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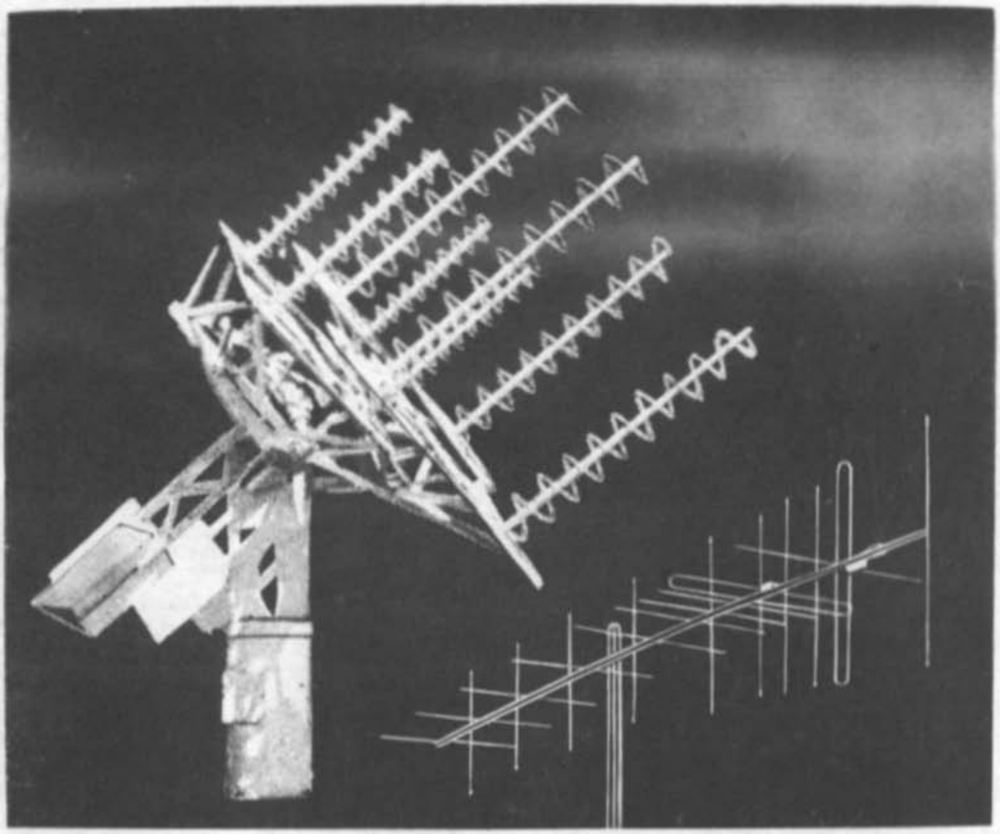

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In spite of increasing costs, we have decided to keep the subscription price down at 14.00 DM as it has been for the last three years. Please assist us in keeping our prices as low as possible by introducing VHF COMMUNICATIONS to other VHF/UHF amateurs. It is not our principle to fill the magazine with advertisements in order to make ends meet.

Please do not forget to renew your subscription to VHF COMMUNICATIONS for 1974. There are lots of outstanding designs and articles coming up such as a linear-transverter for 144 MHz/432 MHz, transistorized linear amplifier for 432 MHz, linear-transverter for 144 MHz/1296 MHz, highly selective converter for repeater receivers, foxhunt receiver for 2 m with preset channels, wideband reflectometer 0-1300 MHz for 50 Ohms, etc.

The publishers and staff would like to wish all readers a very Merry Christmas and a happy 1974

DJ 3 QC, DL 3 WR, G 3 JVQ / DJ Ø BQ

9 MHz FM EXCITER MATCHING THE 80-CHANNEL SYNTHESIZER

by J. Kestler, DK 1 OF

Modules DK 1 OF 001 to 006 of an 80-channel synthesizer for 135 to 137 MHz were described in (1). This article is to describe modules DK 1 OF 007 to 010 which have the following functions: 9 MHz FM exciter, AF-preamplifier, 1750 Hz calling tone generator and transmit mixer. Figure 1 gives a block diagram of a FM transceiver comprising these modules. Previously described modules described by other authors can be used for the other modules such as power amplifier, receive converter and IF chain. Details of suitable modules were given in (1).

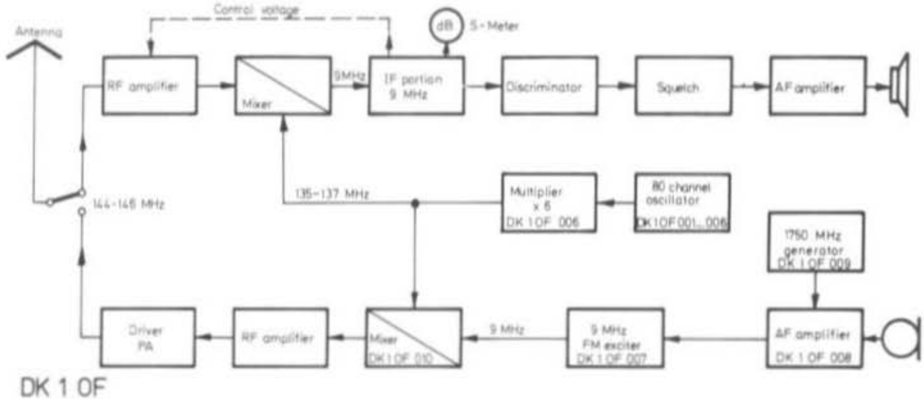

Fig. 1: Block diagram of a FM transceiver with 80-channel synthesizer

1. 9 MHz FM EXCITER DK 1 OF 007

There are two possibilities of generating a frequency-modulated signal:

a) Pulling the frequency of a LC or crystal-controlled oscillator with the aid of a voltage-dependent capacitance (varactor diode) in the frequency-determining circuit.

b) Phase modulation of a crystal-controlled signal.

The disadvantage of phase modulation is that the attainable frequency deviation is very low which means that a frequency multiplication is always necessary subsequent to the phase modulator. As an example, a phase-modulated 9 MHz carrier with a phase deviation of $\pm\pi$ (ideal modulator) and a modulating frequency of 300 Hz would exhibit the following equivalent frequency deviation:

$$\Delta f = f_M \times \Delta\varphi = 300 \times \pi = 940 \text{ Hz};$$

Of course, a frequency deviation of approximately 5 kHz is required when using a 25 kHz channel spacing and the appropriate IF bandwidth is approximately 15 kHz. This means that this type of modulation is not possible.

The frequency stability of an LC oscillator whose frequency is modulated by a varactor diode is not sufficient for the rugged operating conditions encountered, for instance, during mobile operation or when large temperature fluctuations are to be expected. A VCXO (voltage-controlled crystal oscillator) is not usually able to provide a sufficiently large and linear frequency deviation with normal crystals.

The most favourable compromise between frequency stability and the possibility of frequency modulation is provided by the following circuit which uses a ceramic IF filter as frequency dependent circuit. Relatively cheap 4.5 MHz ceramic filters are available for use in the sound intermediate frequency of receivers operating according to the US TV standard. The generated frequency is then doubled to 9 MHz.

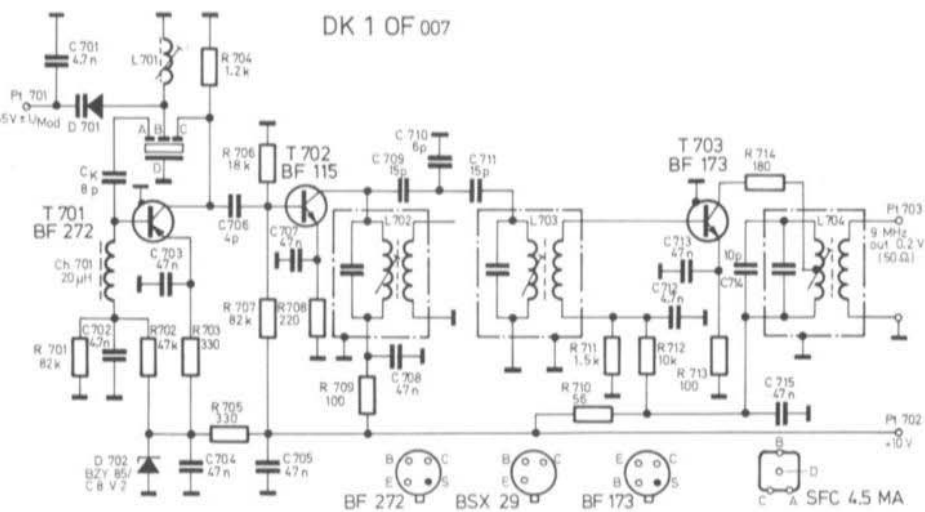

Fig. 2: Circuit diagram of the 9 MHz FM exciter

The circuit diagram of the 9 MHz FM exciter is given in Figure 2. The ceramic filter is in the feedback link of the oscillator transistor T 701. The varactor diode whose mean capacitance is compensated for by inductance L 701, is able to provide a maximum frequency deviation in excess of ± 5 kHz. The operating voltage of the oscillator stage is stabilized with the aid of a 8.2 V zener diode.

The subsequent stage T 702 is loosely coupled via a capacitor of 4 pF to the oscillator circuit and doubles the 4.5 MHz signal to 9 MHz. The following bandpass filter consists of ready-made miniature transformers for 10.7 MHz which are coupled and resonated to 9 MHz with the aid of a T-circuit comprising three capacitors. The output stage is equipped with a low-reactive transistor (T 703) which amplifies the 9 MHz signal to approximately 1 mW. The resonant circuit at the output matches the output stage to 50 Ω.

Figure 3 gives the characteristic curve of the modulator in the FM exciter. The curve is sufficiently linear in the range of ± 10 kHz from the centre frequency.

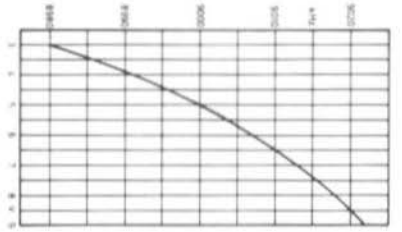

Fig. 3 : Modulation characteristics of the FM exciter

1.1. SPECIAL COMPONENTS FOR DK 1 OF 007

T 701: BF 272 or BSX 29 (Fairchild) silicon PNP UHF transistor
T 702: BF 115, BSX 26 (2 N 708) or similar VHF transistor
T 703: BF 173 or BF 224

D 701: MV 1650 (120 pF at 4 V, Motorola) or two BA 124/65 or BA 150/65 (AEG-Tfk)
D 702: BZY 85/C8V2 or other 8.2 V zener diode

L 701: 40 turns of stranded wire (20 x 0.04 mm dia.) or similar in special coil set.
L 702 - L 704: Miniature 10.7 MHz IF transformers.

Ceramic filter 4.5 MHz Murata SFC 4.5 MA.

Ch 701: Approx. 20 μH miniature ferrite choke.

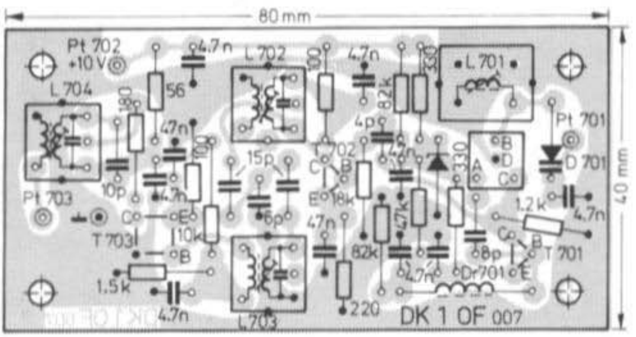

Fig. 4 : Printed circuit board DK 1 OF 007 for the FM exciter

1.2. CONSTRUCTION

This module is accommodated on PC-board DK 1 OF 007 whose dimensions are 80 mm x 40 mm. This PC-board is given in Figure 4 together with the component locations. The board is usually mounted together with module DK 1 OF 008 as shown in Figure 7.

1.3. ALIGNMENT OF THE 9 MHz FM EXCITER

Firstly, the operation of the oscillator is checked by connecting an operating voltage of 10 V to connection Pt 702 and a DC voltage of 5 V to connection Pt 701. This is followed by connecting an oscilloscope to the base of transistor T 702 or by monitoring the signal, and harmonics of the signal with the aid of a receiver. It may be necessary to alter the value of capacitor C_C in order to compensate for the spread of the ceramic filter and transistor T 701. C_C should only be just large enough that the oscillator commences oscillation at all times. The resonant circuits comprising L 702 to L 704 are aligned several times for maximum output. Finally, the alignment core of inductance L 701 should be adjusted so that the output frequency at connection Pt 703 is exactly 9 MHz.

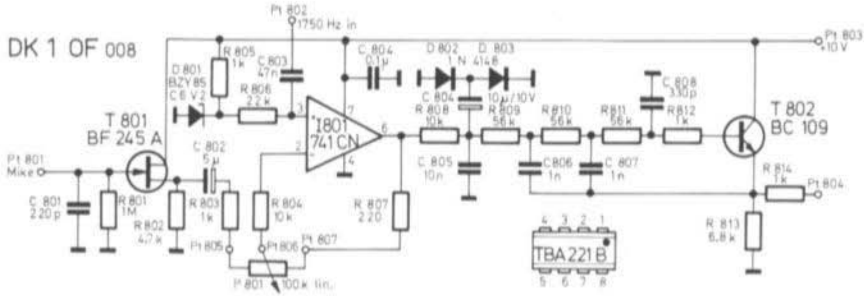

Fig. 5 : Circuit diagram of the AF preamplifier

2. AF-PREAMPLIFIER DK 1 OF 008

The circuit diagram of this module is given in Figure 5. The input stage equipped with the field effect transistor T 801 is used as impedance converter in order to allow connection of high impedance microphones such as crystal types. The whole voltage gain is provided by the integrated amplifier I 801. The gain figure and thus the frequency deviation of the transmitter can be adjusted with the aid of an external potentiometer (P 801). The calling-tone voltage from the 1750 Hz generator DK 1 OF 009 is fed from connection Pt 802 to the non-inverting input of the amplifier.

In order to ensure that the adjacent channels are not interfered with due to incorrect operation of the FM transmitter, a frequency deviation limiter has been provided. When the AF preamplifier is overloaded, silicon diodes D 802 and D 803 will conduct and will clip the AF output signal to a value of ±0.7 V. As can be seen in Figure 3, this modulation voltage corresponds to a frequency deviation of approximately ± 5 kHz. If the exciter is to be used in areas where larger channel spacings and frequency deviation values are used, it is possible for two diodes to be connected in series so that the maximum frequency deviation will be in the order of ± 10 kHz. Any distortion products generated during the clipping process are filtered out in an active lowpass filter, as described by DJ 4 BG in (2), having a cut-off frequency of 3 kHz.

The output DC voltage from the modulator must be stable since it determines the centre frequency of the FM exciter. For this reason, the bias voltage of the non-inverting input of the integrated amplifier I 801 is stabilized with the aid of a 6.2 V zener diode. The temperature coefficient of zener diodes in this voltage range only deviates slightly from zero.

2.1. SPECIAL COMPONENTS FOR DK 1 OF 008

T 801: BF 245 A (Texas Instruments), W 245 A (Siliconix) or similar FET
T 802: BC 109 or similar low-noise silicon NPN AF transistor with B min. 300
I 801: LM 741 CN (National Semiconductor) or TBA 221 B (Siemens)
D 801: BZY 85/C6V2 or similar 6.2 V zener diode
D 802, D 803: 1 N 4148, 1 N 914 or similar silicon diode.

2.2. CONSTRUCTION

A PC-board with the same dimensions as DK 1 OF 007 was also developed for module DK 1 OF 008. The dimensions are therefore 80 mm by 40 mm. This printed circuit board and the component locations are given in Figure 6. Figure 7 shows a photograph of the author's prototype showing how the completed modules DK 1 OF 007 and 008 can be mounted together. The AF preamplifier is located below the exciter.

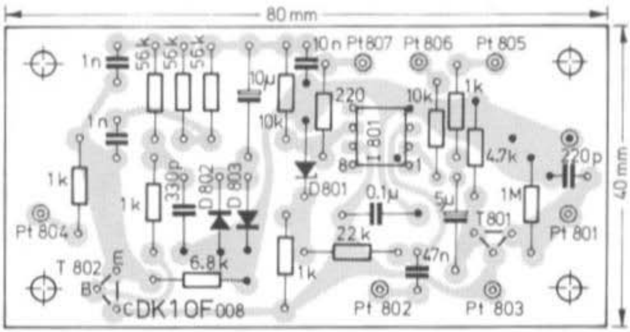

Fig. 6: Printed circuit board DK 1 OF 008 for the AF preamplifier

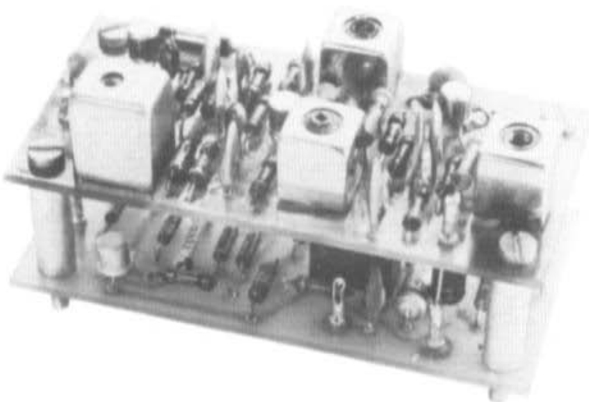

Fig. 7: Author's prototype of PC-board DK 1 OF 007 and 008 (below)

3. 1750 Hz CALLING TONE GENERATOR DK 1 OF 009

The circuit diagram of the calling tone oscillator is given in Figure 8. A Hartley circuit is used with feedback via the source of the oscillator transistor T 901. The resonant circuit is coupled relatively loosely to the field effect transistor and comprises a potted core L 901 and a plastic foil capacitor C 902. An electrolytic capacitor cannot be used in this application.

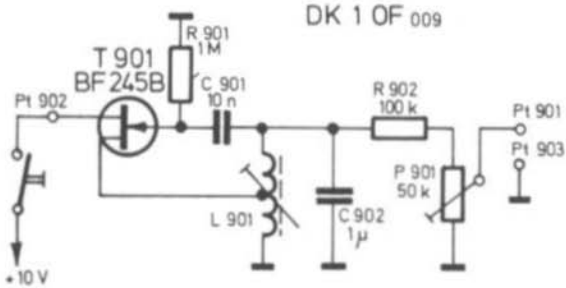

Fig. 8 :
Circuit diagram
of the 1750 Hz
calling tone generator

3.1. COMPONENTS FOR DK 1 OF 009

T 901: BF 245 B or A (Texas Instruments) or W 245 B (Siliconix)

L 901: 227 turns of 0.16 mm dia. (34 AWG) enamelled copper wire (coil tap 50 turns from the ground end) in a potted core pf 14 mm x 8 mm, material N 22, AL value:160.
Or 182 turns of 0.2 mm dia. (32 AWG) enamelled copper wire (coil tap 40 turns) in a potted core of material N 28, AL-value: 250.

C 902: 1 µF plastic foil capacitor, max. 17.5 mm long.

P 901: 50 kΩ linear trimmer potentiometer, horizontal mounting, spacing 17.5/10 mm.

3.2. PC-BOARD

The printed circuit board for the 1750 Hz calling tone generator is 45 mm x 40 mm and has been designated DK 1 OF 009. Figure 9 shows this PC-board and the component locations.

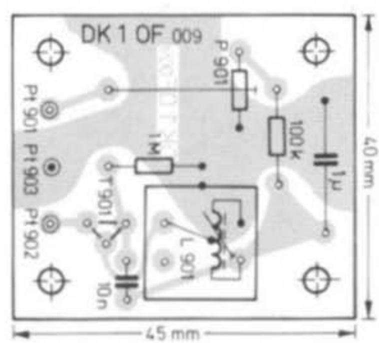

Fig. 9 : PC-board DK 1 OF 009 for the calling tone generator

3.3. ALIGNMENT

The alignment is carried out with the aid of a frequency counter, or an oscilloscope with calibrated deflection, connected to the output (Pt 901). The exact frequency can be aligned with the aid of the core of inductance L 901. Since the alignment range is only in the order of ± 50 Hz, it may be necessary for several capacitors to be tried for C 902, if the tolerances of inductance and capacitor are unfavourable.

4. TRANSMIT MIXER DK 1 OF 010

The 9 MHz FM signal from the exciter is mixed with the output frequency of the synthesizer DK 1 OF 001 to 006 (135 MHz to 137 MHz) in the transmit mixer. The sum of both frequencies results in 80 transmit channels in the

2 m band (144 MHz to 146 MHz). Since the spacing between the local oscillator frequency and the output frequency is relatively small, a balanced mixer circuit has been used in order to obtain the maximum suppression of spurious signals. It is advisable for such a mixer to be equipped with field effect transistors because their square characteristic curves will ensure that no conversion products of higher order are generated.

Figure 10 gives the circuit diagram of this module. The local oscillator signal from the synthesizer is fed to connection Pt 1001 after which it is amplified in transistor T 1001 and passed through a bandpass filter for suppression of any spurious signals. A signal is then passed via a coupling capacitor of 2.2 pF in push-push to both gates of the mixer transistors T 1002 and 1003. A portion of the local oscillator voltage is tapped off at the secondary of the bandpass filter and fed to connection Pt 1002 where it is available for the receive mixer.

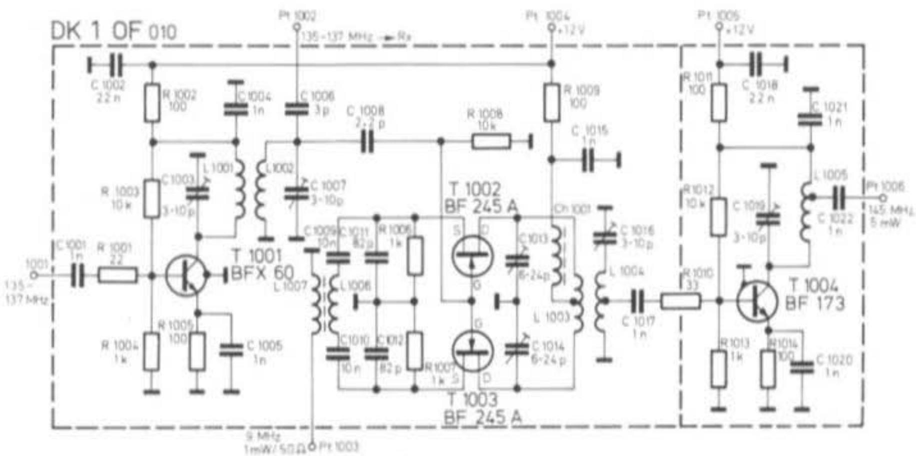

Fig. 10 : Circuit diagram of the transmit mixer

The 9 MHz FM signal is fed to connection Pt 1003 and drives the source connections of the mixer transistors in push-pull. The required phase reversal is obtained in a balancing transformer (L 1007, L 1006) whose secondary has been provided with capacitors in order to form a resonant circuit for 9 MHz. This resonant circuit does not need to be aligned since it is dampened by the input impedance of the transistors and is therefore very wideband. The transformation ratio of L 1006 to L 1007 has been selected so that the input impedance at Pt 1003 is approximately 50 Ω. The two mixer transistors operate in a common gate circuit for the 9 MHz signal. This is the reason why a relatively high drive power of approximately 1 mW is required.

