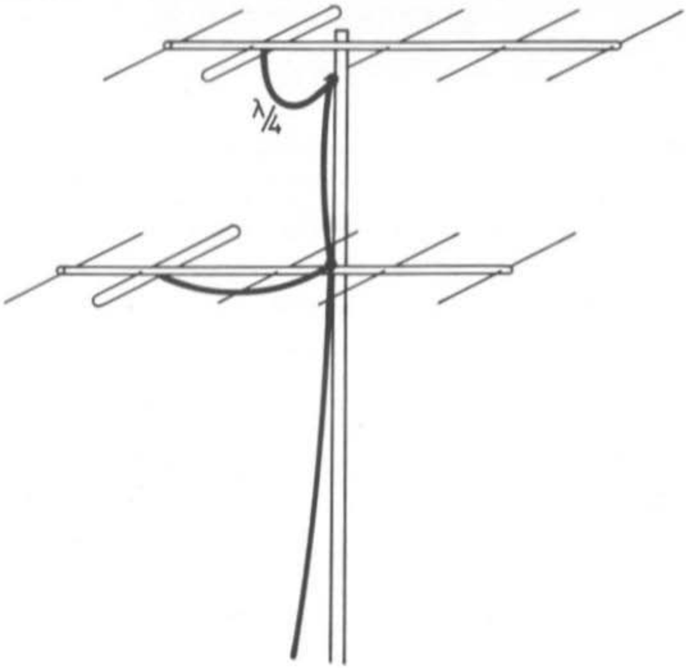


Fig. 1: Construction of an antenna for maximum front-to-back ratio

In practice complete cancellation of the unwanted signal is not achieved since it is seldom possible to obtain the correct phase relationships as the signal can arrive not completely horizontally at the antennas, and there is always a certain amount of signal scattered back from hills, trees, buildings etc. However, the described system is able to increase the front-to-back ratio up to 50 - 60 dB without difficulties. This can, of course, be increased further by really optimizing the phase relationship in the field by altering the length of the calculated phase line, or by simply shifting the stagger of the two antennas.

The following Antenna Notebook will describe various antennas that consist of stacked arrays. These are the colinear group antennas and the skeleton-slot types of arrays and the variants which have originated from these types of arrays.

3. REFERENCES

(1) T. Bittan: Antenna Notebook
VHF COMMUNICATIONS 6 (1974), Edition 2, Pages 82-84

(2) Karl Rothammel: Antennenbuch, Edition 4, Page 161.

A HELICAL ANTENNA FOR 70 cm

by W. Stich, OE 1 GHB

A helical antenna is to be described for clockwise circular polarization (RHCP). It is designed for a centre-frequency of 435 MHz, but since such helical antennas are extremely wideband, all dimensions with the exception of those of the matching transformer are not critical. A photograph of the antenna is given in Figure 1. Several articles have been published which have indicated the advantages of circular polarization (1), (2). The future launch of the Oscar 7 satellite is a further reason for using circular polarization. Helical antennas have been used for some time in professional telecommunications circles.

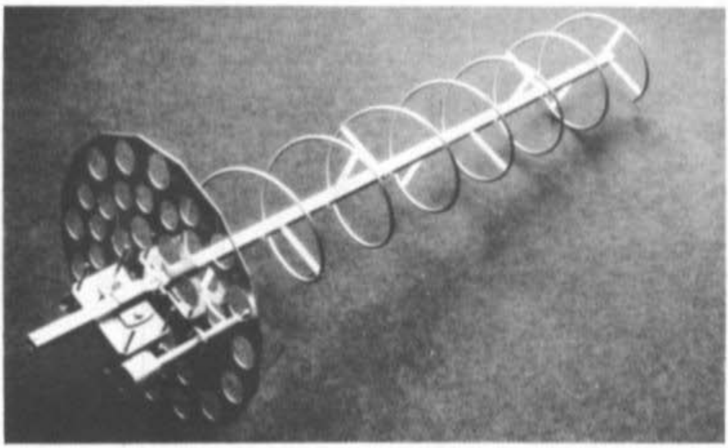


Fig. 1:

Photograph of the 70 cm helical antenna

1. CONSTRUCTION

The antenna was designed according to information given in (1) and (3). For mechanical reasons, the number of turns was limited to 7; the active part of the antenna is approximately 127 cm which corresponds to 1.8 λ. The overall dimensions of the antenna are 148 cm long (boom) and 45 cm in diameter (reflector). Figure 2 gives a cross-section of the antenna.

The ohmic component of the feedpoint impedance of approximately 150 Ω is transformed to approximately 52 Ω using a $\lambda/4$ transformer having an impedance of 88 Ω. This means that 50 Ω or 60 Ω coaxial cable can be used for feeding the antenna.

The reactive component of the feedpoint impedance is compensated by shifting the reflector somewhat back from the antenna. On shifting the reflector back and forward along the boom experimentally, it was found that this caused a deterioration of the standing wave ratio. The given spacing results in the most favourable standing wave ratio and is identical to information given in (3).

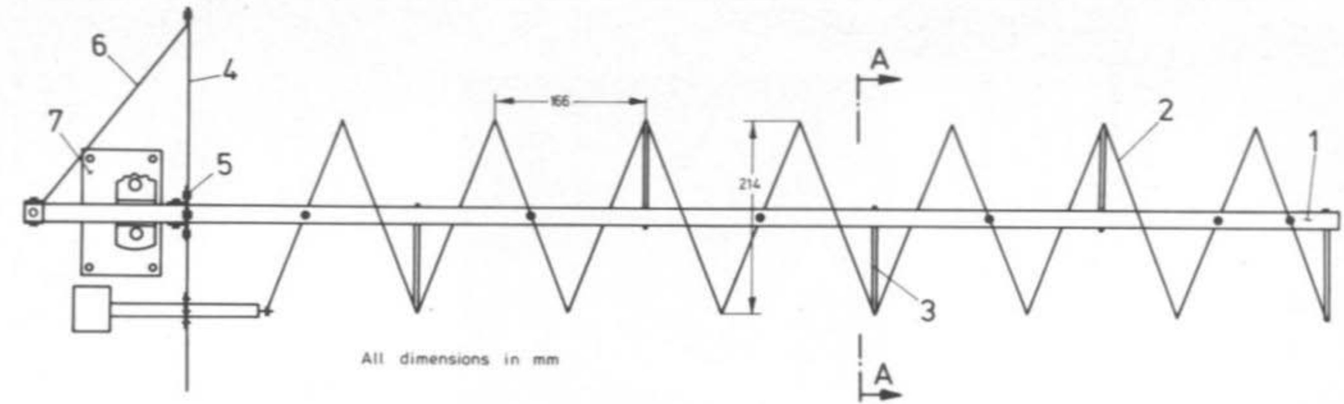
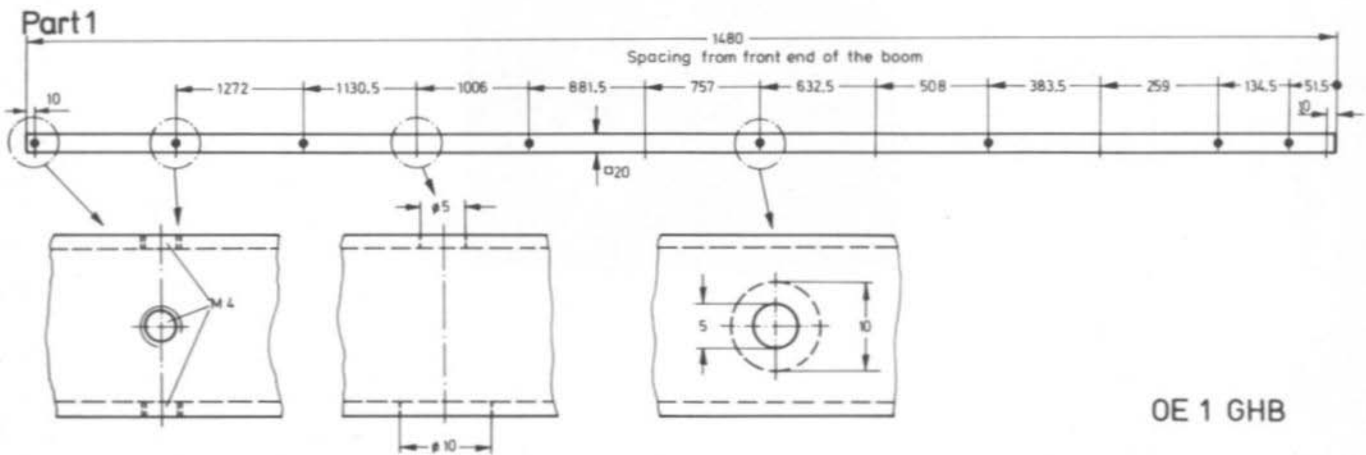


Fig. 2: Cross-section of the complete helical antenna



OE 1 GHB

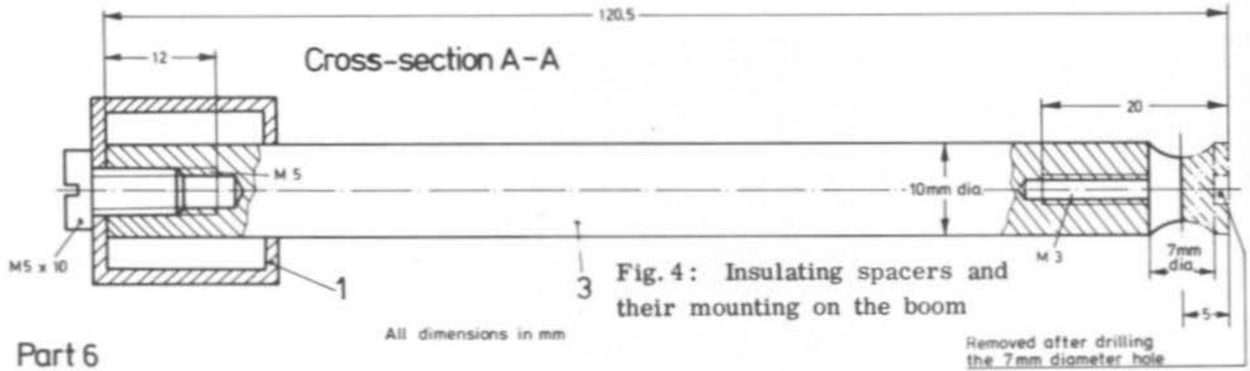
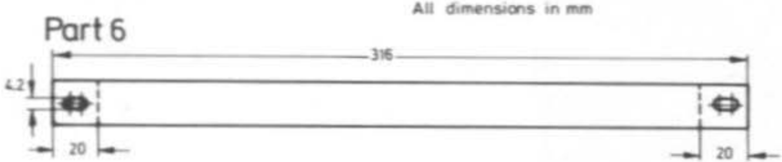


Fig. 4: Insulating spacers and their mounting on the boom



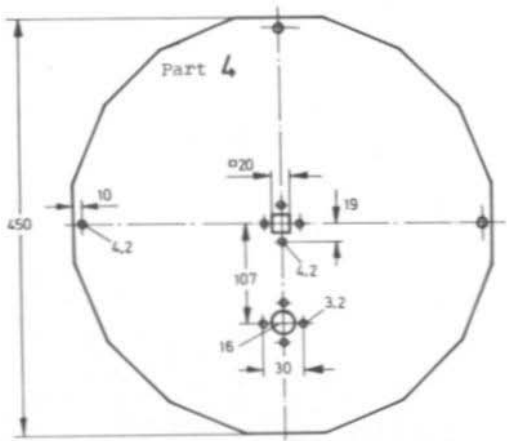


Fig. 6: Dimensions of the reflector

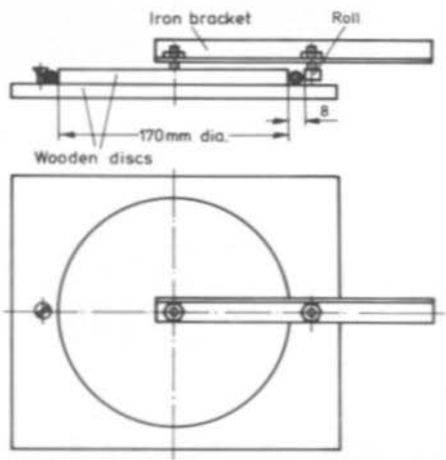


Fig. 8: Bending tool for the helical element

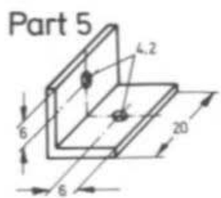
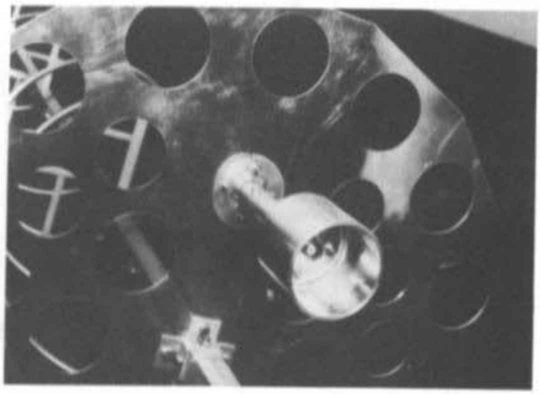


Fig. 9: Brackets and support for mounting the reflector

1.1. BOOM AND SPACERS

The boom is 1480 mm long. It should be drilled as shown in Figure 3 where the given spacings are measured from the front end of the boom. Information as to the material used is given in Section 2. The 11 spacers (part 3) can be made from any insulating material that is weatherproof and easy to work with. The spacers are 120.5 mm in length and have a diameter of 10 mm. It is recommended that they should be provided with an approximately 3 mm diameter thread (M 3) at one end and a 5 mm thread (M 5) at the other. A 7 mm hole is drilled at right angles to the M 3 thread. After this, the spacer is sawn off at 5 mm so that the support is now provided with a semicircular cross-section which is then able to accept the helical element. The side with the M 5 thread is finally passed through the 10 mm hole in the boom and screwed to the other side.

Fig. 5 : Photograph of the reflector plate and matching transformer



1.2. THE REFLECTOR

The reflector consists of a 2 mm thick aluminum plate and can be provided with holes in order to reduce the wind loading. The holes should not be more than $\lambda/10$, e.g. no more than 7 cm. The dimensions of the reflector as well as the cut-outs for the boom, matching transformer and supports (part 6) are given in Figure 6.

1.3. MATCHING TRANSFORMER

Figure 7 shows all individual parts of the $\lambda/4$ transformer. Firstly, the two identical parts 9 are soldered onto the tube part 8. This is followed by soldering or glueing (dual-component adhesive) the weather protection (part 11) onto the disc at the end of the transformer. The (BNC) connector is mounted onto part 10 after the nut has been filed so that it fits into tube 8. The two polystyrol discs (part 13) are glued to the inner conductor (part 12) and the inner conductor soldered to the centre pin of the connector. The inner conductor is then placed into tube (part 8) and possibly glued to the front insulating disc so that the transformer is completely water-tight. The matching transformer is now screwed to the reflector with disc (part 9) at the centre.

