135 — 137 MHz

×6

22.500 — 22.833 MHz

VCO

144 — 146 MHz

80 Channels; 25 kHz Spacing

:4

:N

φ

:960

1MHz

VHF
COMMUNICATIONS

VHF COMMUNICATIONS,

the international edition of the German publication UKW-BERICHTE, is a quarterly amateur radio magazine especially catering for the VHF/UHF/SHF technology. It is published in February, May, August and November. The subscription price is DM 14.00 or national equivalent per year. Individual copies are available at DM 4.00, or equivalent, each. Subscriptions, orders of individual copies, purchase of P. C. boards and advertised special components, advertisements and contributions to the magazine should be addressed to the national representative.

We would be grateful if you would address your orders and queries to your representative :

The continuous revaluation of the DM and the devaluation of several currencies has meant that most of our prices have increased abroad. Since several of our customers seem to think that this is extra profit for us, we would like to explain that we do not receive any extra money due to the revaluation since we receive everything in DM the same as before, even if the customer has to pay more in his local currency.

The constant revaluation of the DM makes it difficult to make ends meet, since we are sure that you will have noticed that our kit prices have remained virtually constant since 1969, even though the prices of components have increased considerably. The subscription price should also increase due to the higher costs of preparing and printing the magazine, however, we do not wish to do this as the revaluation has already made the magazine expensive even without us increasing our prices.

You can help us keep our prices stable by introducing VHF COMMUNICATIONS to your friends who are not subscribers to our magazine so that we are able to increase circulation. We think that VHF COMMUNICATIONS is worth the price, even after revaluation, and we are sure that you do too. It is only necessary for you to compare the amount of technical articles, the quality of same, as well as the lack of advertising space to that of any other magazine.

FM TRANSCEIVER WITH MULTICHANNEL SYNTHESIZER

Part I: 80-Channel Synthesizer for 25 kHz Spacing

by J. Kestler, DK 1 OF

INTRODUCTION

The large growth of FM activity at VHF makes it extremely expensive to pro-
vide crystals for all of the 80 or more channels. A frequency synthesis principle
of selecting a given number of channels is more favourable if more than a few
channels are to be used.

In the case of a frequency synthesis oscillator, the channel is selected digitally
with the aid of coding switches. This means that any channel can be selected
as required. A FM transceiver is to be described that can be provided with a
frequency synthesizer as laid down in IARU-region 1, or with a frequency syn-
thesizer with 10 kHz spacing which is suitable for operation outside Europe.
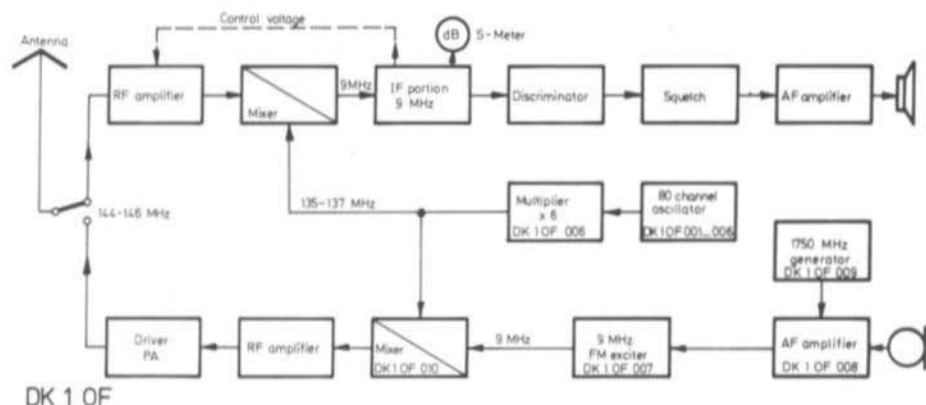
Fig. 1: Block diagram of a FM transceiver with 80 channel synthesizer

The first part of this article is to describe the 80-channel (25 kHz) frequency
synthesizer for this transceiver.

As has already been mentioned, a channel spacing of 25 kHz has been laid down
for FM transmissions in the 2 m band by the IARU region 1. This results in
80 FM channels between 144 MHz and 146 MHz.

A block diagram of a FM transceiver using an 80 channel oscillator is shown
in Figure 1. In order to ensure that the same oscillator can be used for trans-
mit and receive, a superheterodyne principle is also used in the transmitter.
If an intermediate frequency of 9 MHz is to be used, the output frequency of
the local oscillator must be in the range of 135 MHz to 137 MHz. This frequency
plan has been used for previous SSB transceivers such as DJ 9 ZR, DC 6 HL,
and somewhat differently by DL 6 HA. This means that a number of these
modules can be used together with the described 80-channel oscillator in order
to obtain a complete FM transceiver.

This article is now to describe modules DK 1 OF 001 to 006 of the 80-channel oscillator, which forms the main part of the transceiver. The transmitter modules DK 1 OF 007 to 010, as well as the second synthesis oscillator will be described in a later edition of VHF COMMUNICATIONS. Of course, the frequency synthesizer can, as has already been mentioned, be used with other modules. It would then be possible for any converter to be used if the crystal oscillator is disabled and whose IF output provides or can be modified to provide a 9 MHz output (DC 6 HL, DJ 9 ZR 006, DL 6 HA 001, DL 6 SW 004). Several modules are also available for the IF circuit: Either selectivity using a crystal filter and demodulation using a crystal discriminator (DJ 4 BG 008) or a combination of crystal filter and coincidence demodulator (DK 1 PN 005) or further conversion to a second intermediate frequency of 455 kHz with selectivity provided by a ceramic filter and subsequent demodulation using a coincidence demodulator (DC 6 HL 007). Any amplifier (linear or class C operation) can be used after the transmit mixer that can be driven by a few mW (e.g. DC 6 HL 001).
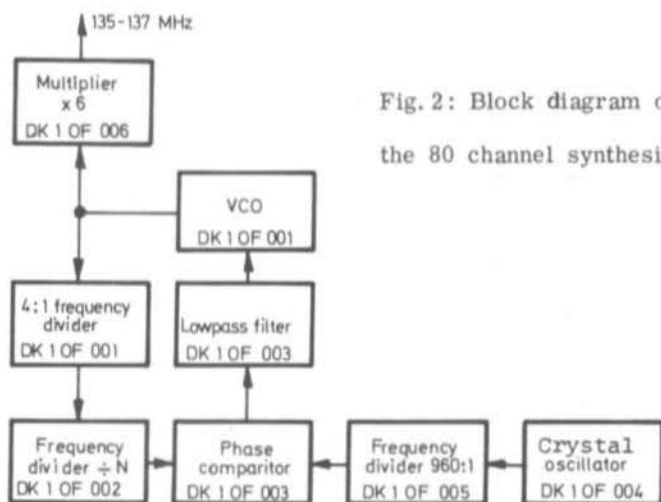
Fig. 2: Block diagram of

the 80 channel synthesizer

1. THE 80-CHANNEL FREQUENCY SYNTHESIZER

Figure 2 gives a block diagram of the frequency synthesizer. With the exception of the multiplier, the synthesizer operates in a phase-locked loop (2). The voltage-controlled oscillator (VCO) oscillates at one sixth of the output frequency so that normal, inexpensive TTL/ICs of the SN 74.. N series can be used in the subsequent frequency divider. The frequency of the oscillator is firstly divided by four. This is followed by a further frequency divider whose frequency division factor N (N = input frequency divided by output frequency) is variable between 5400 and 5479. The required channel is obtained by selecting the required frequency division factor.

The output frequency of the variable frequency divider is compared in a phase comparitor to a crystal-controlled reference frequency. The output voltage of the phase comparitor is proportional to the phase difference between both signals. This voltage adjusts the frequency of the VCO via a lowpass filter. When in its unsynchronized state, the phase comparitor will produce a sawtooth volt-

age that sweeps the VCO over the whole tuning range. When the selected fre-
quency is obtained, the phase difference between the divided frequency from the
VCO and the reference frequency is zero and the control circuit will lock in.

The lowpass filter will suppress the alternating voltage component of the phase
comparitor voltage which would otherwise cause an unwanted frequency modu-
lation of the voltage controlled oscillator. The hold range of the control circuit
is mainly dependent on the cut-off frequency of this lowpass filter which means
that it must be carefully determined.

The long-term stability of a phase-locked oscillator is practically identical to
that of the reference oscillator used (1 MHz crystal oscillator). The short-
term stability is usually less since noise and interference AC voltages in the
control circuit cannot usually be completely suppressed. This means that the
design of the low-pass filter therefore represents a compromise between the
hold characteristics and interference deviation. With the described frequency
synthesizer, the unwanted frequency deviation is less than 100 Hz and the syn-
chronizing time less than 0.1 s, which should be completely satisfactory for
practical operation.

1.1. VOLTAGE-CONTROLLED OSCILLATOR
AND 4 : 1 FREQUENCY DIVIDER

Figure 3 shows a circuit diagram of the voltage-controlled oscillator and 4 : 1
frequency divider. The frequency of the oscillator (T 101) is in the range of
22.500 MHz and 22.833 MHz. With the aid of the moderately close coupling of
the varactor diode D 101, it is possible for this tuning range to be covered with
the aid of a tuning voltage of 4 V to 8 V. The tuning voltage is fed to connec-
tion point Pt 103 and filtered in a RC-link comprising R 108, C 108, C 109 in
order to suppress any interference voltage. The two filter capacitors, espe-
cially C 109, should not be electrolytic capacitors, but plastic foil types. The
leakage current of electrolytic capacitors would represent too large a load for
the tuning voltage. An alternating voltage of a few mV can be connected to
connection Pt 105 for direct frequency modulation of the oscillator. This re-
presents the simplest manner of frequency modulation. However, the frequency
deviation and AF frequency range are limited by the phase-locked loop and its
low-pass filter. Very low frequencies will be filtered out, and if the frequency
deviation is too great it is possible that the hold range of the control circuit
will be exceeded and the synchronization lost.

The resonant circuit of the oscillator is coupled to a subsequent buffer stage
(T 103) which is biased for class B operation. This buffer provides square,
half waves having the frequency of the VCO, which are divided in the subse-
quent double flip-flop (I 101) by 4. This means that the output frequency at
connection Pt 104 is in the range of 5625.0 kHz and 5708.25 kHz.

The integrated frequency divider and its driver (buffer) stage are fed with an
extra stabilized voltage of 5 V. The voltage stabilizer (D 102, T 102) and the
oscillator are fed with a pre-stabilized voltage of 10 V, which is also required
for the other modules.
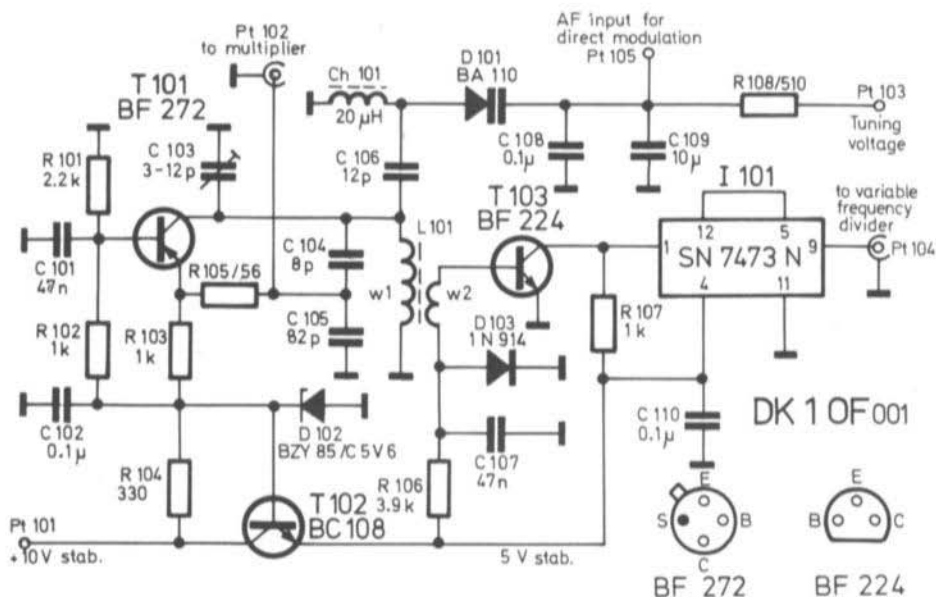
Fig. 3: Circuit diagram of the VCO with 4:1 frequency divider

1.1.1. SPECIAL COMPONENTS FOR THE VCO

T 101: BF 272 (SGS) silicon PNP- UHF transistor
T 102: BC 108 or similar
T 103: BF 224, BF 173 or similar

D 101: BA 110 (ITT), BA 121 (AEG-Tfk) or similar (10 pF/2 V)
D 102: BZY 85/C5V6 or similar 5.6 V zener diode
D 103: 1 N 914 or any silicon diode

I 101: SN 7473 N

L 101: w 1: 10 turns, w 2: 2 turns of 0.4 mm dia. (26 AWG) enamelled
       copper wire in a potted core (14 mm dia. x 8 mm) (Siemens)

Ch 101: 20 μH, ferrite wideband choke (value not critical)

C 103: 3 - 12 pF ceramic disc capacitor 10 mm dia.
C 102, C 108, C 110: 0.1 μF ceramic capacitor, 7.5 mm spacing
C 109: 10 μF, plastic foil capacitor, spacing max. 25 mm

All other capacitors: ceramic types, for max. 7.5 mm spacing.