The sum frequency formed in the mixer is filtered out in the push-pull drain circuit, which forms a bandpass filter together with the subsequent circuit comprising L 1004/C 1016. This suppresses the image (difference) frequency of 126 MHz to 128 MHz. The output stage equipped with the low reactive transistor T 1004 provides an output power of approximately 5 mW which can be taken from Pt 1006 and fed to the subsequent power amplifier.

4.1. COMPONENTS FOR DK 1 OF 010

T 1001: BFX 60 (Siemens), BF 173
T 1002, T 1003: BF 244 A (or BF 245 A inserted by bending the connections)
T 1004: BF 173

C 1003, C 1007, C 1016, C 1019: 3-10 pF or 3-12 pF ceramic disc trimmers of 10 mm dia. or plastic foil trimmers
C 1013, C 1014: 6-24 pF ceramic disc trimmers of 10 mm dia. or plastic foil trimmers

L 1001, L 1002: 5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 6.5 mm former, self-supporting, coil length corresponding to the spacing on the PC-board.
L 1003, L 1004: 6 turns with centre tap, otherwise as L 1001
L 1005: 5 turns, coil tap 0.5 turns, otherwise as L 1001
L 1006: 20 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound together with:
L 1007: 8 turns of 0.4 mm dia. (26 AWG) enamelled copper wire wound in a potted core of 11 x 7 mm (material K 12, AL-value = 16).

4.2. CONSTRUCTION

The transmit mixer is accommodated on the PC-board DK 1 OF 010 whose dimensions are 120 mm x 45 mm. Figure 11 shows this printed circuit board and the component locations.

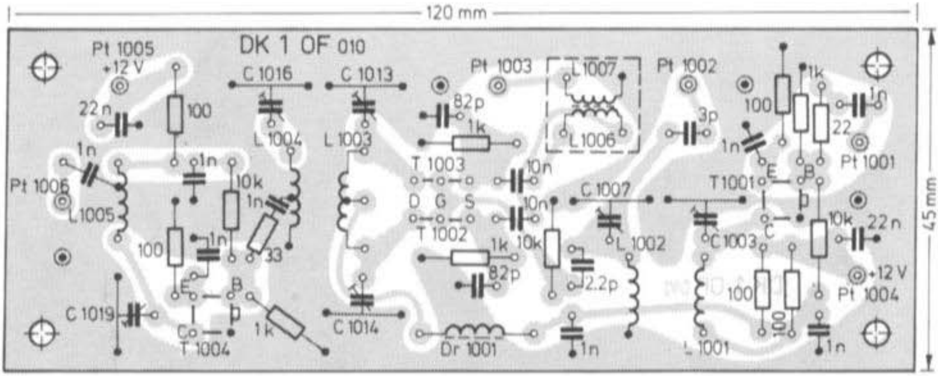

Fig. 11 : PC-board DK 1 OF 010 for the transmit mixer

4.3. ALIGNMENT

The alignment of the transmit mixer DK 1 OF 010 is carried out after installation into the complete unit. Firstly, all circuits are aligned for maximum output at the centre of the band (channel 40). Attention should be paid that both trimmers of the balanced drain circuit (C 1013, C 1014) should have approximately the same position. Finally, these two trimmers are aligned for maximum suppression of the oscillator frequency at the output. This alignment is carried out by careful, alternate adjustment of the trimmers.

5. DIRECT MODULATION OF THE SYNTHESIZER

It was stated in (1) that it was possible to directly modulate the 80 channel synthesizer but that it had certain disadvantages. A suitable circuit for this is to be given for those readers that would prefer this type of modulation.

It was mentioned in the description of the voltage-controlled oscillator DK 1 OF 001 that a tuning range of 135 MHz to 137 MHz is obtained with a tuning voltage of 4 V to 8 V. This corresponds to a tuning slope of 2 V/MHz. This means that a modulating voltage of ± 10 mV is required at input Pt 105 for a mean frequency deviation of ± 5 kHz. Since capacitor C 109 of 10 µF is connected in parallel with this input, higher modulation frequencies will be suppressed. It is therefore necessary for the corresponding emphasis to be made in the AF preamplifier.

The disadvantages given in (1) can be compared with the advantage of the simplification: The 9 MHz exciter DK 1 OF 007 together with the AF preamplifier DK 1 OF 008 could be replaced by a simple 9 MHz crystal-controlled oscillator in the described circuit (e.g. DK 1 PN 002).

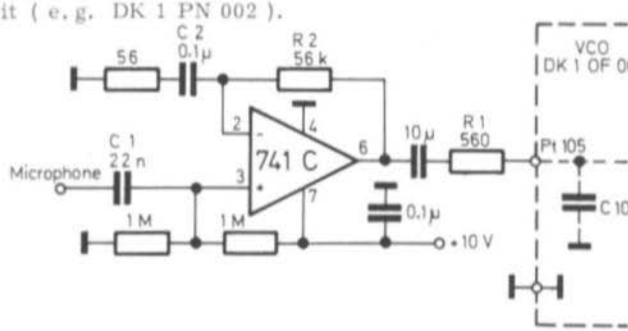

Fig. 12: Circuit diagram of an AF preamplifier for direct modulation of the synthesizer

A circuit diagram of a suitable AF preamplifier is given in Figure 12. In order not to affect the dynamic behaviour of the synthesizer, the modulator is connected via the decoupling resistor R 1. This resistor results in a lowpass filter together with capacitor C 109 whose effect on the frequency response is cancelled out by the RC-link R 2, C 2 in the feedback link of the integrated circuit. The coupling capacitor at the output of the integrated circuit should be able to accept both polarities which means that a bipolar electrolytic capacitor or plastic foil capacitor is required. Since the voltage gain of the whole circuit is about 1, approximately 7 mV (RMS) are required at the input for a frequency deviation of ± 5 kHz. The input impedance is approximately 400 kΩ.

EDITORIAL NOTES

A new frequency synthesizer with a channel spacing of 10 kHz is to be described in one of the next editions of VHF COMMUNICATIONS. This synthesizer can be used with the described exciter in a similar manner to that of the 80 channel oscillator. Such an oscillator is very suitable for use, for instance, in North America where a 30 kHz spacing is used.

6. REFERENCES

(1) J. Kestler: FM Transceiver with Multi-Channel Synthesizer
VHF COMMUNICATIONS 5 (1973), Edition 3, Pages 130-145

(2) D. E. Schmitzer: Active Audio Filters
VHF COMMUNICATIONS 1 (1969), Edition 4, Pages 218-235.

DIGITAL VOLTMETER

Short Description of Three Different Principles

by K. Wilk, DC 6 YF

Digital voltmeters (DVM) have the advantage over analog meters that there is far less possibility of reading the incorrect value. In addition to this, it is possible for the accuracy of the readings and resolution to be increased by a power of 10.

The following measuring methods have found to be successful for workshop and laboratory instruments; however they operate according to entirely different principles:

Frequency method
Sawtooth method
Dual-integration method.

Other methods which are especially fast or accurate are more complicated and therefore limited to special applications.

The basic digital voltmeters are usually designed for measurement of DC voltages. However, as with vacuum tube voltmeters (VTVM), a DVM can be extended so that alternating voltages, direct and alternating currents as well as resistance values can be measured.

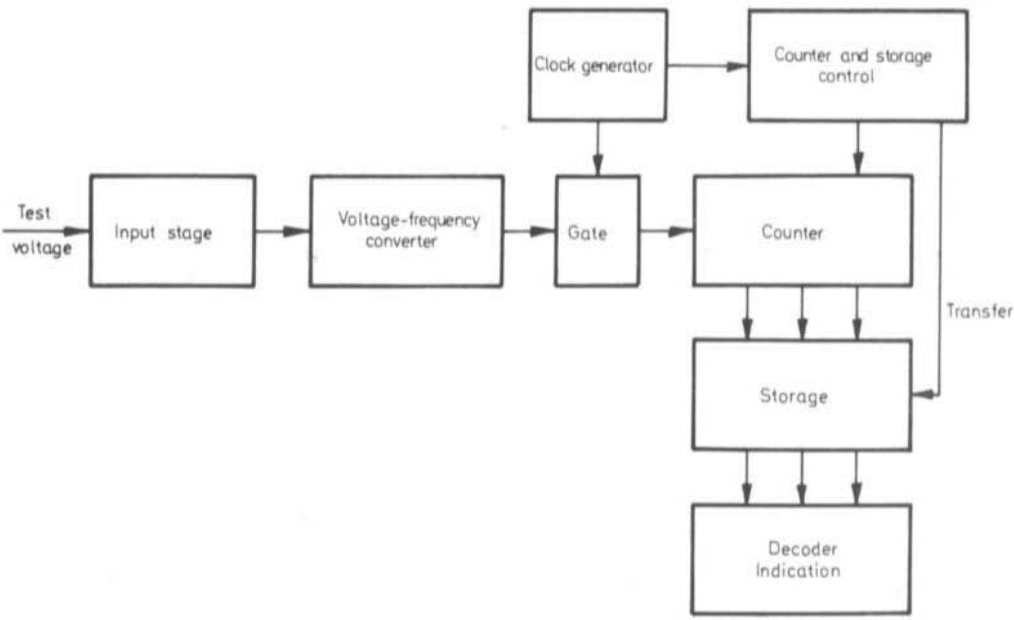

Fig. 1 : Block diagram of a digital voltmeter operating according to the frequency method

1. THE FREQUENCY PRINCIPLE

In the frequency principle, the test voltage is converted into impulses whose frequency is directly proportional to the test voltage. The evaluation and indication is made with the aid of a circuit that is similar to that of conventional digital frequency counters.

The voltage-frequency converter is mainly an integrator with subsequent comparator. The test voltage is fed to the integrator whose output voltage increases linear in time if a DC voltage is present. The rise time is proportional to the input voltage. The comparator compares the output voltage of the integrator with a constant reference voltage. If the voltage is equal, a voltage jump will appear at the output of the comparator and the integrating capacitor will be discharged via an electronic switch.The discharge time is inversely proportional to the input voltage to the integrator, which corresponds to the test voltage. This means that the impulse frequency corresponds to the test voltage.

Figure 1 shows a block diagram of such a configuration. The voltage-frequency converter is fed from an input stage that can be switched to the various input values and usually consists of an impedance converter with switchable voltage divider in the case of direct voltage measurements.

The clock generator controls a gate circuit in front of the counter that allows the pulses of the voltage-frequency converter to pass for a predetermined time during each measuring process. According to the test voltage and corresponding pulse frequency, a number of pulses will be passed to the counter that are proportional to the test voltage. The state of the counter is stored after each measurement and then reset to zero. A control circuit supplies these signals that operate in synchronous with the clock generator for controlling the gate. The BCD signals at the output of the storage are recoded and passed to the indicator tubes. The storage ensures that a flicker-free reading results that will only change when a variation of the test voltage occurs. This part of the circuit does not differ greatly from the principle used in frequency counters.

The accuracy of the indication is dependent on the timing generator and the voltage-frequency converter. With the frequency method of realizing a DVM the input magnitude is converted into a time interval in an integrator. The integration principle ensures a good interference voltage suppression. This is especially of value for symmetrical interference voltages, such as a superimposed AC hum.

If an automatic circuit is not provided, it will be necessary to calibrate the zero point and gain at regular intervals against a calibration voltage source built into DVM, and to adjust when necessary.

2. THE SAWTOOTH METHOD

With the sawtooth method, the connected test voltage is compared to a sawtooth voltage which increases linear in time. The time taken from the zero pass until voltage equality, is proportional to the test voltage and will be indicated. This means that the measuring voltage cannot be evaluated as it is but must be converted into a time interval that is counted and indicated digitally.

The operation of such a meter is to be described with the aid of the block diagram given in Figure 2. The test voltage is fed to the measuring system via the input stage. The sawtooth generator provides a time-linear voltage that, for instance, increases from a negative value linearly to a positive value during a measuring period. A signal in the form of a voltage jump will appear at the output of the zero-comparator during the zero pass. When the sawtooth voltage has reached the same value as the measuring voltage at the comparator, a second voltage jump will appear, in this case at the output of the comparator. The time between both signals is proportional to the test voltage. A clock provides impulses of a constant frequency during this period to the counter. The number of pulses present at the counter is then a measure of the connected test voltage.

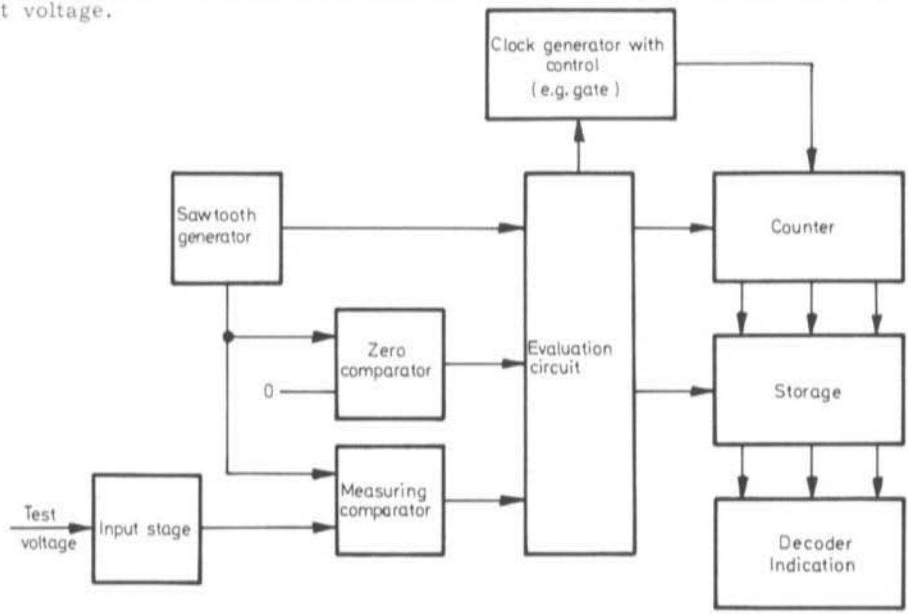

Fig. 2: Block diagram of a digital voltmeter operating according to the sawtooth method

An evaluation circuit controls the clock, the reset of the counter before the voltage-time conversion, and the subsequent acceptance of the measuring value by the storage. The value from the storage is then decoded and indicated in the same manner as with the frequency method.

The sawtooth method possesses a very simple, clear circuit and the measuring accuracy is mainly dependent on the two generators.

With respect to the calibration, the same is valid as was stated for the frequency method.

3. THE DUAL-INTEGRATION METHOD

The main parts of a DVM circuit operating according to the dual-integration (slope) method are the integrator, signal evaluation, counter and reference voltage source.

The measuring cycle is commenced after resetting the counter and the measuring voltage is fed to the input of the integrator. The counter is fed with pulses of a constant frequency from a clock. After a certain period, e.g. when position (1) 000 is reached in the case of a three-digit counter (transfer from hundreds to thousands), the integrator input is switched to a constant voltage source (reference voltage) that has the opposite polarity to the test voltage. The integrator is now discharged linear with time. The time taken from the connection of the constant voltage until the zero potential is reached at the output of the integrator is a measure of the test voltage.

The measuring cycle comprises the charging process from zero potential with the test voltage and the discharge process back to zero potential with the reference voltage, from which the designation dual integration method has been obtained.

It is only necessary for the clock to possess a short-term stability, which means that it is only necessary for its frequency to remain constant during one measuring cycle. A long-term stability is difficult to obtain and is not required in this principle; the same is valid for the timing elements of the integrator. The accuracy of the circuit is mainly dependent on the stability of the reference voltage. As with the frequency method, an integrating method has a good noise and hum suppression.

The block diagram (Fig. 3) shows the various stages of such a DVM.

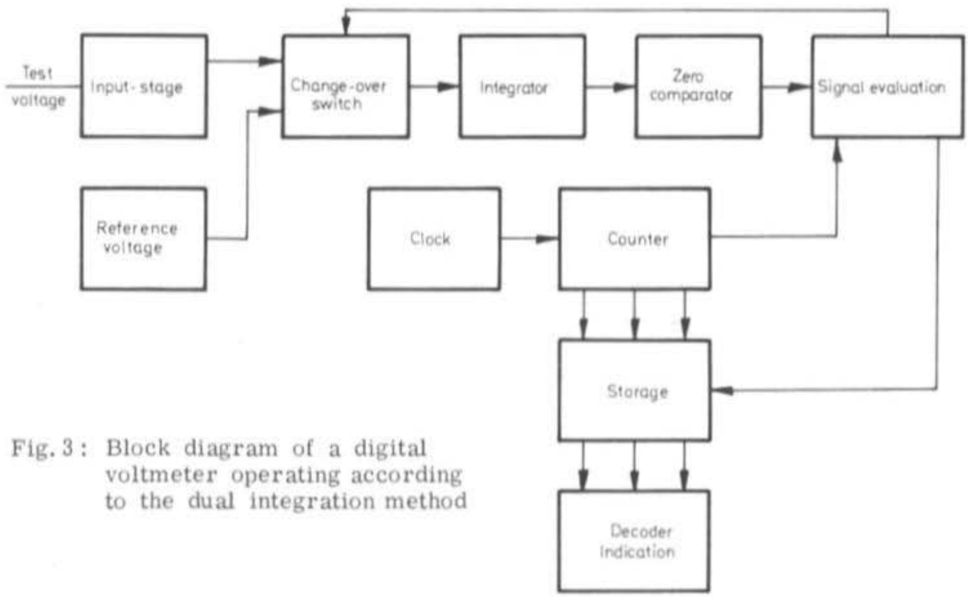

Fig. 3 : Block diagram of a digital voltmeter operating according to the dual integration method

The integrator is fed with either the test voltage (via the input stage) or the reference voltage, according to the position of the electronic switch. The signal evaluation circuit recalls the overflow of the counter and causes the electronic switch to change over. A signal from the zero comparator will be received on reaching zero potential which will cause it to pass a pulse to the storage for acceptance of the momentary counter state. The contents of the storage are indicated as the result of each measurement. The next measuring cycle commences after the subsequent reset of the counter. With the dual integration (dual slope) method, the accuracy depends mainly on the stability of the reference voltage source which is a demand that can be more easily met than the provision of exact times or frequencies as is the case with the frequency and sawtooth methods. Since a more stable DVM that is easier to operate results with the same extent of circuitry, practically all digital voltmeters offered on the market operate according to the dual-integration method.

4. CONSTRUCTION OF A SIMPLE DIGITAL VOLTMETER

Principally speaking, no great difficulties are to be expected during the construction of a digital voltmeter using any of the previously mentioned methods.

In the case of the frequency method, it is not easy to obtain a sufficiently linear voltage-frequency converter.

The dual-integration method can be classed as the most favourable method technically, but fault-finding is not very simple if the unit does not work immediately due to the double integration of analog and digital portion and the synchronizing and storage stages.

The sawtooth method seems especially suitable for a virtually foolproof construction. If the circuit is designed correctly, one is able to avoid complicated logical interactions, storage and delay stages.

A digital voltmeter suitable for workshop use that operates according to the sawtooth method, with simple construction will be published in one of the next editions of VHF COMMUNICATIONS.

AN INTEGRATED RECEIVER SYSTEM FOR AM, FM, SSB AND CW
PART IV: AF AMPLIFIER AND CW FILTER
by H. J. Franke, DK 1 PN

The AF amplifier with active CW filter to be described has been especially designed for the integrated receiver system described in (1). This module is also enclosed in a TEKO box in a similar manner to the SSB-IF module (2) and carrier oscillator (3). In principle, it is possible for any AF amplifier to be used that can be driven with an AF voltage of 7 mV. However, the more expensive AF integrated circuit SL 630 possesses the following advantages:

The upper frequency limit is adjustable with the aid of only 1 external capacitor.

The gain is adjustable with the aid of a DC voltage line.

The amplifier can be switched off by a DC voltage line with very low current drain.

Figure 1 shows the characteristic curves of the integrated AF amplifier for voice reproduction as available at the loudspeaker output as well as the frequency response of the CW filter as present at the headphone output.

1. CIRCUIT DETAILS

The circuit diagram of the AF amplifier module given in Figure 2 shows how the voice reproduction is provided over the loudspeaker from the integrated AF amplifier SL 630 whereas CW reception is made via the narrow band filter and headphone output.