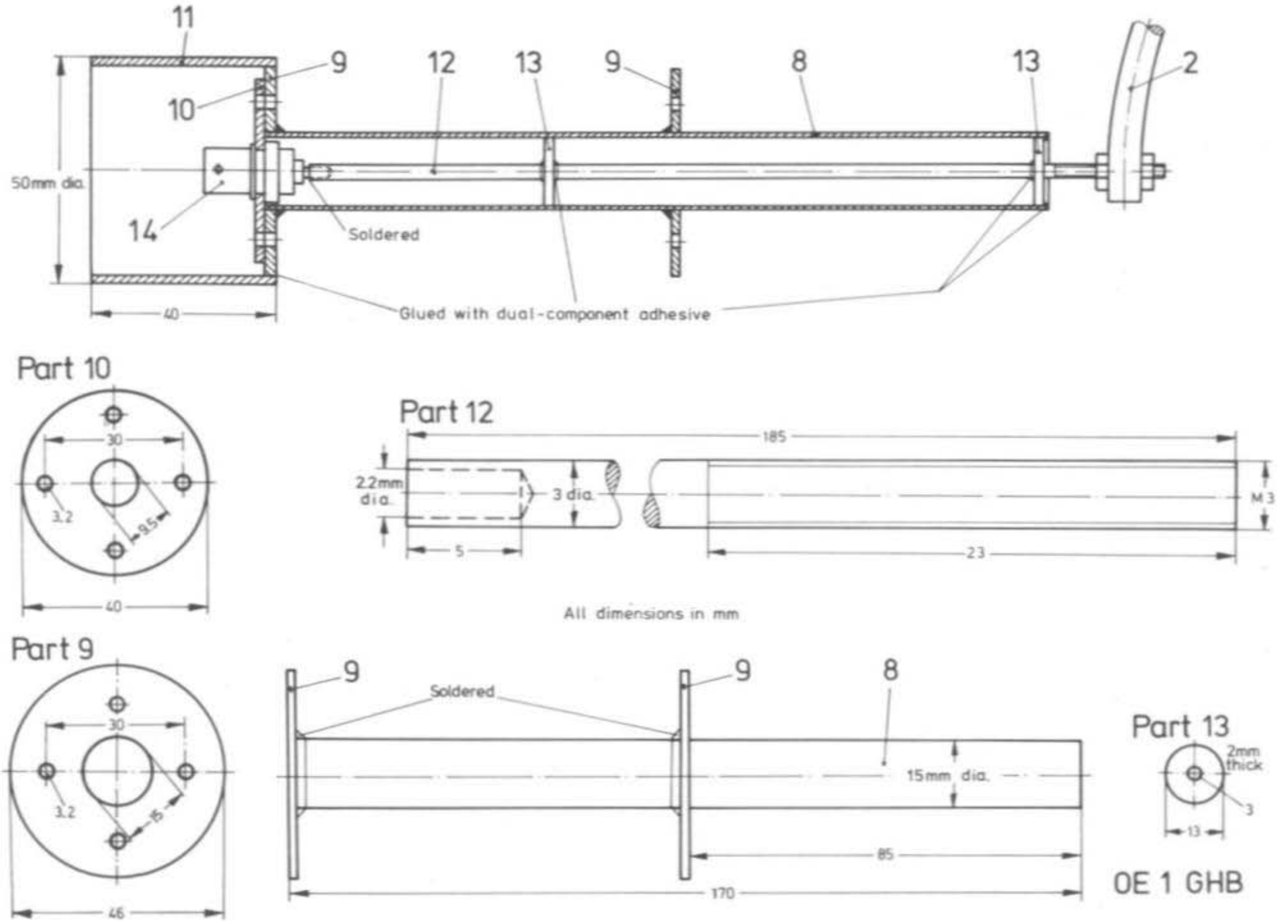


Fig. 7: Individual parts of the matching transformer

1.4. THE HELICAL ELEMENT

The helical element can be made with the aid of the bending tool shown in Figure 8. It is made from 7 mm diameter aluminum wire. One end of the wire is held with the aid of the large countersunk screw (shown on the left). The wire is then bent around the wooden disc with the aid of the lever after which the countersunk screw is released and the next section pushed into place and bent, etc. Attention should be paid that a clockwise thread results.

The screwshaped helical element is then pulled out to the required spacing between turns. The final corrections are made after placing the helical element over the boom and spacers, and placing these into the mounts so that virtually no lateral pressure is present.

The length of wire used for the helical element has been given as 8 meter so that the two (usually scratched) ends can be sawn off. A 3.5 mm hole is now drilled at the centre of the helical element where it is screwed to the appropriate support. The other mounting holes are then marked and drilled from the centre in both directions after which the helical element is screwed into place. The end which is to be connected to the matching transformer is bent straight and also provided with a 3.5 mm diameter hole (Fig. 7 above, right).

The reflector and built-in $\lambda/4$ transformer can now be screwed onto the boom. This is done with the aid of four brackets (part 5) and three supports (part 6) as shown in Figure 9.

MATERIAL REQUIREMENTS

Part no.	Number	Material	Total Requirements
1	1	square aluminum tubing 20 x 20 x 1.5 mm	1.6 m
2	1	7 mm dia. soft aluminum wire	8.0 m
3	11	plastic spacers, 10 mm dia.	1.5 m
4	1	2 mm thick aluminum plate	50 x 60 cm
5	4	aluminum brackets 15 x 15 x 2 mm	10 cm
6	3	aluminum tape 20 x 4 mm	1.2 m
7	1	mast clamp	
8	1	15 mm dia. x 1 mm aluminum tubing	20 cm
9	2	2 mm thick aluminum plate	
10	1	2 mm thick aluminum plate	
11	1	50 mm dia. x 2 mm aluminum tubing	5 cm
12	1	3 mm dia. brass rod	20 cm
13	2	2 mm thick polystyrol or similar	10 cm^2
14	1	BNC connector UG-1094/4	
	10	M3 nuts	
	8	M3 x 8 mm screws	
	11	M3 x 15 mm screws	
	7	M4 nuts	
	7	M4 x 5 mm screws	
	7	M4 x 10 mm screws	
	11	M5 x 10 mm screws	

3. MEASURED VALUES

The standing wave ratio was measured to be 1.4 over the whole 70 cm band which means that the matching is completely satisfactory (0.1 dB loss when compared to true matching).

The gain was determined by measurement against an antenna with a known gain. It amounts to 9.5 dB over a circular-polarized crossed dipole. The -3 dB beamwidth amounts to 50°. The helical element of this antenna can, of course, be extended as required in order to increase the gain. It will not be necessary to change the matching, but the mounting of the antenna to the mast will have to be changed.

4. REFERENCES

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(2) T. Bittan: Circular Polarization on 2 Metres
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(3) K. Rothammel: Antennenbuch
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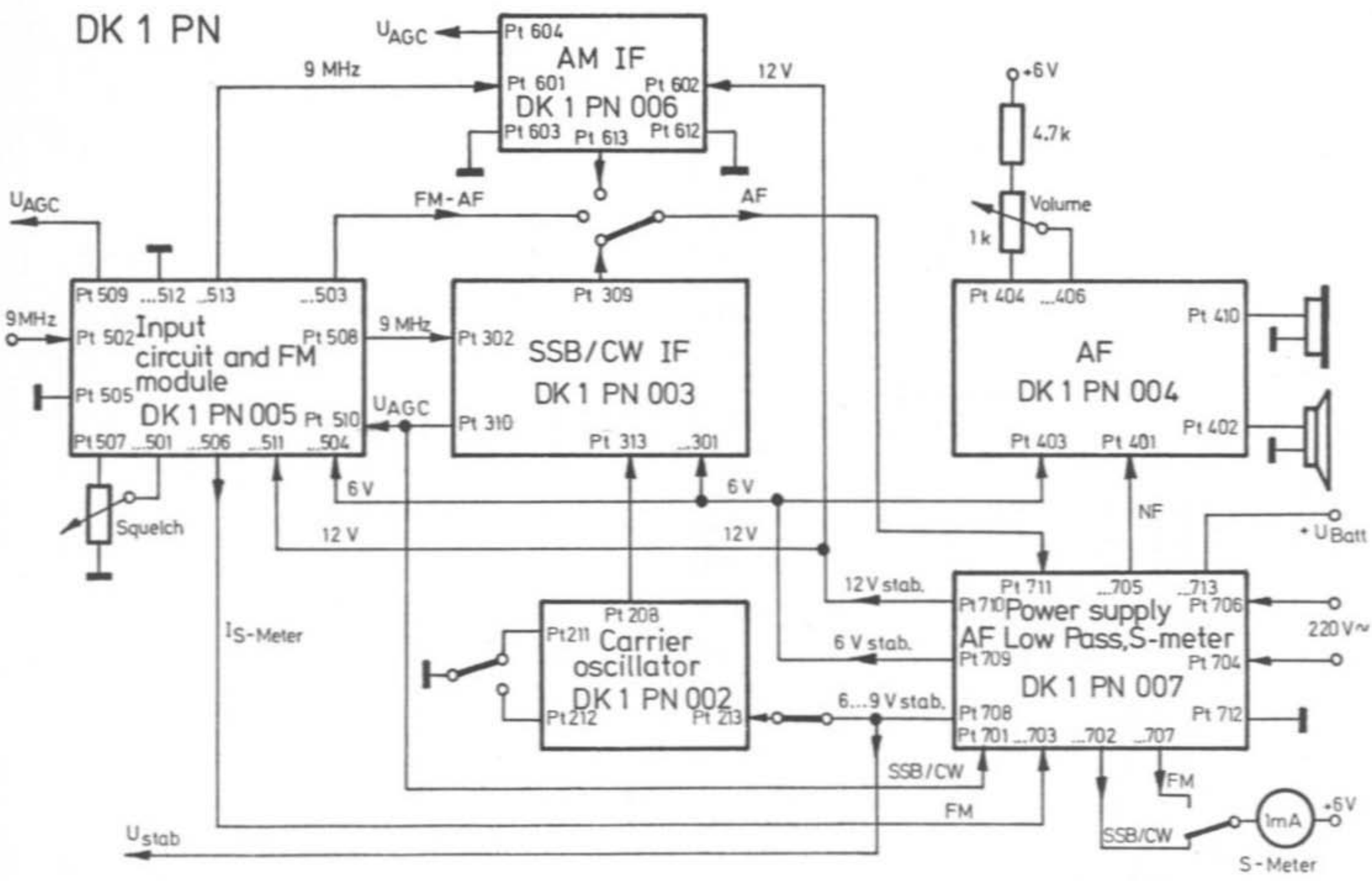

Fig. 1: Block diagram of the complete, integrated 9 MHz receiver system

AN INTEGRATED RECEIVER SYSTEM FOR AM, FM, SSB and CW

PART VII: THE AM PORTION

by H. J. Franke, DK 1 PN

This article describes the sixth module of the 9 MHz integrated receiver. This means that the IF-portions of the complete receiver have now been described completely for all modes. In order to simplify construction of the complete unit, a so-called system board is being developed that provides connectors for all six modules, as well as connections for potentiometers, switches, S-meters, the loudspeaker and power-line cable. The block diagram of all IF-portions of the receiver is given in Figure 1.

1. CHARACTERISTICS AND CIRCUIT DETAILS

Since AM (A 3) has lost popularity for VHF and UHF communications after being mainly replaced by SSB and FM, the AM IF-portion should be small and inexpensive. For this reason, no separate, 6 kHz wide crystal filter was used, and the selectivity is provided by the 12 kHz bandwidth of the FM crystal filter in the input circuit DK 1 PN 005. The noise bandwidth which determines the sensitivity is limited in the active AF filter in module DK 1 PN 007 to approx. 2.5 kHz. The two resonant circuits of the AM IF-portion DK 1 PN 006 shown in Figure 2 are only used for matching.

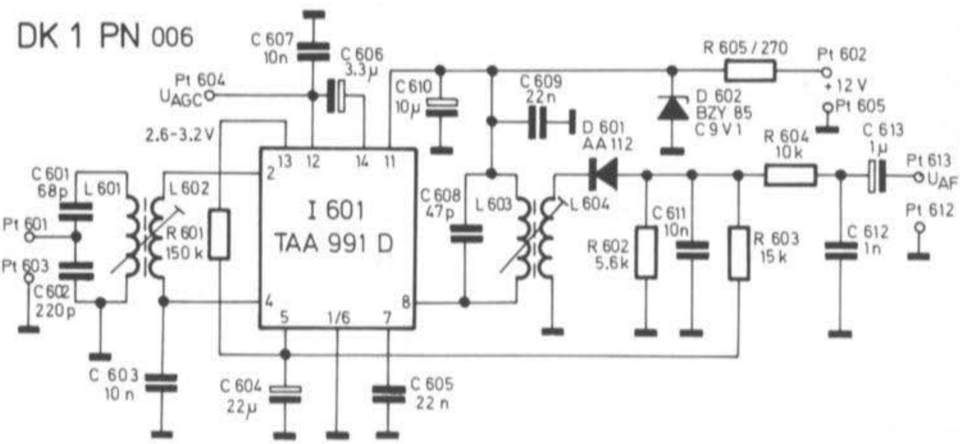

Fig. 2: Circuit diagram of the AM-IF-module DK 1 PN 006

The input impedance is relatively high with approximately 1.5 $k\Omega$ so as not to load the source-follower output (Pt 513) of module DK 1 PN 005. For this reason, the interconnection should be made with a short wire and not with coaxial cable.

The integrated circuit TAA 991 used in the AM-portion is a combined AM/FM IF-amplifier for use in broadcast receivers. The nominal operating voltage is 9 V. Since the operating voltage should not exceed 11 V, the operating voltage is stabilized at 9.1 V in module DK 1 PN 006 with the aid of a zener diode. A stabilized voltage of between 2.6 and 3.2 V (pin 13) is generated within the integrated circuit; this is used as the bias current for an internal transistor (pin 5) and for demodulator diode D 601.

The circuit provides a high sensitivity: S + N/N = 3 dB is provided for a 9 MHz input voltage of 0.4 μV. AM signals will be readable with 9 MHz input voltages in excess of approximately 1 μV (Fig. 3). An overload condition will not take place up to approximately 25 mV.

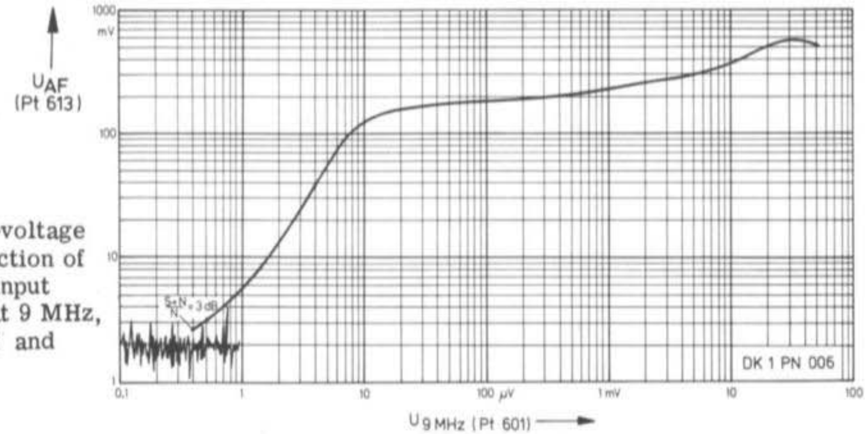

Fig. 3 :
The AF-voltage as a function of the RF-input voltage at 9 MHz, 80 % AM and 1 kHz

The integrated circuit TAA 991 also generates a control voltage (Fig. 4) which is not used in the described receiver system, since the gain control is made in module DK 1 PN 005 via the FM-channel. However, this control voltage of the AM module is available at connection Pt 604 for other applications.