1.1.2. CONSTRUCTION OF THE VCO

The previously described module VCO and 4 : 1 frequency divider are accommodated on PC-board DK 1 OF 001 whose dimensions are 65 mm x 40 mm. The component locations on this board are shown in Figure 4. Figure 5 shows a photograph of the author's prototype. This module is very uncritical if the construction details are followed. The preliminary alignment of the module is made with the aid of a frequency meter, a low-impedance DC source and voltmeter. The given frequency range is adjusted so that it can be covered with a tuning voltage in the range of 4 V to 8 V. One fourth of the VCO frequency should be measured at the output connection Pt 104.
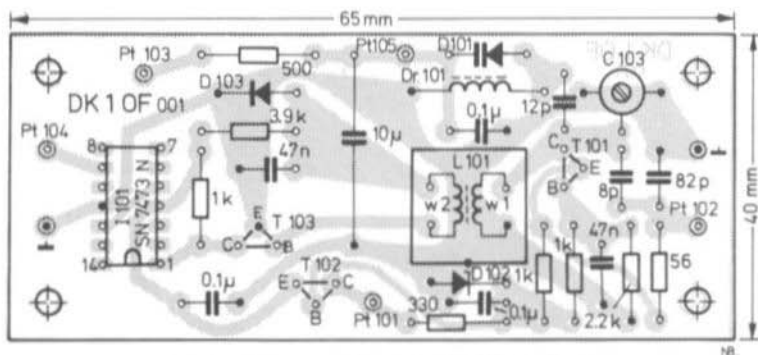
Fig. 4 : PC-board DK 1 OF 001 (VCO)
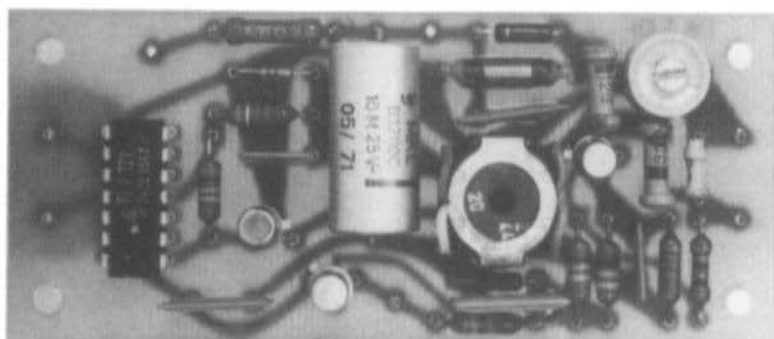
Fig. 5 : Author's prototype of the VCO module

1.2. VARIABLE FREQUENCY DIVIDER

The circuit diagram of this digital module is given in Figure 6. The upper row
in the circuit diagram comprises a 4-stage decimal counter using the wellknown
integrated circuits SN 7490 N. The four integrated circuits of the lower row are
4 Bit comparitors. An output signal is provided when the logic signals coincide
at their two sets of four inputs. The setting inputs (A to D) are programmed
and the other set of inputs are connected to the BCD outputs of the counting
decades. If the state of the counters (BCD code) coincides to the programmed
numbers (also in BCD coding), the comparitor will generate an impulse. When
all four counting decades coincide with the preselected (channel) numbers, the
common output impulse of the comparitors will reset the four-stage counter to
zero. This impulse is also fed to the phase comparitor. Since the input fre-
quency is between 5625.0 kHz and 5708.25 kHz and the numbers between 5400
and 5479 are present at the programming inputs A to D, coincidence will occur
at a frequency of 1.041 kHz. This is therefore the frequency of the output signal.
The programming inputs D 1 to D 4 and C 1 to C 4 receive a fixed programming
(D = 5; C = 4) whereas inputs B 1 to B 4 and A 1 to A 4 are connected to two
preselector switches where the required channel (00 to 79) is selected.

The preselector switches are constructed so that the selected number is directly
available in BCD code (on 4 connections). Some versions of these switches
have also inverted BCD outputs. These are required in our application.

The programming inputs D and C are not wired on the board in order to keep this module as universal as possible. The variable frequency divider operates up to input frequencies of 12 MHz without selected integrated circuits. If different frequencies are to be used for transmit and receive operation (repeater operation), it will be necessary for two preselector switches to be provided that are then switched with the transmit-receive relay. Since the programming leads only carry DC voltage, no problems are involved here.
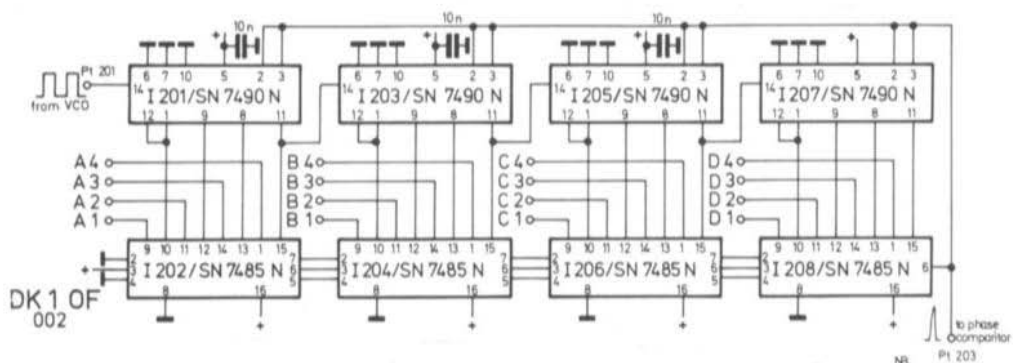
Fig. 6 : Circuit diagram of the variable frequency divider

1.2.1. SPECIAL COMPONENTS FOR THE VARIABLE FREQUENCY DIVIDER

I 201, I 203, I 205, I 207: SN 7490 N
I 202, I 204, I 206, I 208: SN 7485 N

3 ceramic disc capacitors 10 nF (value uncritical) spacing 5 mm

4 pieces 10-stage preselector switches for direct wiring with BCD complimentary (inverted) output (available from the publishers).

1.2.2. CONSTRUCTION OF THE VARIABLE FREQUENCY DIVIDER

Figure 7 shows the component locations and conductor lanes on PC-board DK 1 OF 002 whose dimensions are 115 mm x 50 mm. This board is double-coated and possesses through contacts. It is only necessary for the eight ICs, three bypass capacitors and the connection pins to be soldered into place. Figure 8 shows a photograph of the author's prototype. The preselector switches are mounted on the front panel of the transceiver and are wired. No alignment is required and a frequency counter can be used to check the operation.

1.3. PHASE COMPARATOR

The circuit diagram of this module is given in Figure 9. The flip-flop comprising transistors T 301 and T 302 is used as phase comparitor. The rise slope of the reference voltage (via Pt 302) deflects the flip-flop to the working position and the output impulse of the variable frequency divider (via Pt 301) resets it. This results in a square-wave signal whose pulse length is dependent on the time spacing (phase difference) of these two signals. The RC-link comprising R 310/C 306 provides a DC voltage by integration whose amplitude is dependent on the pulse width and thus the phase difference.
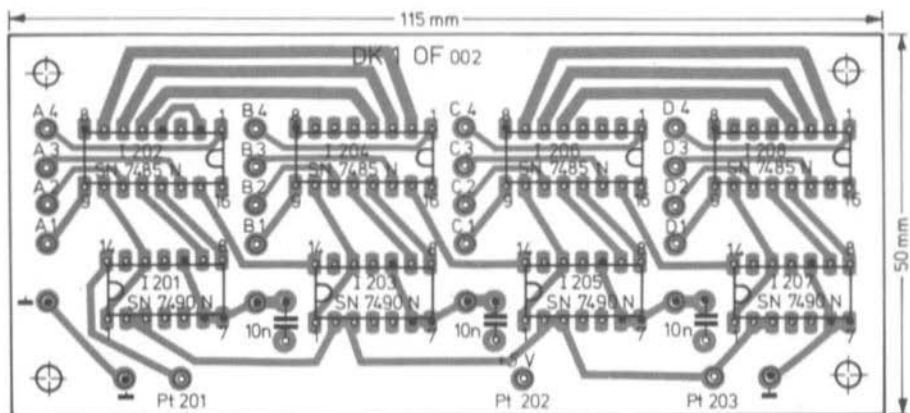
Fig. 7: PC-board DK 1 OF 002
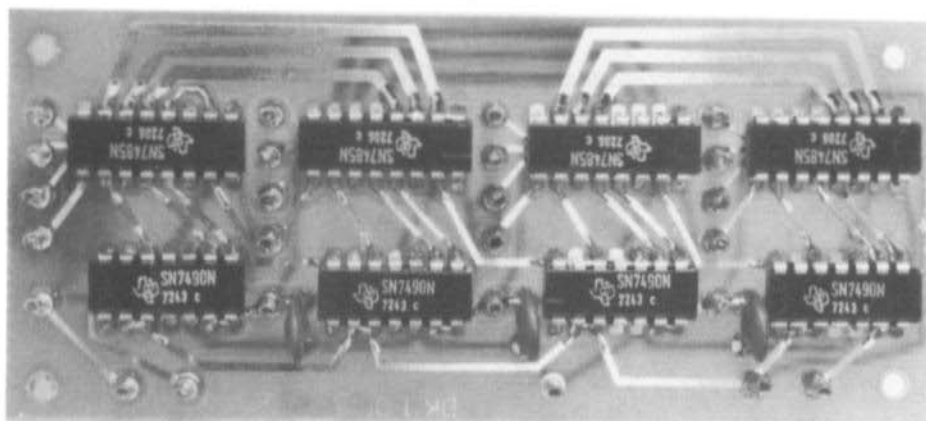
Fig. 8: Photograph of the author's prototype (variable frequency divider)

The subsequent source follower equipped with the field effect transistor T 303 ensures that no loading of the integration link occurs. The subsequent operational amplifier increases the value of the direct voltage signal to a suitable value for the varactor diode of the VCO. The two-stage lowpass filter at the output has a cut-off frequency of 300 Hz and is used to suppress any residual phase comparitor frequency of 1041 Hz from the tuning line. The low-pass filter is designed for an impedance of 500 Ω.

1.3.1. SPECIAL COMPONENTS FOR THE PHASE COMPARITOR

T 301, T 302: BC 108 or similar, however, both should be
            from the same current gain category.

T 303: BF 245 A (TI) W 245 A (Siliconix) or similar FET

D 301, D 302: 1 N 914, 1 N 4148 or similar

I 301: LM 741 CN (National Semiconductor), TBA 221 B (Siemens)

L 301, L 302: 350 turns of 0.1 mm dia. (38 AWG) enamelled copper
            wire wound in a potted core of 14 mm dia. x 8 mm
            (N 30, A_L = 4200) (Siemens)

Fig. 9: Circuit diagram of the phase comparitor

Fig. 10: PC-board DK 1 OF 003 (phase comparitor)

Fig. 11: Photograph of the author's prototype (phase comparitor)

C 301, C 307: 0.1 μF plastic foil capacitor, spacing 7.5 mm or 10 mm
C 306, C 308, C 310: 1 μF plastic foil capacitor, spacing max. 20 mm
C 309: 2.2 μF plastic foil capacitor, spacing max. 22.5 mm.

All other capacitors: ceramic disc or tubular capacitors.

R 312: 50 kΩ thermistor (NTC-resistor).

1.3.2. CONSTRUCTION OF THE PHASE COMPARITOR

This module is accommodated on the single-coated PC-board DK 1 OF 003 whose
dimensions are 115 mm x 50 mm. The component locations are given in Fig. 10.
Due to the low frequencies involved, the construction is extremely uncritical.
However, any hum voltages of more than a few mV should be avoided. Figure 11
shows the author's prototype.

The two input signals are required in order to carry out a functional check of
the module. With the control loop open, (Pt 304 should not be connected to the
next module but connected via 510 Ω to ground), the frequency of the VCO can
be tuned over its range with the aid of a temporary tuning voltage. The voltage
at Pt 304 can be measured with a high impedance voltmeter or DC oscilloscope:
It must be in the range of 4 V and 8 V. The voltage amounts to 2.6 V to 5.4 V
at the source connection of transistor T 303. If any difficulties are encountered,
the lowpass filter should be checked with the aid of an AF generator. Its cut-
off frequency (-3 dB) should be approximately 300 Hz. The generator should
be connected via R 314 in order to maintain the impedance of 500 Ω.

1.4. REFERENCE OSCILLATOR

The circuit diagram of the reference oscillator is given in Figure 12. In this
circuit, a 1 MHz standard frequency crystal operates at parallel resonance in
a clapp-circuit. The alignment trimmer C 401 allows the crystal to be pulled
to the exact nominal frequency. The subsequent stage (T 402) converts the
sinusoidal voltage into the required impulses to drive the TTL ICs. This is
the reason why the operating voltage of this module amounts to 5 V.