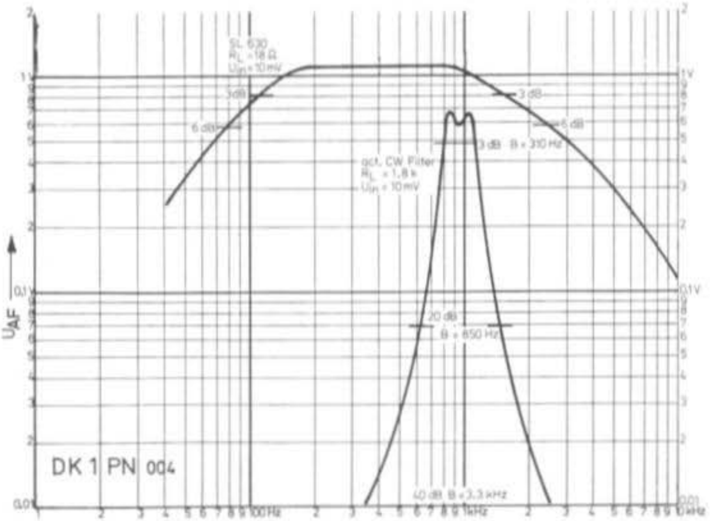

Fig. 1: Frequency response of the AF amplifier DK 1 PN 004 at the loudspeaker output (wide) and headphone output (narrow)

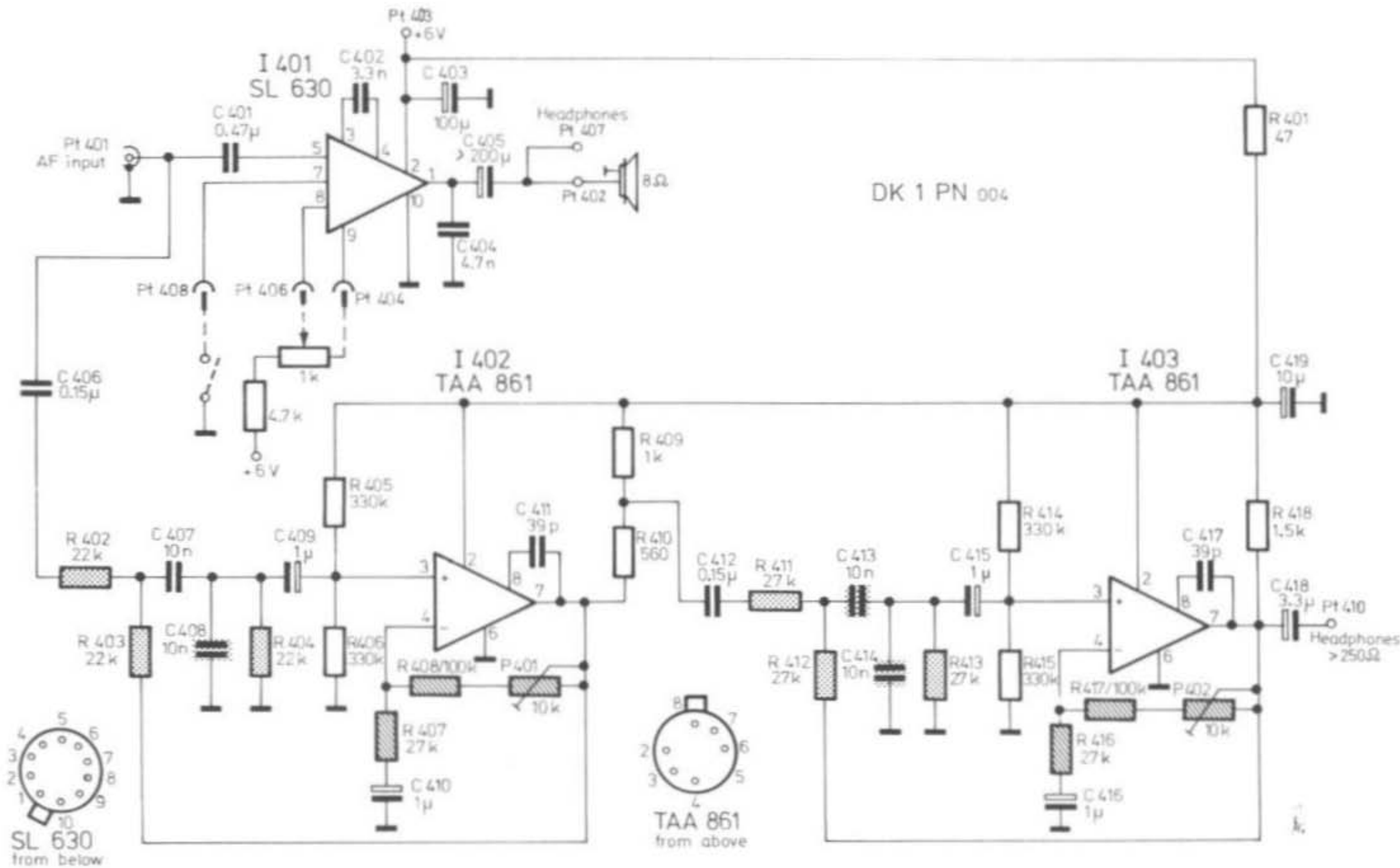

Fig. 2: Circuit diagram of the AF amplifier with active CW bandpass-filter

1.1. VOICE CHANNEL

The upper frequency limit f_c is determined by capacitor C 402 which can be calculated according to the following formula:

$$\omega_c = 2 \pi f_c = \frac{10^8}{C\ 402} \qquad \text{with C 402 in pF and } f_c \text{ in Hz}$$

As has already been mentioned, connection 7 of the integrated circuit can be used for switching off the amplifier and has therefore been fed to the connector. In the case of CW reception, this point is grounded.

The impedance of the loudspeaker should be at least 8 Ω. In this case the output power will amount to approximately 50 mW, which is usually sufficient for home and portable operation. The AF amplifier is operated from 6 V as are the other ICs manufactured by Plessey.

1.2. CW CHANNEL

Two active filters are operative in addition to the AF amplifier during CW reception. The resonant frequencies of these filters are somewhat offset from another at somewhat over 800 Hz and somewhat over 1000 Hz. The frequency-determining components are marked by dots in the circuit diagram. The tolerances of these components should be less than ± 10% and preferably only ± 5%.

The shaded resistors determine the gain of the two operational amplifiers I 402 and I 403 and thus the Q of the resonant circuits. The voltage divider comprising resistors R 410 and R 409 ensures that the second amplifier limits large signal amplitudes before the first amplifier. With the given values, the CW channel will process signals up to a voltage of 15 mV.

Both amplifier stages will break into oscillation when the resistor combination R 408/P 401 or R 417/P 402 has too large a value. For this reason, the adjustment range of the potentiometers was limited to 10% with the aid of fixed resistors.

In addition to this, the first amplifier will oscillate if the input of the circuit is of too high an impedance. This means that the input must be terminated by a resistor whose value is far less than that of R 402. In the described circuit, this is done by the input resistor of amplifier I 401.

An AF sweep generator is required for an exact alignment of the CW filter. If such an unit is not available, potentiometers P 401 and P 402 can be replaced by fixed resistors of 4.5 kΩ. In this case, the 3 dB bandwidth of 300 Hz will remain, however, it is possible that the passband curve will be unsymmetrical or rounded.

2. CONSTRUCTION

The AF amplifier with CW bandpass filter is accommodated on the PC-board DK 1 PN 004 which is shown in Figure 3. The PC-board is suitable for mounting in a TEKO box 3A and can be provided with a 13-pin connector.

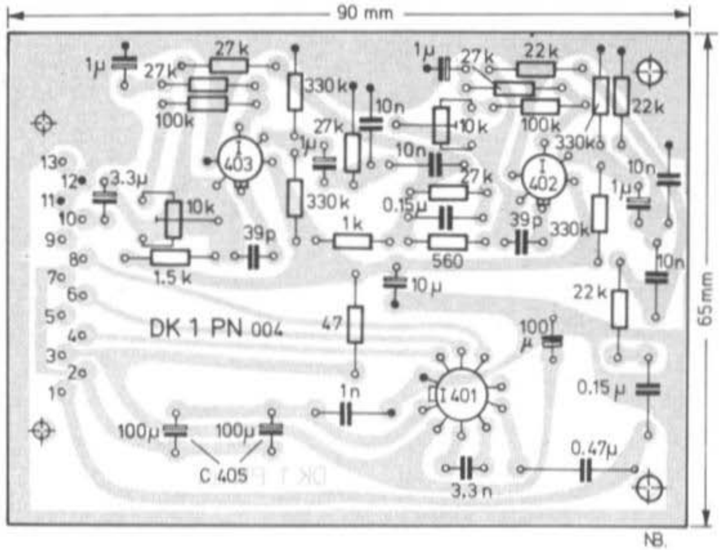

Fig. 3: Component locations on PC-board DK 1 PN 004

2.1. SPECIAL COMPONENTS

I 401: SL 630 C (Plessey)
I 402, I 403: TAA 861 (Siemens)

C 402: 3.3 nF styroflex capacitor
C 411, C 417: 39 pF ceramic capacitors
C 407, C 408, C 413, C 414: 10 nF ± 5% plastic foil capacitors

13 pin connector
TEKO box 3A.

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AN INTEGRATED RECEIVER SYSTEM FOR AM, FM, SSB AND CW
PART V: INPUT MODULE AND FM PORTION
by H. J. Franke, DK 1 PN

The following article describes the fifth module of the 9 MHz receiver system. When completed, this modular receiver will have a separate channel for each of the modulation modes AM, FM and SSB/CW. It is mainly equipped with integrated circuits. The block diagram of the complete receiver is given in Fig. 1. The connection point numbers (Pt) provide information as to the interconnection of the receiver.

The following modules have already been described in VHF COMMUNICATIONS: SSB/CW IF-portion (1), the carrier oscillator (2) and the AF amplifier and CW filter (3).

1. CHARACTERISTICS OF THE 9 MHz MODULE WITH FM-PORTION

The most important output values of the module are given in Figure 2 in the form of a diagram. The block diagram also given in this illustration shows the filters and active components as well as the gain values. In the SSB/CW mode, the signal is passed through a resonant circuit which acts as a transformer to the (FM) crystal filter and the Plessey IC SL 610. This IC amplifies the signal by approximately 20 dB as long as no control voltage is provided from the SSB module DK 1 PN 003. When fully controlled, the integrated circuit SL 610 will decrease the signal by 50 dB. The transformation of the resonant circuit and the attenuation of the filter lead to overall gain values of 31 dB without control and - 21 dB with full control of this module. In the FM mode, the signal is amplified in a dual-gate MOSFET subsequent to the crystal filter. A junction FET is then used as buffer so that the output to the AM module (Pt 513) and the integrated FM system do not load or detune the resonant circuit. The gain up to this point amounts to 42 dB with weak signals and approximately 12 dB when the MOSFET is controlled by the control voltage from the FM module.

The integrated FM circuit CA 3089 provides a noise voltage U_n of approximately 250 mV_{pp} to the AF output Pt 503 when the squelch is open and the input signals are less than 2 μV. With input signals between 1.5 μV and 3 μV, the noise voltage falls quickly towards zero and the demodulated AF voltage U_{mod} will be provided. The amplitude depends on the frequency deviation, and the following values have been measured.

Frequency Deviation kHz	AF Voltage peak to peak
± 2	30 mV
± 3	45 mV
± 4	60 mV

The integrated circuit CA 3089 also provides a current for the S-meter. A quescent current is available until the noise is quietened; after this, the current flowing through the S-meter is practically dB-linear, which means that the scale can be calibrated linearly in dB. Since this indication is dependent on the RF carrier, it can also be used for AM. The indication of the quiescent current will be suppressed in a module that still has to be developed.

DK 1 PN

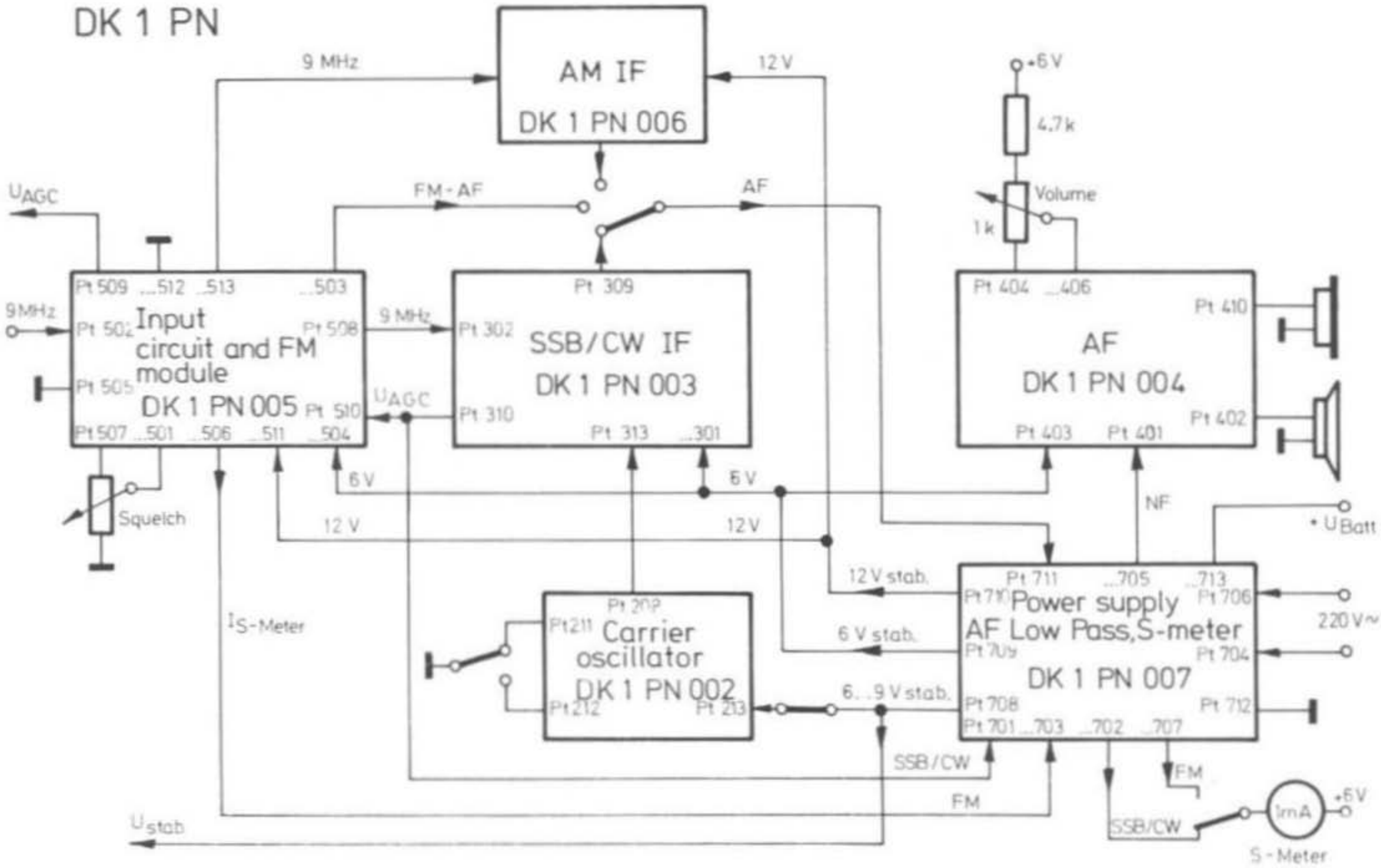

Fig. 1: Block diagram of an integrated receiver for AM, FM, SSB and CW

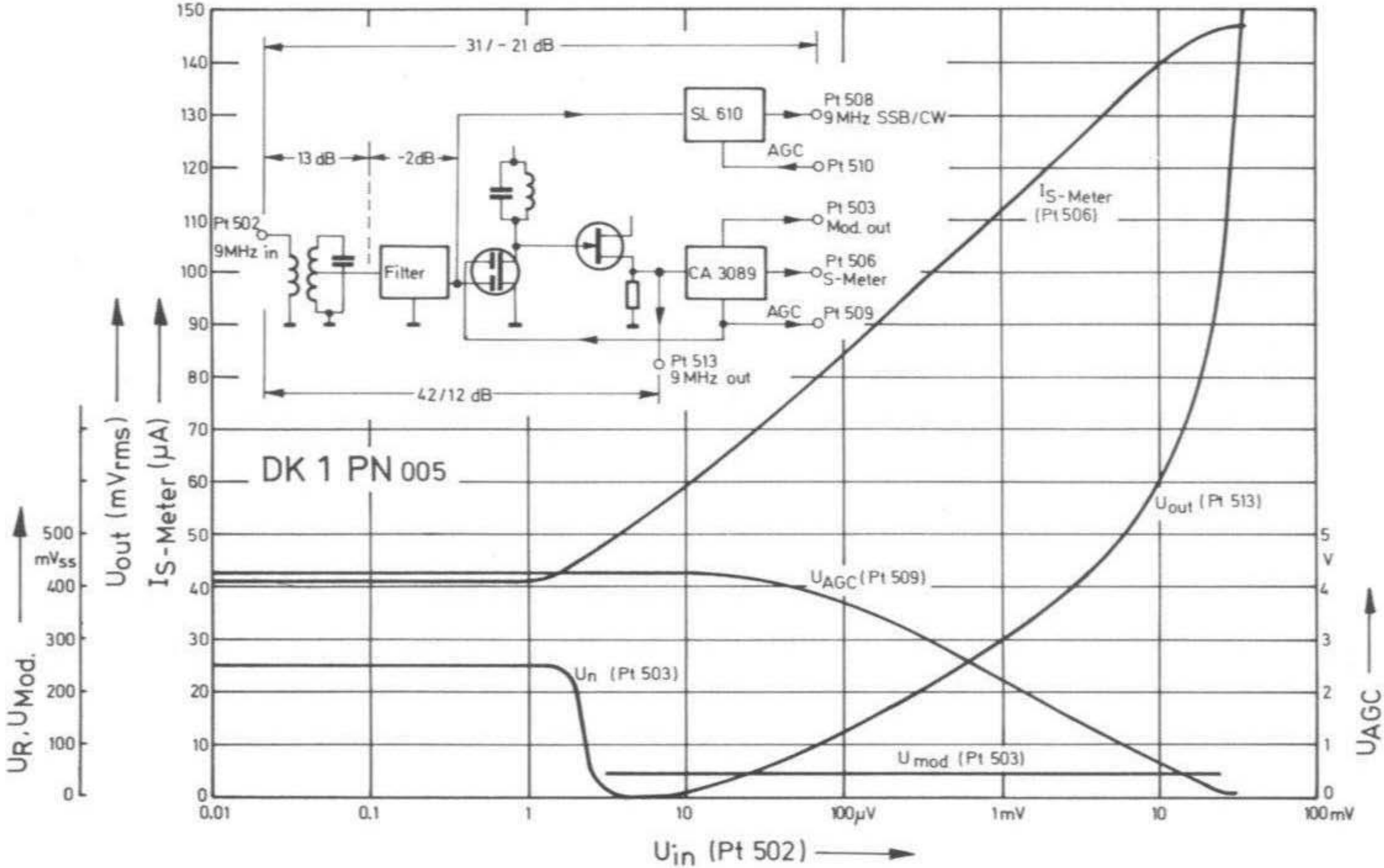

Fig. 2: Block diagram, gain curves and measured values of the input circuit and FM module DK 1 PN 005

Finally, a control voltage U_{AGC} is generated whose action is delayed by approx. 20 dB. This is to ensure that the gain of the previous stages is only controlled when the signal has full quietening (when the noise is completely suppressed). This voltage is in the range of approximately 4 V to 0 V, and is thus suitable for direct control of the MOSFETs. In this module, the gain of the MOSFET directly subsequent to the crystal filter is controlled thus ensuring that the RF voltage for the AM module at connection Pt 513 remains within limits. The overall gain of the module up to this output will then be reduced from 42 dB to 12 dB with full control.

2. CIRCUIT DETAILS

Figure 3 gives the complete circuit diagram of the input circuit and FM portion. The input coupling is designed for connection of a coaxial cable. The tapping on the resonant circuit is dimensioned so that the crystal filter XF-9E is presented with an impedance of 1.2 kΩ. However, it is necessary for the input of the module to be loaded with 50 Ω to 60 Ω in order to achieve this.

The SSB and FM link are connected in parallel at the output of the crystal filter. Capacitor C 517 replaces the usual trimmer for alignment of the passband curve of the crystal filter for minimum ripple (max. 2 dB), which cannot be correctly aligned without the corresponding swept-frequency equipment. The module can, of course, also be constructed for only SSB or only FM reception. In this case, it is only necessary to leave out the corresponding components: Capacitor C 517 should be increased to approximately 25 pF and resistor R 511 reduced to approximately 1.4 kΩ if the SL 610 is deleted. On the other hand, if the FM portion is not to be used, no modifications will be required.

In case of the Plessey IC SL 610, all measures have been taken to ensure neutralization. Resistor R 515 is especially required in order to avoid oscillation when loading the output with a long coaxial cable. The operating voltage of 6 V which is required for all Plessey ICs should be stabilized in the previously mentioned additional module.

The operating point of the dual-gate MOSFET T 501 in the FM circuit is stabilized by the source resistor and a positive bias voltage. Any further spread of the transistors can be compensated for by selecting the correct value for resistor R 503 so that the drain current obtains the given value of 9 mA $\pm$ 10%. The subsequent buffer stage equipped with the junction FET is not critical. Any RF FET can be used whose drain current is at least 3 mA at a gate voltage of 0 V. This current is necessary to obtain a sufficiently low output impedance. The load exhibited by the FM module amounts to 390 Ω to which the load of the AM module must also be added. However, the input impedance of such a module will be high.

The resonant circuit comprising L 503 and C 525 is used for the quadrature demodulation. With this type of FM demodulation, the phase curve of a resonant circuit is utilized which is fundamentally of "S" shape that appears as the amplitude curve with a ratio detector. The spacing between the humps of the phase-curve and thus the slope of the straight portion between them is dependent on the Q of the resonant circuit. A frequency deviation of several kHz is very small when compared with the intermediate frequency of 9 MHz or 10.7 MHz. In order to obtain a sufficiently high AF voltage in spite of the low frequency

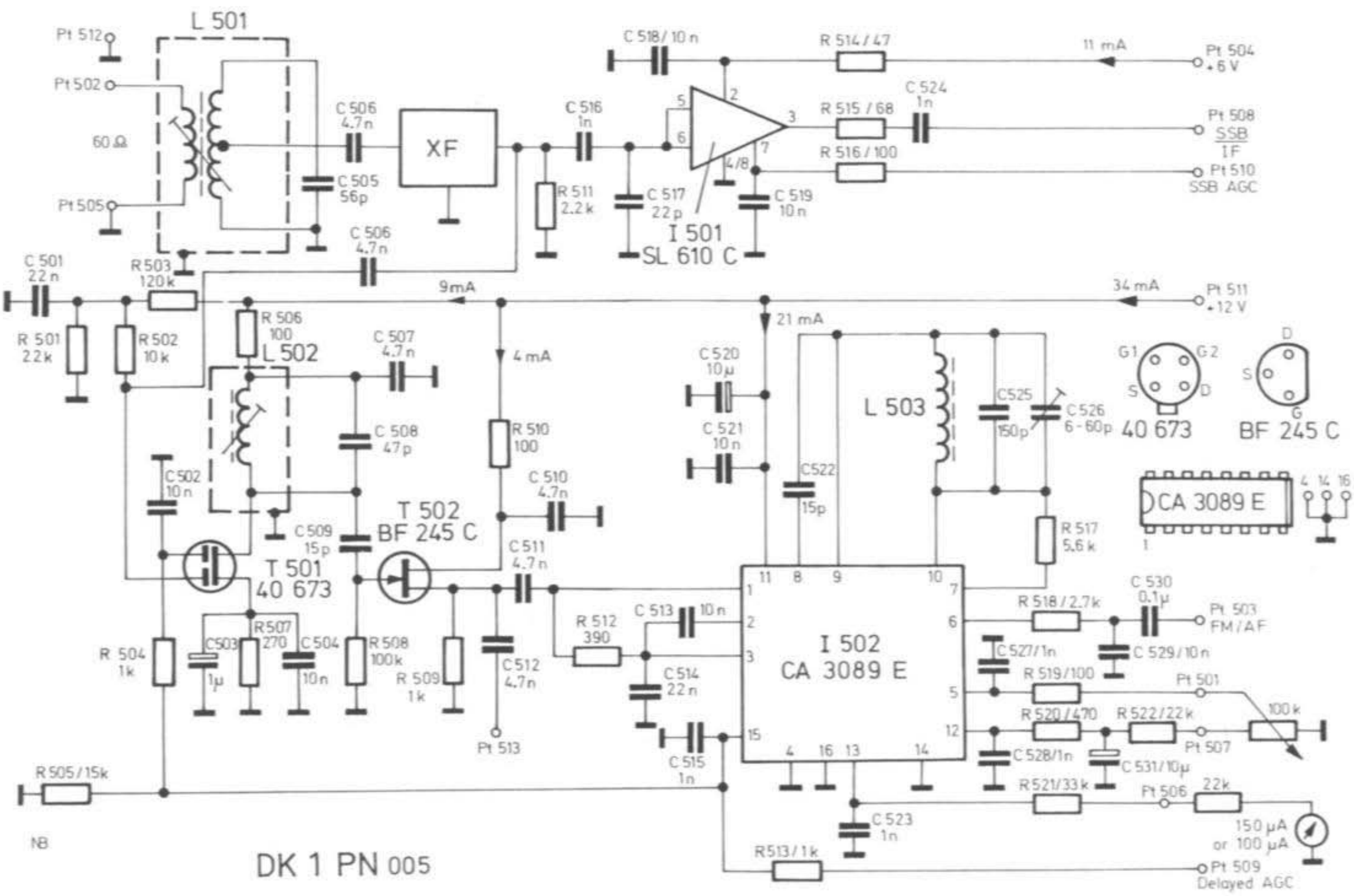

Fig. 3: Circuit diagram of the input circuit and FM module

deviation, it is necessary for a steep phase curve, e.g. a high Q of the resonant circuit, to be used. Inductance L 503 is therefore in the form of a ferrite toroid which allows a non-loaded Q of at least 150 to be obtained even when the winding with enamelled copper wire is not made perfectly. With silk-covered enamelled copper wire, Q-values of up to 230 have been measured. It is not possible to obtain such high Q-values even when wound carefully with the coil sets provided for L 501 and L 502, or with potted cores.