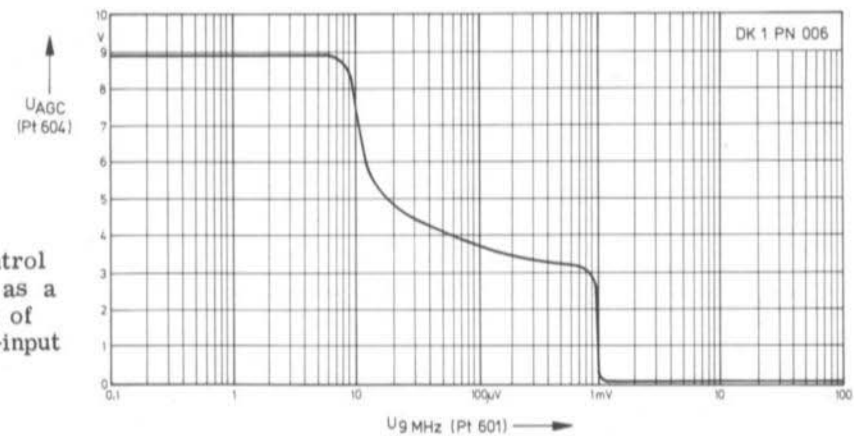

Fig. 4 :
The control voltage as a function of the RF-input voltage

2. CONSTRUCTION AND COMPONENT DETAILS

The AM-module DK 1 PN 006 is also of modular construction and is enclosed in a TEKO box. However, since the module is extremely small, it is only necessary to use the small TEKO box 2 A. PC-board DK 1 PN 006 is 65 mm x 45 mm and is double-coated. Due to the high gain of the integrated circuit, the continuous ground surface on the component side of the board is absolutely necessary. The component locations are shown in Figure 5. Those connections that are not grounded are drilled and a certain amount of the ground surface should be removed around these connections to ensure that no short-circuits take place. The grounded connections are shown as black points in the component location plan and these are directly soldered to the ground surface on the component side.

I 601: TAA 991 D (Siemens)

D 601: AA 112, AA 116, 1 N 87 A or similar germanium demodulator diode
D 602: BZY 85/C9V1 or similar 9.1 V zener diode

L 601: 30 turns, L 602: 6 turns of enamelled copper wire in special coil set
L 603: 30 turns, L 604: 21 turns of enamelled copper wire in special coil set as L 601.

With the exception of the inductances and capacitors used in the resonant circuits, all other components are not critical.

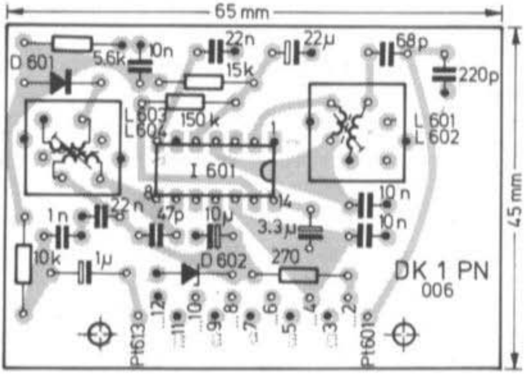

Fig. 5 : Component locations on PC-board DK 1 PN 006

3. ALIGNMENT

A 9 MHz signal is required for alignment and this can be provided by module DK 1 PN 002 (carrier oscillator) via a voltage divider to input Pt 601 of the AM module. It is also possible for the alignment to be made together with module DK 1 PN 005 when the signal is fed to Pt 502. A voltmeter is connected to connection Pt 604 and the RF-input voltage adjusted so that the control voltage corresponds to a mean level of approximately 5 to 7 V (Fig. 4). It is now only necessary for both resonant circuits to be aligned for minimum control voltage. If the control voltage should drop below approximately 3 V during the alignment process, it is necessary for the RF input voltage to be decreased.

4. OTHER APPLICATIONS OF THE AM MODULE

The AM-module DK 1 PN 006 is just as suitable for use at other intermediate frequencies. Any frequency can be used from a few hundred kHz up to at least 10.7 MHz. The upper frequency limit is not given by the manufacturer. It is only necessary for the two resonant circuits to be recalculated for the required frequency. The transformation ratios, e. g. turns ratio and the capacitance ratio C 601/C 602, as well as the L/C ratio should be maintained.

The input of the module can be designed for 50 to 60 Ω by decreasing the value of C 601 to approximately 51 pF and C 602 to approximately 1 nF.

If the module is to be used in a (portable) receiver where the operating voltage is only 9 V, zener diode D 602 should be deleted and resistor R 605 bridged. The operating voltage can be in the range of 4.5 V to 11 V. Of course, the drive range decreases at lower operating voltages than at 9 V. The control voltage at Pt 604 is suitable for controlling MOSFET's.

Since there is still quite a bit of AM activity in many countries, e.g. England and France, this module is also very suitable for modifying FM equipment for AM reception. Since most of the FM receivers use an IF of 10.7 MHz which is then usually converted down to 450 kHz, this AM module could be connected either to the 10.7 MHz or 450 kHz IF.

5. REFERENCES

(1) H. J. Franke, R. Lentz: An Integrated Receiver System for AM, FM, SSB and CW
Part II: The SSB IF Portion
VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 47-53

(2) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW
Part III: The Carrier Oscillator
VHF COMMUNICATIONS 5 (1973), Edition 3, Pages 167-168

(3) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW
Part IV: AF Amplifier and CW Filter
VHF COMMUNICATIONS 5 (1973), Edition 4, Pages 208-211

(4) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW
Part V: Input Module and FM Portion
VHF COMMUNICATIONS 5 (1973), Edition 4, Pages 212-219

(5) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW
Part VI: Power Supply, AF-Lowpass Filter and S-Meter Stages

(6) VHF COMMUNICATIONS 6 (1974), Edition 2, Pages 107 - 113

SECOND FRENCH EDITION OF VHF COMMUNICATIONS

AN INTEGRATED RECEIVER SYSTEM FOR AM, FM, SSB and CW

PART VIII: THE SYSTEM BOARD

by H. J. Franke, DK 1 PN

This article is to describe the system board which has been mentioned in conjunction with the last two modules of the 9 MHz receiver. In addition to this, several modifications are to be described for the individual modules.

This means that the 9 MHz, all-mode receiver is thus complete. The great advantage of this system is that it requires only a minimum of alignment. The individual description and possible modifications are again listed in the references (1) to (6).

It is, of course, possible for 10.7 MHz filters to be used and the receiver aligned to this intermediate frequency. Any receive converter providing a suitable intermediate frequency can be used. Two different control voltages are available for controlling the preamplifier stages of the receive converter: One having suitable characteristics for controlling the Plessey ICs and the second suitable for controlling dual-gate MOSFETs.

If the 9 MHz receiver is only to be used for shortwave operation where frequency modulation is not required, it is possible for module DK 1 PN 005 to be equipped with an AM crystal filter instead of the FM crystal filter and for the integrated circuit CA 3089 E to be deleted. Gate 2 of transistor T 501 should then be fed via a voltage divider via Pt 509 with approximately 4.5 V. The author and editors would be very interested in a design of a suitable, switchable receive-converter for the five amateur shortwave bands. For single-band operation, it is possible to use module DJ 4 BG 011 (7); in this case, the first crystal filter should be in this module instead of in module DK 1 PN 005.

This combination is also suitable for use with UHF converters having an intermediate frequency of 28 to 30 MHz. However, for 2 m operation, a single-conversion superhet is preferable with respect to the capability of handling strong signals. A 144 MHz/9 MHz receive converter is to be described in one of the next editions of VHF COMMUNICATIONS that has been designed by DK 1 OF. It is extremely suitable for use with this system.

1. THE SYSTEM BOARD DK 1 PN 008

As can be seen in the photograph of the author's prototype given in Figure 1, the six individual modules are mounted back-to-back in two rows on the system board. The connection pins are to be found on three sides of this board. 13-pole connectors are used throughout. All modules are mounted so that the alignment elements are accessible when the covers have been removed.

The system board has the dimensions 240 mm x 87 mm and has been designated DK 1 PN 008. It is double-coated so that the conductor lanes do not radiate and so that good ground conditions exist. The component locations are given in Figure 2. The upper surface is not given since it only consists of a continuous ground surface which has only been etched around the connection pins and around the few components. With the exception of the six 13-pole connectors, only a few ferrite chokes and bypass capacitors are provided. The 10 kΩ resistor is

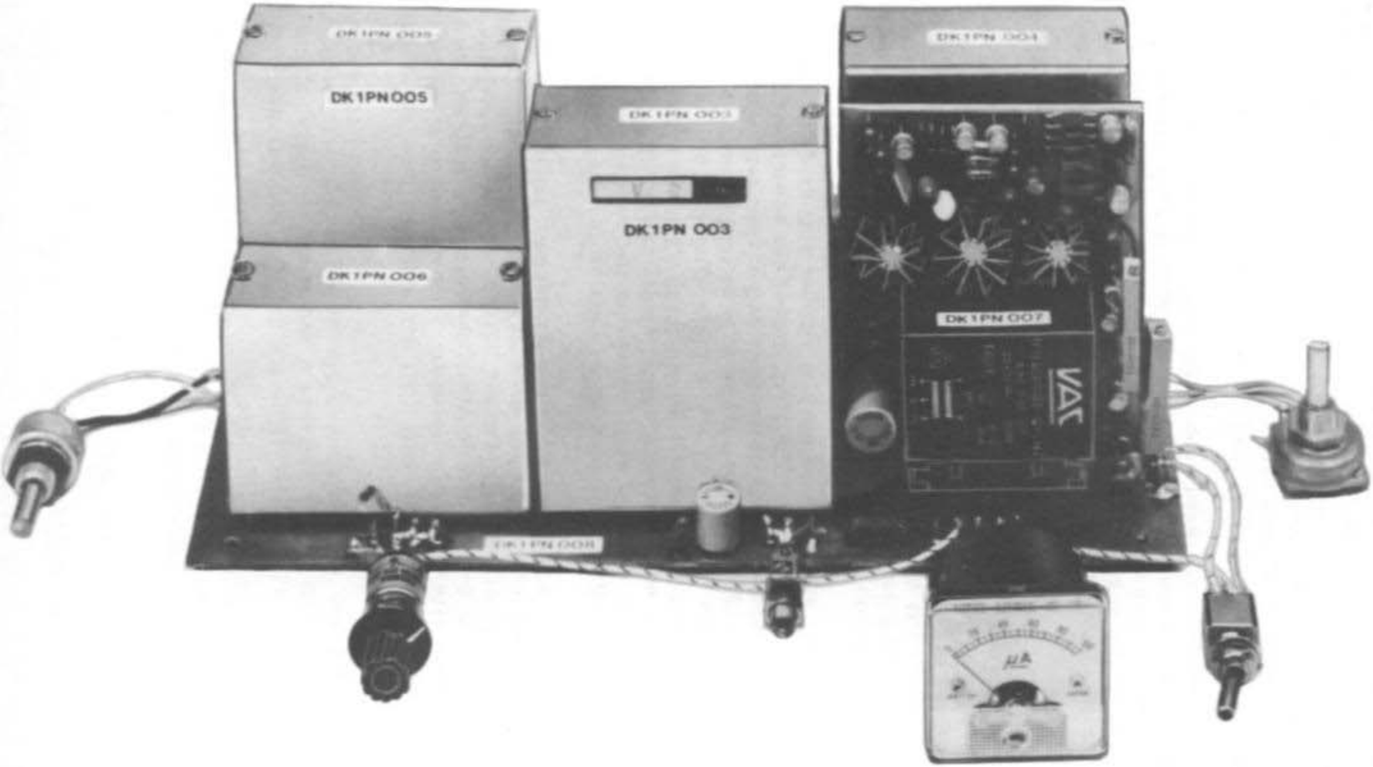

Fig. 1: Author's Prototype of the DK 1 PN 9 MHz Receiver - Mounted on the System Board