Fig. 12: Circuit diagram
of the reference
oscillator
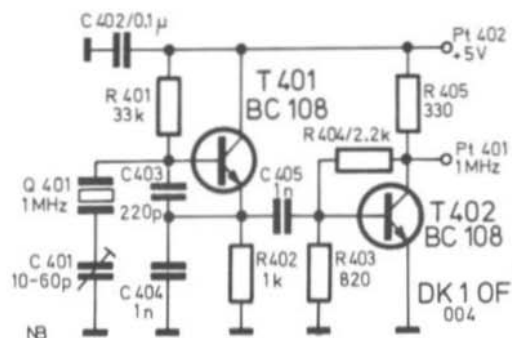
1.4.1. SPECIAL COMPONENTS FOR THE REFERENCE OSCILLATOR

T 401, T 402: BC 108 or similar

C 401: 10 - 60 pF ceramic disc trimmer, 10 mm dia.
C 403: 220 pF styroflex capacitor
C 404: 1000 pF styroflex capacitor

Crystal holder for horizontal mounting.
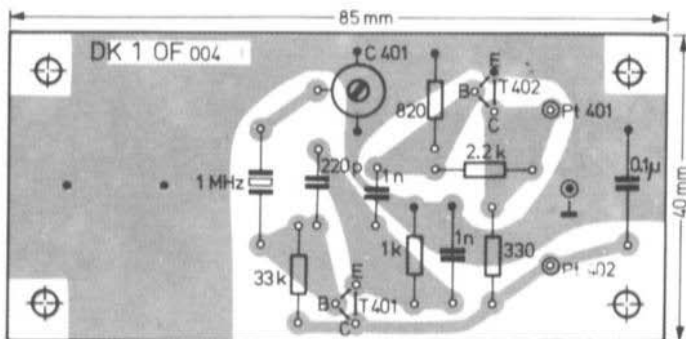
Fig. 13: PC-board DK 1 OF 004 (reference oscillator)
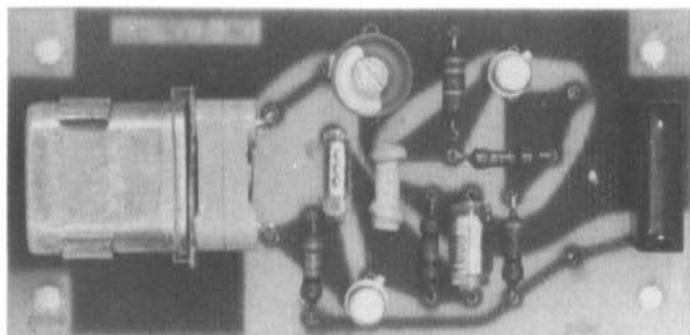
Fig. 14: Author's prototype of the reference oscillator

1.4.2. CONSTRUCTION OF THE REFERENCE OSCILLATOR

This module is accommodated on the 85 mm x 40 mm large PC-board DK 1 OF 004 which is shown in Figure 13. This board is single-coated. Figure 14 shows a photograph of the author's prototype. A functional check can be made on this module with the aid of a frequency counter, oscilloscope or by simply monitoring the harmonics.

1.5. 960 : 1 FREQUENCY DIVIDER

Figure 15 shows the circuit diagram of this digital module. It is equipped with three frequency dividers: I 501 divides the input frequency of 1 MHz by 10; I 502 is wired so that the frequency is divided by 8, and I 503 finally divides by 12. The total frequency division factor is 960 which means that an input frequency of 1 MHz results in an output frequency of 1041 Hz which is then fed to the phase comparitor.

1.5.1. COMPONENTS AND CONSTRUCTION
OF THE 960 : 1 FREQUENCY DIVIDER

I 501, I 502: SN 7490; I 503: SN 7492 N

1 plastic foil capacitor or tantalium electrolytic 1 μF (value uncritical)
spacing 7.5 - 20 mm.
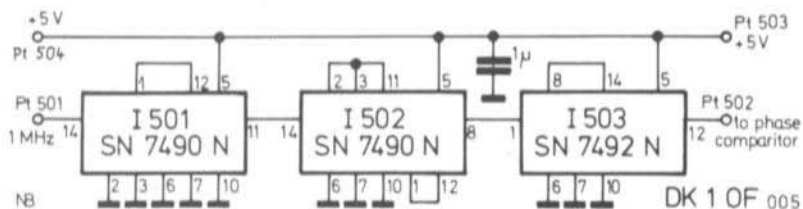
Fig. 15 : Circuit diagram of the 960 : 1 frequency divider

The module can be checked with the aid of a frequency counter, or oscilloscope with calibrated horizontal deflection. A photograph of the author's prototype is given in Figure 17.
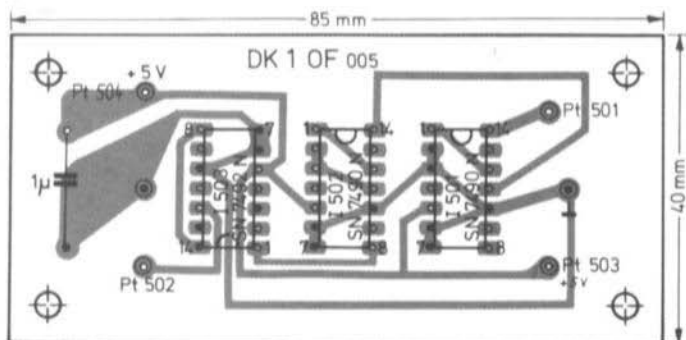
Fig. 16 : PC-board DK 1 OF 005 (960 : 1 frequency divider)
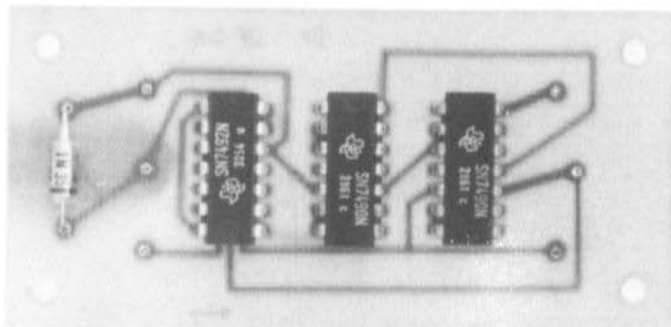
Fig. 17 : Author's prototype of the 960 : 1 frequency divider

1.6. FREQUENCY MULTIPLIER

As previously mentioned, the VCO oscillates at approximately 22.5 MHz which means that a frequency multiplication factor of six is required when used together with a two metre transceiver. Figure 18 shows a circuit diagram of this module. The first stage (T 601) doubles the frequency to approximately 45 MHz; the second stage amplifies the signal so that sufficient drive is available for the tripler stage (T 603) that operates in class C. Bandpass filters are provided after each frequency multiplication process to ensure sufficient suppression of the fundamental wave and unwanted harmonics.
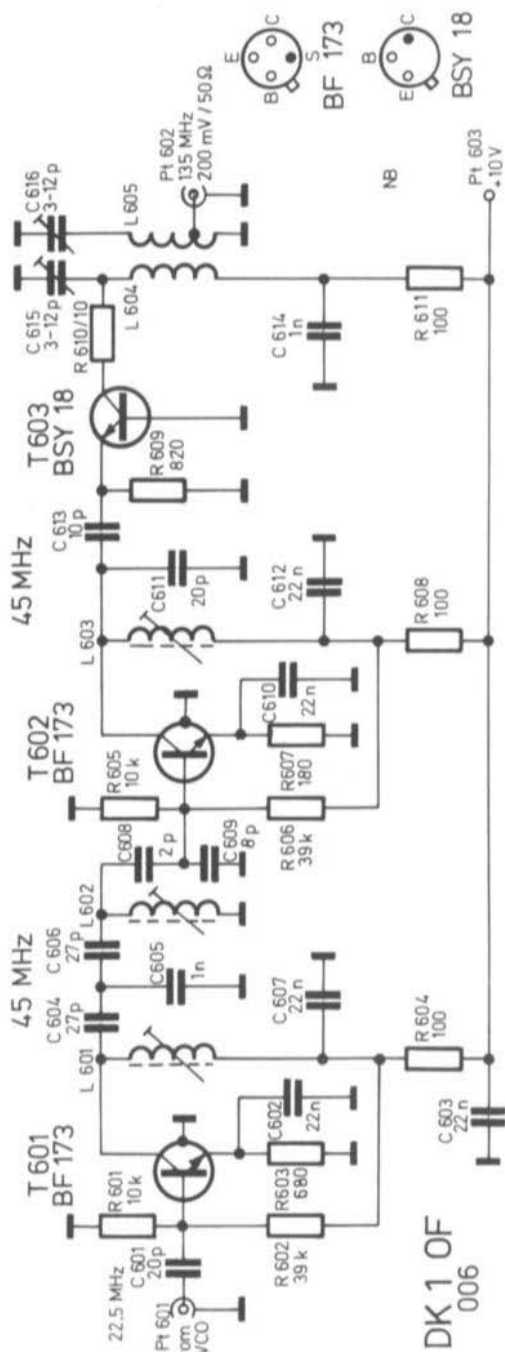
Fig. 18 : Circuit diagram of the frequency multiplier
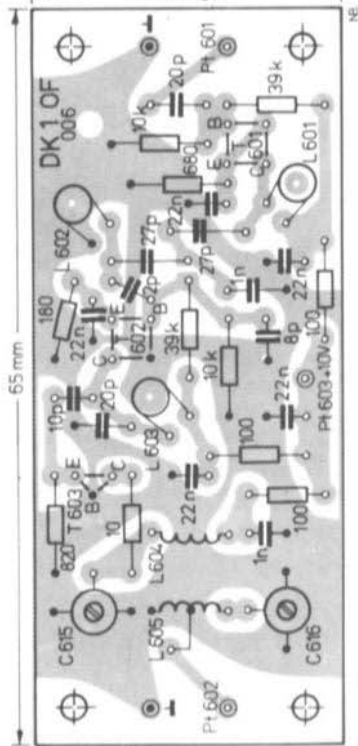
Fig. 19 : PC-board DK 1 OF 006 for the frequency multiplier
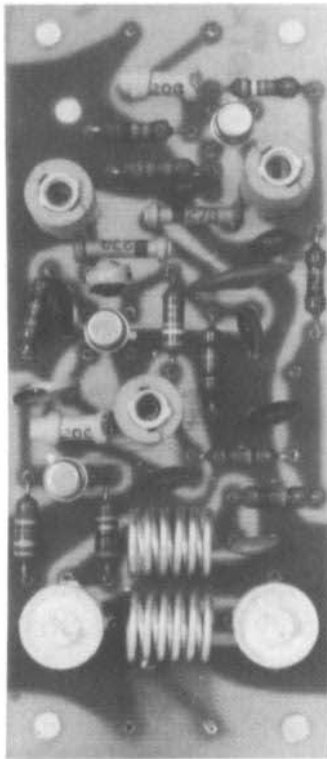
Fig. 20 : Author's prototype of the frequency multiplier

The spectral purity of the oscillator signal between 135 MHz and 137 MHz is extremely high due to the frequency processing method used. Discrete harmonics only exist in the form of harmonics of the VCO. However, these are virtually suppressed in the module which has just been described. A further suppression of unwanted signals is carried out in the subsequent transmit amplifier and is extremely effective due to the large frequency spacing from the required signal.

1.6.1. SPECIAL COMPONENTS FOR THE FREQUENCY MULTIPLIER

T 601, T 602: BF 173 (or BF 224) BF 167
T 603: BSY 18, 2 N 918

L 601 to L 603: 7.5 turns of 0.4 mm dia. (26 AWG) enamelled copper wire
              wound on a special coilformer (available from the publishers)
L 604: 6 turns of 1 mm dia. (18 AWG) silver-plated copper wire
    wound on a 6.5 mm former, self-supporting.
L 605: As L 604 but with coil tap 1.5 turns from the cold end.

C 615, C 616: 3 - 12 pF ceramic disc trimmer 10 mm diameter.

All other capacitors: ceramic disc types.

1.6.2. CONSTRUCTION OF THE FREQUENCY MULTIPLIER

This module is accommodated on the printed circuit board DK 1 OF 006 whose dimensions are 65 mm x 40 mm. The component locations of this board are given in Figure 19. The construction should be made in a suitable manner for VHF circuits. A dipmeter is used for the preliminary alignment. The first two stages should be aligned for maximum collector current of the third stage. The bandpass filter at the output is aligned with the aid of a temporary connected demodulator circuit or in the complete unit for maximum conversion gain. The coupling between the two bandpass filters is designed so that no voltage drop is noticeable at the band limits (135 MHz and 137 MHz). Figure 20 shows a photograph of the author's prototype.

1.7. INTERCONNECTION OF THE 80-CHANNEL SYNTHESIZER

Figure 21 gives the electrical connections of the six described modules of the synthesizer. The wiring plans of the four preselector switches are given in Figure 22. In order to allow the two groups of two switches to be set independently to transmit and receive as required, it is only necessary for the connection pins "3" to be grounded by a change-over contact on the transmit-receive switch. With the aid of the reversal switch, it is possible to receive, for instance, on the input frequency of a repeater and transmit on the transmit frequency, without having to change the channel number. This additional application is useful for establishing whether direct communication is possible when using a repeater, or for direct communications to another mobile station who is only equipped for repeater operation (of course, when out of range of the repeater).