The AF output of the FM module is provided with a lowpass filter for filtering out any residual RF voltage and to ensure that no RF-injection occurs from local broadcasting stations. The same is valid for the output and input of the squelch circuit. The connection leads to the adjustment potentiometer of the squelch (100 kΩ), which is not on the PC board, only carry a DC-voltage. The integrated circuit CA 3089 will only provide a sufficiently large DC-voltage for opening the squelch when the signal-to-noise ratio allows the signal to be demodulated (4). The threshold voltage is thus not dependent on the amount of gain, but on the signal-to-noise ratio. This means that it is not the threshold voltage that is adjusted with the potentiometer but the amount of quietening. The squelch will only operate correctly when the Q of inductance L 503 is high enough for use with the low frequency deviation used in modern FM communications.

The voltage at pin 7 of the CA 3089 can be used for automatic frequency control of an oscillator. This has not been utilized in our application. The value of the external resistor in series with the S-meter depends on the impedance of the meter used. The value of this resistor should be selected so that the highest signal to be expected produces full scale deflection. In contrast to the control voltage of the Plessey ICs, this S-meter indication will follow fluctuations of the signal strength and gain variations without noticeable delay so that it can be used, most favourably, when rotating the antenna, whilst tuning up, and for the alignment.

3. CONSTRUCTION

The circuit given in Figure 3 is accommodated in a TEKO box 3A. The PC-board DK 1 PN 005 is 90 mm x 65 mm, is double-coated. It possesses through-contacts, since the high gain of ghe FM-portion can only be handled without difficulties with such a compact construction with a continuous ground surface on the component side. Figure 4 gives this PC-board (without ground surface) and indicates the component locations. A photograph of a prototype is given in Figure 5. It shows that only miniature components can be accommodated on this PC-board. Further details are given in Section 2.1. Sufficient room has been provided around the integrated circuit CA 3089 so that a 16-pin socket can be provided. The use of a socket does not bring any disadvantages in our application and could be of advantage for fault-finding and repair.

No difficulties are to be expected during construction when the well-proved PC-board is used. Special attention must only be paid to the correct winding and connection of inductances L 501 and L 502. The connections of the crystal filter must be short and directly made to the conductor lanes that are fed up to the crystal filter. It would have had no advantage to make these conductor lanes longer since the connections of the crystal filter are designed for wire connections. The case of the crystal filter should be directly in contact with the ground surface. The resistors, on the other hand, should maintain a small spacing if they are not capless types in order to avoid short-circuits due to insufficient insulation . The toroid core L 503 can be mounted on the PC-board with the aid of a drop of dual-component glue. All coupling and bypass capacitors should be soldered into place with the shortest possible connections.

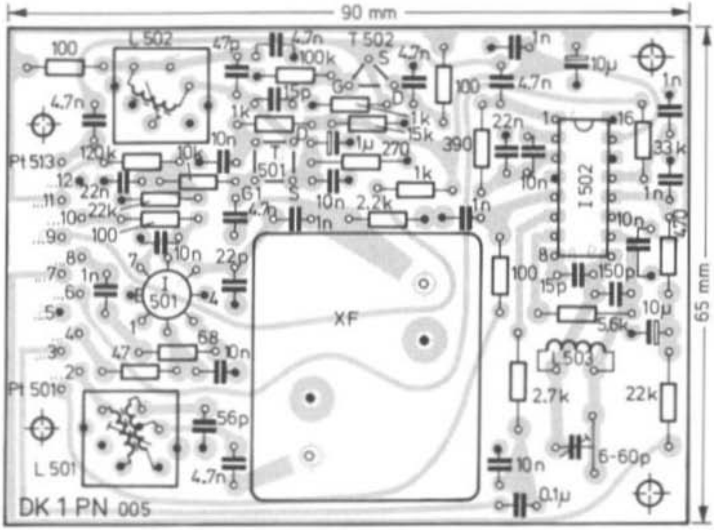

Fig. 4: Printed circuit board DK 1 PN 005 with component locations

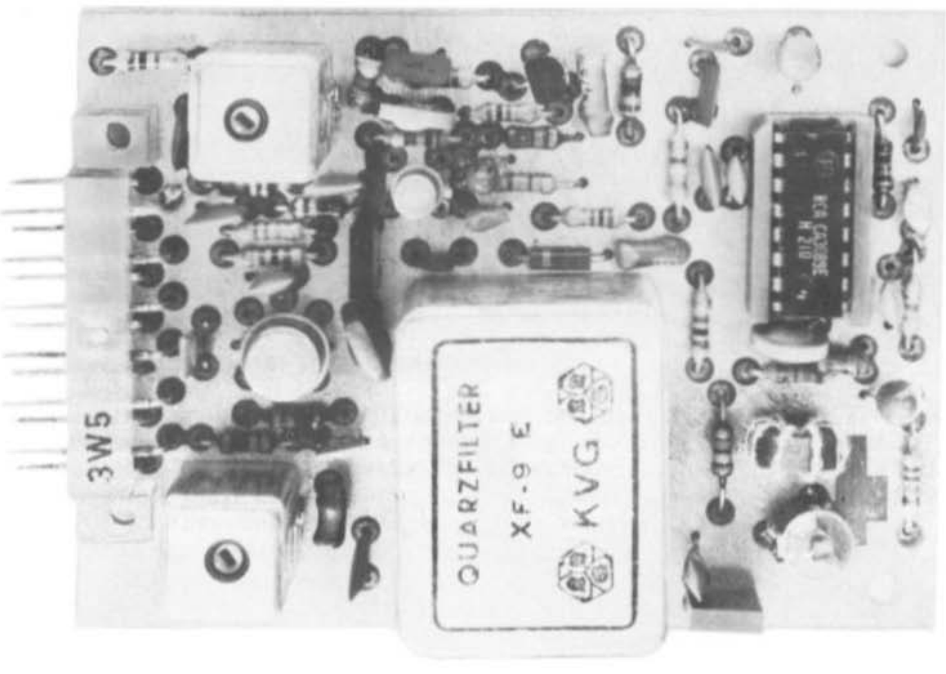

Fig. 5: Prototype of module DK 1 PN 005

3.1. COMPONENTS

I 501: SL 610 C (Plessey)
I 502: CA 3089 E (RCA)

T 501: 40673 or other similar dual-gate MOSFET
T 502: BF 245 C, BF 246 (TI) or W 245, E 300 (Siliconix)

L 501: Coupling winding 3 turns, coil 30 turns, coil tap 13 turns from the cold end. Wire: 0.25 dia. (30 AWG) enamelled copper wire using special coil set.
L 502: 30 turns, wire and special coil set as for L 501
L 503: 7 turns of 0.25 mm (30 AWG) (silk-covered) enamelled copper wire (1.7 μH) wound on a toroid core R8-M7 (004-05), A_L = 40 or R 10 - K 1, A_L = 40

C 526: 6 - 40 pF or 6 - 60 pF ceramic or plastic foil trimmers of 7 mm dia.
C 520, C 531: 10 μF/16 V tantalium electrolytic capacitors, drop-type
C 503: 1 μF tantalium electrolytic capacitor, drop-type

All other capacitors: Ceramic disc types for 5 mm spacing.
A plastic-foil or tantalium capacitor can also be used for C 530.
A spacing of 10 mm or 12.5 mm is available for all resistors.

1 crystal filter XF-9E:
A high degree of distortion will occur with this 12 kHz filter if the frequency deviation of the station to be received is greater than ± 5 kHz. For this reason, a special 15 kHz version of this filter will be available shortly for those areas where a greater frequency deviation is usually used.

One 16-pin DIL IC socket, one 13-pole connector, one TEKO box 3A.

4. ALIGNMENT

The current values of each stage are given in the circuit diagram and should be checked immediately during the construction. The drain current of the MOSFET T 501 may have to be adjusted to approximately 9 mA by varying the value of R 503. After this has been carried out, the alignment is very simple.

The S-meter to be used should be connected to Pt 506 and the carrier oscillator module DK1 PN 002 to input Pt 502. It is now only necessary for inductances L 501 and L 502 to be aligned for maximum reading. After this, a high impedance voltmeter (greater than 10 kΩ/V, 10 V range) should be connected to Pt 507 and trimmer capacitor C 526 aligned for minimum reading on the voltmeter. After this, these three alignment settings should be corrected.

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Part IV: AF Amplifier and CW Filter
In this edition of VHF COMMUNICATIONS

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ANTENNA NOTEBOOK

by T. Bittan, G 3 JVQ / DJ Ø BQ

Since articles on VHF/UHF antennas tend to be rather neglected nowadays it was thought that a regular series of antenna articles would be of interest.

1. CIRCULAR POLARISATION

The basic fundamentals of circular polarisation were given in (1) and (2). This article aroused interest in this type of polarisation, also in commercial VHF communications and mobile radio circles. Recent measurements over a 200 km path of rolling countryside has shown that the circular polarisation exhibited an additional gain of 12 dB over the same antenna when switched to linear polarisation. The measured gain over this distance, and a distance of 275 km have shown that the reduced path loss of circular polarisation using the author's homemade 10 element crossed Yagi provided a gain of 22-24 dB over a vertical dipole. In the case of the 275 km path, the other station was using linear (vertical) polarisation. Experience gained by DK 4 MV with a group of eight five-element crossed Yagis has shown that circular polarisation does not show any special features for communication with linear polarised stations upto about 200 km, after which a considerable improvement is noticeable. However, if the location of the station or partner is poor, the gain of circular polarisation over linear polarisation will be noticed immediately, even with local stations.

1.1. CLOCKWISE OR ANTICLOCKWISE ?

For communication with linearly polarised stations, it is immaterial whether clockwise or anticlockwise polarisation is used. However, due to the growth of circular polarisation, and this trend will most certainly continue, it is important that a common standard polarisation is agreed.

Clockwise circular polarisation is usually used in commercial practice where an extra degree of isolation is not required between two systems. For this reason, the author would like to forward clockwise circular polarisation for all amateur transmissions in the VHF, UHF and microwave bands. The designation clockwise or anticlockwise is the rotation direction as seen from behind the antenna looking in the direction of radiation. A crossed dipole, for instance, transmits clockwise circular polarisation in one direction and anticlockwise circular in the other. However, the direction of transmission is known in the case of a Yagi antenna.

The question is now how do we know whether we are transmitting clockwise or anticlockwise circular polarisation? This indeed a problem that is not too easily to explain. Anyway, let us assume a crossed dipole as seen from behind. The crossed dipole (or Yagi) should now be mounted so that the two "hot ends" (to which the inner conductor of the coaxial cable is connected) point upwards in the form of a "V" as shown in Figure 1. The asymmetrical phase line is now connected so that the shorter arm "A" is connected to dipole half "A" and the longer arm "B" to dipole half "B". The resulting circular polarisation will be circular clockwise.

Of course, the same can be done in the shack if two equal lengths of coaxial cable are fed to the shack. In this case, it is possible to select either the vertical or horizontal antenna separately, or connect the phase line to both

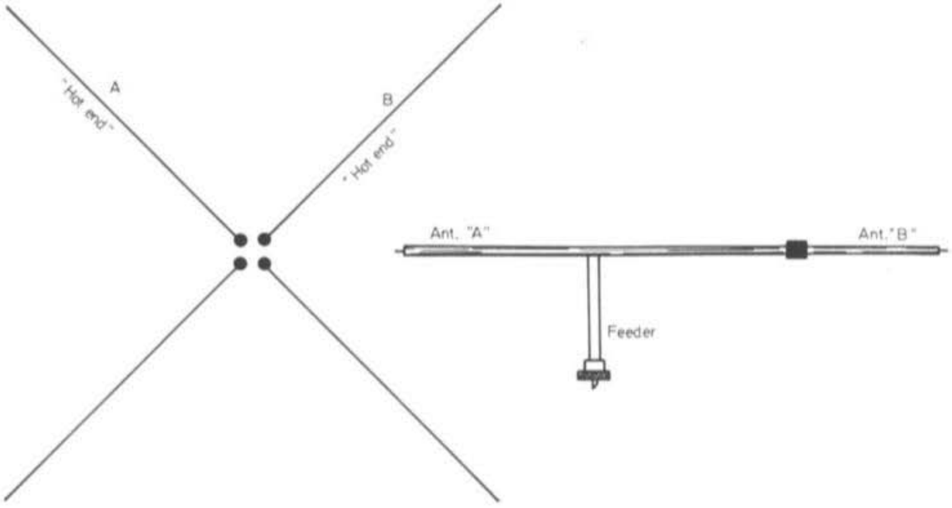


Fig. 1: Connection of a crossed Yagi to obtain clockwise circular polarisation

antennas to obtain circular polarisation. Since the author's introductory article on circular polarisation appeared in (1), the manufacturers of the J-Beam MOONBOUNCER antenna have changed their range of crossed Yagis so that they are similar to the author's G3JVQ Twister as described in (1). This means that the new range of antennas have the two dipoles of the crossed Yagi directly adjacent to another. Therefore the information given in this article is also valid for the new range of MOONBOUNCER antennas, which now comprise crossed Yagis of five, eight and ten elements. Of course, it is also valid for any other homemade antennas constructed similar to the "Twister".

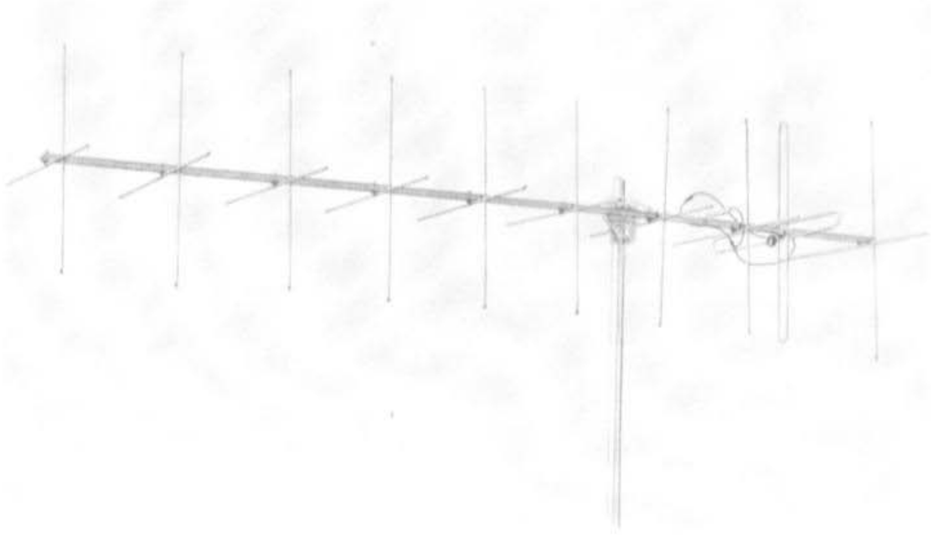


Fig. 2: Photograph of the J-Beam Moonbouncer

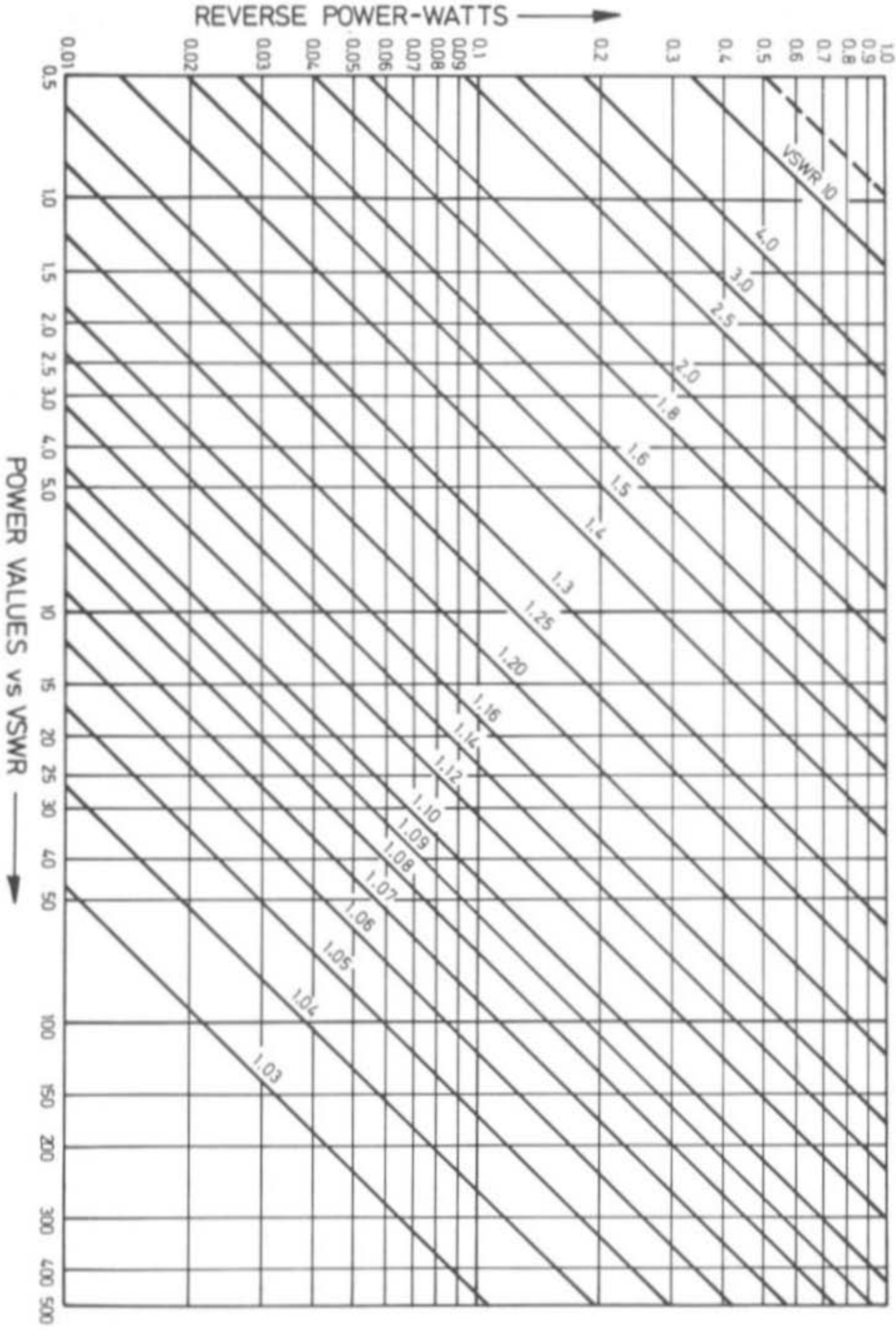


Fig.3: Power loss as a function of the VSWR

If only circular polarisation is to be used, it will be adviseable to mount the crossed Yagi so that the elements are mounted diagonally in the form of an "X". This ensures that the mast has the least effect. However, this will not be possible if the polarisation is to be switched.

Figure 2 shows a photograph of the new ten-element crossed Yagi of the MOONBOUNCER series of J-Beam antennas which has now been reduced in length from 4.92 meters to 3.65 meters.

2. POWER LOSSES DUE TO STANDING WAVES

Ever since reflectometers have become readily available to the amateur there has grown a completely exaggerated relationship between the standing wave ratio and the performance of an antenna. The author knows of several amateurs who would return an antenna to the manufacturers if the SWR was 2.1 : 1 instead of their maximum permissible 1.1 : 1. When it is considered how much output power is actually reflected by the antenna and when this is correlated to the output power, it will be seen that it will probably not be measureable at the receive end let alone be audible. This is also assuming that this power is actually lost, which is not true.

Let us assume the following with the aid of Figure 3: Given are an output power of 5 W, as well as a "terrible" standing wave ratio of 2 : 1. It will be seen in Figure 3 that the reverse power is 0.5 W so that the "lost" power is of this value and 4.5 W will be radiated. It is known that the power must be doubled or halved to produce a power ratio of 3 dB. In our case, the power ratio is 1.11 for a 2 : 1 standing wave ratio which is in the order of 0.5 dB loss. The power loss for a SWR of 1.5 : 1 is only 0.17 dB. This means that standing wave ratios of less than 1.5 : 1 are completely satisfactory.

The above information is, of course, based on the SWR at the antenna and not that measured in the shack at the end of a long feeder (3). The author does not wish to underestimate the importance of low standing wave ratios, but wishes them to be seen in their correct perspective.

3. REFERENCES

(1) T. Bittan: Circular Polarisation on 2 Metres
VHF COMMUNICATIONS 5 (1973), Edition 2, Pages 104-109

(2) Dr. A. Hock: Theory, Advantages and Type of Antennas for Circular Polarisation at UHF
VHF COMMUNICATIONS 5 (1973), Edition 2, Pages 110-115

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

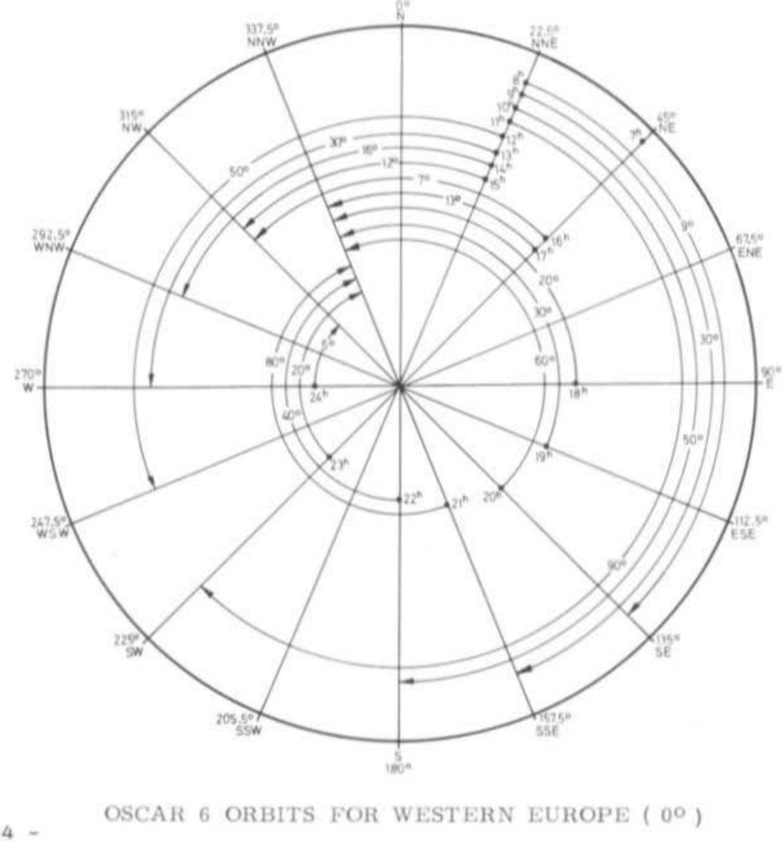
OSCAR 6: ANTENNA DIRECTION AS A FUNCTION OF TIME FOR A GROUND STATION IN WESTERN EUROPE

The following drawings allow the approximate antenna direction to be established for communication over the amateur satellite OSCAR 6 from ground stations in Western and Central Europe. All times are given in Central European Time, which is one hour ahead of GMT. The maximum elevation is also given.