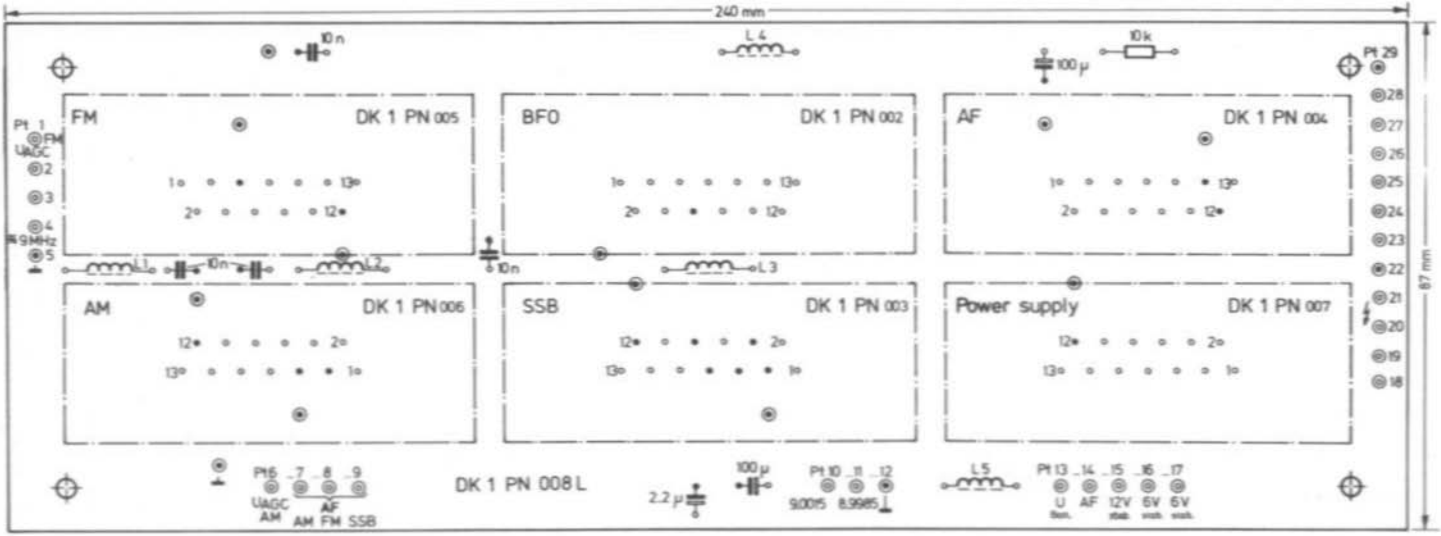

Fig. 2: Conductor Lanes and Component Locations on the System Board DK 1 PN 008

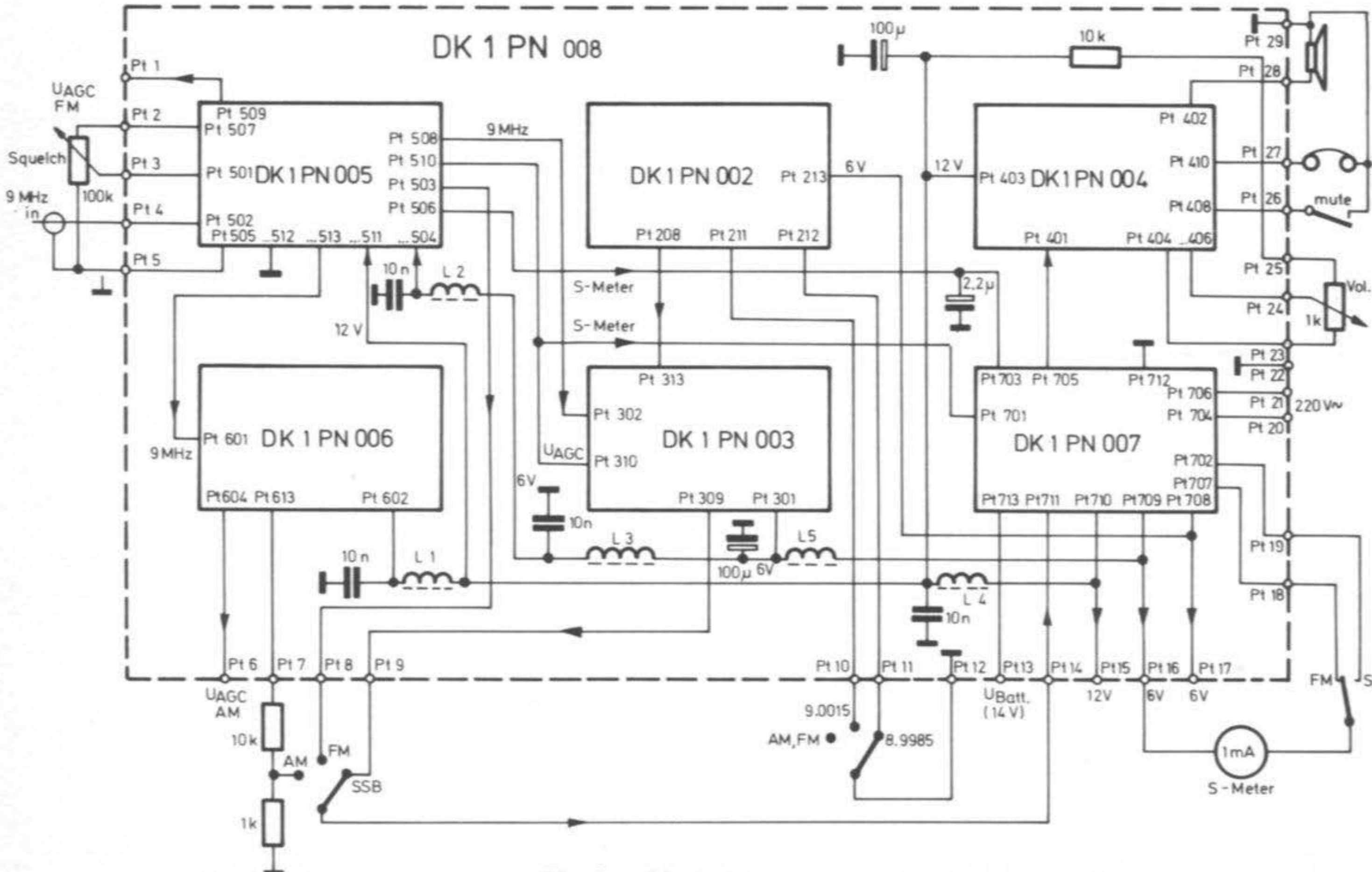

Fig. 3: Block Diagram of the Complete 9 MHz Receiver Showing Required External Components

is used as volume control for module DK 1 PN 004. Connections are made to the ground surface at eleven points on the board, which are designated by a ringed point on the component location plan. These connections should be made on the upper and lower side of the board, since the components, (connectors, capacitors) cover this and can therefore not be soldered to the upper side of the board.

Figure 3 gives the complete block diagram of the 9 MHz receiver with all components that are accommodated on the system board and also indicatesthe required external components. A coaxial cable should be connected to the 9 MHz input (Pt 4/Pt 5); all other connections should be made with normal insulated wire and are not critical. The individual switches and potentiometers can be combined if required, as can two S-meters be provided instead of one. The voltage divider at connection Pt 7 matches the high AF-voltage of the AM-module to the output of the other modules.

Battery operation is possible via connection Pt 13; this point and the receive converter are fed with a stabilized voltage of 12 V via Pt 15. The 6 V voltage at connection Pt 17 should only be used for the oscillators. Connection Pt 16 is available for all other 6 V consumers. Higher level AF-amplifiers should be supplied from the unstabilized voltage available at Pt 13, however, attention should be paid that the built-in power transformer is not overloaded (5 VA).

1.1. COMPONENTS LOCATED ON THE SYSTEM BOARD

6 13-pole connectors for solder mounting
5 wideband ferrite chokes 2.5 turns in a 6-hole core
2 electrolytic capacitors 100 µF/16 V
1 electrolytic capacitor 2.2 µF/6 V (tantalium drop type)
4 ceramic capacitors of 10 nF or more
1 resistor of 10 kΩ

1.2. PREPARATIONS ON THE SYSTEM BOARD

After drilling the system board, the 29 connection pins and the 11 ground connections are mounted. The through-connections should not protrude on the upper side (ground surface) of the board. After this, the components and the six 13-pole connectors are soldered into place. The connectors should also be screwed into place so that no pressure is made on the solder connections. It is now possible for all external connections to be made.

2. MODIFICATIONS TO THE INDIVIDUAL MODULES

It is assumed that each of the individual modules have been built up as described in VHF COMMUNICATIONS. A preliminary alignment is not necessary since the final alignment can be made easily on the system board. However, it is necessary for the lower mounting holes of the TEKO boxes to be increased to a diameter of 3.2 mm and for M 3 nuts (or similar) to be glued into place with a dual-component adhesive. The two upper holes per box can remain as they are for self-tapping screws, or also be changed. After hardening of the adhesive, the individual modules are screwed to the system board with the aid of 10 mm long M 3 (or similar) screws. If required, M 4 nuts can be used as spacers. This mounting also provides an additional ground connection.

2.1. CARRIER OSCILLATOR DK 1 PN 002

With the given dimensioning of the base voltage divider comprising R 202/R 203, the collector current will be so high that virtually no collector-emitter voltage will remain due to the voltage drop across the collector and emitter resistor. This results in an output voltage which is too low and may even be distorted. This fault can be avoided by increasing the resistance value of R 202 to 33 kΩ.

2.2. AF AMPLIFIER DK 1 PN 004

According to the latest information published by Plessey, the AF amplifier SL 630 should be operated from 12 V, (Maximum permissible voltage: 18 V). This fact has been taken into consideration on the system board. The dropper resistor for the volume control should therefore be increased from 4.7 kΩ to 10 kΩ. The loudspeaker should have an impedance of at least 8 Ω. The most favourable impedance would be approximately 40 Ω; in this case, the amplifier would provide an output power of approximately 200 mW.

For adjustment of the frequency response, Plessey also recommended an additional 100 pF capacitor in addition to capacitor C 402 which is then connected between connection 4 of the SL 630 and ground. In addition to this, two modifications are required within module DK 1 PN 004 which have already been mentioned in VHF COMMUNICATIONS:

Capacitor C 404 should be increased to 22 nF and an additional direct-ground connection should be made on the PC-board from connection 10 of the SL 630 to Pt 11/12 of the plug-in connector.

2.3. POWER SUPPLY DK 1 PN 007

The supply voltage of 12 V was found to have too high an impedance for operating the AF amplifier. For this reason, two modifications are to be given each of which is suitable:

The dropper resistor R 701 for the zener diode D 703 should be reduced from 560 Ω to 390 Ω; or the parallel capacitor C 702 can be increased from 1 μF to 10 μF or 22 μF.

3. CHECKING THE OPERATION

Firstly insert only module DK 1 PN 007 and connect to the power line. The voltages at connections Pt 13 and Pt 15 to Pt 17 should be measured. If these are correct, it is possible for the other modules to be inserted. The extremely simple alignment of modules DK 1 PN 005 and 006 should be carried out according to the information given in the individual articles. Finally, potentiometer P 301 of module DK 1 PN 003 should be adjusted so that the SSB S-meter just indicates a slight noise level with converter and antenna connected. The full scale deflection can be adjusted with the aid of resistor R 709.

In the case of the FM S-meter, the full scale deflection can be adjusted with the aid of resistor R 711. Further details are given in the individual articles describing these modules.

4. REFERENCES

(1) H. J. Franke, R. Lentz: An Integrated Receiver System for AM, FM, SSB and CW - Part II: The SSB IF Portion
VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 47-53

(2) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW Part III: The Carrier Oscillator
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(3) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW Part IV: AF Amplifier and CW Filter
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(4) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW Part V: Input Module and FM Portion
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(5) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW Part VI: Power Supply, AF-Lowpass Filter and S-meter Stages
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In this edition of VHF COMMUNICATIONS.

(7) D. E. Schmitzer: Shortwave Receiver Module
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2 m CONVERTER WITH EXTREMELY HIGH SELECTIVITY

by H. Sütterlin, DL 1 LS

The author demonstrated the above 2 m converter at the German VHF Convention in Weinheim. The extremely high selectivity, and good large-signal handling capabilities of the converter made it extremely suitable for repeater stations. Of course, such characteristics are just as desirable for home and mobile stations. For this reason, the converter is now to be described. The high selectivity is provided using bandpass filters using coaxial lines with helical inner conductors. Since the construction is made in individual chambers, it is more extensive than would be the case when using printed circuit boards. However, it is not too extensive for amateur means. A certain degree of measuring equipment is necessary if the selectivity is to be used to the full. The described converter does not possess a local oscillator, but such an oscillator strip can be taken from one of the many converters that have been described in this magazine.

1. SPECIFICATIONS

Bandwidth and selectivity: According to the alignment of the bandpass filter:

Bandwidth	Selectivity at approx. 1 MHz spacing
0.5 MHz	20 dB (Fig. 1)
1 MHz	16 dB (Fig. 2)
2 MHz	

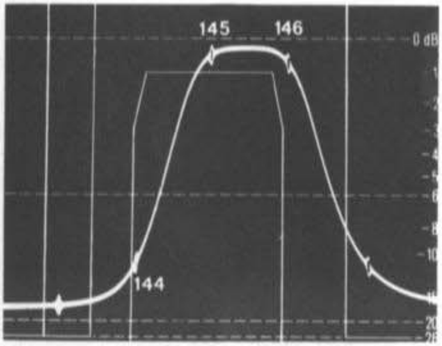


Fig. 1: Passband curve of the DL 1 LS converter

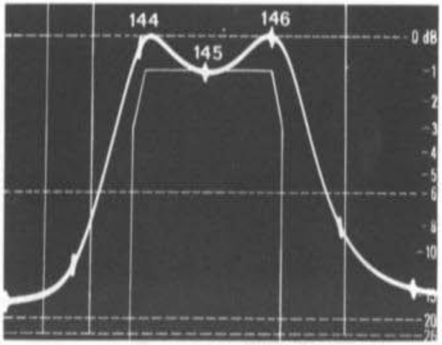


Fig. 2: Passband curve with fixed coupling

Gain: 22 dB with a bandwidth of 0.5 MHz;
17 dB with a bandwidth of 2 MHz;
measured with a local oscillator voltage of 1.5 V at the mixer.

Input matching: Reflected power less than 1%.

The other specifications were measured with an IF strip of 10.7 MHz.

Image rejection: 105 dB (with 21.4 MHz spacing).

Sensitivity: Signal-to-noise ratio = 20 dB for U_{in} = 0.25 μV
(measured at 60 Ω with $\pm$ 8 kHz deviation)

Large-signal-capabilities: With a required signal of 1 μV at the input of the receiver an increase of noise by 10 dB was observed when an unwanted carrier was present at the given frequency spacings and at the given power ratios in excess of the wanted signal:

Frequency spacing	Strength of the interferring signal (referred to 1 μV into 60 Ω)
1.6 MHz	96 dB (156 dB with additional 3-link filter)
600 kHz	90 dB (135 dB with additional filter)
100 kHz	70 dB
50 kHz	60 dB

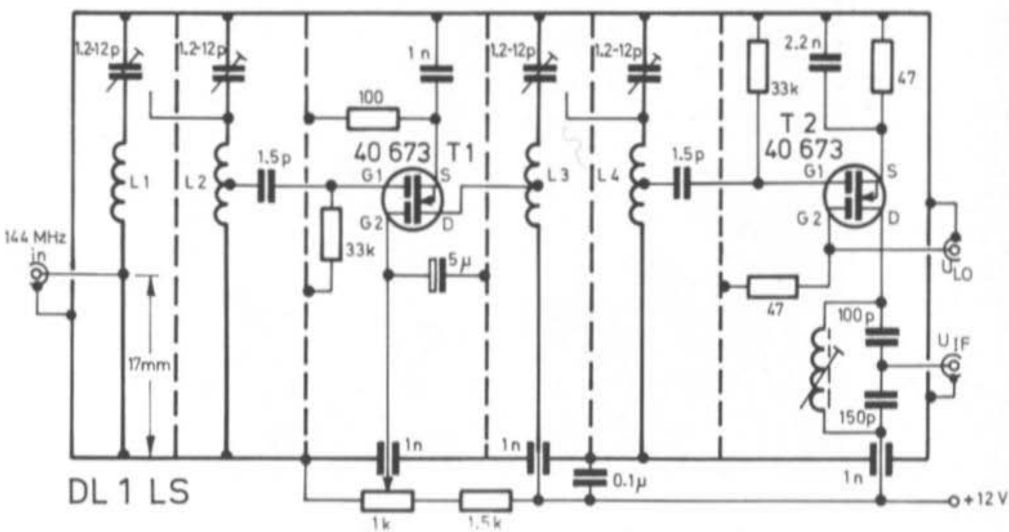


Fig. 3: Circuit diagram of the 2 m converter DL 1 LS

2. CIRCUIT DETAILS

The circuit diagram of the converter is given in Figure 3. One bandpass filter is provided before the RF amplifier and between the mixer stage and IF amplifier. The four VHF circuits are resonant helical lines, they are tuned with the aid of a high-quality tubular ceramic trimmer. The input is designed and tuned for power matching. Wires of 1.5 mm diameter (15 AWG) and 30 mm in length are used for coupling the bandpass filters and are placed near to the hot end of the circuits. This type of coupling allows the converter to be adjusted to a bandwidth of between 0.5 MHz and 2 MHz according to the application.

Gate 1 of the dual-gate MOSFET is loosely coupled. The gain of this transistor can be varied with the aid of the DC-voltage at gate 2 of this transistor. The voltage required for this is between approximately + 5 V and 0 V and can be adjusted manually or obtained from a rectifier circuit in the IF chain and used as AGC. A 5 μF capacitor with short connections is mounted in the vicinity of the intermediate panel in order to increase the control-time constants and for filtering.

The mixer is also equipped with a protected dual-gate MOSFET transistor. Gate 2 should be fed with a low-harmonic local oscillator voltage of 1.5 V for full gain. Gate 2 is grounded via 47 Ω for DC-voltages. This resistor is used simultaneously for terminating the local oscillator voltage so that a 50 Ω coaxial cable can be used for interconnecting the converter to the local oscillator module.

The resonant circuit at the drain is designed for a frequency of 10.7 MHz in the given application. The capacitive tap is selected so that the required bandwidth is obtained under load. It is only necessary for this one resonant circuit to be changed when used in conjunction with other intermediate frequencies.