A few tips are necessary regarding the mechanical construction:

The printed circuit boards and solder tags are arranged so that two PC-boards can be mounted one above the other in the following manner:

004	above	003	above	006	above
005	below	002	below	001	below
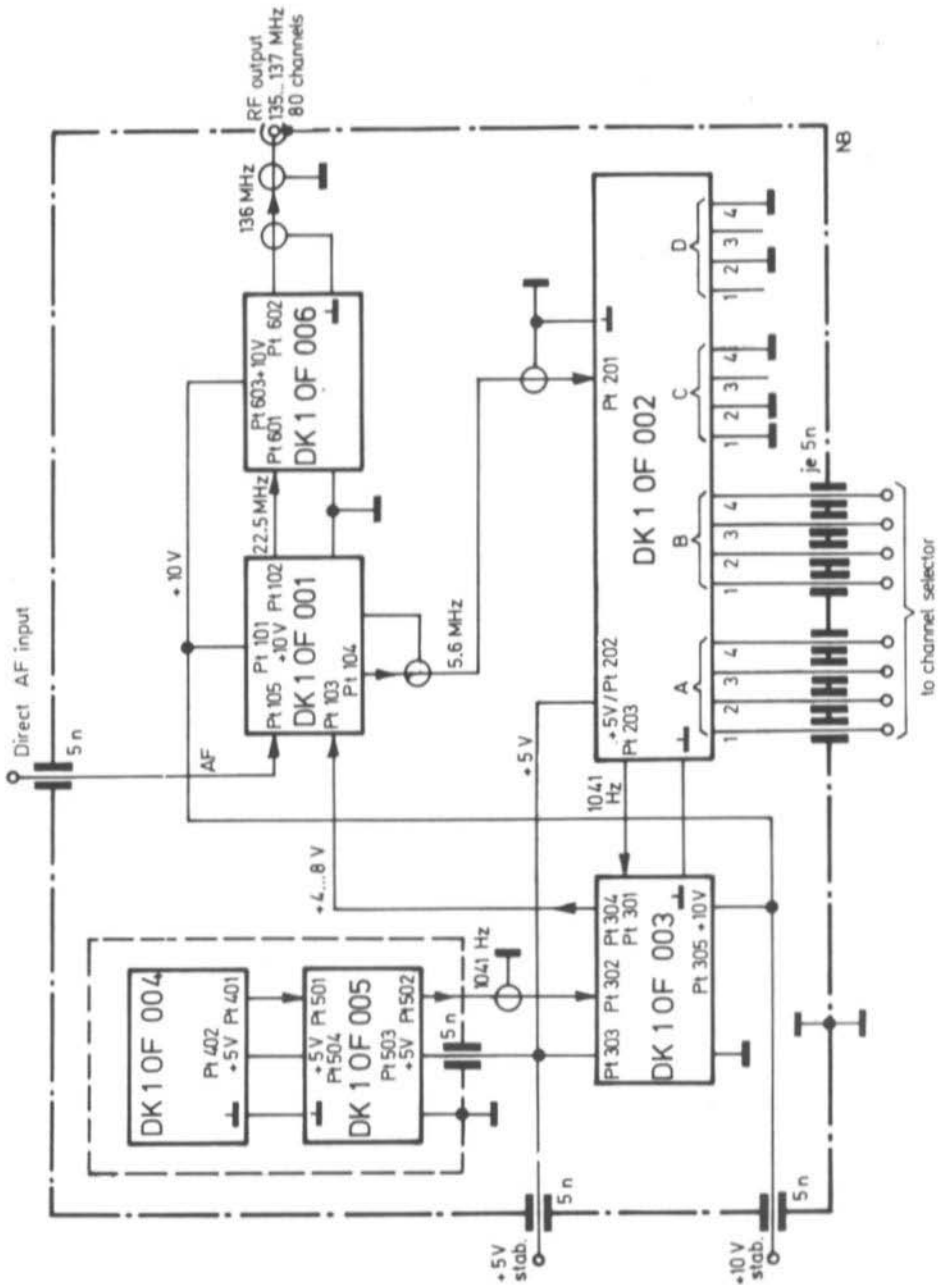
Fig. 21 : Interconnections between the six modules of the 80 channel synthesizer

DK 1 OF

D1 ... D16 :
any silicon diodes mounted
on the preselector switches
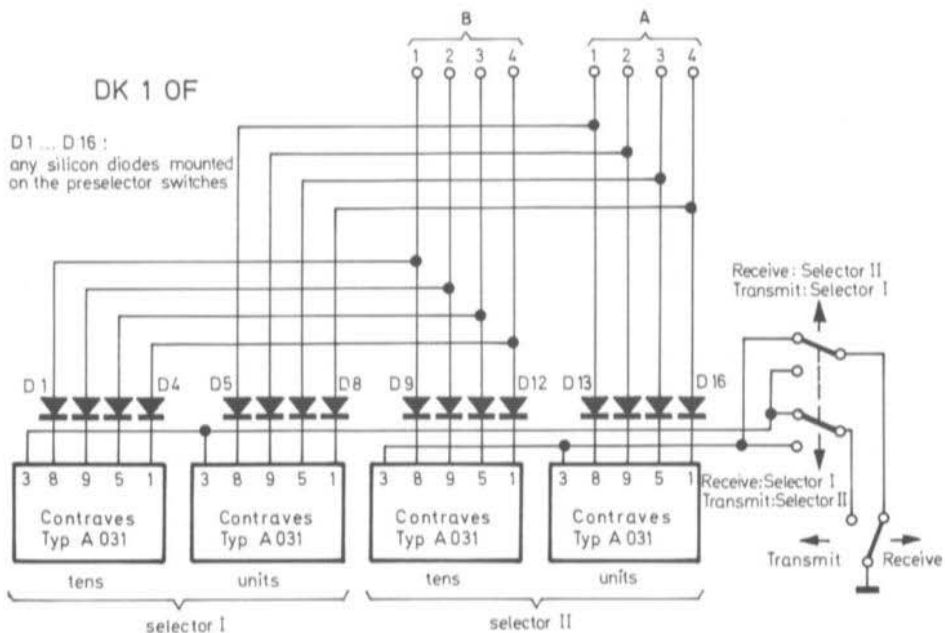
Fig. 22: Wiring plan for the channel selectors
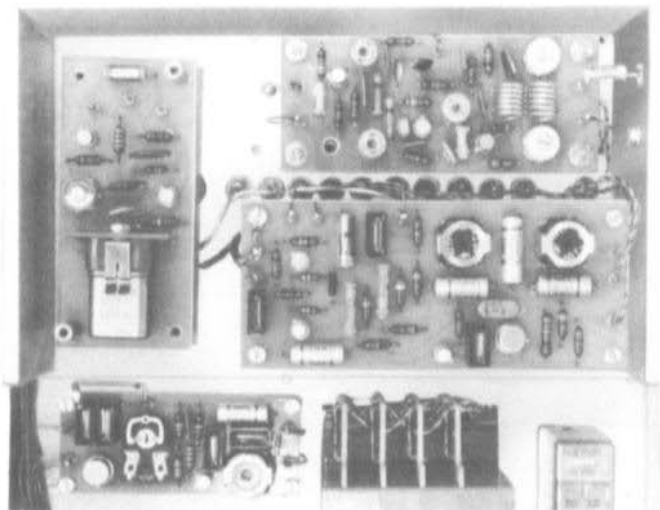
Fig. 23: Author's prototype of the complete
80 channel synthesizer

The two PC-boards 004 and 005 should be well screened since the harmonics of the steep 1 MHz pulses would otherwise cause interference to the 9 MHz IF of the receiver.

In addition to this, the whole of the synthesis oscillator should be carefully screened. Feedthrough capacitors should be used for all connections, and a coaxial connector for the RF output. Figure 23 gives a photograph of this section of the author's transceiver.

1.8. REFERENCES

(1) IARU Region 1 Conference 1972: 144 MHz Bandplan

(2) T. Schad: Phase-locked Circuits
VHF COMMUNICATIONS 4 (1972), Edition 2, Pages 80-87.

COMPLETE KIT FOR A SIX-DIGIT 250 MHz FREQUENCY COUNTER
comprising the following kits: (special transformer on request)
DL 8 TM 002   Six-Digit Frequency Counter
DL 8 TM 003   Dual-Input Preamplifier
DJ 1 JZ 001   1 MHz Frequency Standard with Crystal Oven
DJ 6 PI 001   250 MHz Prescaler
DL 3 YK 002   Universal Power Supply
Total price :   DM 900. --
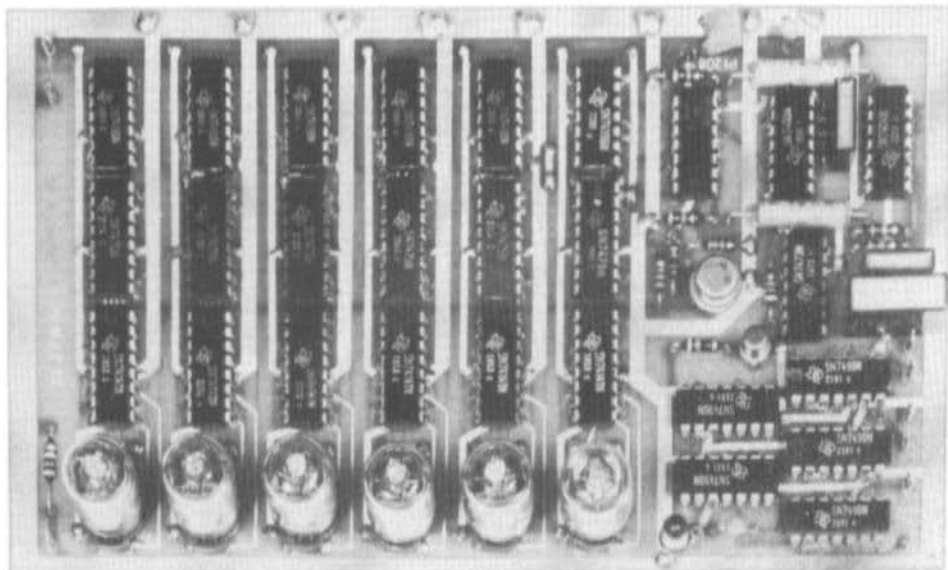
ADJUSTING THE OPERATING POINT
OF FIELD EFFECT TRANSISTORS

by D. E. Schmitzer, DJ 4 BG

The principles of adjusting the operating points of field effect transistors, especially with dual-gate MOSFETs are not understood as well as that of conventional transistors. This article is therefore to examine these devices with examples so that they can be more easily understood.

1. JUNCTION FIELD EFFECT TRANSISTORS

Figure 1 shows the simplest possible circuit for a junction FET. The gate is grounded for direct current via a resistor (or inductance). The source is directly grounded. This is the circuit mainly given in data sheets by which the drain current is completely unstabilized and can fluctuate within the limits determined by the semiconductor. The following table shows how great these fluctuations can be for several well-known N-channel transistors:

Type	I_D at $U_{GS} = 0$ V
BF 245 (non-selected)	2 - 25 mA
BF 245 A	2.0 - 6.5 mA
BF 245 B	6.0 - 15 mA
BF 245 C	12.0 - 25 mA
BC 264 (non-selected)	2 - 12 mA
BC 264 A	2.0 - 4.5 mA
BC 264 B	3.5 - 6.5 mA
BC 264 C	5.0 - 8.0 mA
BC 264 D	7.0 - 12.0 mA
2 N 4416	5 - 15 mA
2 N 3823	4 - 20 mA
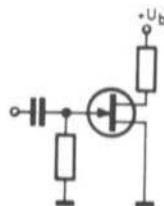
Fig. 1: Simple amplifier circuit with a N-channel junction FET

With the non-selected BF 245 transistors, it will be seen that the current can vary at a ratio of 1 : 12.5 and the variation range is an average of 1 : 3 even with the selected types (A, B, C).

Fig. 2: Selecting the operating point with the aid of a source resistor
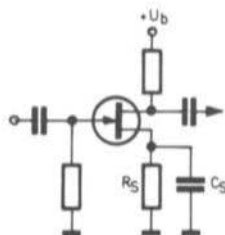
1.1. OPERATION WITH A SOURCE RESISTOR

The operating point can be stabilized by inserting a resistor in the source circuit (Fig. 2). The drain current will then always be lower than when operating without this resistor. As is known in vacuum tube technology, the source, or drain current causes a voltage drop that produces a positive bias at the source or drain. This means that the drain will have a negative bias with respect to the source.

In order to avoid a feedback link for the alternating voltage that is to be amplified, it is necessary for the source resistor to be bridged with a capacitor. The reactance of this capacitor must be equal to or less than the parallel connection of source resistor and the input impedance of the field effect transistor in a common-gate circuit ($1/S$).