The commencement of each pass is assumed to be at the full hour; if this is not the case, it will be necessary to interpolate the information.

The diagram is based on the following information:

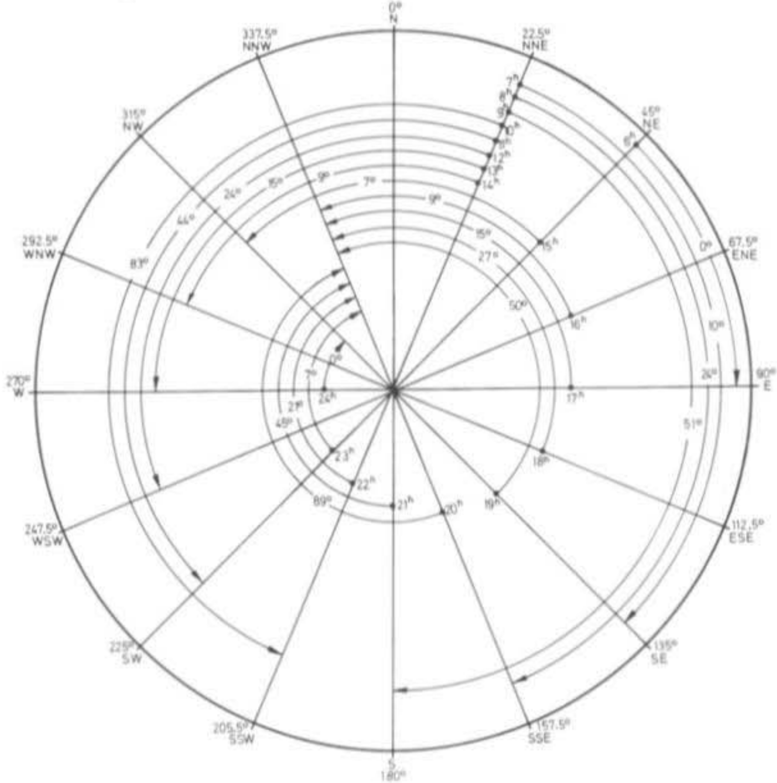
Range:	4000 km	Calculation:	G 3 WPO
Orbit:	115 min.	Design:	OE 6 TH
Pass commencement with respect to equator:	102°	Calculation for C. Europe:	DJ 3 RV
Daily advance of equatorial crossing:	28.75° westerly		



OSCAR 6 ORBITS FOR WESTERN EUROPE (0°)

OSCAR 6 ORBITS FOR CENTRAL EUROPE (11° E)

1 segment = 22.5° = 3.4 min. duration



MOSFET 10 metre receiver AR 10: DM 193,32

Available for 28—30 MHz as IF strip for 2 m converters, or for 26—28 MHz for reception of the citizen band or for use with the 26—28 version of the AC 2 converter below. High sensitivity ensures excellent shortwave reception. The good large-signal behaviour makes it suitable for use as an IF strip for converters. Double superhet with crystal controlled second conversion. 7-stage second IF of 455 kHz. Built-in SSB and CW demodulator. Connections for S-meter, noise limiter/squelch. Prepared for installation of a piezo-ceramic or mechanical filter. A discriminator module is available so that the receiver can be switched to AM, FM, CW or SSB. A matching AF-module AA 1 is also available.

Specifications:

Input impedance:	50 Ohm	Image and spurious	
Sensitivity:	1µ V for 10 dB S/N	suppression:	60 dB
Selectivity:	4.5 kHz (—6 dB)	Operating voltage:	11—15 V/15—22 mA
	12 kHz (—40 dB)	Dimensions:	200 mm x 83 mm x 32 mm

Also extremely suitable for use with the Microwave Modules series of converters. See inner rear cover.

SUMMARY OF AMSAT-OSCAR-B SPACECRAFT SYSTEM

by P. Klein, K 3 JTE

1. AMSAT DEUTSCHLAND REPEATER
(designed by Karl Meinzer, DJ 4 ZC)

Input frequency passband between 432.125 and 432.175 MHz.
Output frequency passband between 145.975 and 145.925 MHz.
Power output (high power mode) is 14 W PEP.
Downlink passband is inverted from uplink passband.
Repeater is 45% efficient using envelope elimination and restoration technique.
Linear Operation -- SSB and CW are preferred modes.
Repeater is commandable to either 3.75 or 14 W PEP output.
Telemetry beacon at 145.980 MHz (200 mW).
Uplink power requires - 300-400 W EIRP.

2. AMSAT TWO-TO-TEN METER REPEATER
(designed by Perry Klein, K 3 JTE)

Input frequency passband between 145.85 and 145.95 MHz.
Output frequency passband between 29.40 and 29.50 MHz.
Power output is 2 W PEP.
Downlink passband is not inverted from uplink passband.
Linear Operation -- SSB and CW are preferred modes.
Telemetry beacon at 29.50 MHz (not same as OSCAR 6).

3. MORSE CODE TELEMETRY ENCODER
(designed by John Goode, W 5 CAY)

24 analog input channels.
Converts each analog value into a two-digit Morse code number or "word".
A third digit precedes the telemetry value and gives the line number in which the word is located.
Format is arranged 4 words per line, six lines per telemetry frame.
Morse code rate is commandable to 10 w.p.m. or 20 w.p.m.

4. TELETYPE TELEMETRY ENCODER
(developed by Peter Hammer, VK 3 ZPI and Edwin Schoell, VK 3 BDS)

60 analog input channels.
Converts each analog channel to a three-digit number transmitted in Baudot code.
Each three-digit value is preceded by its channel number, making a five-digit telemetry word.
The data is arranged 10 words per line by six lines per telemetry frame.
Two lines of status information follow the analog matrix and give the spacecraft time (i.e., time in "counts" from launch, 1 count = 96 minutes).
Output keys 435.1 MHz beacon in FSK: 850-Hz shift; 45.5 Baud: (reversed from U.S. standard). Also keys 145.98 MHz and 29.50 MHz beacons as AFSK, on command.

5. 435.1 MHz BEACON TRANSMITTER
(developed by Larry Kayser, VE 3 QB and Bob Pepper, VE 2 AO)

Beacon output frequency is 435.10 MHz
Power output is 0.4 W at an efficiency of 45%
Beacon is FSK modulated 850-Hz shift.

6. 2304 MHz SMALL BEACON TRANSMITTER
(developed by San Bernardino Microwave Society)

0.1 W at 2304 MHz.
Turned on by command only for 30-min. periods.
CW keyed -- HI followed by 30-sec. carrier. Also keyed with Morse code telemetry on command.

7. CODESTORE -- MESSAGE STORE-AND-FORWARD SYSTEM
(built by John Goode, W 5 CAY)

896 bit memory capacity using COS/MOS shift register memory.
Loaded via command link.
Output code speed is 13 w.p.m.

8. EXPERIMENT CONTROL LOGIC
(designed by Jan King, W 3 GEY)

Selects the spacecraft operating modes.
Protects satellite against excessive battery drain by reducing repeater output power or by shutting if off completely.

9. INPUT SOLAR POWER / BATTERY CHARGE REGULATOR
(developed by Karl Meinzer, DJ 4 ZC and Werner Haas, DJ 5 KQ)

Converts 6.4 V at arrays to 14 V to charge battery or to supply the spacecraft experiments.
Senses overcharge of battery and reduces charging current.
Senses failure of either of the two redundant regulators and switches to the opposite regulator automatically.

AMSAT-OSCAR-B SPACECRAFT

10. A-O-B (to be known as OSCAR 7 after launch) is an international effort now involving four nations. The A-O-B systems developed in each country are as follows:

Germany:	AMSAT Deutschland Repeater, Spacecraft Structure, Battery Charge Regulator, 28 V Power Regulator, Antenna System - DJ 4 ZC, DJ 5 KQ.
Australia:	Two Redundant Command Decoders, Teletype Telemetry Encoder - VK 3 ZPI.
Canada:	435.1 MHz Beacon Transmitter - VE 3 QB and VE 2 AO.
United States:	2 M/10 M Repeater, Morse Code Telemetry Encoder, Experiment Control Logic, Instrumentation Switching Regulator, Solar Panels, Battery - K 3 JTE, W 3 GEY, WA 4 DGU, W 3 DTN, Marie Marr. Codestore - W 5 CAY. S-Band Beacon Transmitter - K 6 HIJ.

AN 8 W SSB TRANSMITTER SUITABLE FOR OPERATION OVER OSCAR 6 AND OSCAR 7

by K. P. Timmann, DJ 9 ZR

INTRODUCTION

The 5 W SSB transmitter DJ 9 RZ 001 was described in (1), and the matching receiver comprising DJ 9 ZR 005 and 006 was described in (2). Due to the increased interest in SSB transmitters since the launch of OSCAR 6, it was decided to update the DJ 9 ZR 001 transmitter.

This article is now to describe an improved version of the original SSB transmitter. It is accommodated on a new printed circuit board of the same dimensions that has been designated DJ 9 ZR 004. The circuit of the new transmitter does not differ greatly from that of the original. However, several improvements have been made which will be the subject of this article.

1. BLOCK DIAGRAM

The block diagram of the DJ 9 ZR 004 transmitter is given in Figure 1. The AF-signal from the microphone is fed via a 1 : 4 microphone transformer (suitable for 200 Ω microphones) to the AF-preamplifiers comprising transistor T 1, where it is amplified and fed to the balanced modulator comprising D 1 and D 2. In contrast to the original circuit, the sideband oscillator signal is not generated by an internal crystal-controlled oscillator but obtained from an external source such as from the DJ 9 ZR 005 receiver (2) or from the DC 6 HL 002 carrier oscillator (3). The sideband oscillator signal is fed via Pt 2 and buffer T 2 to the balanced modulator equipped with the two varactor diodes D 1 and D 2. The 9 MHz DSB-signal at the output of the balanced modulator is fed via buffer T 3 to the crystal filter XF-9A or XF-9B where the unwanted sideband is suppressed. The resulting 9 MHz SSB signal is only slightly amplified in buffer T 4 and fed to the push-pull mixer comprising T 5/T 6, where it is converted to 144 MHz-146 MHz with the aid of a local oscillator signal of

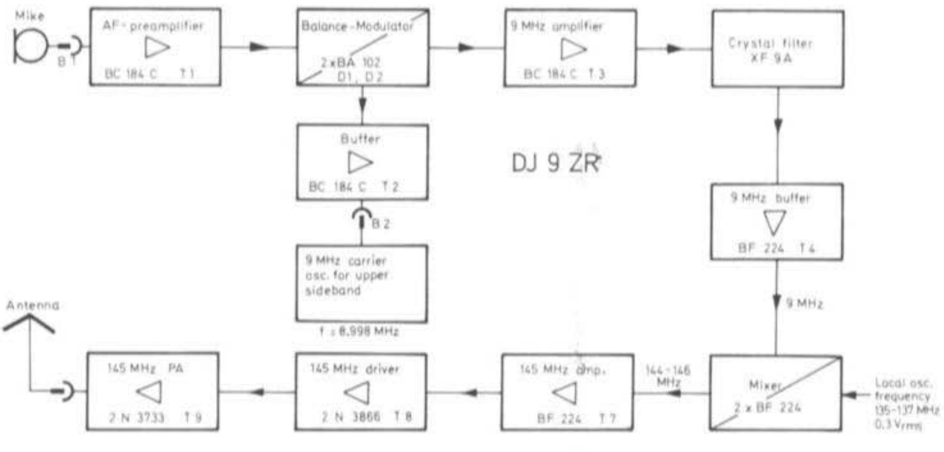


Fig. 1: Block diagram of the DJ 4 ZR 004 SSB transmitter

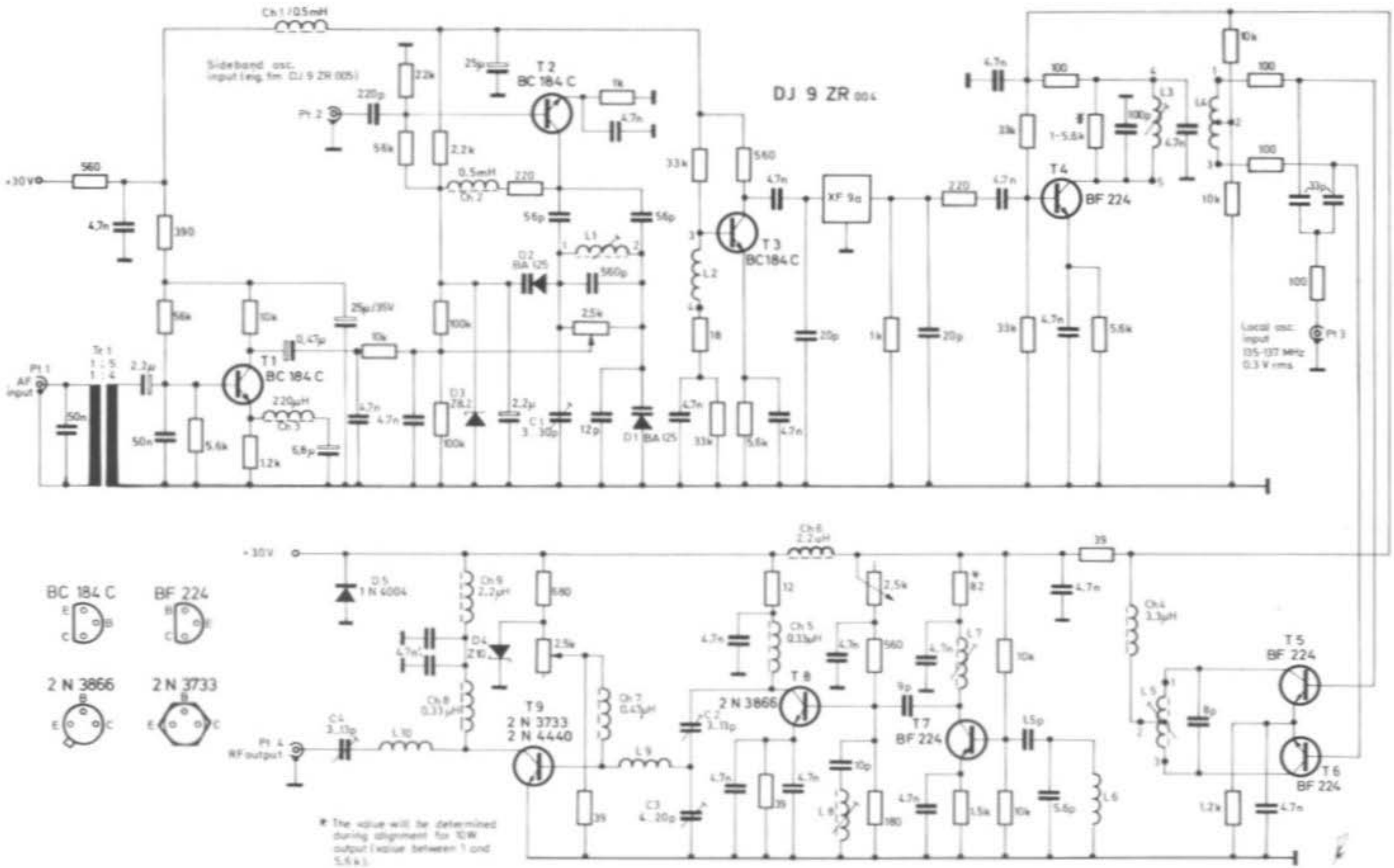


Fig. 2: Circuit diagram of the SSB transmitter

135 MHz to 137 MHz. The local oscillator signal is fed into the transmitter via Pt 3; it can be obtained from the DJ 5 HD 001/002 synthesis oscillator (4), or from the DJ 9 ZR 002 VXO (5) if only a limited frequency range is required. The 9 MHz signal is fed via the buffer transistors T 7 and T 8 to the output transistor T 9 where it is amplified and fed to the output connector Pt 4. The output signal in the order of 8 W can be fed to an antenna, or subsequent linear amplifier.

2. CIRCUIT DETAILS

The circuit diagram of the new SSB transmitter is given in Figure 2. The main improvements to the original circuit, are to be described in the following sections.

2.1. AF-PREAMPLIFIER (T 1)

A microphone transformer (Tr 1) has been provided at the AF-input of the transmitter in order to improve the decoupling and reduce RF-injection. A ratio of 1 : 4 will be suitable for microphones of approximately 200 Ω impedance. A 50 nF capacitor has been provided across the primary of Tr 1 and from the base of T 1 to ground. A 220 μH choke is provided as RF-feedback path between the emitter of transistor T 1 and the 6.8 μF bypass capacitor. The original transistor type 2 N 3707 has been replaced by a BC 184 C.

2.2. BUFFER STAGE (T 2)

The buffer T 2 has been provided on the PC-board instead of the original sideband oscillator. It is also equipped with a BC 184 C transistor. The required sideband signal is now obtained from an external source (see section 1.). Choke Ch. 2 (0.5 mH) has been provided in the base circuit of T 2 in order to increase the impedance.

2.3. 9 MHz BUFFERS (T 3, T 4) AND MIXER (T 5, T 6)

A further BF 184 C is used for T 3. The 220 Ω series resistor in the base circuit of T 3 has been reduced to 18 Ω. The coupling from the crystal filter to the base of transistor T 4 has also been altered. This is now made with the aid of a 220 Ω resistor in series with a 4.7 nF capacitor (previously 680 Ω and 30 pF). The resistor between the collector of T 4 and inductance L 3 has been deleted, and the gain is reduced in the new transmitter by connecting a resistor in parallel with L 3. Due to the fluctuations of the BF 224 transistors, it is necessary for the value to be found experimentally; it will be in the range of 1 kΩ to 5.6 kΩ. The decoupling between the local oscillator input Pt 3 and the push-pull mixer T 5/T 6 has been improved by providing a series resistor of 100 Ω.

2.4. DRIVER AND OUTPUT TRANSISTOR

The resonant circuit comprising inductance L 8 and the 10 pF capacitor connected between the base of transistor T 8 and ground has been re-dimensioned so that it actively assists the operation of the Pi-filter at the input of the output transistor T 9. The resonant circuit comprising L 8 operates in parallel resonance at 145 MHz, but acts as an absorption circuit for the image frequency. The collector circuit of transistor T 8 has also been modified so that the coupling to the output transistor T 9 is now made with the aid of a Pi-filter. The collector of T 8 is connected to choke Ch 5 as well as via two series connected

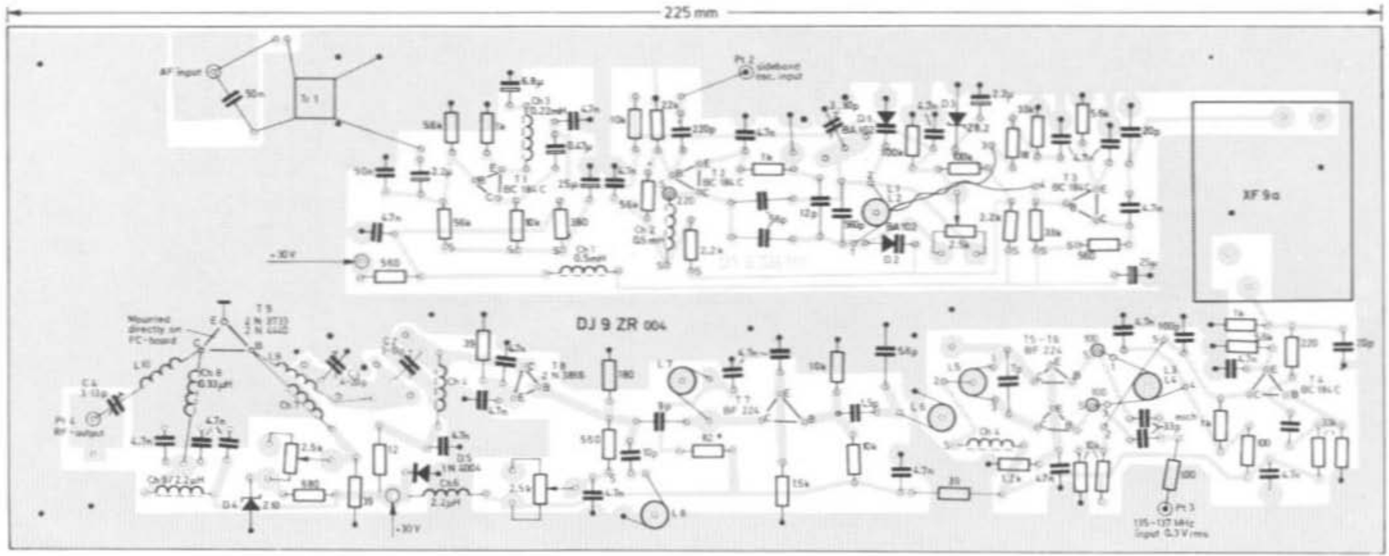


Fig. 3: Printed circuit board DJ 9 ZR 004 with component locations

trimmers to ground. The 145 MHz signal is taken from the interconnection of these two trimmers (C 2, C 3) via inductance L 9 to the base of the output transistor T 9. This Pi-filter at the input of T 9 allows a better power matching to be achieved than was the case with the original transmitter and allows a greater range of input parameters of the output transistor to be aligned. The original 2 N 3375 transistor is replaced by a 2 N 3733 or 2 N 4440 type. However, it has been found that the cheaper 2 N 3632 can also be used and will provide 10-11 W output.

Since the quiescent current of the output transistor will alter with fluctuations of the operating voltage, the base voltage has now been stabilized using the zener diode D 4 so that the base voltage remains constant with operating voltage fluctuations in the range of 21 V and 31 V. The output signal is fed to the output connector Pt 4 via the series circuit comprising L 10 and C 5. Choke Ch 8 forms a parallel-resonant circuit for twice the output frequency together with the output capacitance of T 9. This allows the parametric effects to increase the output power so that an efficiency of approximately 80% is achieved.

3. CONSTRUCTION

The transmitter is accommodated on a printed circuit board with the dimensions 85 mm by 225 mm; this board has been designated DJ 9 ZR 004. The printed circuit board and component locations are given in Figure 3. Figure 4 shows a photograph of one of the prototypes.