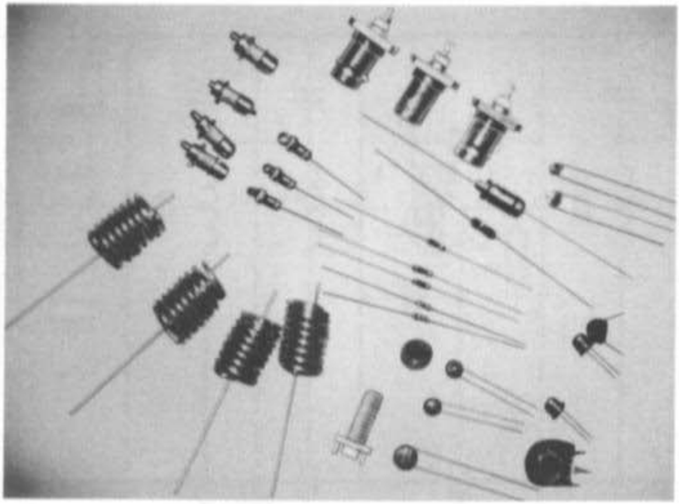


Fig. 4: Components prepared for installation

3. COMPONENTS (Fig. 4)

T 1, T 2: 8 turns of 1.5 mm dia. (15 AWG) silver-plated copper wire, self-supporting, outer diameter 14 mm, coil-length 20 mm, long end 34 mm.
Coil tap on L 2: 4 turns from hot end (trimmer)
Coil tap on L 3 and L 4: 5 turns from hot end.

L 5: 20 turns of 0.4 mm dia. (26 AWG) enamelled copper wire wound on a 5 mm coil former with short-wave core.

Fig. 3 : Holes on the boom (dimensions from the front end)

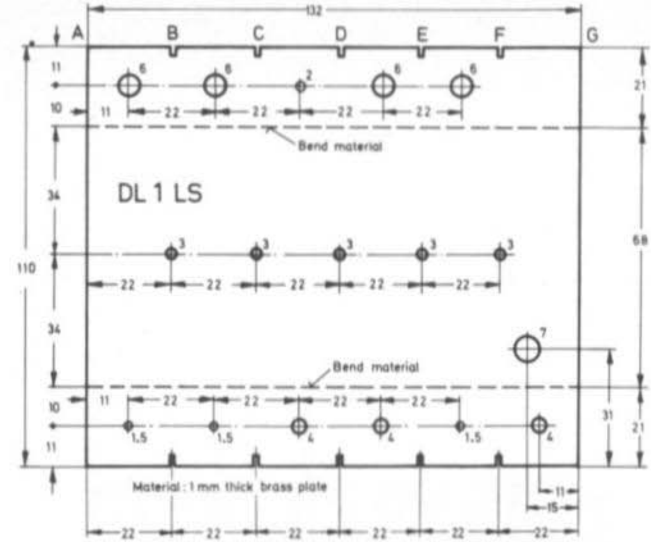


Fig. 6 : Dimensions of the chassis

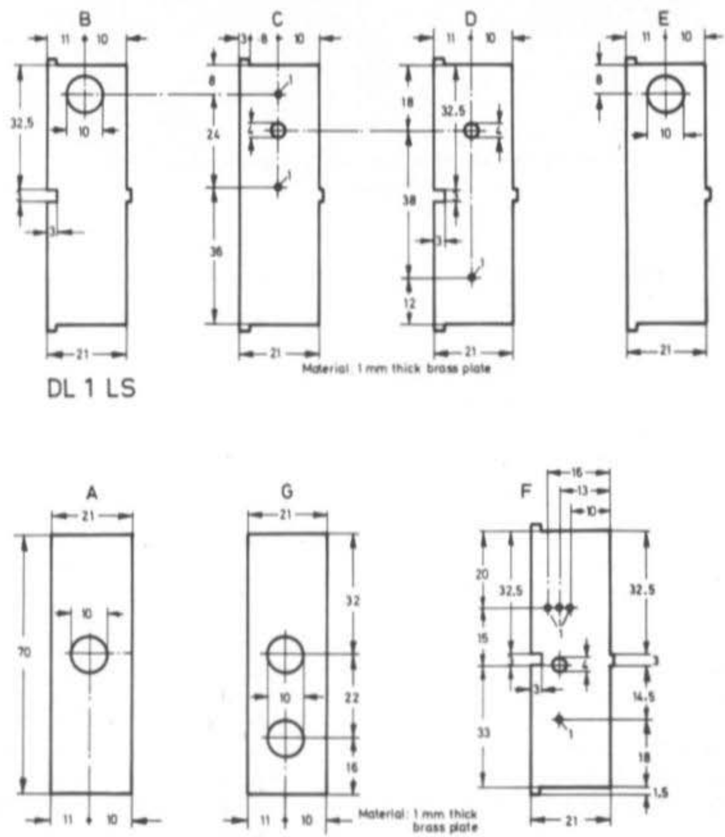


Fig. 7 : Dimensions of the intermediate and end panels

4 ceramic tubular trimmers 1.2 - 12 pF for single-hole or solder mounting.
3 feedthrough capacitors for solder mounting. Approx. 1 nF
(value uncritical 7
3 BNC connectors with square flanges.

4. CONSTRUCTION

Figure 5 shows a photograph of the completed converter with the cover removed. The case is constructed by cutting a 1 mm thick brass plate as shown in Figure 6, which is then drilled and bent up as indicated by the dashed lines. Finally, the five intermediate panels and two side panels are prepared as shown in Figure 7. The intermediate walls should be placed into the appropriate holes or slots in the U-shaped parts and soldered into place. The end pieces are then soldered to the two ends of the U-shaped chassis.

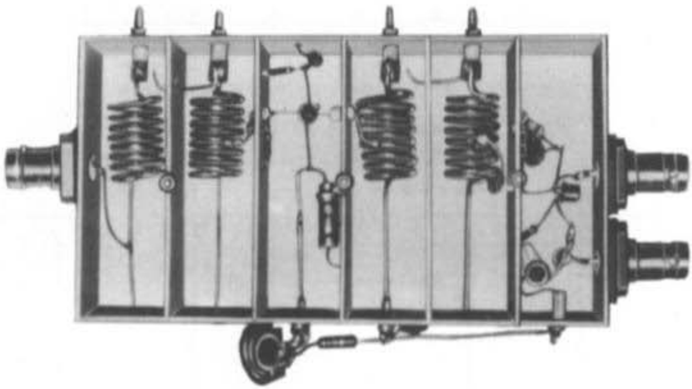


Fig. 5: Highly-selective 2 m converter DL 1 LS

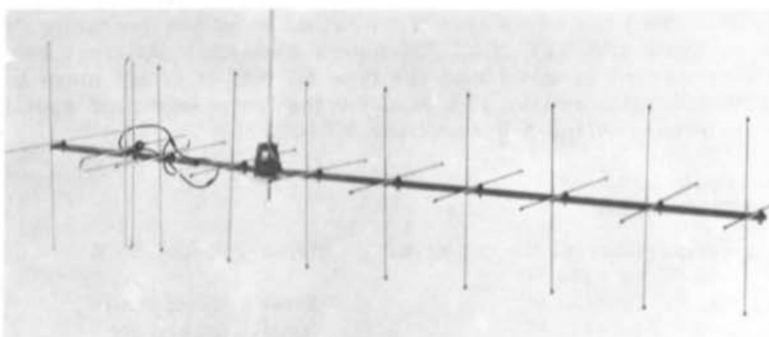
Brass screws should be soldered to the tops of the intermediate panels as shown in Figure 5 and used for mounting the cover (134 mm x 70 mm x 1 mm). In order to completely screen the converter, a layer of metal foil is placed between the case and cover in a similar manner to that used with TV-tuners.

In order to ensure that the BNC connectors do not protrude too far into the case, two spacing plates are mounted between the outer side of the case and the flange of the connectors. It is recommended that the case be silver-plated, but this is not absolutely necessary.

The sources of both transistors are bypassed by connecting the capacitors without connection leads or similar capacitors with very short connections to the nearest screening panel. The coupling capacitors of the bandpass filters are constructed from thick, silver-plated copper wire which is soldered to a trimmer and fed through a hole in the screening panel and then brought into the vicinity of the other trimmer capacitor. The coupling wires can be clearly seen in Figure 5. The IF output coil L 5 is glued to a suitable position in the case.

5. ALIGNMENT

At full gain, the current requirements of the converter are approximately 14 mA (10 mA for the amplifier stage). This value should be reduced on varying the gate 2 voltage of the first transistor and should increase slightly on connecting the oscillator voltage. Attention should be paid during the alignment that the bandpass curve is symmetrical. At a bandwidth of 2 MHz, the amplitude-response is less than ± 0.5 dB. The alignment for the lowest bandwidth is most favourable for repeater applications and when the activity is limited to a small portion of the band.



THE NEW J-BEAM MOONBOUNCERS

INTEGRATED 5 V VOLTAGE STABILIZER FOR 1 A

by U. Tillmann, DJ 5 UO

A voltage stabilizer is available for the power supply of integrated circuits that possesses an prealigned control circuit, a driver amplifier and pass transistor completely integrated in a single case. When this integrated circuit is complemented with two external capacitors and fed with approximately 10 V DC, an output voltage of 5 V at a current of max. 1 A will be available. The IC is also provided with integrated protection circuits against over-current and thermal overload.

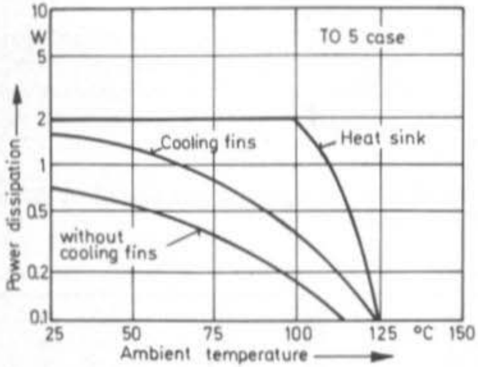
Whereas the variable voltage stabilizer described in (1) equipped with the LM 305/SG 305 and external power transistor possesses slightly better specifications for laboratory power supplies and for providing the operating voltage for transistor circuits, the integrated 5 V stabilizer equipped with the LM 309/ SG 309 (109, 209) has advantages in the provision of the operating voltage of modules equipped with TTL-ICs. Although a number of external components and construction time is saved with the type SG 309, it is not more expensive than the SG 305. This article is to describe the more important specifications and the application of the 5 V stabilizer SG 309.

1. SPECIFICATIONS

1.1. LIMIT VALUES

Operating temperature of the junction:		Input voltage: 35 V
SG 109:	-55 °C to +150 °C	
SG 209:	-25 °C to +150 °C	Power dissipation:
SG 309:	0 °C to +125 °C	limited internally

The above stabilizers are available both in a TO 5 and TO 3 case. Figure 1 and Figure 2 show which current can be provided at a given voltage drop, heat sink and ambient temperature with both versions. Typical values are: Overvoltage 10 V, e.g. voltage drop 5 V; ambient temperature 50 °C. This means that the TO 5 version can be used up to a power dissipation of approximately 1.2 W with appropriate cooling fins which then corresponds to approximately 250 mA. Under the same conditions, the TO 3 version equipped with a heat sink KS 100 (75 mm long) provides a power dissipation of approximately 7 W which then corresponds to $I_{max} \approx 1.4$ A.



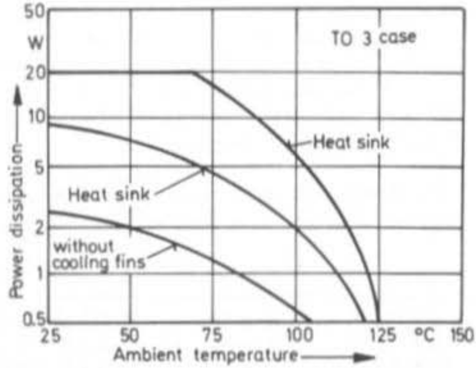


Fig. 1 + 2: Maximum permissible power dissipation of the integrated circuit SG 109/209/309

1.2. OPERATING VALUES

The following specifications are valid, where not otherwise given, for a junction temperature $T_j = 25\ ^oC$; $U_{in} = 10\ V$; $I_{out} = 0.1\ A$.

Parameter	Measuring conditions	min.	typ.	max.	Unit
Output voltage SG 109/SG 209 SG 309		 4.9 4.8	 5.0 5.0	 5.1 5.2	 V V
Fluctuation of the output voltage with an input voltage fluctuation	$7\ V \leq U_{in} \leq 25\ V$	-	4	50	mV
Fluctuation of the output voltage with output current fluctuations TO 5 version TO 3 version	 $5\ mA \leq I_{out} \leq 0.5\ A$ $5\ mA \leq I_{out} \leq 1.5\ A$	 - -	 20 50	 50 100	 mV mV
Total tolerance of the output voltage TO 5: $I_{max} = 0.2\ A$; $P_{max} = 2\ W$ TO 3: $I_{max} = 1.0\ A$; $P_{max} = 20\ W$	$7\ V \leq U_{in} \leq 25\ V$ $5\ mA \leq I_{out} \leq I_{max}$ $P \leq P_{max}$ $\Delta T_j - \Delta T_{jmax}$	4.75		5.25	V
Quiescent Current	$U_{in} \leq 25\ V$	-	5	10	mA
Suppression of interference voltage at the input	$10\ Hz \leq f \leq 10\ kHz$	-	75	-	dB
Interference component of the output voltage	$10\ Hz \leq f \leq 100\ kHz$	-	40	-	μV rms
Output impedance	$10\ Hz \leq f \leq 10\ kHz$	-	0.1	-	Ω
Longterm voltage deviation		-	-	10	mV

Figure 3 shows the output voltage fluctuation measured by the author.

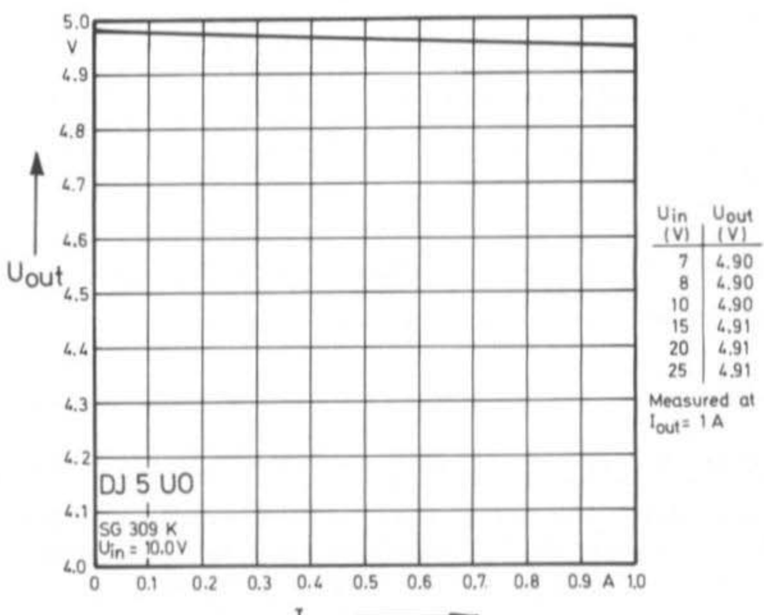


Fig. 3: Measured fluctuations of the output voltage

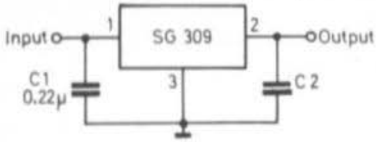


Fig. 4: Application as 5 V stabilizer

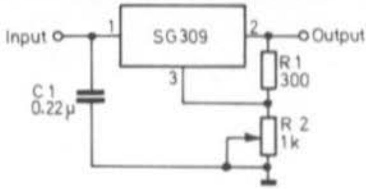


Fig. 5: Circuit for a variable output voltage

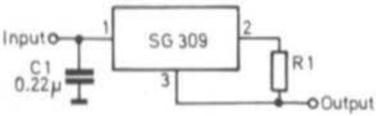


Fig. 6: Circuit as current stabilizer

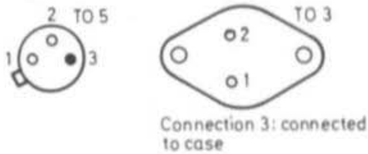


Fig. 7: Connections

2. APPLICATION

Figures 4, 5 and 6 show how easily the integrated circuit SG 309 (109, 209) can be used as a 5 V voltage stabilizer (Fig. 4), as a stabilizer with variable output voltage (Fig. 5) and as stabilized current source (Fig. 6). The input capacitor C 1 is only necessary when a long cable exists between the filter capacitor of the power supply and the integrated circuits. The output capacitor C 2 is only used to improve the dynamic impedance. Resistor R 1 determines the output current.