The effect of such DC-feedback can be seen in the I_D/U_G family of curves in which the characteristics of various source resistors are given. At larger source resistances, the difference between the maximum (intersection with curve I_{Dmax}) and minimum (intersection with curve I_{Dmin}) is considerably smaller. Figure 3 shows the limit characteristics derived from the data sheet values of I_{Dmin} and I_{Dmax} at $U_{GS} = 0$ V, as well as U_{GSmin} and U_{GSmax} for $I_D = 0$ mA for the non-selected BF 245 as well as two resistance lines corresponding to 100 Ω and 1 kΩ.
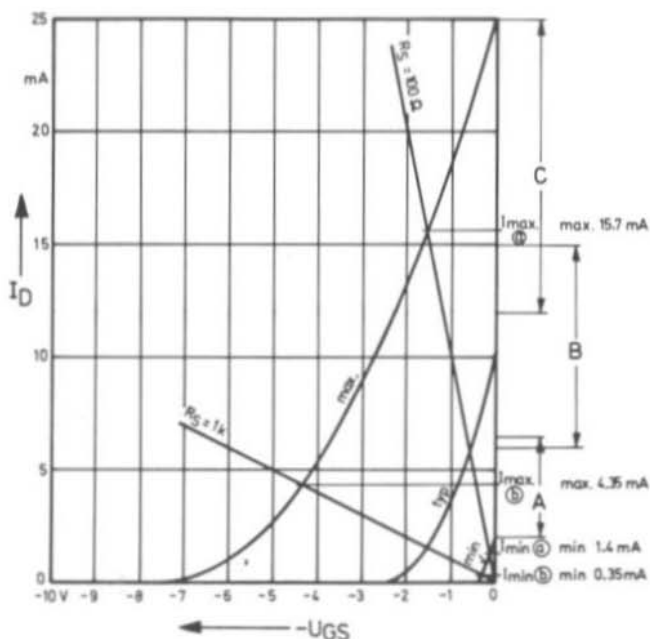
Fig. 3 :
Resistance characteristics of the source resistor in the family of curves for the BF 245

At $R_S = 100\,\Omega$, the drain current can be in the range of 1.4 mA to 15.7 mA; at $R_S = 1$ kΩ, the values will be $I_{Dmin} = 0.35$ mA and $I_{Dmax} = 4.35$ mA. In both cases, the tolerance range is somewhat less than the circuit given in Figure 1. In order to decrease the tolerance range still further, the non-selected transistor should be replaced by one of the selected types. In this case, the current will be in the lower third for the A-type, in the centre for the B-type and in the upper third of the area between the limit lines for the C-type.

1.2. OPERATION WITH SOURCE RESISTOR AND ADDITIONAL BIAS

An improved stabilization of the operating point is obtained with higher source resistance values in conjunction with a positive gate voltage. The positive bias voltage is necessary in order to increase the reduction of drain current caused by the high value of the source resistor. Figure 4 shows three suitable circuits.

The resistor through which the bias voltage is fed (Fig. 4a, b) will reduce the input impedance of the stage. This can be disadvantageous in many cases and can be avoided if the circuit given in Figure 4c is used. The use of the bias voltage will mean that the drain current is dependent on the operating voltage, which is not important in most cases.
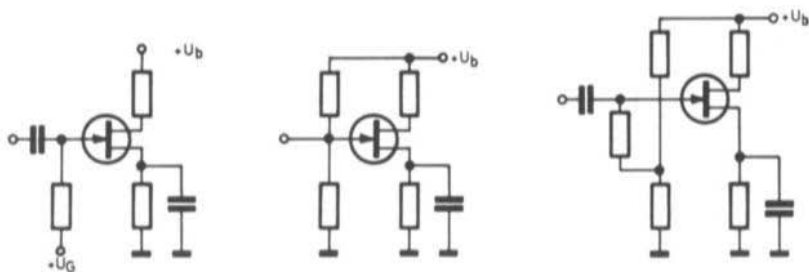
Fig. 4 : Three possible circuits for operation with a positive gate bias voltage

Even a low gate bias voltage of 1 V will produce a considerable reduction of the tolerance range. The limit characteristics given in Figure 5 show that at this value, which has been designated with "a", the drain current can now only fluctuate between 0.5 mA and 2.33 mA. This corresponds to a ratio of 1 : 4.64. The characteristics designated by "b" correspond to a bias voltage of 3 V and a source resistor of 4.7 kΩ which reduces the tolerance range still further to 0.7 mA to 1.82 mA (only 1 : 2.6). In both cases, the non-selected transistor type BF 245 is considered to have a required mean operating point of 1 mA.
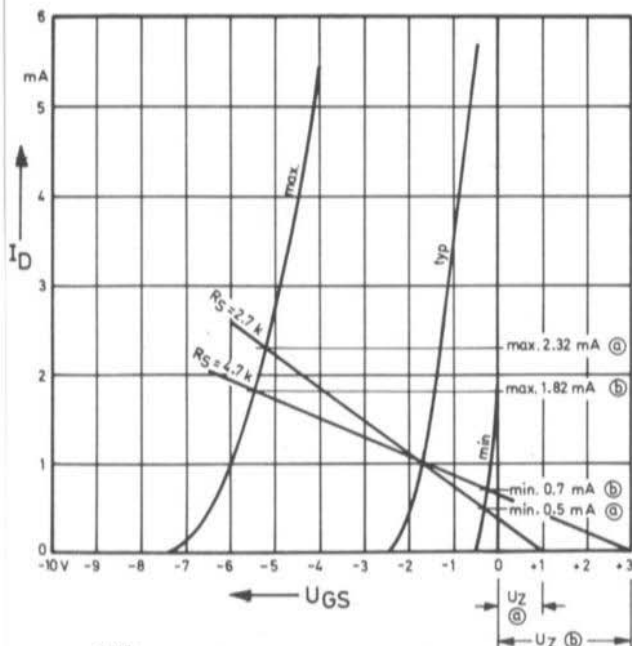
Fig. 5 :
Setting the operating point with a source resistor and positive gate bias voltage

It should be pointed out when operating with a positive gate bias voltage that the voltage between gate and source (U_{GS}), which is decisive for the drain current, is negative. As long as the FET is not used with too low a drain current, the positive voltage drop across the source resistor will be greater than the bias voltage.

At very low drain currents, U_{GS} can become positive. If a value of +0.5 to +0.7 V is exceeded or if this value is exceeded together with the alternating voltage at the input, the gate-source junction will conduct and cause current to flow. This will cause a high degree of distortion and a relatively low input impedance of the stage. Such an adjustment must therefore be avoided.

Stabilizing circuits by which the gate bias voltage is taken directly from the drain will cause a voltage feedback in simple circuits which will reduce the input impedance considerably. This is the reason why such circuits have not been considered here, or such circuits where the stabilization of the operating point is made over several stages.

2. MOSFETs

The same characteristics are virtually valid for MOS - field effect transistors as for junction types. When adjusting the operating point, it is permissible for the gate-source voltage U_{GS} to be positive since no junction can be biased into the forward current region. However, the permissible positive gate-source voltage is usually quite low in the data sheets which means that practically the same limitations are valid as for junction FETs.

2.1. DUAL-GATE MOSFETs

The current values of the popular dual-gate MOS field effect transistors fluctuate so considerably that a stabilization must be provided. Several well-known types manufactured by RCA are to be given as an example.

Type	Current range at $U_{G1} = 0$ V; $U_{G2} = 4$ V	Fluctuation ratio
3 N 140 3 N 159 3 N 187	5 - 30 mA	1 : 6
40 673	5 - 35 mA	1 : 7
3 N 200	0.5 - 12 mA	1 : 24
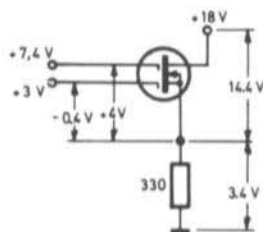
Fig. 6 : DC-voltage values on a mean transistor type 40 673 in a stabilized circuit

A simple stabilization using a source resistor as is often found in circuit dia-
grams will result in values that differ greatly from those given in the data
sheet even at low resistance values in conjunction with transistors in the lower
portion of the tolerance range. This means that typical transistors will not
conform to their calculated values. This can be seen in Figure 7 which gives
the limit characteristics for the transistor type 40673.

$$\frac{I_{max.}}{I_{min.}}: \quad \begin{array}{l} ⓐ : 4.14 \\ ⓑ : 5.01 \end{array}$$
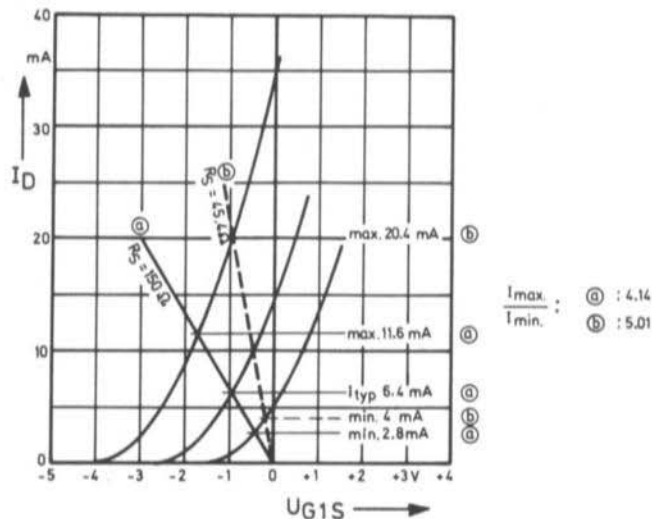
Fig. 7:
Stabilizing the
operating point
of a dual-gate
MOSFET;
case a, b.

At a source resistance of 150 Ω, the drain current will be in the range of
2.8 mA and 11.6 mA (case a). If even the medium value transistors are to
reach a current of 10 mA, which is important for all values given in the data
sheet, it was necessary for the resistance value to be decreased to 47 Ω
(case b). The tolerance range is then 4.0 mA to 20.4 mA, which corresponds
to a tolerance range of 1 : 5. Since the tolerance range without source resistor
corresponds to 1 : 7, the improvement is not very great. This can be clearly
seen in conjunction with MOSFET type 3 N 200 which possesses an extremely
wide tolerance range which cannot be limited without external aids.

2.1.1. POSITIVE BIAS VOLTAGE AT GATE 1

Figure 8 repeats the family of characteristics for the transistor type 40 673.
Case c is given with a source resistor R_S = 330 Ω with a positive bias voltage
at gate 1 of 3 V. An operating voltage is now required at gate 2 that is gener-
ally approximately 4 V higher than the source voltage. The resulting or re-
quired voltage values are given in Figure 6. It is not necessary for separate
voltage dividers to be used for setting both gate voltages. This can be achieved
by adding the gate 2 voltage to that at gate 1, as is shown in Figure 9.

This circuit will obtain a drain current of 10.4 mA (Fig. 8, case d) with
mean-value transistors even at a bias voltage of only 1 V at gate 1 and a source
resistor R_S = 130 Ω. The tolerance range will then be in the range of 5.4 mA
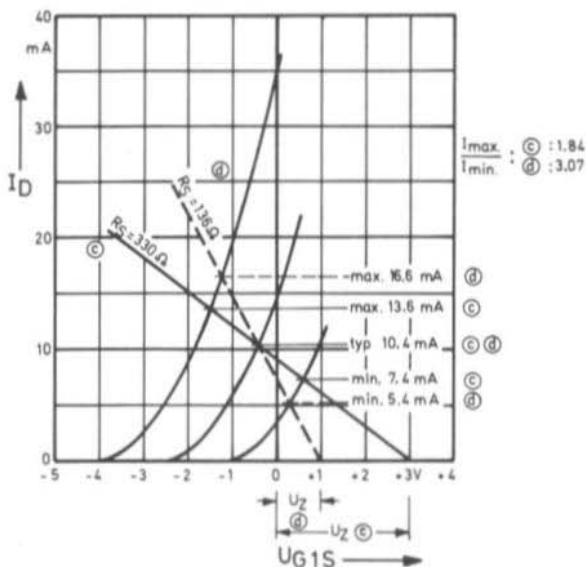
Fig. 8:
Stabilizing the
operating point
of a dual-gate
MOSFET;
case c, d.

and 16.6 mA, which corresponds to a ratio of 1 : 3. When comparing this to
the values without bias voltage (Fig. 7, cases a and b), this represents a
considerable improvement.
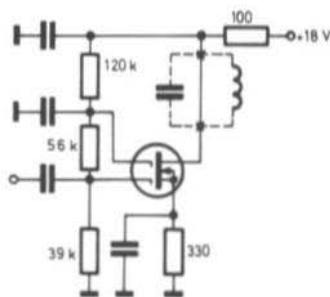
Fig. 9:
Setting the operating point
with a common voltage
divider for the gate voltages
(using standard resistance
values)

3. MIXER STAGES WITH DUAL-GATE MOSFET
3.1. DC-OPERATING POINT

Unfortunately, not very much information is given on the use of dual-gate
MOSFETs as mixers. However, the brief data that is available shows that the
voltage values present at gate 1 and gate 2 are practically equal and are in the
range of +0.6 V and +1.2 V according to the transistor type. Even the measur-
ing circuits given in the data sheets give an increased source resistor and
positive gate bias voltage. In order to provide reproduceable characteristics,
the measuring circuits used separate voltage dividers with low tolerance resis-
tors. The circuit given in Figure 10 is simpler for practical applications. The
voltage is fed from a common voltage divider via a decoupling resistor to gate 1
and gate 2. This means that the voltage balance at both gates can be main-
tained simply and without precision resistors.

Attention should be paid not only here but also during the construction that no great tolerances are present. Of course, a large number of circuits are un-critical so that even a tolerance of $\pm$ 20% will not cause any difficulties. How-ever, since it is not always possible to state which resistors are not critical and which will have a considerable effect on the reproduceability of the values determined by the author, attention should be paid to the tolerance of the pur-chased components. A tolerance of $\pm$ 5% should represent a favourable compro-mise between the requirements and price. Such resistors are usually not very much more expensive than types with $\pm$ 10% or $\pm$ 20% tolerance.
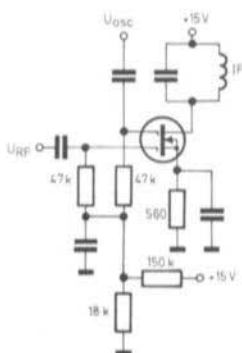
Fig. 10:
Setting the operating point
of a dual-gate MOSFET
mixer

3.2. LOCAL OSCILLATOR AMPLITUDE

Since virtually no data is available regarding the required amplitude of the local oscillator voltage at gate 2 of a dual-gate MOSFET when used as a mixer, the conversion gain was determined as a function of the amplitude of the local oscillator voltage in a circuit as given in Figure 10. Since the measurement was made in the shortwave region (20 m band, the determined values may not be valid for 2 m applications. However, the same basic principles will be valid.