All components are now located on the component side of the PC-board. It is now possible for the balanced modulator to be aligned from above.

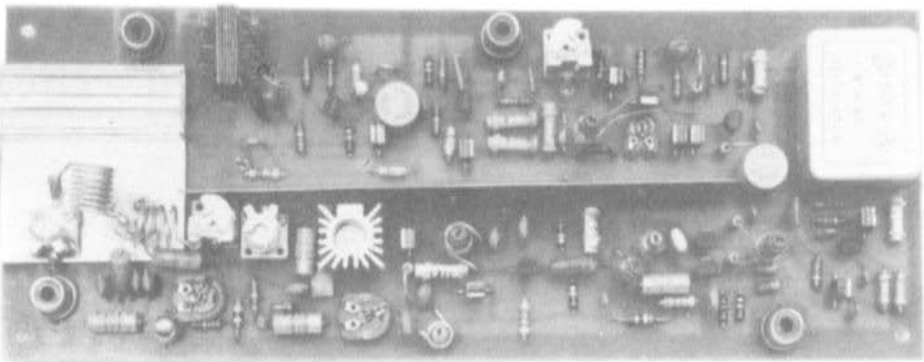


Fig. 4: Photograph of one of the prototypes

3.1. COMPONENTS

Most of the inductances remain as given in (1) for the original transmitter DJ 9 ZR 001. L 9 has been deleted, and L 8 and L 11 (now L 10) have been altered:

L 8: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 4 mm dia. coilformer with SW core.

L 11: 8 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 8 mm dia. former, self-supporting, fairly close-wound.

Chokes:

Ch 1, Ch 2	: 0.5 mH	Ch 5, Ch 8	: 0.33 µH
Ch 3	: 0.2 mH	Ch 6, Ch 9	: 2.2 µH
Ch 4	: 3.3 µH	Ch 7	: 0.47 µH

Trimmer capacitors:

C 1, C 3: 3-30 pF

C 2, C 4: 3-13 pF

Two or four pin types with 7.5 mm spacing.

Semiconductors:

T 1 - T 3: BC 184 C, BC 109 C

T 4 - T 7: BF 244

T 8: 2 N 3866

T 9: 2 N 3733 or 2 N 4440

D 1, D 2: BA 125 or similar 40 pF/2 V varactor diode

D 3: BZY 85/C8V2 or similar 8.2 V zener diode

D 4: BZY 85/C 10 or similar 10 V zener diode

4. MEASURED VALUES

An output power of 8 W was measured at an operating voltage of 27 V, which increased to 10 W to 12 W when the operating voltage was increased to 32 V. If the transmitter is to be used without linear amplifier it would be advisable to place a parallel-resonant circuit aligned to the output frequency across the output connector Pt 4. This allows the harmonic suppression to be increased to a total of 60 dB.

5. AVAILABLE PARTS

There will be a slight delay in obtaining the PC-board etc. for this transmitter. However, supplies will be available approximately four to six weeks after publication of the magazine.

6. REFERENCES

(1) K.P.Timmann: A 5 Watt Transistorized SSB Transmitter for 145 MHz VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 73-82

(2) K.P.Timmann: A 9 MHz IF-AF Portion Using Integrated Circuits VHF COMMUNICATIONS 1 (1969), Edition 3, Pages 136-150

(3) G.Otto: A Portable SSB Transceiver for 144-146 MHz with FM-Attachment VHF COMMUNICATIONS 4 (1972), Edition 1, Pages 2-15

(4) G.Bergmann, M.Striebel: A Synthesis VFO for 144-146 MHz or 135-137 MHz VHF COMMUNICATIONS 3 (1971), Edition 1, Pages 44-55

(5) K.P.Timmann: A Variable Frequency Crystal Oscillator (VXO) VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 87-94.

MINI-MOSFET CONVERTER FOR 2 METRES

by B. Lübbe, DJ 5 XA

The requirements for a miniature 2 m converter with good sensitivity and large signal characteristics led to the development of the described converter. Since the converter was to be inexpensive, it was decided to avoid extensive circuitry. For this reason, the converter is only equipped with one preamplifier stage, which is followed by a bandpass filter and mixer, and complemented either with a single or two-stage local oscillator. The converter can be installed in a TEKO-box 2A and modified for any intermediate frequency that may be required.

1. CHARACTERISTICS

Supply voltage	12 V
Current drain (for 2-stage oscillator)	approx. 16 mA (22 mA)
Overall gain	20 dB
3 dB bandwidth	3 MHz
Noise figure	3 dB to 3.5 dB
Image rejection (IF = 30 MHz)	66 dB

Large-signal behaviour:

A required signal of 50 μV is desensitized by 3 dB when an interference signal of 160 mV is present at a spacing of 500 kHz. The same required signal will obtain modulation from the interference signal (AF voltage at the output of the shortwave receiver equal to the peak value of the noise voltage) when the interference signal obtains a value of 100 mV at a frequency spacing of 500 kHz. The depth of modulation of the interference transmitter amounted to 80%.

Measuring equipment:

μA-multizet (Siemens); Noise Generator SKTU (Rohde u. Schwarz);
Selective Microvoltmeter USVH (Rohde u. Schwarz);
Calibrated Attenuator 0 - 110 dB (Rohde u. Schwarz);
Signal generator for receivers and power-signal generator (Rohde u. Schwarz).

2. CIRCUIT DETAILS

Figure 1 gives the circuit diagram of this simple converter. The preamplifier stage is equipped with a dual-gate MOSFET type 40673. Of course, other MOSFET types can be used such as the 40822 or 40841 especially if the operating points are adjusted for each type. In spite of the operating point stabilization according to (1), the fluctuations were still greater than was permissible for a linear and low-noise preamplifier. For this reason, the drain current was adjusted to 8 mA by exchanging the value of resistor R 1. The given value of 22 kΩ is only an approximate value.

The mixer is also equipped with a MOSFET type 40673. The same is valid for this stage as was previously mentioned for the preamplifier. The mixer stage used is very conventional. With respect to the value of the local oscillator voltage, the author would like to confirm the information given in (1). A voltage of approximately 800 mV (RMS) at gate 2 represents the most favourable compromise between conversion gain and intermodulation rejection. Since the

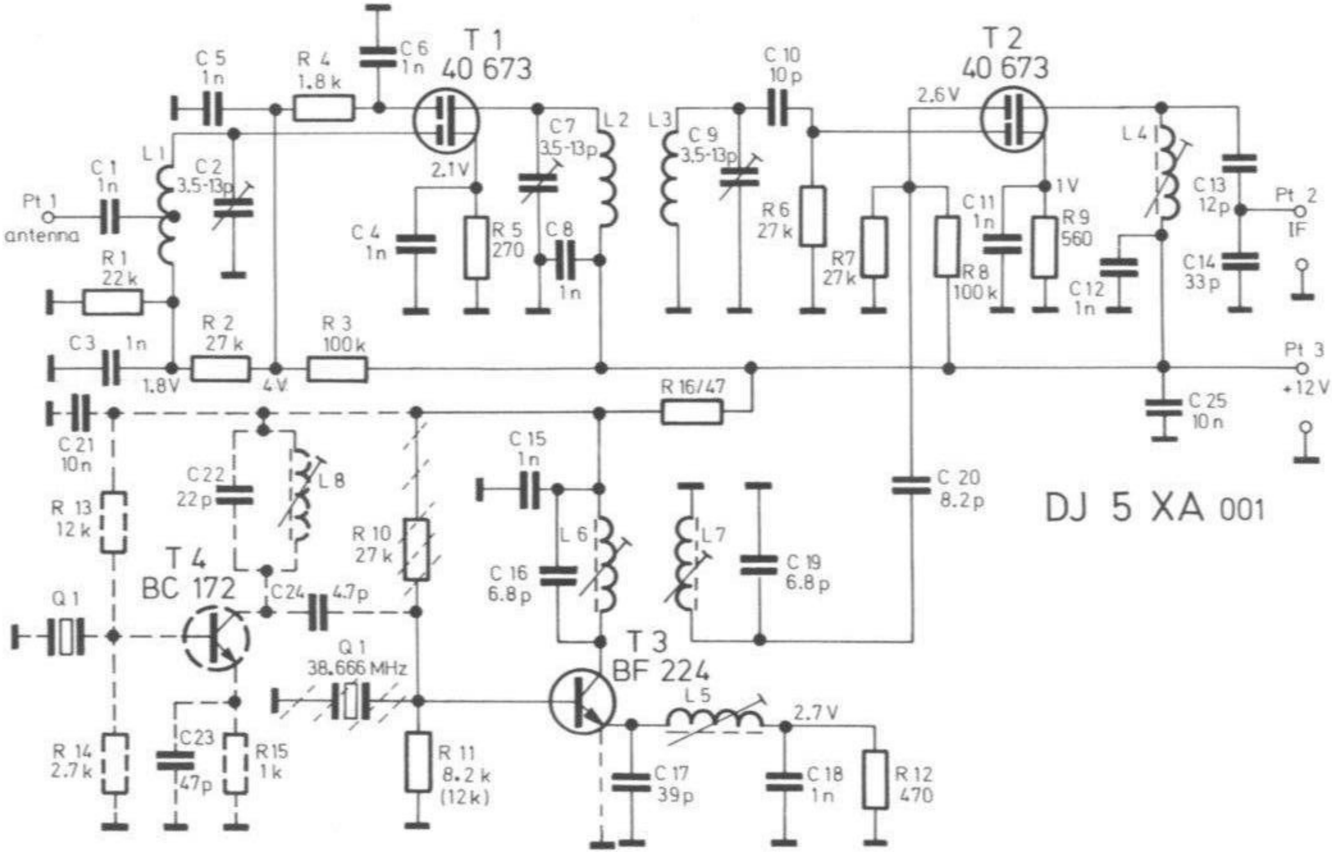


Fig. 1: Circuit diagram of the mini-MOSFET converter for 2 metres

drive of the MOSFET (with the exception of the circuit losses) is practically made at zero power, it is only necessary for the oscillator to provide very low power output. For this reason, the original design of the author only possessed a single-stage in the local oscillator. The resonant circuit at the emitter is tuned to the crystal frequency, where the collector circuit filters out the required harmonic (in our case 116 MHz). The subsequent bandpass filter suppresses the fundamental frequency and unwanted harmonics.

This simple oscillator has, however, some disadvantages: Firstly a certain amount of alignment was required with some crystals until the oscillator commences operation (after which reliable operation was always obtained). Another disadvantage is that the crystal will oscillate approximately 2 kHz to 4 kHz above its given frequency in this circuit which means that conversion errors of 6 kHz to 12 kHz will result. If this is inconvenient, a 2-stage local oscillator circuit can be provided. The additional stage and the modifications to the original oscillator circuit are given as dashed lines in Figure 1. The emitter of transistor T 3 should be grounded and the components usually connected to this point and resistor R 10 should be removed. The crystal is now connected to the base of transistor T 4. The value of resistor R 11 should be increased to 12 kΩ.

This will mean that all the disadvantages of the single-stage local oscillator will be avoided. However, it is necessary for the 2-stage local oscillator to be supplied with a stabilized operating voltage so that no fluctuation of the amplitude occurs.

3. CONSTRUCTION

The construction of the converter is not very critical if the well-proved circuit board is used. Figure 2 shows the component locations and conductor lanes of the PC-board which has been designated DJ 5 XA 001. The dimensions are 65 mm x 52 mm and the board is single-coated. No difficulties were encountered whatsoever with the 5 prototypes that were constructed.

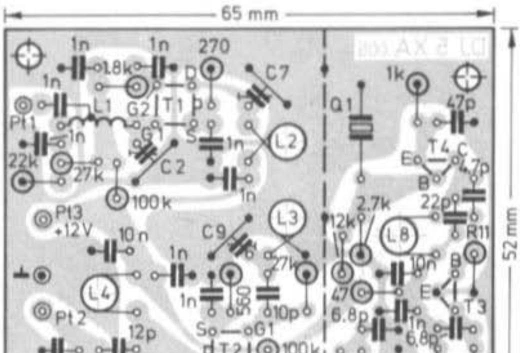


Fig. 2: Component locations and conductor lanes of PC-board DJ 5 XA 001

The printed circuit board is designed for accommodation of the 2-stage local oscillator. However, the single-stage local oscillator can also be equipped without difficulty. The spacing between the components is so small, that they can remain in their planned position and it is only necessary for some conductor lanes to be broken and a few wire bridges to be connected. For example, inductance L 8 should be connected to the emitter of transistor T 3 so that it is then L 5. C 22 will become C 17, and C 21 = C 18 as well as R 13 = R 10. The interconnection from L 8 to the operating voltage is broken and resistor R 12 connected to ground instead.

The most favourable sequence of construction is as follows: Oscillator, mixer, preamplifier after which the operation of each part is tested. This allows a continuous check on the operation of each stage and the currents involved. The screening plate between oscillator and preamplifier/mixer is made from single-coated PC-board material or thin metal plate. It is 20 mm in height and is grounded to the PC-board by thick pieces of wire at the three positions indicated in Figure 2.

Figure 3 shows the author's prototype installed in a box. In this case a HC-18/U crystal was used although there was sufficient room for a HC-6/U crystal if it is directly mounted on the PC-board. This prototype used a single-stage local oscillator. Originally coils without core were used for the 116 MHz bandpass filter and trimmers used for alignment. This can be seen in the photograph. The described version of the converter saves both of these trimmers.

The RF-inputs and IF-outputs are on the same side of the board. This means that low-capacity feedthroughs can be mounted on the same side of the box. In the prototype shown in Figure 3, thin coaxial cables are placed through holes drilled in the box. Of course, the same can be done with the 12 V operating voltage, however, it is preferable to use feedthrough capacitors.

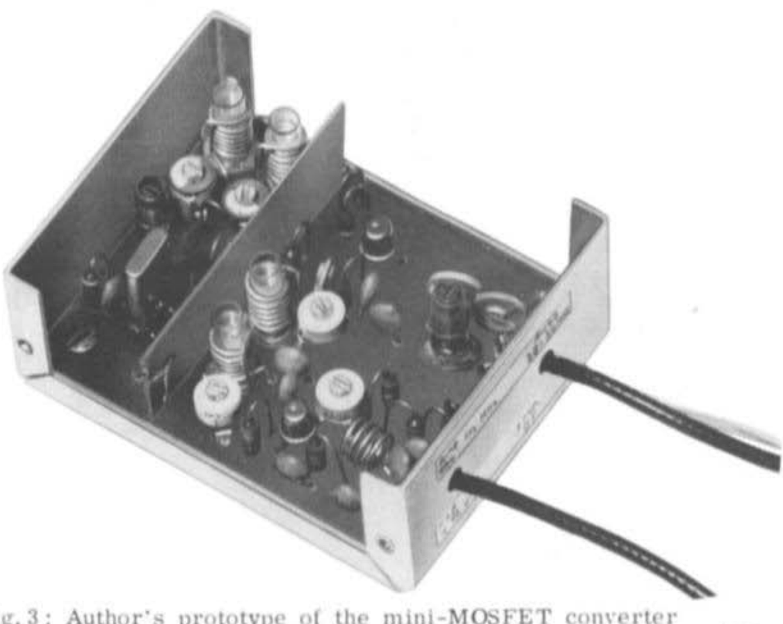


Fig. 3: Author's prototype of the mini-MOSFET converter

3.1. SPECIAL COMPONENTS

T 1, T 2: 40673 (RCA) or similar protected MOSFETs
T 3: BF 224 (TI) or other high-gain UHF transistors
T 4: BC 172 (ITT-Intermetall), BC 183 (TI), BC 108

L 1: 4 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 5 mm former, coil length 5 mm, coil tap 1 turn from the cold end, self-supporting.

L 2, L 3: 6 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire wound on a 5 mm coilformer, coil length 10 mm, without core.

L 4, L 5: 15 turns of 0.25 mm dia. (30 AWG) enamelled copper wire close wound on a 5 mm coilformer, tied together, SW core.

L 6, L 7: 5 or 6 turns of 0.8 mm dia. (20 AWG) silver-plated or enamelled copper wire wound on a 5 mm diameter coilformer, coil length 7 mm, VHF-core.

L 8: 10 turns of 0.4 mm dia. (26 AWG) enamelled-copper wire close wound on a 5 mm coilformer and tied, SW-core.

The four bandpass filter inductances should be wound in the same direction; the cold ends of the coils are at the bottom near the PC-board.

C 2, C 7, C 9: 3.5 - 13 pF ceramic trimmers of 7 mm dia. or plastic foil-trimmers 2 - 22 pF.

A spacing of 5 mm is available for all ceramic disc capacitors and resistors. The resistors are mounted vertically.

Crystal: 38.6667 MHz HC-6/U

1 TEKO Box 2 A

4. ALIGNMENT

After the local oscillator portion has been built up, the operating voltage is connected and the resonant circuit at the collector is coarsely aligned to 116 MHz with the aid of a dipmeter. If other intermediate frequencies are used, the stages should be aligned in a similar manner to the required frequencies. The core of inductance L 5 is now rotated until the oscillator commences oscillation which will be seen when the collector current increases noticeably. The stage can be classed as oscillating reliably when the oscillator commences oscillation immediately after reconnecting the operating voltage and without readjusting the core of L 5. If difficulties are encountered, the value of capacitor C 17 should be varied. Fundamentally speaking, the willingness of the crystal to commence oscillation depends mainly in the Q, that is on the series-resonance impedance of the crystal used. The bandpass filter of the oscillator is aligned for maximum amplitude with the aid of an absorption wavemeter. The collector current should finally correspond to approximately 6 mA. In the case of the 2-stage oscillator, the current drain to the oscillator portion should amount to approximately 12 mA.

Afterthis, the mixer is equipped and brought into operation. The additional current drain to the mixer amounts to approximately 2 mA. A signal of approx. 145 MHz (centre of the band) is now tuned in and the IF circuit, bandpass filter in front of the mixer, and finally the bandpass filter of the local oscillator aligned for maximum reading on the S-meter.

Finally, the preamplifier stage is equipped and connected, and the input connected to a signal generator. As has already been mentioned, the drain current should be adjusted to a value of 8 mA by exchanging R 1. Considerable deviations from the adjusted value are possible. With the aid of a signal at the centre of the band, all circuits should be aligned for maximum S-meter reading. With the power matching at the input made in this manner, the noise figure will correspond to approximately 5 - 6 dB. For noise matching, it is necessary for trimmer C 2 to be increased to a slightly larger value which will soon be found with the aid of a noise generator. The alignment should be checked after installing the converter into the box. If the noise at the band limits is weaker than at the centre of the band, the bandpass filter between the preamplifier and mixer should be aligned at the centre of the band whilst damping the other coils.

5. REFERENCES

(1) D. E. Schmitzer: Adjusting the Operating Points of Field Effect Transistors VHF COMMUNICATIONS 5 (1973), Edition 3, Pages 146-153

LINEAR REPEATER OK Ø A

The linear repeater OK Ø A is the first Czechoslovakian repeater and one of the few linear repeaters that are in operation in Europe. A linear repeater differs from normal FM repeaters in that all operating modes can be used similar to the balloon and satellite carried translators.

The repeater receives a 20 kHz segment of the 2 metre band between 145.090 MHz and 145.110 MHz and transposes this to 145.690 to 145.710 MHz. Identification with the callsign OK Ø A is made in CW at 145.700 MHz which also allows stations to easily find the centre of the output frequency segment. It also indicates when the repeater is open.

It is possible to open the repeater with a 5 s long calling tone of 1750 Hz transmitted in AM or FM. The on time of the repeater is then dependent on the signal strength of the received signal. A signal strength of more than 2 μV will switch on the repeater for 150 s. It is therefore necessary for weaker stations to make relatively short transmissions or monitor the output frequency so that the repeater is not switched off. If the partner station is stronger than 2 μV, then he will be able to reopen the repeater with a calling tone of 1750 Hz.

Specifications:

Input centre frequency:	145.100 MHz
Output centre frequency:	145.700 MHz
Input level for full output:	2 μV
AGC range:	80 dB
Output power:	15 W
Width of transposed band:	20 kHz
Identification:	OK Ø A in CW on 145.700 MHz
Output power of beacon:	0.5 W
Keying speed:	20 wpm
Repetition time:	45 s
Calling tone:	1750 Hz $\pm$50 Hz (AM, FM)
Min. calling tone duration:	5 s
Range:	approx. 200 km
Antennas:	Turnstile, horizontal polarisation
Recommended modes:	AM, CW, FM, SSB

REFERENCES

S. Blazka: Der DX-Umsetzer OK Ø A
CQ DL, November 1973, Pages 665-666

VARIABLE FREQUENCY OSCILLATOR MODULE FOR THE MODULAR RECEIVER SYSTEM

by D. E. Schmitzer, DJ 4 BG

A stable, universal variable frequency oscillator which can be used in the frequency range of approximately 3 MHz to 30 MHz has been designed for use with the modular receiver system described in VHF COMMUNICATIONS in (1) to (5). A printed circuit board has also been developed which allows a VFO to be constructed for a large number of applications, and instructions are given on how this oscillator can be dimensioned and constructed to form oscillators for other applications in this frequency range. Several, well-proved examples are given; however, each application will require a different dimensioning and it is not possible that details for all of the applications to be given here. Anyway, the information given provides a good basis for constructing VFOs for other applications.

1. OSCILLATOR CIRCUIT

A well-known common-base oscillator circuit (Fig. 1a) has been extended by an emitter follower in the feedback link. This results in a high current gain arrangement similar to a differential amplifier which allows a looser coupling of the resonant circuit of the oscillator (Fig. 1b). Due to the use of PNP transistors, the circuit was placed "upside-down" so that the oscillator circuit could be grounded on one side. This is an advantage especially when tuning with varactor diodes.

The circuit was further extended by another transistor to form a cascode circuit to allow low-reactive output coupling of the oscillator voltage. This method was found to be very effective in (2). Figure 1c shows the basic principle of the arrangement.