3. REFERENCES

(1) H. Kahlert: A Universal Power Supply Using an Integrated DC-Voltage Stabilizer
VHF COMMUNICATIONS 4 (1972), Edition 2, Pages 121-126

RSGB HANDBOOKS NO LONGER AVAILABLE

Unfortunately, the RSGB have now informed us that both the Radio Communications Handbook and the VHF-UHF Manual are out-of-print and are no longer available.

HIGH-IMPEDANCE PREAMPLIFIER FOR FREQUENCY COUNTERS FROM DC TO 60 MHz

by H. U. Schmidt, DJ 6 TA

TTL integrated circuits are now inexpensive and allow radio amateurs to construct digital frequency counters inexpensively. Several articles have been written and several kits have been offered. These inexpensive TTL ICs can be used upto approximately 60 MHz.

Schottky TTL ICs are available for use upto approximately 100 MHz. However, the construction is complicated since the inputs and outputs of the TTL ICs produce virtually a complete reflection of the signal which can cause large overshoot effects and multiple pulses even when short connections are used. Furthermore, it is often difficult to obtain the high input levels required at 100 MHz. This means that ECL-circuits, whose inputs and outputs can be matched to 50 Ω, are preferable at frequencies over approximately 60 MHz.

An upper frequency limit of 60 MHz is also valid for a high-impedance preamplifier, since a conventional input of 1 MΩ//20 pF only exhibits an input impedance of approximately 150 Ω (capacitive shunt) at 60 MHz. For this reason, preamplifiers for higher frequencies should be designed for lower impedances (50 - 75 Ω).

A digital frequency counter should preferably also be able to make period and pulse measurements. Such a frequency counter for DC to 60 MHz should have a high-impedance input and be provided with a preamplifier having an input sensitivity of 20 mV over the whole frequency range. Unfortunately, the only suitable preamplifiers that have been described are equipped with expensive ECL III circuits.

A large number of preamplifiers use separate input circuits for AF and RF-voltages, usually of medium or low impedance and the lower frequency limit is often too high for period and pulse measurements. For this reason, a suitable preamplifier has been designed to satisfy these demands, which is also equipped with inexpensive components.

1. CIRCUIT DESCRIPTION

As can be seen in Figure 1, the preamplifier is equipped with a high-impedance source-follower input circuit comprising the field effect transistor BF 245 C (T 1). The two antiphase silicon diodes D 1 and D 2 together with resistor R 2 protect the FET against overvoltages of upto 200 V. The majority of the amplification is provided by the integrated circuit I 1. The Motorola type MC 1035 used for I 1 belongs to the ECL II series. It is an emitter-coupled type for frequencies of upto 120 MHz. The MC 1035 P (14 pin dual in-line case) is not a digital logic circuit, but comprises three differential amplifiers with balanced inputs and outputs (one of the amplifiers possesses a single-ended output). This circuit can be used as amplifier, comparator, Schmitt-trigger or level converter. Each amplifier possesses a gain of six times for push-pull. The -3 dB-bandwidth is 50 MHz, and a usable gain is provided upto approx. 100 MHz. The common-mode suppression amounts to 80 dB and the output voltage coincides with the ECL-level of approximately 850 mV. The IC is also

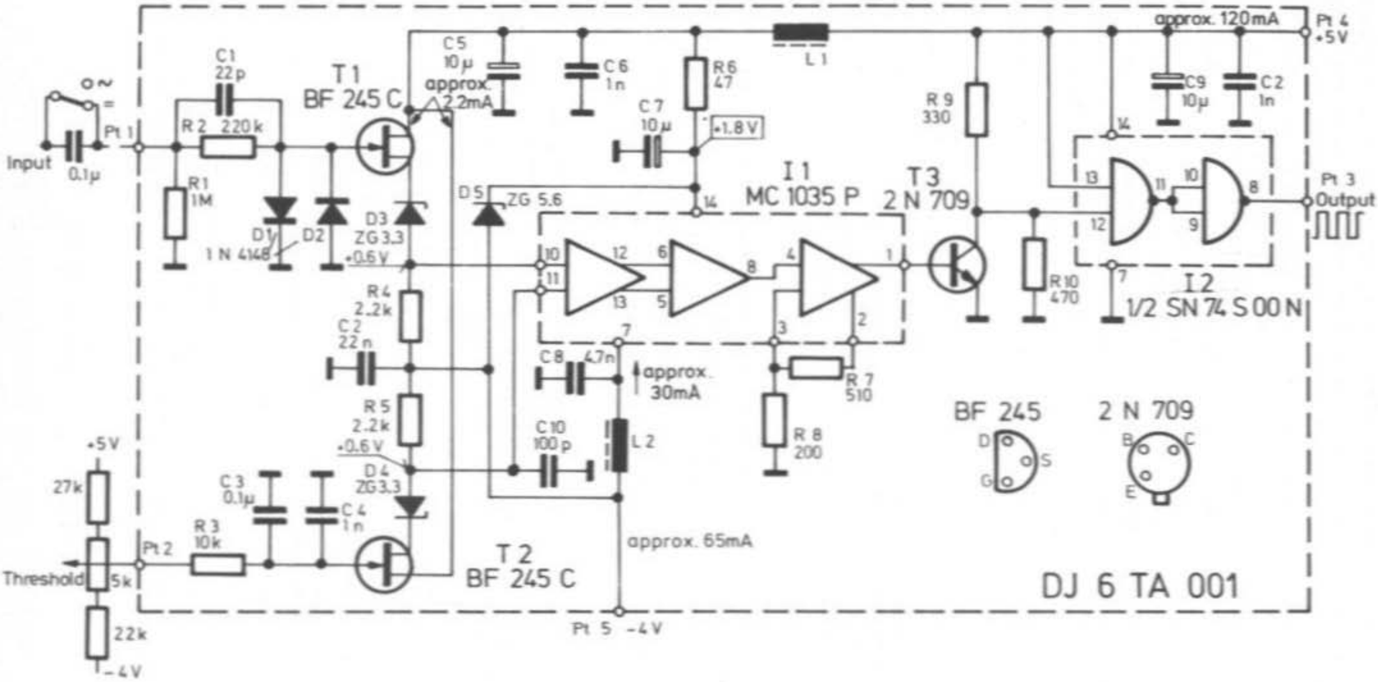


Fig. 1: High-impedance preamplifier for frequency counters from DC to 60 MHz

equipped with a constant voltage source which feeds the base transistor of the differential amplifiers (constant current sources) and also provides a voltage-U_{bb} (approx. - 1.2 V ref. to U_{cc}). One input of a differential amplifier can be connected here if single-ended operation is required.

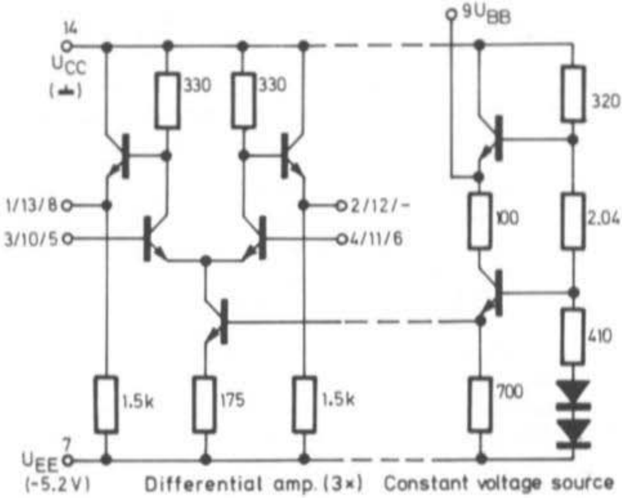


Fig. 2: Internal circuit of the Motorola MC 1035

In the described application, two differential amplifiers are used in series and the third operates as Schmitt-trigger via the feedback voltage divider comprising R 7/R 8. The unused differential input of I 1 is connected via a second source-follower (T 2) to an adjustable voltage of approximately 0 V for temperature compensation. This voltage allows the switching threshold of the Schmitt-trigger to be adjusted for various input voltages which is very favourable for measuring the frequency of voltages with a harmonic content. If possible, transistors T 1 and T 2 should be selected to have the same electrical data, e.g. the voltage of both sources should be identical with the gate grounded, the absolute value is not important.

Normally, the MC 1035 operates with an operating voltage of - 5.2 V. The voltages U_{cc} and U_{ee} are shifted with respect to ground according to application notes (5) so that output 1 can directly drive a common-emitter NPN transistor (T 3). This serves as the level converter to TTL-level. Since connection 14 must be fed with + 1.8 V, connection 7 ($-U_{ee}$) is connected to - 4 V and the operating voltage for the IC is generated with the aid of the zener diode D 5. In order to keep transistor T 3 within the TTL operating range even under unfavourable conditions, it is important that the voltage of + 1.8 V to ground at pin 14 is maintained exactly. This can be achieved by slight variation of the negative voltage at Pt 5.

The slope of the squarewave output voltages at the upper frequency limit are steepened in the subsequent pulse shaper comprising the two Schottky TTL-gates (1/2 SN 74 S 00 N). The second, and other two that remain unused can be used for the logic circuit of the actual counter if required.

The limiting, wideband amplifier with subsequent Schmitt-trigger provides a squarewave output voltage even at the lowest frequencies. The high gain ensures that the input sensitivity of 20 mV is maintained upto 60 MHz.

2. CONSTRUCTION

The complete preamplifier including pulse shaper is accommodated on a double-coated PC-board with the dimensions 73 mm by 50 mm. This board has been designated DJ 6 TA 001 and is shown in Figure 3 complete with component locations. All connections are made on the conductor lane side of the board. The component side is in the form of a broken ground surface.

The preamplifier should be installed into the frequency counter so that it is as near as possible to the input connector. If this is not possible, the interconnection should be made at low capacitance using a (90 Ω) screened cable. It is often necessary for the whole preamplifier to be screened, since steep pulses could be introduced into the high impedance input from the TTL-circuits.

An isolating capacitor of 0.1 μF should be provided for measurements of superimposed DC-voltages. This capacitor can be bridged when galvanic coupling is required. Such a capacitor will, however, reduce the lower frequency limit to 10 Hz. Since input voltages in excess of $\pm$ 0.7 V will cause diodes D 1 and D 2 to conduct in forward direction and will alter the input impedance, it is also advisable for a 10 : 1 or 100 : 1 voltage divider or probe to be used. This allows the threshold potentiometer to be used even with higher input voltages.

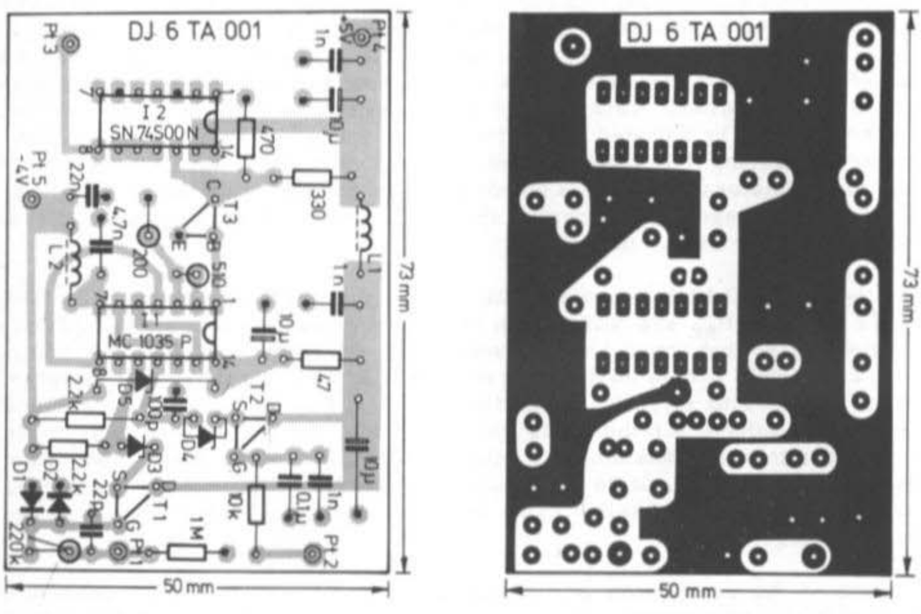


Fig. 3 : Component locations on PC-board DJ 6 TA 001

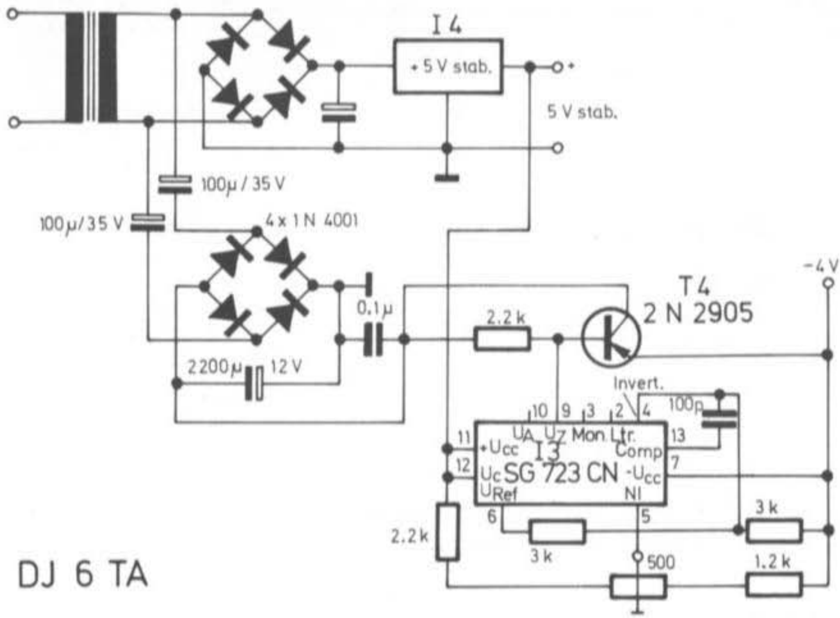


Fig. 4: Circuit suitable for the negative power supply using just one transformer winding (cannot be short-circuited)

3. COMPONENTS

I 1: MC 1035 P (Motorola)
I 2: SN 74 S 00 N
I 3: µA 723 C (Fairchild), SG 723 (Silicon General),
LM 723 CD (National Semiconductor), or TBA 281 (Siemens/Philips)
I 4: SG 309 or LM 309

T 1, T 2: BF 245 C (TI) or similar FET
T 3 : 2 N 709 (very fast switching transistor)
T 4 : 2 N 2905 A (silicon PNP in TO 5 case)

D 1, D 2: 1 N 4148 or similar diode (1 N 914)
D 3, D 4: BZX 83/C3V3 (Siemens), ZG 3.3 or similar 3.3 V zener diode.
D 5 : BZX 83/C5V6 (Siemens), ZG 5.6 or similar 5.6 V zener diode.