The measured curves are given in Figure 11. 50% of the maximum conversion gain is achieved with a drive of only 1.1 V_{pp} (corresponding to 310 mV RMS). This means that the actual value is very uncritical at local oscillator voltages of more than 1 V_{pp}. No noticeable increase of the gain will be present at con-siderably higher values, however, the non-linearities of the conversion charac-teristic will become more and more noticeable which will in turn, cause a considerable increase of spurious signals and "birdies".

The drain current or voltage drop across the source resistor can serve as an accessible measure for determining the correct drive value. However, the variation is very low, namely only by 11% from 0 V to full drive in the case in question, however, the run of the ID - curve (which remains constant bet-ween 0 and 1 V_{pp} and will only increase at values over 1 V_{pp}) allows a coarse indication when more than 50% of the maximum conversion gain has been achieved. This is indicated when ID increases slightly. In the example in question, it will be seen in Figure 11 that 75% of the maximum conversion gain is achieved at a local oscillator voltage of 2.5 V_{pp} (corresponding to approximately 0.9 V RMS), which causes an increase of I_D from 2.14 mA to 2.20 mA, e.g. by approximately 3%.
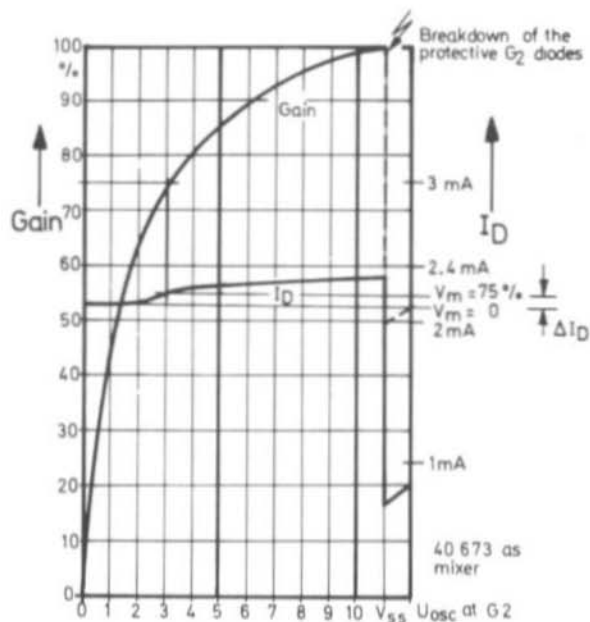
Fig. 11 :
Relative conversion
gain of a mixer
equipped with the
40 673 adjusted
according to Fig. 10
(f_{in} = 14 MHz,
f_{lo} = 5 MHz,
f_{IF} = 9 MHz)

4. REFERENCES

Data sheets and applications notes of: Texas Instruments, Valvo, RCA.

FET 2 metre converter AC 2:
The matching converter to receiver AR 10. Neutralized FET input stages, push-pull FET mixer.
Crystal 38.6667 MHz (HC-25/U).

Specifications:

Input frequency:	144—146 MHz	Noise factor:	1.8 dB
Output frequency:	28—30 or 26—28	Image suppression:	> 70 dB
Gain:	22 dB ± 2 dB	Operating voltage:	12—15 V/15—20 mA
Input frequency:	50 Ohm	Dimensions:	120 mm x 50 mm x 25 mm

Verlag UKW-BERICHTE, H. Dohlus oHG, D-852 Erlangen, Gleiwitzer Str. 45

A 10:1 PRESCALER AND PREAMPLIFIER WITH AN UPPER FREQUENCY LIMIT OF 250 MHz FOR USE WITH FREQUENCY COUNTERS

by J. Grimm, DJ 6 PI

Frequency counters equipped with Schottky TTL-circuits, are, at present, able to count up to frequencies of approximately 120 MHz (1), (2). It is not possible to make measurements on the 2 m band even if these integrated circuits are selected. ECL integrated circuits are suitable for counting up to higher frequencies. The input frequency is then usually divided down to a frequency that can be processed by cheaper TTL circuits. A frequency division ratio of 10:1 is most favourable since the output of the counter need only be multiplied by 10.

The described prescaler uses the Fairchild 10:1 ECL divider 95 H 90, which is relatively cheap in comparison to other similar types. The upper frequency limit of this integrated circuit varies in the range of 220 MHz to 320 MHz (according to the data sheet). With four such prescalers constructed by local amateurs, one operated correctly up to 235 MHz, whereas the other three were able to measure 265 MHz. The prototype built up by the editorial staff operates up to frequencies of 280 MHz within the first minute after switching on, and then falls to 240 MHz when the integrated circuit gets hot.

The triple differential amplifier Fairchild 9582 used for preamplification is also an ECL circuit. This circuit is very sensitive: 5 mV is sufficient in the short-wave range for a stable frequency indication, whereas 35 mV are required at 145 MHz and 100 mV at the upper frequency limit.

The prescaler and preamplifier are accommodated on a small single-coated PC-board that is also equipped with an ECL/TTL level converter. This means that it is possible for the counting gate of a TTL counter to be directly connected. The module requires a voltage of 5 V at approximately 220 mA.

1. CHARACTERISTICS

Lower frequency limit:
approx. 100 kHz;
Upper frequency limit:
typically 250 MHz;
Input voltage requirements: See Fig. 1;
The input voltage requirements increase below approx. 1 MHz and will reach 100 mV at 100 kHz.

Input impedance: less than 100 Ω with capacitive component above 50 MHz.

Output voltage:
Either ECL or TTL level.

Operating voltage:
+5 V stab. at approx. 220 mA.
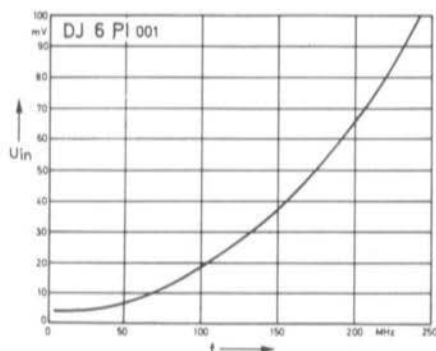
Fig. 1: Input voltage requirements of the 10:1 divider module DJ 6 PI 001

Preamplifier   Divider   Level converter   Level converter ECL/TTL

I 1 : 9582

I 3 : SN 7404 N
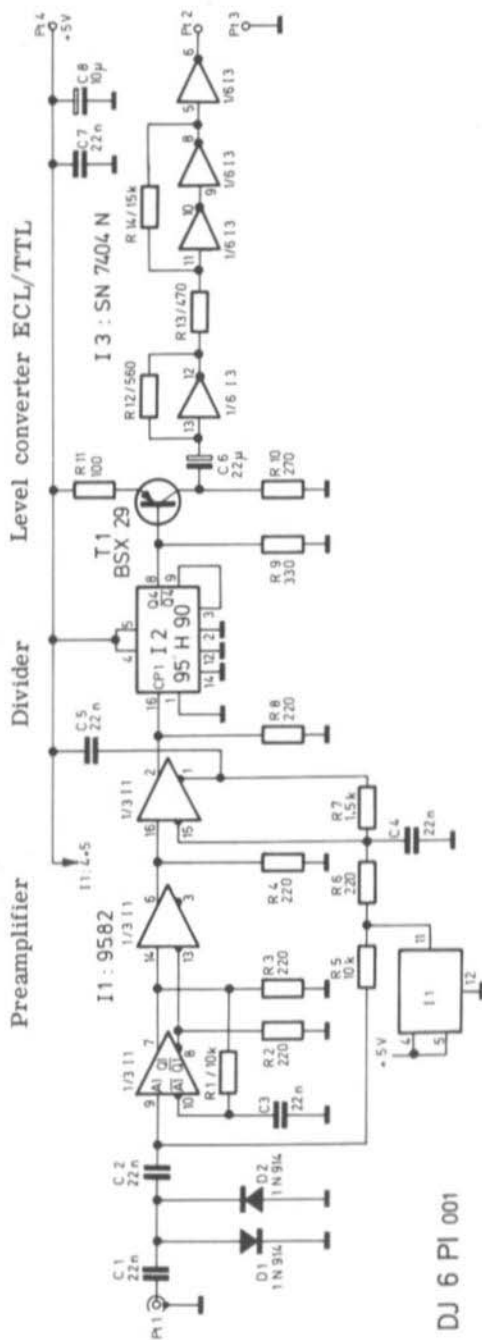
Fig. 2: Circuit diagram of the 10:1 prescaler with amplifier and level converter

DJ 6 PI 001

2. CIRCUIT DETAILS

Figure 2 shows the circuit diagram of the complete module. The three differential amplifiers accommodated in the integrated circuit 9582 are connected in series. The operating points are stabilized by an internal circuit by use of external resistors. Resistor R 1 has a great effect on the input sensitivity. The given circuit guarantees reliable operation and high sensitivity. Two anti-phase silicon planar diodes protect the input of the first amplifier against overvoltages. During normal operation, these diodes represent a total capacitance of 4 pF to 8 pF according to type. Together with the input capacitance of the amplifier and the circuit, this represents the previously mentioned capacitive component of the input impedance.

The values of the coupling and bypass capacitors are not critical. It is presumed that the input voltage requirements at frequencies of less than 1 MHz would be reduced if the values of the input coupling capacitors were increased (the author originally planned to use 68 nF). However, this is not important and it is sometimes difficult to obtain such high values in a form suitable for VHF.

The integrated circuit I 2, which is the actual 10:1 divider, only requires a few connections. The output (divided) frequency in the ECL voltage range (approx. 1 V peak-to-peak) can be taken from the output, at resistor R 9, or better still from the amplifier. The preamplifier of a frequency counter can be connected via coaxial cable to capacitor C 6. In this case, integrated circuit I 3 can be deleted.

The 6-stage integrated inverter SN 7404 N converts the pulses to TTL level. The voltage between connection points Pt 2 and Pt 3 can be directly connected to the counting gate or via a selector switch.

2.1. COMPONENT DETAILS

I 1: 9582 (Fairchild)
I 2: 95 H 90 or 9590 DC (Fairchild)
I 3: SN 7404 N

T 1: BSX 29 (Fairchild), BCY 78 (Valvo, Siemens), 2 N 711 (TI),
    BC 213 (TI) or similar silicon PNP transistor.

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

C 1 to C 5, C 7: 10 nF or more, ceramic disc capacitor 5 mm spacing.
C 6, C 8: 10 μF or more, tantalium electrolytic.

All resistors: Capless, non-inductive carbon resistors for a spacing of 10 mm.

3. CONSTRUCTION

The circuit given in Figure 2 is accommodated on a single-coated PC-board whose dimensions are only 85 mm x 35 mm. This board has been designated DJ 6 PI 001 (Fig. 3). If another PC-board is to be used, attention should be paid that short, straight and low-capacitive lanes are used with sufficient ground surface. Sockets for the ECL.-circuits and long wires for the resistors and capacitors should be avoided. Figure 4 shows a prototype of the module.
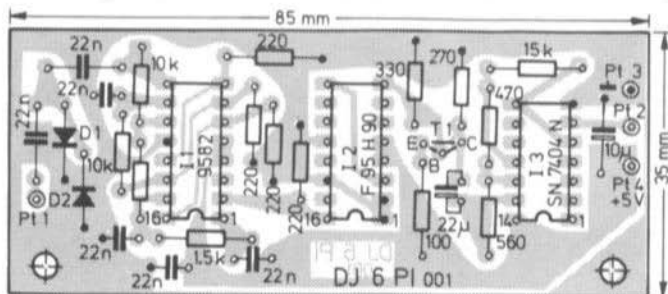
Fig. 3 : Component location plan of module DJ 6 PI 001
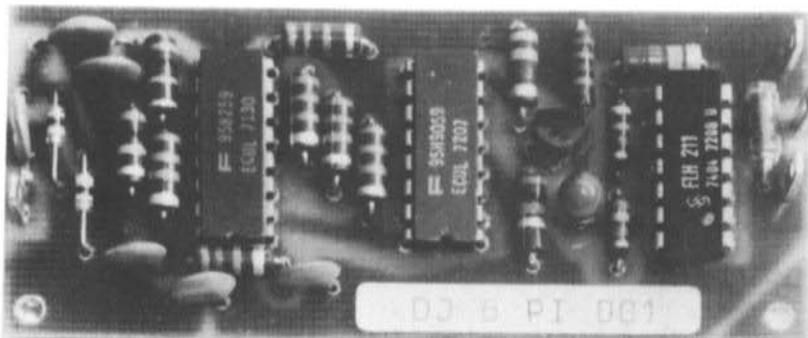
Fig. 4 : Prototype of the prescaler as given in Fig. 2

Since the two ECL-circuits each dissipate approximately 500 mW in heat (class A operation of the internal transistors) they will get so hot that it is hardly possible to touch them.

It is therefore favourable for module DJ 6 PI 001 to be mounted vertically to increase the air flow past these integrated circuits.

4. INTERCONNECTION WITH FREQUENCY COUNTERS

If the prescaler DJ 6 PI 001 is to be connected in front of an operative frequency counter, it should be installed in a small case and its ECL output connected to the input of the counter via coaxial cable. However, the module is usually installed in the counter.