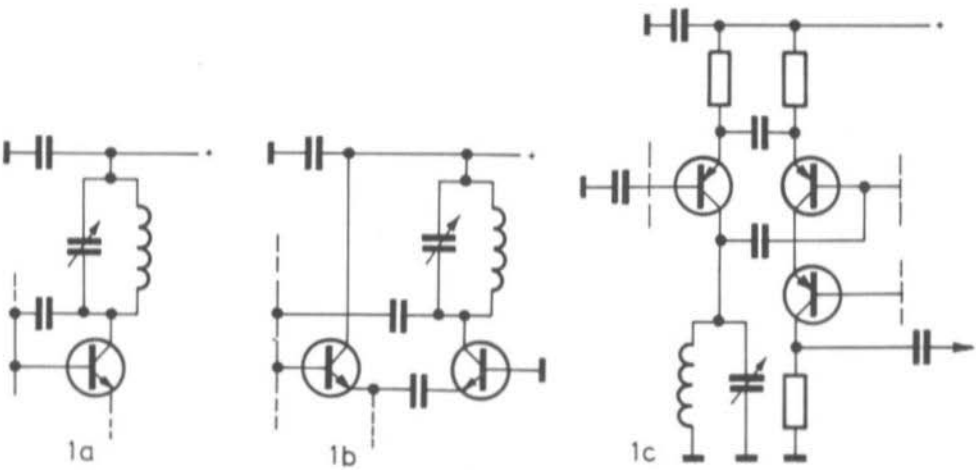


Fig. 1: Development of the oscillator circuit

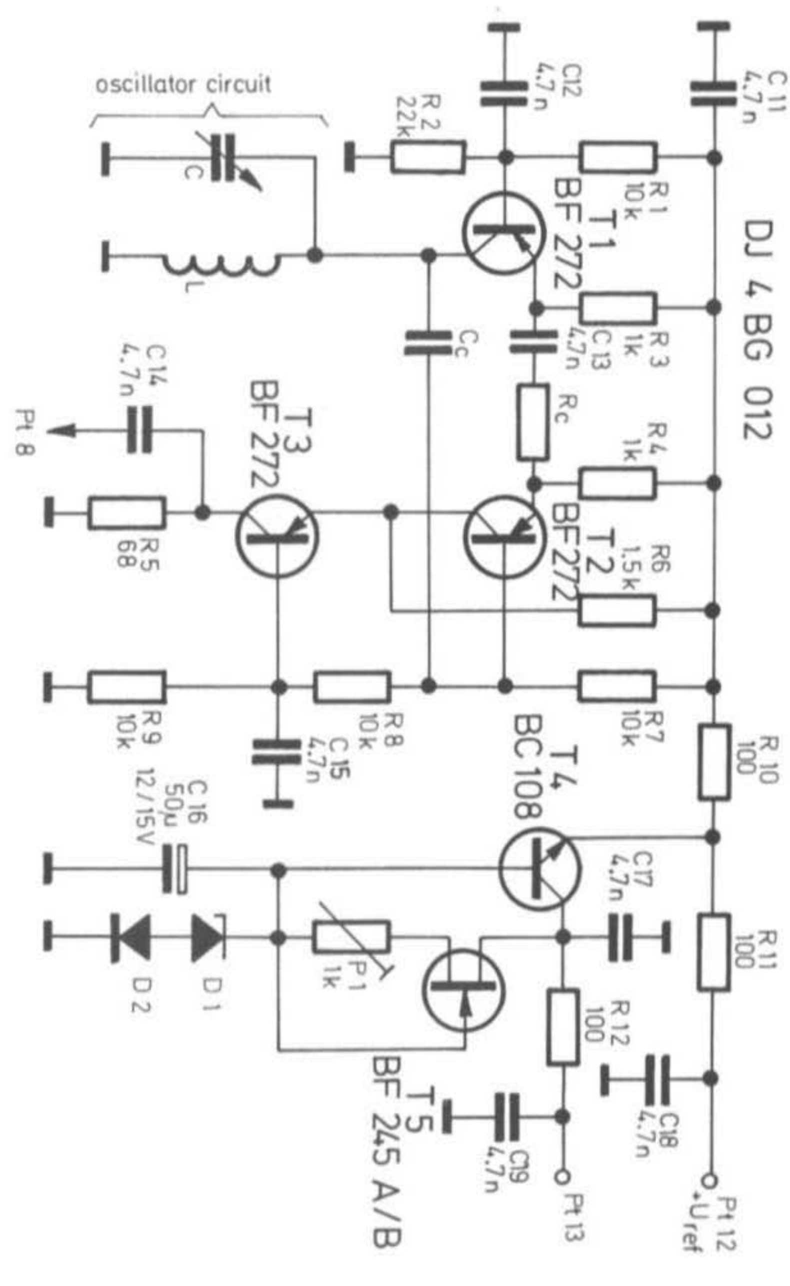


Fig. 2: Complete circuit diagram of the variable frequency oscillator module

The transistors used in the circuit should possess a low capacitance and the highest transit frequency possible. These demands are satisfied by the transistor type BF 272.

Figure 2 gives the complete circuit diagram. The dimensioning of the resonant circuit and stabilizing circuit is to be described in the following sections. Components without values should be dimensioned according to each application by following the given information.

1.1. RESONANT CIRCUIT OF THE OSCILLATOR

The oscillator circuit can be tuned both with the aid of a variable capacitor or varactor diode. The circuit given in Figure 3 shows several possibilities. The maximum possible complement is given in all cases for the printed circuit board described in section 3. In order to have a free hand in the combination of capacitors with various temperature coefficients for temperature compensation of the oscillator, space has been provided for up to six individual capacitors of various dimensions. In addition to this, a possibility is shown by which the tuning arrangement uses a variable capacitor for coarse tuning and fine alignment with a varactor diode. This can be used, for instance, with advantage for frequency modulation, fine tuning during transceive operation or for frequency-shift keying with CW (so that the oscillator can continue to oscillate in the pauses and to obtain an especially stable frequency) or for RTTY applications.

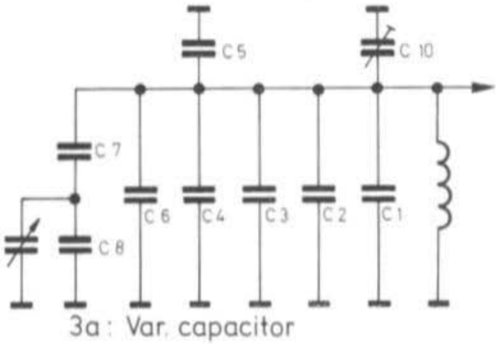
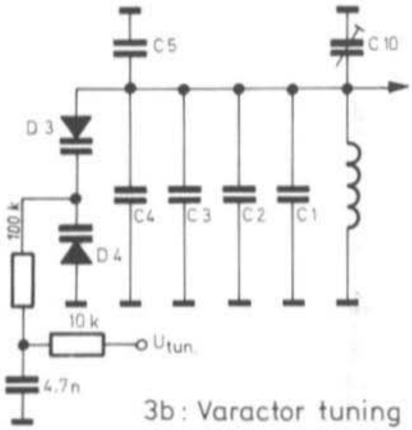
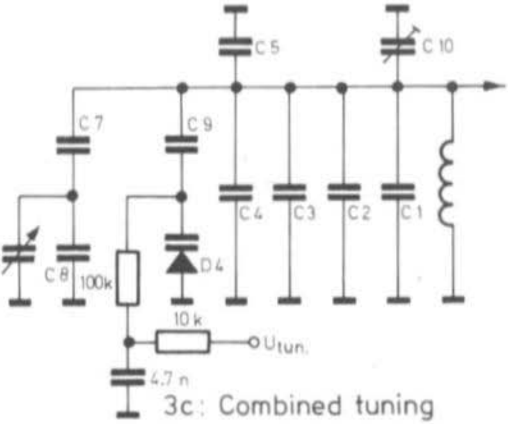


Fig. 3 :
Possibilities of tuning the oscillator circuit





Of course, only the highest quality components should be used when constructing a variable frequency oscillator. Ceramic capacitors with suitable temperature coefficients and/or mica capacitors should be used for the resonant circuit. Styroflex or metal-paper capacitors cannot be used when high demands are placed on the frequency stability. Ceramic formers with vapourized windings comprising a silver layer are ideal, but it is very difficult to find the required value of inductivity.

However, good inductances can be made by winding thick enamelled or non-enamelled silver-plated copper wire firmly onto a matching glass or ceramic rod and gluing it into place with a dual-component adhesive (e.g. UHU-plus). After complete hardening, if possible at increased temperature, the individual windings will be fixed firmly in place. The temperature coefficient of such an inductance will only be dependent on the material of the coilformer if only a minimum of adhesive has been used. A temperature coefficient of approximately $+100 \times 10^{-6}/{}^{o}C$ was exhibited when using a glass tube. However, this is only an approximate value since individual materials can differ. An extensive description of measures for compensating the temperature response are too extensive to be explained here.

Fundamentally speaking, a low LC-ratio should be used when dimensioning the resonant circuit components. There are two reasons for this: Firstly, a large capacitance of the circuit ensures a correspondingly lower effect of the voltage and temperature dependent capacitances of the transistors on the frequency. This means less drift on switching on and therefore less chirp on keying the oscillator, as well as a lower dependance of the frequency on fluctuation of the operating voltage. Secondly, the inductance will require less turns so that it can be constructed from thicker wire and thus be more stable mechanically. It is then possible for two wires to be wound together in parallel after which one is removed. The inductance is then provided with one diameter of the wire spacing between turns. Due to the reduced capacitance of the winding and reduced eddy-current losses, high Q values can be obtained in spite of the use of lower inductance values.

2. VOLTAGE STABILIZATION

In order to keep the oscillator module as independent as possible from external influences, a simple voltage-stabilizing circuit is provided for the operating voltage. It exhibits a very high filter factor. The circuit comprising pass transistor, zener diode and constant voltage source equipped with a field effect transistor exhibits, in spite of its simplicity, a stabilizing factor with respect to operating voltage fluctuations of more than 1000. Load variations need not be taken into consideration if the oscillator runs continuously. The stabilized voltage is also fed to a connection point on the PC-board where it can be taken for feeding the tuning diodes. Since zener diodes exhibit different TC-values according to the zener voltage, various combinations of zener diode and compensating diodes can be provided in order to obtain the required temperature dependance. They can be used for temperature compensation when using varactor tuning. Figure 4 gives various combinations. The current flowing through the zener diode can be adjusted with the trimmer potentiometer (for example to 5 mA).

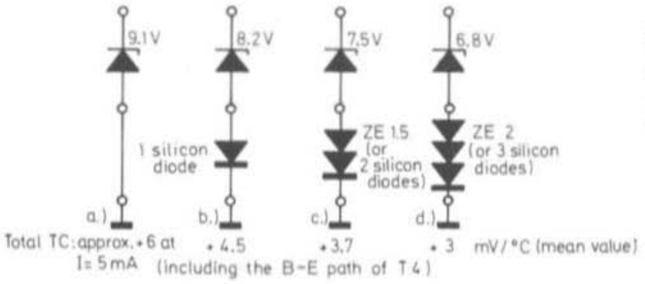


Fig. 4: Possible combinations of D 1 and D 2 for adjustment of the TC of U_{ref}

3. PRINTED CIRCUIT BOARD DJ 4 BG 012

Several special features of printed circuit board DJ 4 BG 012 have been described in the previous sections. The dimensions of this board are 65 mm x 90 mm matching the other boards of the modular receiver system. It is single-coated in order to keep circuit capacitances as low as possible (Fig. 5). Further details regarding component locations of the oscillator circuit are given in Figure 6.

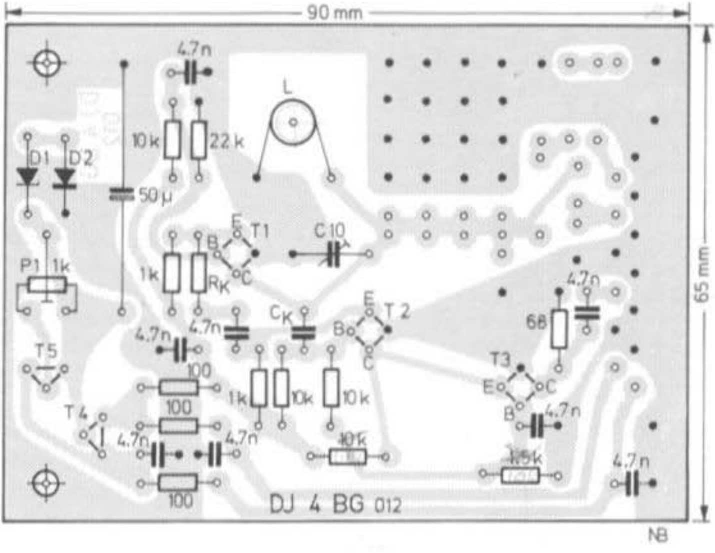


Fig. 5: Printed circuit board DJ 4 BG 012 with component locations (not oscillator circuit)

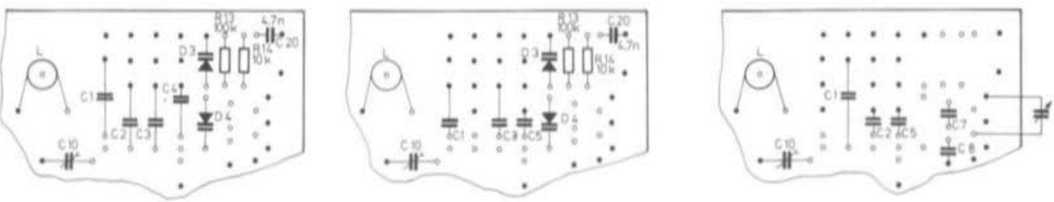


Fig. 6 a... c: Components of the oscillator circuits as given in table 2

3.1. COMPONENT DETAILS

T 1 to T 3: BF 272 (SGS) or similar PNP-UHF silicon transistor.
Data of BF 272: $f_{Tmin.}$ 700 MHz; F max. 6 dB at 800 MHz:
V_{pe}typ. 13 dB/800 MHz and capacitance typ. 0.3 pF when used in a common emitter circuit or 0.05 pF in common base circuit.

T 4: BC 108 B, BC 413 B (Siemens) or similar silicon NPN AF transistor with B min. 100.

T 5: BF 245 B (TI) or similar FET

D 1: Silicon zener diode, selected as shown in Fig. 4

D 2: 1 N 914, 1 N 4148 or similar silicon diode

D 3, D 4: Varactor diode, according to application, see table 2.

Trimmer: Air-spaced trimmers for PC-mounting with 2 connection pins. Capacitance according to application, see table 2.

Bypass capacitors: 4.7 nF ceramic disc types

Resistors: For 10 mm spacing.

4. MECHANICAL CONSTRUCTION

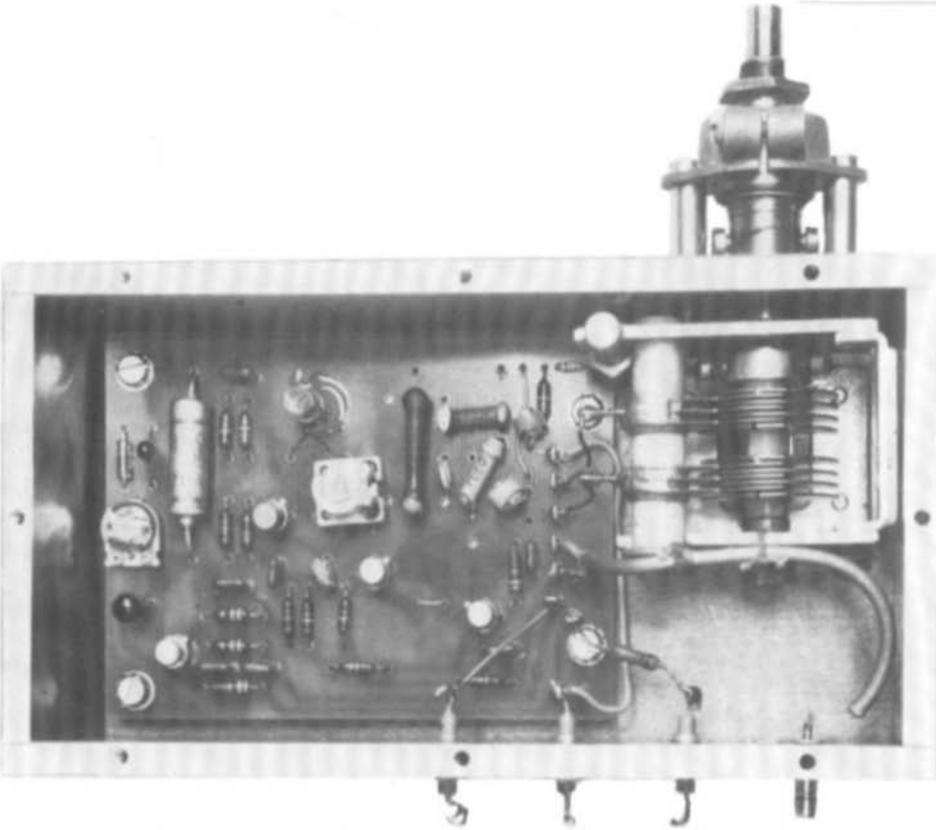
The stability of an oscillator is directly dependent on the quality of the mechanical construction. Of course, it is possible to mount PC-board DJ 4 BG 012 in a TEKO box size 3 A or 3 B and the results will be very good. However, if the dimensions, weight and other mechanical possibilities allow it, a more stable case should be used. The author used a rectangular box of suitable size that was constructed from 5 mm thick aluminum plates. It is true that this requires a certain amount of extra work and care on tapping the holes, but such a case with thick walls provides excellent results with rapid temperature fluctuations and operates as a "cold crystal oven". Further details are given in the following table 1:

Material: Aluminum plate 5 mm thick	Small case for varactor tuning	Large case for capacitor tuning
Short side	80 mm x 25 mm	70 mm x 40 mm
Long side	100 mm x 25 mm	150 mm x 40 mm
Top and bottom plates	80 mm x 110 mm	150 mm x 80 mm

The photographs of the two prototypes (Fig. 7 and 8) show further details of the construction. The DC-voltages are fed in via feed-through capacitors and the output voltage of the oscillator is fed out via a low-capacitance feedthrough or miniature connector. It is advisable for a coaxial connector to be used due to the extra screening.

5. PRACTICAL EXAMPLES

Several prototype oscillators were constructed using PC-board DJ 4 BG 012. They were dimensioned for the various frequency and tuning ranges and use partly varactor and partly variable capacitor tuning. Even if the selected ranges cannot be directly used for the special application in question, they do provide proved dimensionings and further details.



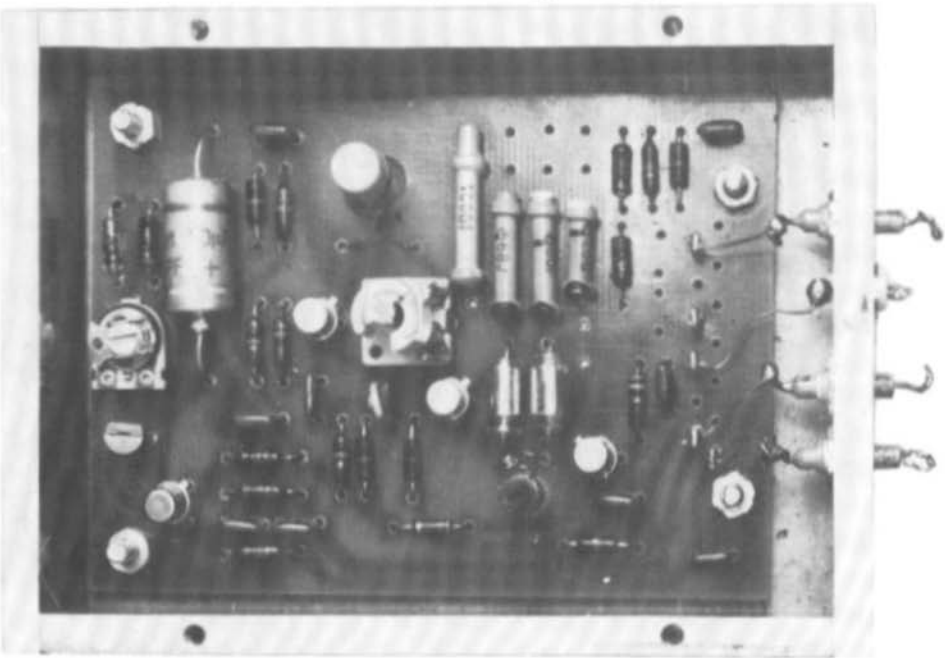


Fig. 7 and 8: A prototype of the oscillator for variable capacitor and varactor tuning

Table 2 gives a list of the components that have been proved in various prototypes:

	5.0 MHz to 5.1 MHz	19.0 MHz to 21.0 MHz	14.0 MHz to 14.1 MHz
C 1	100 pF N 150	35 pF N 150	100 pF N 150
C 2	68 pF NPO	-	6.8 pF N 150
C 3	60 pF NPO	25 pF N 150	-
C 4	25 pF NPO	-	-
C 5	-	2 pF N 033	2 pF N 150
C 6	-	-	-
C 7	-	-	35 pF N 033
C 8	-	-	40 pF N 150
C 9	-	-	-
C 10	13 pF air spaced trimmer	5 pF air spaced trimmer	5 pF air spaced trimmer
C_c	6.8 pF	5 pF	10 pF
R_c	68 Ω	68 Ω	220 Ω
D 3	BA 111 (ITT)	BA 112 (ITT)	variable capacitor
D 4	BA 124, BA 150 (Tfk)	MV 1650 (Motorola) or 2 x BA 124/65, BA 150/65	2 - 14 pF
L	35 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound on a 6 mm dia. ceramic rod L = approx. 10 μH	17 turns of 0.5 mm dia. (24 AWG) enamelled copper wire wound on a 5 mm dia. glass rod L = approx. 0.58 μH	19 turns of 0.35 mm dia. (27 AWG) enamelled copper wire wound on a 5 mm dia. glass rod L = approx. 1 μH
	Best oscillator circuit: Fig. 6a	Fig. 6b	Fig. 6c

Figure 9 shows the frequency of the 5 MHz oscillator as a function of the voltage. Figure 10 gives the temperature dependance of the same oscillator.

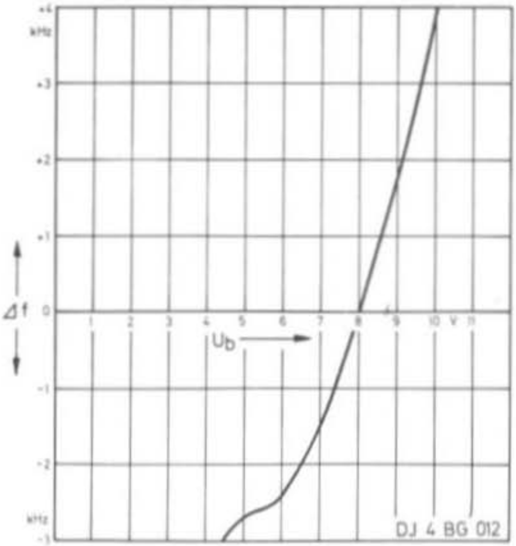


Fig. 9 : Voltage dependence of the 5 MHz oscillator frequency (measured without stabilizing circuit)

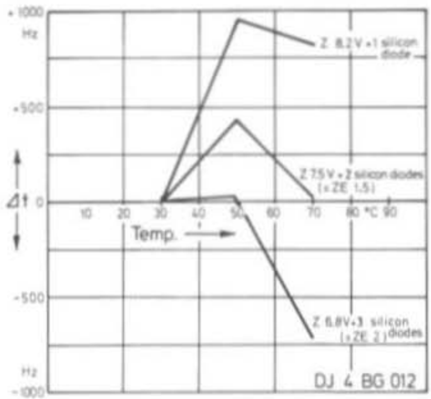


Fig. 10 : Temperature response of the 5 MHz oscillator

6. FURTHER DETAILS

The resonant circuit between transistors T 2 and T 3 that can be seen in the photograph of the prototypes allows an especially low-harmonic output voltage to be generated for certain applications. However, since it has an effect on the oscillator frequency, it was deleted at a later date. When using this variable frequency oscillator together with the shortwave converter module DJ 4 BG 011 (5), the oscillator voltage is filtered by a resonant circuit at the input. A similar method should also be used when using the VFO for other applications.