Bridge rectifier: 4 diodes 1 N 4001 or similar

L 1, L 2: Wideband ferrite chokes

The capacitor values are uncritical with the exception of C 1 (22 pF ceramic disc capacitor), and C 10 (100 pF ceramic disc capacitor).

A spacing of 10 mm is available for the resistors.

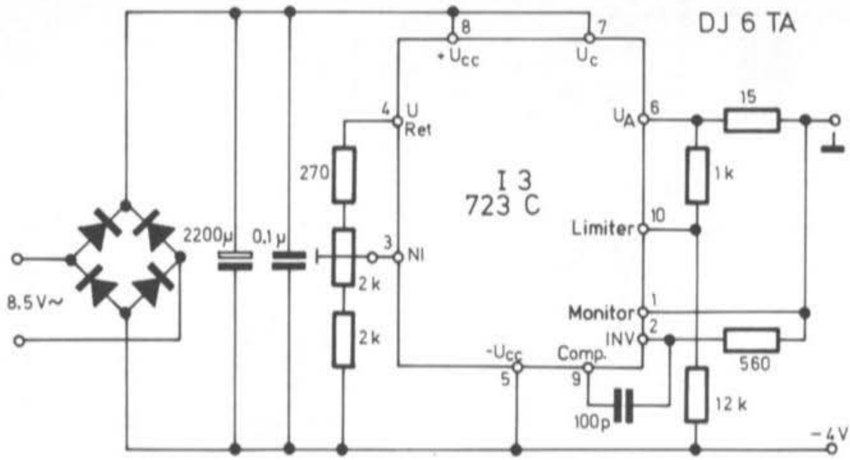


Fig. 5 : Circuit for a power supply if a separate transformer winding is available (may be short-circuited)

4. POWER SUPPLY

A suitable power supply with -4 V is given in Figure 4. An integrated stabilizer 723 CN is used and the circuit assumes that only one transformer winding is available for both voltages. If, however, the transformer provides two secondary windings of approximately 8.5 V each, the circuit given in Figure 5 can be used. Of course, a simpler circuit with two PNP transistors and a zener diode would also be possible; however, the voltage should be well-stabilized and adjustable within certain limits in order to ensure a voltage of + 1.8 V at pin 14 of I 1 (6).

5. REFERENCES

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VHF COMMUNICATIONS 3 (1971), Edition 3, Pages 159-171

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VHF COMMUNICATIONS 5 (1973), Edition 2, Pages 95-103

(3) W. R. Kritter: A Wideband Preamplifier for Frequency Counters upto 60 MHz
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(4) W. R. Kritter: A Dual-Input Preamplifier with 2 : 1 Prescaler for Frequency Counters from 1 Hz to Minimum 100 MHz
VHF COMMUNICATIONS 5 (1973), Edition 2, Pages 91-94

(5) Motorola Data Sheet ECL II

(6) Data sheet of SG 723 C/CN (Silicon General).

NEW WEST GERMAN REPEATER ALLOCATIONS

by T. Bittan, G 3 JVQ / DJ Ø BQ

The West German repeater network is in the process of changing over to the 600 kHz spacing agreed for IARU Region 1. Since it is known that a large number of amateurs use such repeaters as beacons, and since reciprocal licencing allows operation in other countries it was thought that a list of the planned allocations would be of interest. Approximately half of the repeaters have changed over to the new frequencies and the rest will be changing over before mid 1975.

1. VHF REPEATER CHANNELS FOR IARU REGION 1

R Ø	145.000 / 145.600 MHz	R 5	145.125 / 145.725 MHz
R 1	145.025 / 145.625 MHz	R 6	145.150 / 145.750 MHz
R 2	145.050 / 145.650 MHz	R 7	145.175 / 145.775 MHz
R 3	145.075 / 145.675 MHz	R 8	145.200 / 145.800 MHz
R 4	145.100 / 145.700 MHz	R 9	145.225 / 145.825 MHz

1.2. VHF REPEATER ALLOCATIONS

Location		Callsign	Location		Callsign
Channel I 0					
Altenwalde	EN 14f	DB Ø XA	Ludwigsburg	EJ 76f	DB Ø YY
Berlin	GM 47a	DB Ø SP	Ochsenkopf	FK 80f	DB Ø ZB
Feldberg/Ts	EK 63h	DB Ø UF			
Channel I 1					
Aschberg	EO 49g	DB Ø ZA	Friedrichshafen	EH 17c	DB Ø WV
Bamberg	FJ 05a	DB Ø UB	Heidelberg	EJ 44e	DB Ø ZH
Bremen	EN 75g	DB Ø WU	Merzig	DJ 33c	DB Ø XS
Detmold	EL 05g	DB Ø WT	Northeim	FL 21g	DB Ø YN
Duisburg	DL 44c	DB Ø WW	Winterberg	GI 62j	DB Ø WB
Channel I 2					
Feldberg/Ts	EK 63h				
Channel I 3					
Baltic Coast	FO 74b	DB Ø XB	Karlsruhe	EJ 72j	DB Ø UK
Black Forest	EH 21b	DB Ø YH	Nordhelle	DL 69d	DB Ø VR
Cham	GJ 74c	DB Ø YC	Oldenburg	EN 26f	DB Ø UD
Goslar	FL 03f	DB Ø WS			
Channel I 4					
Coburg	FK 55c	DB Ø UC	Leer	DN 68a	DB Ø WO
Deggendorf	GI 15j	DB Ø XD	Luechow	FN 65j	DB Ø ZL
Kalmit	EJ 51j	DB Ø XK	Muenster	DL 09h	DB Ø WM
Knuell	EK 08f	DB Ø XU	Ochsenwang	EI 38j	DB Ø WN
Koblenz	DK 49j	DB Ø ZK			
Channel I 5					
Zugspitze	FH 46g	DB Ø ZU			
Bocksberg	FL 12b	DB Ø XY			

Locations		Callsign	Locations		Callsign
Channel I 6					
Berlin	GM 47a	DB Ø WF	Kassel	EL 57e	DB Ø XE
Bergheim	DK 04a	DB Ø XO	Konstanz	EH 26d	DB Ø WK
Frankfurt/M.	EK 64e	DB Ø VF	Muenchen	FI 78a	DB Ø ZM
Hamburg	EN 40d	DB Ø XH	Nuernberg	FJ 47a	DB Ø ZN
Hannover	EM 49d	DB Ø WH	Osnabrueck	EM 61j	DB Ø ZO
Kaiserstuhl	DI 79j	DB Ø ZF	Stuttgart	EI 17d	DB Ø WR
Channel I 7					
Bad Koenig	EJ 15d	DB Ø VB	Greding	FJ 77c	DB Ø XG
Bentheim	DM 56c	DB Ø UG	Hagen	DL 58a	DB Ø WH
Bredstedt	EO 25d	DB Ø XN	Koeln	DK 05j	DB Ø VK
Elm	FM 65.	DB Ø XC	Lahr	DI 60a	DB Ø WL
Fulda	EK 39j	DB Ø UE	Trier	DJ 24h	DB Ø UT
Goeppingen	EI 30g	DB Ø WG	Wuerzburg	EJ 20e	DB Ø WZ
Channel I 8					
Aachen	DK 11j	DB Ø WA	Dortmund	DL 48b	DB Ø ZR
Augsburg	FI 55b	DB Ø UA	Grab	EJ 78c	DB Ø ZZ
Bad Hersfeld	EK 19a	DB Ø YB	Kaiserslautern	DJ 47e	DB Ø YK
Berlin		DB Ø YL	Loerrach	DH 30a	DB Ø XR
Bremerhaven	EN 33c	DB Ø WC	Triberg	EI 72a	DB Ø WX
Darmstadt	EJ 24h	DB Ø VD	Weiden	GJ 22c	DB Ø ZW
Deister	EM 58j	DB Ø WD			
Channel I 9					
Essen	DL 45d	DB Ø WE	Pforzheim	EI 04d	DB Ø UP
Freising	FI 39g	DB Ø XF	Pirmasens	DJ 69g	DB Ø VP
Hoher Meissner	EL 70h	DB Ø XM	Siegen	EK 01h	DB Ø YS
Nuernberg	FJ 46c	DB Ø UN			

2. UHF REPEATER CHANNELS

R 70	431.050 / 438.650 MHz	R 80	431.300 / 438.900 MHz
R 72	431.100 / 438.700 MHz	R 82	431.350 / 438.950 MHz
R 74	431.150 / 438.750 MHz	R 84	431.400 / 439.000 MHz
R 76	431.200 / 438.800 MHz	R 86	431.450 / 439.050 MHz
R 78	431.250 / 438.850 MHz		

2.1. UHF REPEATER ALLOCATIONS

Channel R 70					
Nuernberg	FJ 46c	DB Ø VN	Wetzlar	EK 23e	DB Ø UJ
Oldenburg	EN 26f	-			
Channel R 72					
Duisburg	DL 44c	DB Ø UD	Merzig	DJ 33e	-
Channel R 74					
Dortmund	DL 48b	DB Ø ZV	Feldberg/Ts	EK 63h	DB Ø VE
Channel R 76					
Osnabrueck	EM 61j	DB Ø UM	Darmstadt	EJ 24h	DB Ø UU
Elm	FM 65.	DB Ø XX	Ochsenkopf	FK 80f	

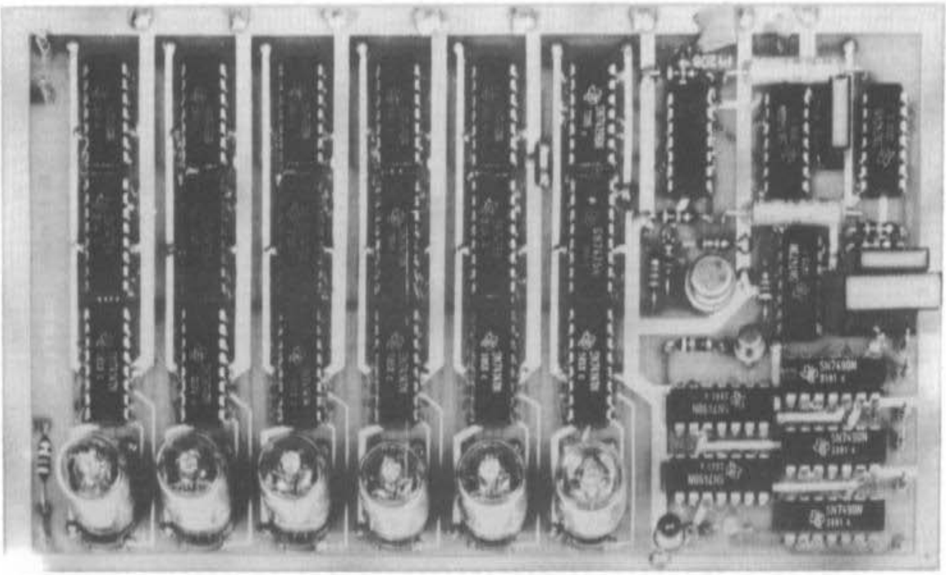
Channel R 78					
Erbeskopf	DJ 16e	DB ∅ VV	Zugspitze	FH 46g	DB ∅ ZS
Goch-Kleve	DL 11c	DB ∅ VG			
Channel R 80					
Bamberg	FJ 05a	-	Haltern	DL 16e	DB ∅ UR
Feldberg	EH 11h	DB ∅ VS	Luebeck	FN 04h	DB ∅ XL
Goslar	FL 03f	-	Mayen	DK 47f	-
Channel R 82					
Marburg	EK 14.	DB ∅ UI	Damme	EM 32g	DB ∅ US
Channel R 84					
Hannover	EM 69a	DB ∅ VH	Nortorf	EO 70g	DB ∅ UL
Knuell	EK 08f	DB ∅ UQ	Stuttgart	EI 17d	-
Channel R 86					
Berlin	EM 47.	-	Pirmasens	DJ 69d	-
Lingen	DM 38e	-			

NEW AUSTRIAN REPEATER ALLOCATIONS

Channel	Location	Callsign	Channel	Location	Callsign
I ∅	GH 39e	OE 5 XKL	I 6	FH 57h	OE 7 XTI
I 2	GI 67a	OE 5 XUL	I 8	HI 78e	OE 3 XPA
I 4	HI 53a	OE 5 XLL	I 8	HG 23h	OE 8 XMK
I 6	GH 23h	OE 5 XGL	I 9	GH 32.	OE 7 XKI

Transponders:

Input MHz	Output MHz	Callsign	Location	
432.000	145.600	OE 2 XSG	Salzburg	GH 16j
144.375	145.575	OE 7 XZI	Zugspitze	FH 46g



COMPLETE KITS for a 250 MHz SIX-DIGIT FREQUENCY COUNTER

Total . . . . DM 775. -- excluding crystal oven

Total . . . . DM 879. -- including crystal oven and special crystal

A DOMESTIC TV-RECEIVER AS VIDEO MONITOR

by K. Wilk, DC 6 YF

A domestic TV-receiver can be used as an inexpensive monitor for video signals. With older, tubed models, some difficulties can be encountered since the chassis is often directly connected to one side of the power line. However, a large number of transistorized, monochrom TV-receivers possess a built-in transformer which isolates the set galvanically from the power line. These latter receivers are suitable for use in an amateur radio station. Since television receivers are not expensive, it is even worth while to purchase such a unit exclusively for experimental use.

This article described how an additional unit can be built into a conventional domestic TV-receiver. A receiver modified in this manner does not only possess an input allowing it to be used as monitor, but also an output from which the video signal from the programme being received can be taken. This signal corresponds to the television standard and can be used for various alignment and test purposes. When the modified receiver is to be used for conventional, domestic TV-reception, it is only necessary for output and input to be interconnected (Fig. 1).