Due to the very wide frequency range and the extremely high sensitivity, some readers might consider that it is only necessary for this module to be used as the input circuit of a counter and to do away with the previously used preamplifier. However, it should be noted that the counting time for a resolution of 100 Hz would be increased from 10 ms to 100 ms due to the division of 10:1, or from 1 s to 10 s for a resolution of 1 Hz. The demands on the stability and accuracy of the time base will also increase by the same value, namely by a

factor of 10. This means that, with the exception of special, limited applications, the module DJ 6 PI 001 will be used together with another preamplifier module such as DL 8 TM 003 (2), and connected as required to the counting gate. Figure 5 shows a simple circuit for this. The operating voltage is also switched in order to increase the life of the ECL-circuits.
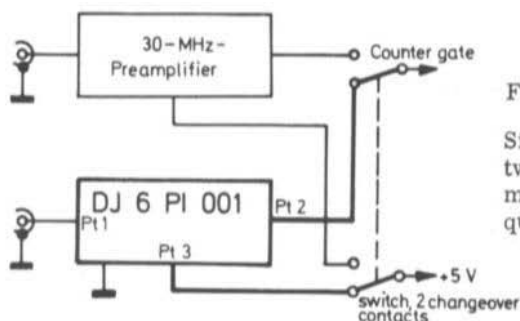
Fig. 5:

Simple circuit for connecting two separate preamplifier modules to a common frequency counter

If the lower frequency preamplifier only operates up to 30 MHz, a simple plastic switch will be sufficient. However, if higher frequencies are present, a ceramic switch or small RF relay should be used. Of course, both outputs could be switched electronically via an OR-circuit to the counting gate. Such a circuit is described with module DL 8 TM 003 (2). In this case, the three NAND-gates represent a NOR-gate in negative logic. The integrated circuit SN 7400 N will be sufficient up to frequencies of approximately 30 MHz, whereas the Schottky-type SN 74 S 00 N should be used for considerably higher frequencies.

The installation of this prescaler into a frequency counter DL 8 TM (1), (2), was carried out by the editorial staff as follows:

The output of the prescaler (Pt 2) was connected using a short, thick wire with the vacant flip-flop input Pt 308 of module DL 8 TM 003. After frequency division by 2 : 1 (which is necessary since the counter module DL 8 TM 002 additionally divides the clock frequency by 2), connections Pt 309 and Pt 304 are connected to the switch-over contacts of a relay. Another short, thick wire is passed from the change-over contact to connection Pt 208 of the counter board. An encapsuled miniature relay type RH-12 (as used in the linear amplifier module DC 6 HL 009) was used, mainly because it was cheaper than a Schottky-gate SN 74 S 00 N. A two-pole switch switches on the energizing voltage of the prescaler. The DC-voltage for the energizing coil of the relay is taken from the 12 V heater voltage of the crystal oven. This is rectified and filtered by any suitable silicon diode and an electrolytic capacitor of approximately 25 μF. Figure 6 shows the interconnection of such a frequency counter system in the form of a block diagram.

Finally we would like to state that the module DJ 6 PI 001 can also be used as a non-dividing preamplifier if the 10:1 divider 95 H 90 and resistor R 8 are deleted and a connection is made between pin 2 of the 9582 and the base of transistor T 1. The sensitivity is then far higher than any of the other pre-

amplifiers described in VHF COMMUNICATIONS. The upper frequency limit is, however, limited by transistor T 1 and especially by the level converter SN 7404 N.
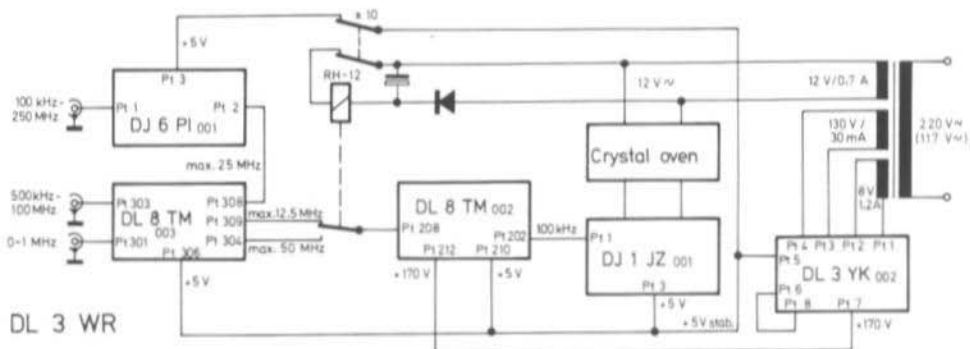
Fig. 6: Block diagram showing the interconnection of the described prescaler in the DL 8 TM frequency counter

For frequency counters that are able to process frequencies in excess of 60 MHz, it is recommended that the Schottky circuit SN 74 S 04 N be used for I 3.

5. REFERENCES

(1) W. R. Kritter: A 6-digit Frequency Counter for Frequencies between 1 Hz and Typically 100 MHz.
VHF COMMUNICATIONS 5 (1973), Edition 2, Pages 95-103

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

FM Limiter and Discriminator AD 4

Input frequency:	455 kHz
Limiting threshold:	100 µV
AM-Suppression:	40 dB
AF Output (dev. ± 3 kHz):	200—300 mV
Operating voltage:	9—15 V

Can be used with any receiver with an IF of 455 kHz and a bandwidth suitable for FM (12 kHz or more). Any IF between 400 kHz and 1 MHz is possible by adjusting coil and/ or changing value of parallel capacitor.

RECEIVE CONVERTER 432 MHz/28 MHz, MATCHING THE TRANSMIT CONVERTER DJ 6 ZZ 002

by J. Dahms, DC Ø DA

A matching receive converter for the transmit converter DJ 6 ZZ 002 described in (1) is to be given that converts the 70 cm band to an intermediate frequency of 28 to 30 MHz. Both modules have identical dimensions and are accommodated on double-coated PC-boards with air-spaced striplines. The receive converter uses the oscillator of the transmit portion so that true transceive operation on exactly the same frequency is guaranteed. Since it had been noted that the oscillator, multiplier and amplifier chain of the DJ 6 ZZ 002 is somewhat critical, a few modifications to the original circuit are to be given directly after this article.

1. CIRCUIT DETAILS

The receive converter was designed to satisfy the following demands:

The UHF selectivity should be good enough that no interference and intermodulation occurs even in the direct vicinity of a television transmitter; the sensitivity and gain should at least be as good as any of the other 70 cm converters described in VHF COMMUNICATIONS. This led to the previously mentioned striplines made from thick wire with air dielectric, using two preamplifier stages equipped with the low-noise and inexpensive germanium transistor AF 239 S with bandpass filter coupling. The circuit diagram of the receive converter is given in Figure 1. The input circuit is a capacitively shortened $\lambda/4$ circuit. In order to avoid difficulties when used with other modules, the ground surface was connected to the minus pole of the operating voltage in spite of the fact that germanium PNP transistors were used. For this reason, the emitters of the two PNP transistors were connected via resistors, which operate as RF chokes, to the positive operating voltage. It is then possible for all striplines to be connected directly to the ground surface at the cold end, and for the collectors of both PNP transistors to be directly connected to the line circuits. Both preamplifier circuits are identical.

The amplified UHF signal is fed via a coupling link to the mixer stage equipped with the silicon RF transistor BF 224; the local oscillator signal is fed via a small capacitor from module DJ 6 ZZ 002. An output coupling link of 20 mm in length made from 1 mm diameter copper wire is mounted at the same height with a spacing of 3 mm to the last 404 MHz circuit (L 5) of the transmit converter. The local oscillator voltage obtained in this manner is sufficient for the mixer of the receive converter and does not affect the conversion gain of the transmit mixer to any extent, presuming that the oscillator circuit is operating correctly.

The mixer is followed by a bandpass filter aligned for 28 MHz to 30 MHz. The 60 Ω output is formed using a capacitive voltage divider. If the inductive coupling of the bandpass filter is too loose with shortwave receivers having too low an input impedance, it would be possible for an additional coupling link or a 1 pF capacitor to be used.
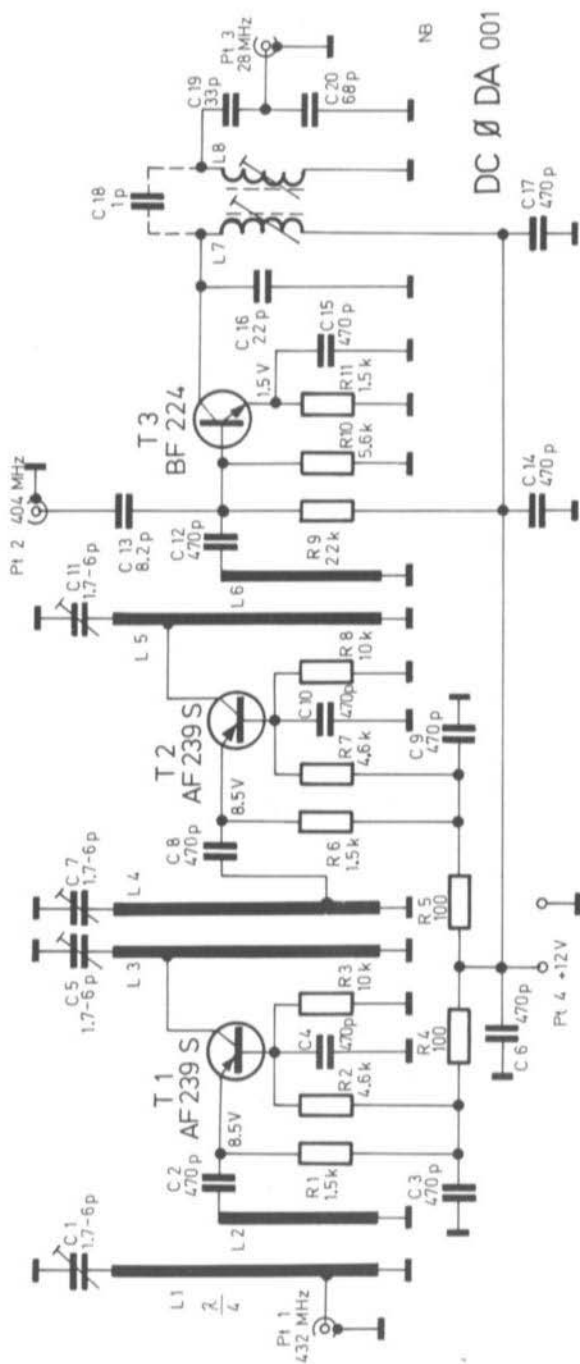
Fig. 1: Circuit diagram of a receive converter 432 MHz/28 MHz, matching transmit converter DJ 6 ZZ 002
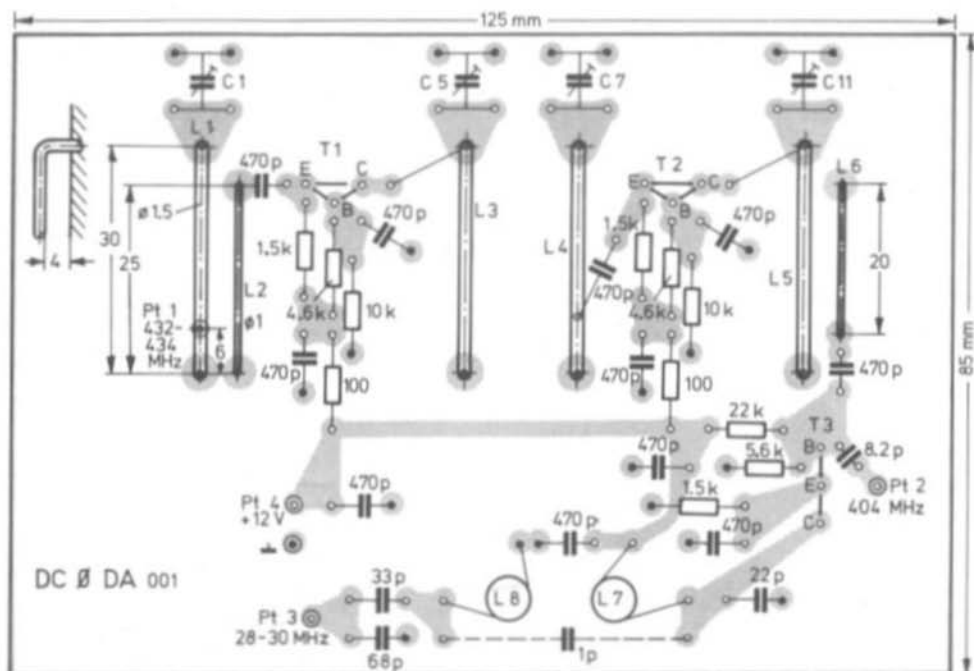
Fig. 2: Component locations of PC-board DC Ø DA 001
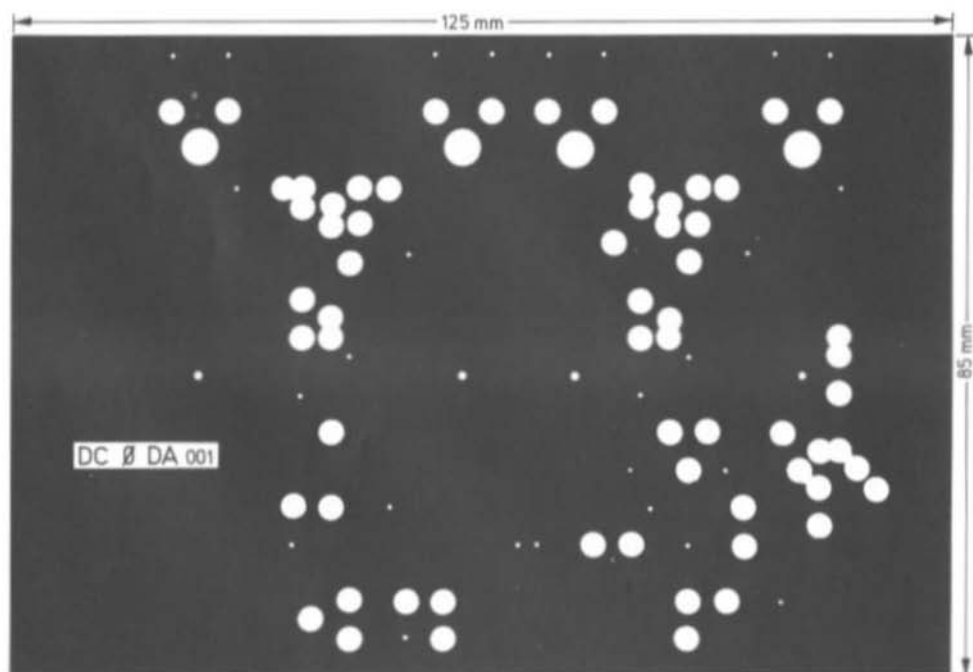
Fig. 3: Ground surface (component side) of PC-board DC Ø DA 001