7. REFERENCES

(1) D. E. Schmitzer: A Modular Receive System
VHF COMMUNICATIONS 3 (1972), Edition 2, Pages 110-114

(2) D. E. Schmitzer: A Crystal Oscillator Module with 3 Independent Oscillators
VHF COMMUNICATIONS 4 (1972), Edition 3, Pages 175-179

(3) D. E. Schmitzer: A 9 MHz IF Module for Frequency Modulation
VHF COMMUNICATIONS 4 (1972), Edition 1, Pages 40-45

(4) D. E. Schmitzer: A Modular 6-Channel FM Receiver
VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 33-36

(5) D. E. Schmitzer: A Shortwave Receiver Module for use with VHF Converters or for Direct Reception
VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 24-32.

TV-PATTERN GENERATOR

Additional Board for Grid and Dot Pattern

by K. Wilk, DC 6 YF

An additional board has been developed to extend the TV-pattern generator described in (1) and (2) for provision of a grid and dot pattern. This article is to describe this module as well as the interconnections between the various modules that have been described, in order to form a complete TV-pattern generator with VHF output. This unit can be used effectively for TV service and for use as the source of a video signal for ATV transmissions.

The module DC 6 YF 004 allows the previously available patterns to be extended to provide a square grid and dot pattern (Fig. 1 and 2).

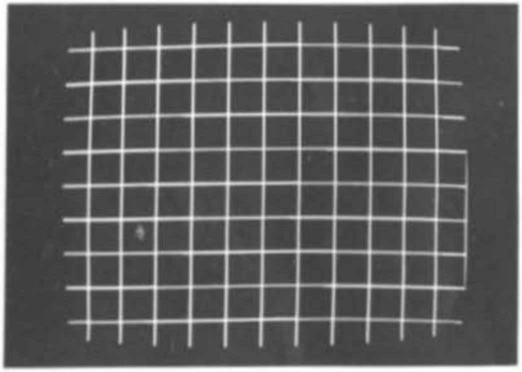


Fig. 1: Grid pattern

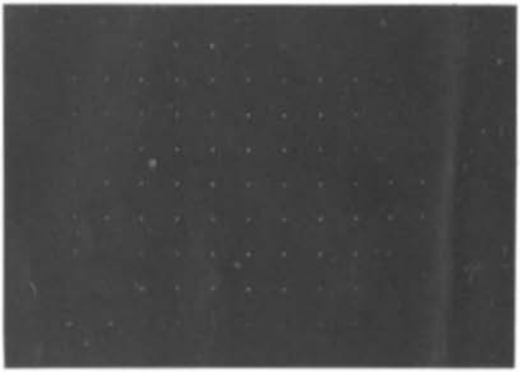


Fig. 2: Dot pattern

This is obtained in a relatively simple manner. The pattern is generated from narrow vertical and horizontal lines. The grid pattern is provided as a white grid of narrow lines on a black background, whereas the dot pattern is formed as a black grid of wide black lines. For this reason, it is necessary for the line signals to be available as positive and negative (black and white) signals.

1. GENERATION OF THE LINE SIGNALS

The signals for generation of the narrow vertical and horizontal lines are generated within the module DC 6 YF 004.

1.1. VERTICAL LINES

The vertical lines are formed from the 250 kHz signal obtained from the pulse centre DC 6 YF 001. A flip-flop is fed with a positive going pulse, and the resulting pulses form the vertical lines.

1.2. HORIZONTAL LINES

The pulses for the horizontal lines cannot be obtained from any signal provided by the pulse centre since no frequency is provided that is suitable for generating a square grid. For this reason, an oscillator is provided which provides

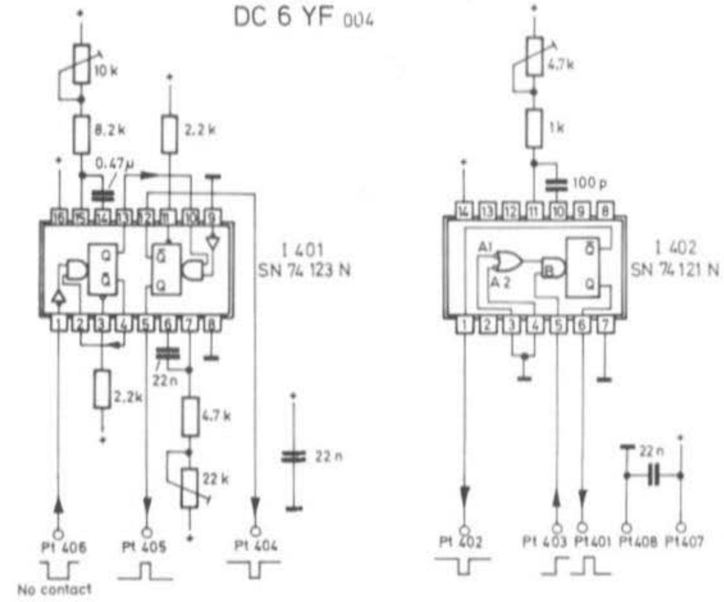


Fig. 3 : Line generator circuit DC 6 YF 004

sufficient stability at the low pulse frequency for the horizontal lines. In order to ensure that a stationary image results, it is necessary for the oscillator to be synchronized to the vertical frequency.

1.3. LINE GENERATOR MODULE DC 6 YF 004

The circuit diagram of module DC 6 YF 004 is given in Figure 3. The integrated circuit I 401 (SN 74 123) contains two flip-flops that are interconnected to form a generator circuit.

The first stage is provided with a feedback link and operates as a self-excited oscillator; the pulse frequency of the output signal is determined by the time constant of the feedback links. The second stage is used as a flip-flop and is driven by pulses obtained from the oscillator. The pulse width is determined by the delay time. The additional components of both stages are selected so that the delay times can be adjusted to be in the order of 1.4 ms and 2.4 ms, or between 50 µs and 100 µs. These times correspond to the required on/off time ratio. This means that the spacing and width of the horizontal lines can be adjusted with the aid of trimmer potentiometers.

The horizontal blanking signal is fed to pin 1 of the integrated circuit SN 74 123 so that the generator commences operation with a defined phase position according to this signal.

The integrated circuit I 402 (SN 74 121) forms a flip-flop whose switching time is adjustable between 0.1 μs and 0.3 μs according to the width of the vertical lines.

All output signals are available as positive and negative going pulses (Q and $\bar{Q}$-outputs).

2. MECHANICAL CONSTRUCTION

The circuit is accommodated on the PC-board DC 6 YF 004 (Fig. 4). A component location plan is printed onto this single-coated PC-board which also gives the values of the required trimmer potentiometers. The dimensions are the same as that of the other boards DC 6 YF 001-003.

The connections are as follows:

Pt 401:	Output 0-1-0	Vertical lines
Pt 402:	Output 1-0-1	
Pt 403:	Input 25 kHz	
Pt 404:	Output 1-0-1	Horizontal lines
Pt 405:	Output 0-1-0	
Pt 406:	Input 50 Hz	
Pt 407:	+ 5 V	Operating voltage
Pt 408:	0 V	

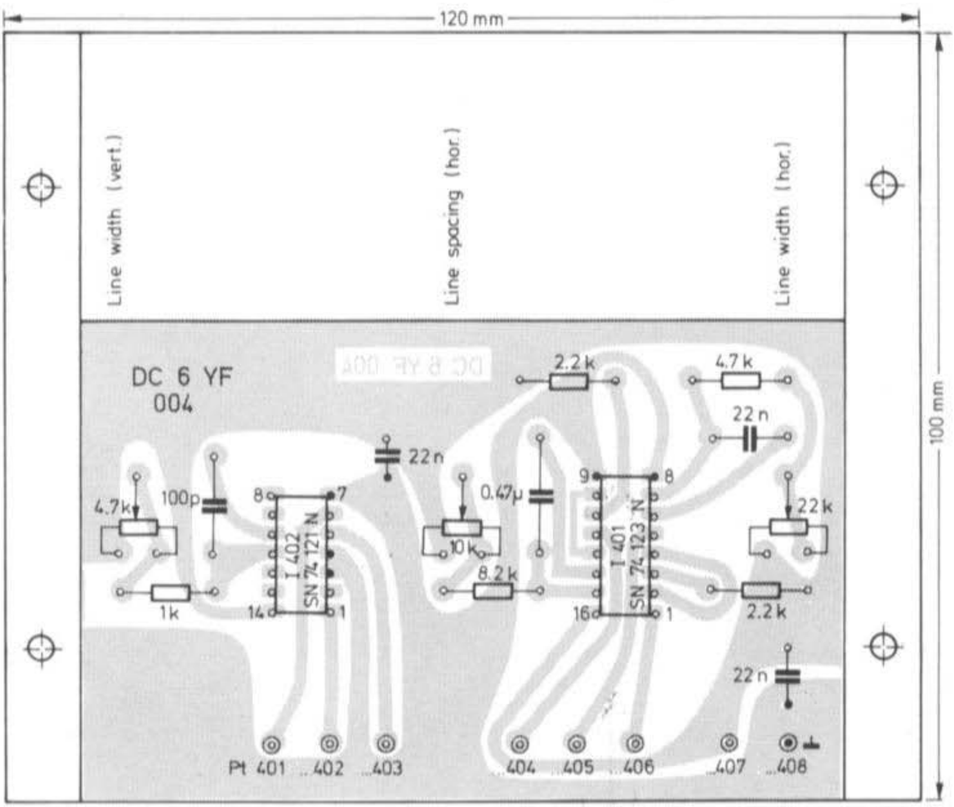


Fig. 4 : PC-board DC 6 YF 004

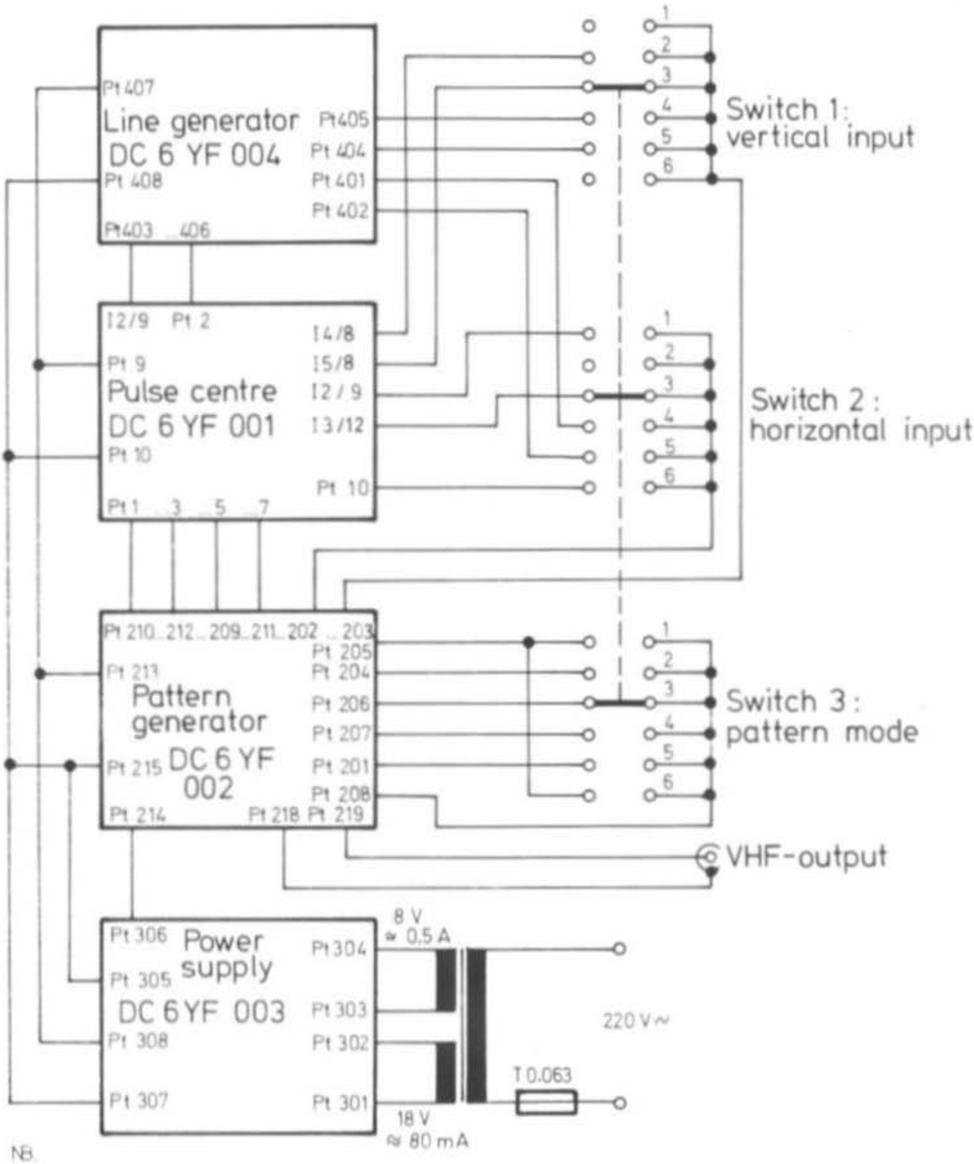


Fig. 5: Block diagram of a TV-pattern generator

2.1. SPECIAL COMPONENTS FOR DC 6 YF 004

I 401: SN 74 123 N
I 402: SN 74 121 N

3 trimmer potentiometers, hor. mounting, spacing 7.5 mm/5 mm
2 ceramic disc capacitors 22 nF, spacing 5 mm
1 styroflex capacitor 100 pF, length max. 10 mm
1 plastic foil capacitor 0.47 μF, spacing 15 mm
1 plastic capacitor 0.022 μF, spacing 10 mm

A spacing of 12.5 mm is available for all resistors.

3. CONSTRUCTION OF A COMPLETE PATTERN GENERATOR

The following TV patterns are recommended from the large number of possibilities:

1. Vertical bars	3. Chessboard	5. Dot
2. Horizontal bars	4. Grid	6. White

These six patterns should be selectable with the aid of a single switch. This results in the arrangement shown in Figure 5. Switches 1, 2, and 3 are the three wafers of the pattern selector switch. The patterns 1, 2 and 3 are given in Figures 13 to 18 in (2).

4. INTERCONNECTION AND ADJUSTMENT

Module DC 6 YF 004 is aligned after interconnecting the complete TV-pattern generator. The interconnection of the other modules of the TV-pattern generator (DC 6 YF 001 to 003) was described in (1) and (2).

In position 4 of the pattern mode switch (grid pattern) the required line width is adjusted with the aid of the 4.7 kΩ trimmer potentiometer. The width of the horizontal lines is selected with the aid of the 22 kΩ trimmer potentiometer.

The spacing between the horizontal lines should correspond to that of the vertical lines in order to provide square rectangles. The easiest method of adjustment (with the 10 kΩ trimmer potentiometer) is provided when the grid pattern is displayed on a TV receiver whose picture has been adjusted correctly according to a known test card. In this case, it is assumed that the aspect ratio and linearity is correct and that the grid pattern will also possess the correct horizontal and vertical relationship.

5. REFERENCES

(1) K. Wilk: An ATV Pulse Centre
VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 54-59

(2) K. Wilk: TV-Pattern Generator
VHF COMMUNICATIONS 5 (1973), Edition 3, Pages 177-189

MATERIAL PRICE LIST OF EQUIPMENT
described in Edition 4/73 of VHF COMMUNICATIONS

DK 1 OF 007	9 MHz EXCITER	Ed. 4/73
PC-board	DK 1 OF 007 (with printed plan)	DM 8.50
Minikit	DK 1 OF 007 (1 ceramic filter, 1 min. ferrite choke, 3 resonant circuits, 1 coil set)	DM 18.50
Semiconductors	DK 1 OF 007 (3 transistors, 3 diodes)	DM 28.00
Kit	DK 1 OF 007 with the above parts	DM 54.00
DK 1 OF 008	MICROPHONE PREAMPLIFIER	Ed. 4/73
PC-board	DK 1 OF 008 (with printed plan)	DM 8.50
Semiconductors	DK 1 OF 008 (1 IC, 2 transistors, 3 diodes)	DM 16.70
Kit	DK 1 OF 008 with the above parts	DM 25.00
DK 1 OF 009	1750 Hz CALLING-TONE GENERATOR	Ed. 4/73
PC-board	DK 1 OF 009 (with printed plan)	DM 7.50
Minikit	DK 1 OF 009 (1 transistor, 1 potted core)	DM 12.50
Kit	DK 1 OF 009 with the above parts	DM 20.00
DK 1 OF 010	PUSH-PULL TRANSMIT MIXER	Ed. 4/73
PC-board	DK 1 OF 010 (with printed plan)	DM 10.00
Minikit	DK 1 OF 010 (4 transistors, 6 trimmers, 1 potted core)	DM 31.00
Kit	DK 1 OF 010 with the above parts	DM 40.00
PRICE REDUCTION		
Kits	DK 1 OF 001-A complete with listed parts	DM 420.00
PC-boards	DK 1 OF 001-006	DM 40.00
DK 1 PN 004	AF AMPLIFIER AND ACTIVE CW FILTER	Ed. 4/73
PC-board	DK 1 PN 004 (with printed plan)	DM 11.00
Semiconductor	DK 1 PN 004 (3 ICs including 1 Plessey IC)	DM 32.00
Minikit	DK 1 PN 004 (6 tantalium caps., 3 electrolytics, 1 styroflex cap., 13 pin connectors, 1 TEKO box 3 A)	DM 24.30
Kit	DK 1 PN 004 with the above parts	DM 66.00
DK 1 PN 005	9 MHz INPUT MODULE AND FM-PORTION	Ed. 4/73
PC-board	DK 1 PN 005 (double coated, through contacts)	DM 28.00
Semiconductors	DK 1 PN 005 (2 transistors, 2 ICs incl. 1 Plessey)	DM 58.50
Minikit 1	DK 1 PN 005 (1 toroid core, 1 trimmer cap., 3 tantalium caps., 21 bypass caps., 1 styroflex cap, 2 coil kits, 1 TEKO case 3 A, 13 pin connectors)	DM 41.50
Minikit 2	DK 1 PN 005 (6 ceramic caps., 22 resistors, coil wire)	DM 11.20
Crystal filter	XF-9E	DM 137.00
Kit	DK 1 PN 005 with the above parts	DM 275.00
DJ 9 ZR 004	SSB TRANSMITTER	Ed. 4/73
PC-board	DJ 9 ZR 004 (with printed plan)	DM 15.00
Minikit	DJ 9 ZR 004 (5 coilformers with core, 4 trimmers)	DM 17.50
Crystal filter	XF-9A with both sideband crystals	DM 106.00
Kit	DJ 9 ZR 004 with the above parts	DM 137.50
DJ 5 XA 001	MINI-MOSFET CONVERTER FOR 2 m	Ed. 4/73
PC-board	DJ 5 XA 001 (with printed plan)	DM 6.50
Semiconductors	DJ 5 XA 001 (4 transistors)	DM 19.30
Minikit 1	DJ 5 XA 001 (3 trimmer caps., 7 coilformers, 5 cores, 12 bypass caps., 1 TEKO case 2 A)	DM 17.10
Minikit 2	DJ 5 XA 001 (9 ceramic caps., 14 resistors, coil wire)	DM 11.00
Crystal	38.6667 MHz (HC-6/U)	DM 13.70
Kit	DJ 5 XA 001 complete with the above parts	DM 67.00
DJ 4 BG 012	UNIVERSAL VFO MODULE	Ed. 4/73
PC-board	DJ 4 BG 012 (with printed plan)	DM 10.00
Semiconductors	DJ 4 BG 012 (5 transistors)	DM 35.00
Kit	DJ 4 BG 012 with the above parts	DM 44.00
DC 6 YF 004	TV PATTERN GENERATOR (Module 4)	Ed. 4/73
PC-board	DC 6 YF 004 (with printed plan)	DM 14.00
Minikit	DC 6 YF 004 (2 ICs, 3 capacitors)	DM 18.30
Kit	DC 6 YF 004 with the above parts	DM 32.00

CRYSTAL FILTERS - FILTER CRYSTALS - OSCILLATOR CRYSTALS

SYNONYMOUS for QUALITY and ADVANCED TECHNOLOGY

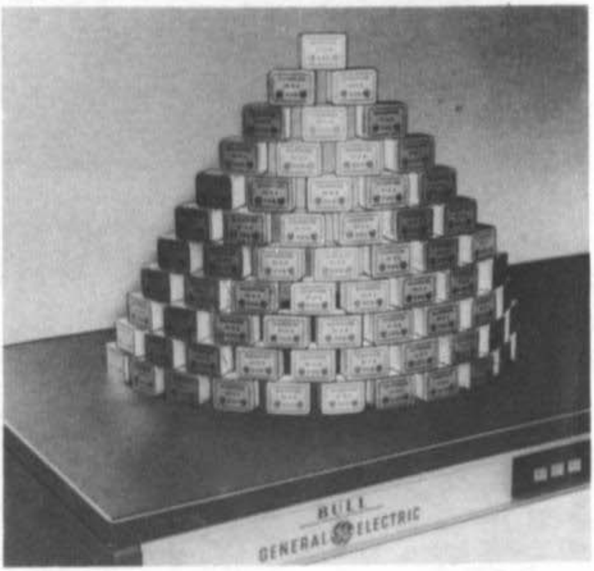
PRECISION QUARTZ CRYSTALS. ULTRASONIC CRYSTALS.

PIEZO-ELECTRIC PRESSURE TRANSDUCERS

Listed is our well-known series of

9 MHz crystal filters for SSB, AM, FM and CW applications.

In order to simplify matching, the input and output of the filters comprise tuned differential transformers with galvanic connection to the casing.



Filter Type		XF-9A	XF-9B	XF-9C	XF-9D	XF-9E	XF-9M
Application		SSB-Transmit.	SSB	AM	AM	FM	CW
Number of Filter Crystals		5	8	8	8	8	4
Bandwidth (6dB down)		2.5 kHz	2.4 kHz	3.75 kHz	5.0 kHz	12.0 kHz	0.5 kHz
Passband Ripple		$<$ 1 dB	$<$ 2 dB	$<$ 2 dB	$<$ 2 dB	$<$ 2 dB	$<$ 1 dB
Insertion Loss		$<$ 3 dB	$<$ 3.5 dB	$<$ 3.5 dB	$<$ 3.5 dB	$<$ 3 dB	$<$ 5 dB
Input-Output	Z_t	500 Ω	500 Ω	500 Ω	500 Ω	1200 Ω	500 Ω
Termination	C_t	30 pF	30 pF	30 pF	30 pF	30 pF	30 pF
Shape Factor		(6:50 dB) 1.7	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:60 dB) 1.8	(6:40 dB) 2.5
			(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:80 dB) 2.2	(6:60 dB) 4.4
Ultimate Attenuation		$>$ 45 dB	$>$ 100 dB	$>$ 100 dB	$>$ 100 dB	$>$90 dB	$>$ 90 dB

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