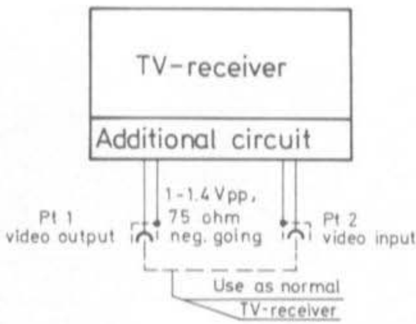


Fig. 1: A domestic TV-receiver modified for use as a video monitor

1. CHARACTERISTICS OF VIDEO OUTPUT AND INPUT

In order to ensure that the receiver can be used universally, it is necessary for the levels etc. of the signals that are available externally to be of the required values. The values required and their characteristics are as follows:

Level: 1 - 1.4 V (peak-to-peak)
Input and output impedance: 75 Ω
Synchronizing pulses: Negative going

DC 6 YF 007

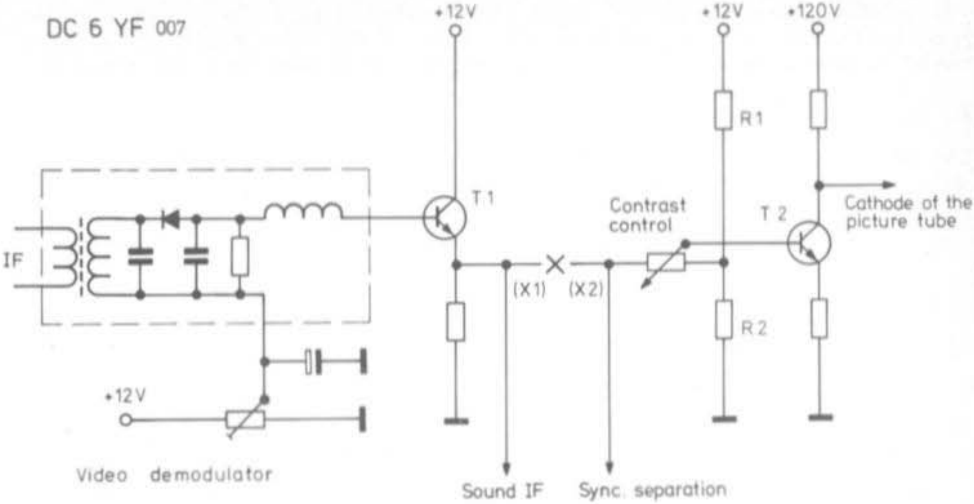


Fig. 2: Circuit of a conventional video amplifier

2. SELECTION OF THE INPUT AND OUTPUT CONNECTIONS

A video amplifier circuit of a TV-receiver is to be described briefly in order to show which connection points are most suitable for connecting the additional module. Such amplifier circuits are to be found in most modern TV-receivers (Fig. 2):

The output signal of the video amplifier usually drives the cathode of the picture tube. The blanking and synchronizing pulses are positive-going and will control the electron beam into the blacker-than-black level when the grid potential remains constant. The last transistor stage (T 2) which drives the picture tube operates as a phase-inverting amplifier, which amplifies the video signal available at the demodulator from 2 - 3 V (peak-to-peak) to the required level of 70 - 100 V (peak-to-peak) required for driving the picture tube.

The gain control for the video amplifier (contrast) is usually connected just before the output stage of the video amplifier. In order to ensure that the output stage is decoupled sufficiently from the video demodulator, an emitter follower (T 1) is provided as impedance converter subsequent to the demodulator and picture tube is DC-coupled. This ensures that the DC-voltage component of the video signal, which determines the brightness level of the picture, is also passed.

The voltage-divider ratio of R 1 and R 2 and the bias voltage adjustment at the video demodulator have been selected so that the DC-voltage component at the base of the output transistor remains practically constant in the various positions of the contrast control. The contrast control only varies the AC-voltage component, and not the brightness of the picture which is dependent on the DC-voltage component.

The position marked with "X" in the circuit diagram given in Figure 2 is especially suitable for use as input and output of the video signal. The video signal at this point is very similar to the characteristics that are required.

3. THE ADDITIONAL CIRCUIT

The additional circuit shown in Figure 3 consists of a single-stage output amplifier and a four-stage input amplifier.

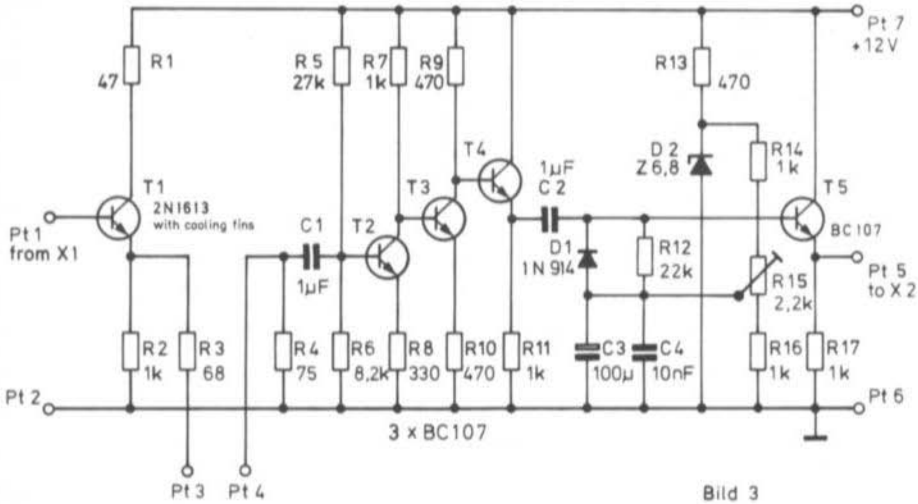


Fig. 3: Circuit of the additional module

3.1. OUTPUT AMPLIFIER

The output amplifier converts the impedance and level characteristics at the emitter of T 1 (Fig. 2) to that required by the 75 Ω system. This means that its output impedance also corresponds to 75 Ω. If an emitter-follower is to be used for isolation and to achieve a higher output power, a non-load voltage of 2 - 2.8 V (peak-to-peak) should be provided in order to obtain the required output level of 1 - 1.4 V (peak-to-peak) which is actually provided by conventional receivers. Since the synchronizing pulses are of the correct polarity (negative-going) a simple circuit comprising only one transistor can be used.

3.2. INPUT AMPLIFIER

The input amplifier increases the level of the video signal to the required level of 2 - 3 V. A gain factor of three times was selected so that a sufficient video level is available for the output transistor even in the most unfavourable conditions.

The external video signal to be viewed on the monitor will possess a DC-voltage component which can differ greatly according to the source. This signal is capacitively-coupled to the amplifier circuit comprising transistors T 2 and T 3 via the input resistor of 75 Ω. After passing the impedance converter comprising transistor T 4, a further stage is provided for adjustment of the black level which will have been lost due to the capacitive coupling. In addition to this, an adjustable DC-voltage is provided that corresponds to the voltage at the output connector of the TV-receiver. This voltage can be varied in the range of 2 V to 5 V with the aid of potentiometer R 15. This is followed by a further impedance converter comprising T 5. In this manner, the same characteristics are provided for the video output stage in the TV-receiver as were provided before the additional circuit was provided.

4. CONSTRUCTION

The additional circuit is accommodated on a single-coated PC-board whose dimensions are 95 mm x 77.5 mm. This PC-board, which has been designated DC 6 YF 007, is shown in Figure 4 together with the component locations.

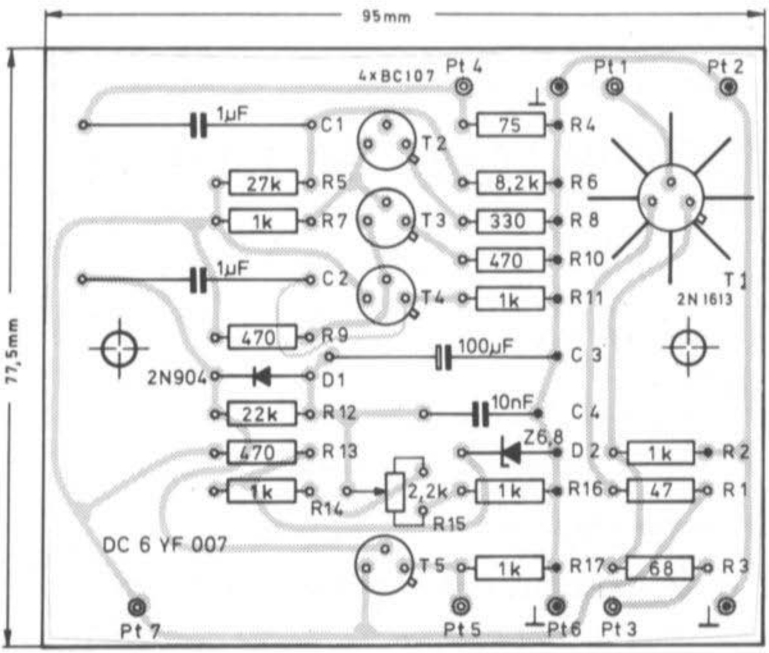


Fig. 4: PC-board DC 6 YF 007 and component locations

5. INSTALLATION AND POWER SUPPLY

Although the position and arrangement of the interconnection leads is relatively uncritical due to the low impedances, there should be no difficulties in mounting the small additional circuit in the vicinity of the video amplifier. The circuit requires approximately 60 mA at an operating voltage of 12 - 15 V. Higher voltages, e.g. 24 V which is used in some units can be reduced to the required value using a suitable zener diode. A 1000 μF electrolytic capacitor should be connected in parallel with the operating voltage connections of the additional circuit. The input and output connectors for the video signal are mounted either on the side or rear panels of the TV-receiver. The author recommends that BNC-connectors be used.

6. ADJUSTMENT

After checking the wiring, it is only necessary for the DC-voltage at Pt 5 without input signal to be adjusted with trimmer resistor R 15 to the value that is measured after the emitter follower stage of the video demodulator. No further adjustments are required on the TV receiver.

7. NOTES

When a video signal is to be taken from a TV-transmission, it is necessary for the input amplifier to be fed with this signal so that the RF and IF circuits of the TV-receiver provide the required control voltage. In this case, resistor R 4 must be switched off; the input amplifier will then have a sufficiently high input impedance.

8. REFERENCES

(1) T. Bittan: Amateur Television
VHF COMMUNICATIONS 4 (1972), Edition 3, Pages 184-190

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 3/74 of VHF COMMUNICATIONS

DK 1 OF	**400 CHANNEL SYNTHESIZER**		**Ed. 3/74**
DK 1 OF 012			
PC-board	DK 1 OF 012	(with printed plan)	DM 10.--
Semiconductors	DK 1 OF 012	(4 ICs, 4 transistors, 1 diode)	DM 52.--
Minikit	DK 1 OF 012	(1 trimmer cap., 2 styroflex cap.)	DM 3.--
Crystal	1.440.000 kHz HC-6/U		DM 22.50
Crystal holder	(horiz. mounting)		DM 4.--
Kit	DK 1 OF 012	with above parts	DM 90.--
DK 1 OF 013			
PC-board	DK 1 OF 013	(double-coated, through-contacts)	DM 16.--
Semiconductors	DK 1 OF 013	(6 ICs, 1 transistor)	DM 68.--
Kit	DK 1 OF 013	with above parts	DM 83.--
DK 1 OF 015			
PC-board	DK 1 OF 015	(with printed plan)	DM 8.50
Semiconductors	DK 1 OF 015	(5 transistors, 1 diode)	DM 32.--
Minikit	DK 1 OF 015	(1 coilset, 1 choke, stranded wire)	DM 10.--
Kit	DK 1 OF 015	with above parts	DM 50.--
Kits	DK 1 OF 012, DK 1 OF 013, DK 1 OF 015		DM 220.--
DK 1 PN 006	**AM-IF-STRIP FOR 9 MHz**		**Ed. 3/74**
PC-board	DK 1 PN 006	(with printed plan)	DM 10.--
Semiconductors	DK 1 PN 006	(1 IC, 2 diodes)	DM 10.--
Minikit	DK 1 PN 006	(2 coilsets, 6 ceramic caps, 4 tantalium caps, 1 TEKO-box 2 A, 1 13-pin connector)	DM 26.--
Kit	DK 1 PN 006	with above parts	DM 45.--
DK 1 PN 008	**SYSTEM BOARD FOR INTEGRATED RECEIVER**		**Ed. 3/74**
PC-board	DK 1 PN 008	(double-coated, no through contacts)	DM 36.50
Minikit	DK 1 PN 008	(5 ferrite chockes, 2 Alu electrolytics, 1 tantalium electrolytic, 4 ceramic caps, 1 resistor, 6 13-pole contact strips)	DM 14.70
Kit	DK 1 PN 008	with above parts	DM 76.--
DJ 6 TA 001	**HIGH-IMPEDANCE COUNTER PREAMPLIFIER**		**Ed. 3/74**
PC-board	DJ 6 TA 001	(double coated, no through contacts)	DM 14.--
Semiconductors	DJ 6 TA 001	(2 ICs, 3 transistors, 5 diodes)	DM 44.--
Minikit 1	DJ 6 TA 001	(2 ferrite chokes, 3 tantalium electrolytics, 9 ceramic caps.)	DM 12.--
Minikit 2	DJ 6 TA 001	(10 resistors)	DM 2.70
Complete Kit	DJ 6 TA 001	with above parts	DM 70.--
DC 6 YF 007	**VIDEO MONITOR FROM TV-RECEIVER**		**Ed. 3/74**
PC-board	DC 6 YF 007	(with printed plan)	DM 11.50

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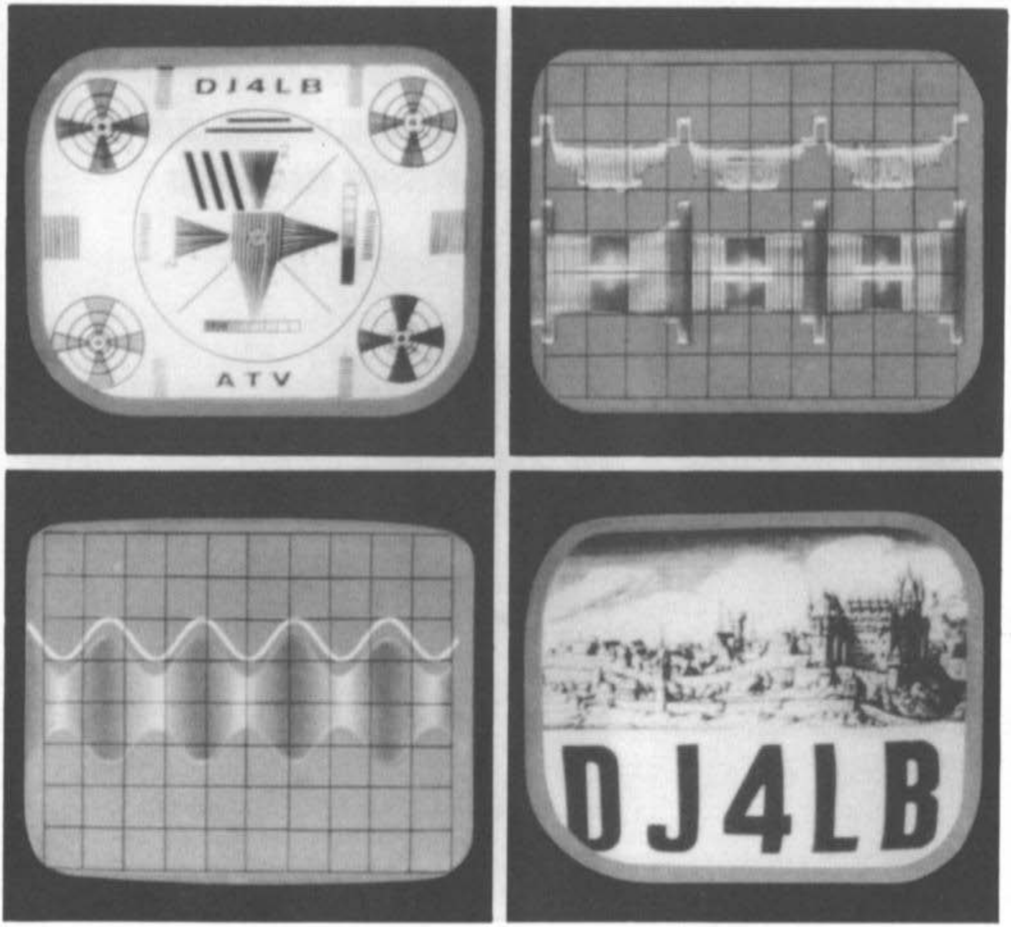

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