2. CONSTRUCTION

.Figure 2 and 3 illustrate the double-coated PC-board DC Ø DA 001 whose dimensions are 125 mm x 85 mm. This module can be accommodated together with the transmit converter DJ 6 ZZ 002 (one above the other) in a common case. The stripline circuits are located sufficiently far from the edges of the printed circuit board that no detuning will occur. The components should be located on the ground-surface side. Since the resistors are partially used as RF chokes, they should be spaced one to two millimeters from the ground surface. Figure 4 shows a photograph of the author's prototype. For the three RF connections, thin coaxial cables should be connected to the connection pins or to the input Pt 1. Of course, sufficient room is provided on the PC-board for the mounting of coaxial sockets (BNC or smaller). The collector leads of the two amplifier transistors should be soldered directly to the stripline circuits.

If the given dimensions and spacing are adhered to, and high-quality small ceramic disc capacitors are used, the construction should represent no difficulties.

2.1. SPECIAL COMPONENTS

T 1, T 2: AF 239 S (Siemens, AEG-Tfk, Philips)
T 3   : BF 224 (TI) or BF 173 (AEG-Tfk, Siemens, Philips)

L 1, L 3, L 4, L 5: 1.5 mm dia. (15 AWG) silver-plated copper wire bent and soldered into place as given in the component location plan, effective length: 30 mm.

L 2, L 6:   1 mm dia. (18 AWG) silver-plated copper wire as given in the component location plan.

L 7, L 8:   16 turns of approx. 0.4 mm dia. (26 AWG) enamelled copper wire close wound on a 5 mm coil former with SW-core, wound in the same direction.

C 1, C 5, C 7, C 11: 1.7 - 6 pF air-spaced trimmer for PC-board mounting.

All other capacitors: Ceramic disc types for 5 mm spacing.
A spacing of 10 mm is available for all resistors.

3. ALIGNMENT

After connecting the operating voltage, the emitter voltages given in the circuit diagram should be checked. These voltages give information as to the currents flowing via the three transistors. This is followed by coarse alignment of the 28 MHz bandpass filter with the aid of a dipmeter. After connecting a shortwave receiver and local oscillator circuit (e.g. DJ 4 LB 003 if a transmit converter is not required, it should be possible to hear an input signal. The four 432 MHz circuits are uncritical and can be aligned for maximum .S-meter reading on the shortwave receiver. Trimmer C 7 is nearly completely inserted, whereas the other trimmers are inserted by approximately half. When receiving a constant signal, preferably at the centre of the band, it is very easy to obtain the most optimum alignment of the output bandpass filter and the UHF circuits.

4. EXPERIENCE WITH THE CONVERTER

Since the author is not able to carry out any measurements under laboratory conditions, he is only able to report on practical experience gained with this

converter. The described converter was compared under the same, difficult conditions of an adjacent television translator with such converters as the DC 6 HY, DJ 8 MF, DL 9 GU and a commercial German converter. Whereas hardly any difference could be found with respect to the sensitivity, the selectivity of the described converter was found to be far superior and lacked the "birdies" of the previously mentioned converters. The gain was more than sufficient for a 10 m transceiver (FT 200).
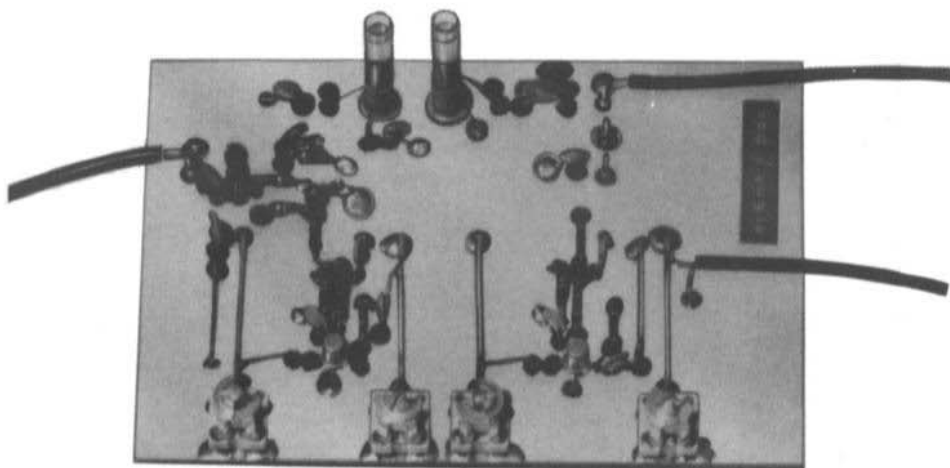
Fig. 4: Author's prototype of the receive converter

5. REFERENCES

(1) F. Weingärtner: A 28-432 MHz Transmit Converter with FET Mixer.
     VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 99-106.

(2) K. Eichel: Stripline Transverter for 70 cm
     VHF COMMUNICATIONS 2 (1970), Edition 4, Pages 224-239

(3) E. Krahe: A 432/144 MHz Converter with Silicon Transistor Complement
     VHF COMMUNICATIONS 1 (1969), Edition 2, Pages 65-72

(4) VHF-UHF Manual, RSGB, London, England.

AWG	mm	AWG	mm	AWG	mm
40	0.08	24	0.51	12	2.05
38	0.10	23	0.57	11	2.31
36	0.13	22	0.64	10	2.59
34	0.16	21	0.72	9	2.91
32	0.20	20	0.81	8	3.26
31	0.23	19	0.91	7	3.67
30	0.25	18	1.02	6	4.12
29	0.29	17	1.15	5	4.62
28	0.32	16	1.29	4	5.19
27	0.36	15	1.45	3	5.83
26	0.40	14	1.63	2	6.54
25	0.45	13	1.83	1	7.35

NOTES ON THE 28 MHz/432 MHz TRANSMIT CONVERTER DJ 6 ZZ 006

by J. Dahms, DC Ø DA

A few tips are to be given regarding some difficulties that have been encountered during the construction of the SSB transmit converter (10 m to 70 cm) that was described by DJ 6 ZZ in (1).

1. SEMICONDUCTORS

The transistor BF 224 only possesses a sufficiently high transit frequency f_T of more than 500 MHz when its collector current is in the range of at least 2 mA and max. 13 mA. This must be taken into consideration when aligning the drive from the local oscillator circuit. This is also important for the adjustment of the operating point of transistor T 7. Fundamentally speaking, the four transistors BF 224 should be selected to have a minimum current gain of at least 30. The construction is also easier with "true" UHF transistors, which are, of course, more expensive. Suitable types are the BFX 62 and BFX 89 (Siemens). Good results were also obtained with transistor type BF 357 (Texas Instruments). However, attention should be paid that this type has different emitter and base connections to that of the BF 224.

The best results in the output stage T 8 were obtained with selected 2 N 3866 transistors manufactured by Motorola. The selection is made for maximum output power in the circuit.

2. INDUCTANCES

The stripline circuit L 10 should be approximately 5 mm longer, since otherwise it is very difficult to align it for resonance. On the other hand, stripline L 11 should be slightly shorter (a few mm) than the details given in the article.

3. ALIGNMENT VALUES

Transistor T 2 can easily be "overdriven" by the oscillator and get hot. In this case, the emitter resistor R 4 should be increased from 15 Ω to 56 Ω - 100 Ω. In this case, the 202 MHz output signal of this stage should not be weaker since the efficiency will increase due to the optimized current-flow angle and the increased transit frequency.

A small ceramic trimmer potentiometer of 500 Ω can be used for resistors R 8 and R 9 in the push-pull mixer stage for optimizing the balance and conversion gain.

The final transistor T 8 is adjusted for a quiescent current of 3 mA to 4 mA at an operating voltage of 18 V with the aid of resistor R 16. The value of R 16 should possibly be decreased to 560 Ω in order to achieve this.

The 28 MHz drive signal should not exceed 400 mV to 500 mV at connection Pt 3 (RMS value with sinusoidal single-tone drive).

4. MEASURED VALUES

The modified transmit converter of the author (T 2 to T 4 and T 7: BF 357) provides an output power of 160 mW. The operating voltage of the final transistor is 18 V; the collector current of this transistor increases from a quiescent current of 38 mA to 75 mA.

The following voltage values were measured on this transmit converter with a tube voltmeter HRV 240 with RF-probe (diode 1 N 34 A):

Testpoint	RF Voltage (RMS)
Base of transistor T 2	2 V to 3 V
Base of transistor T 3	0.4 V
Base of transistor T 4	0.3 V
Line L 5, at branch to C 13	0.16 V
Line L 8, at the bend to C 19	0.2 V
Base of transistor T 7	0.12 V
Line L 10, at the branch to C 27	0.25 V
Line L 11, at the hot end	approx. 1 V

5. REFERENCES

F. Weingärtner: A 28 - 432 MHz Transmit Converter with FET Mixer
VHF COMMUNICATIONS 3 (1971), Edition 2, Pages 99-106.

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

by H. J. Franke, DK 1 PN

An integrated receiver system was introduced in (1) and (2). Figure 1 gives the latest stage of development in the form of a block diagram.
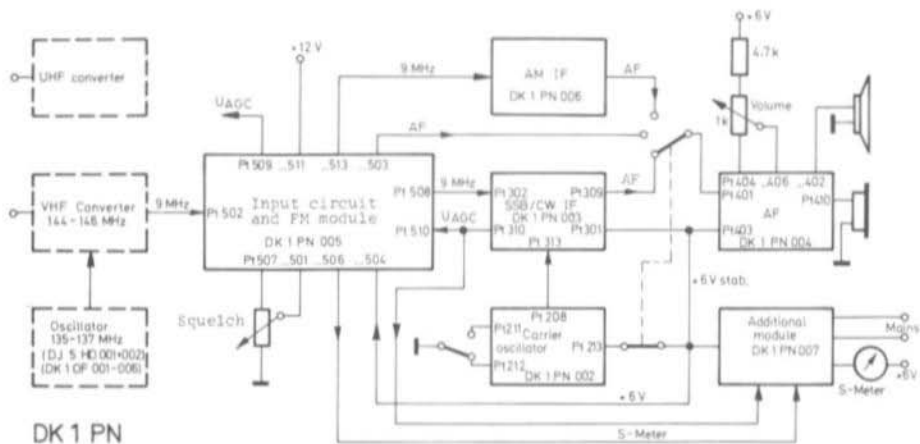
Fig. 1: Block diagram of an integrated receiver for AM, FM, SSB and CW

1. MODIFICATIONS TO THE SSB/CW IF PORTION DK 1 PN 003

The latest application notes issued by Plessey give a number of modifications to the original designs as well as further information. These modifications were made on the prototype constructed by the editorial staff and the following two improvements have been found:

1.1. RISETIME OF THE CONTROL VOLTAGE

The following modifications are necessary in order to guarantee that the control voltage is built up within a few ms without transient effects:

The largest value capacitor for the available room should be taken for capacitor C 322 so that the short-time peak current requirements of the SL 621 can be provided.

The coupling capacitor C 320 should be increased from 0.22 μF to a maximum of 3.3 μF so that the time constants for the input of the control voltage IC is reduced.

The coupling capacitor C 314 and possibly also C 312 should be reduced to an approximate value of 330 pF so that oscillations of the control voltage are avoided especially after the first second after the rise time.

1.2. AF OUTPUT VOLTAGE AND FREQUENCY RESPONSE

The coupling capacitor C 326 should be increased to $0.5 \mu F$ and be directly connected to pin 5 of I 303. This allows lower AF frequencies to be reproduced better and the AF voltage will become independent of the adjustment of potentiometer P 301.

2. REFERENCES

(1) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW
   VHF COMMUNICATIONS 5 (1970), Edition 1, Page 46

(2) H. J. Franke: An Integrated Receiver System for AM, FM, SSB and CW
   VHF COMMUNICATIONS 5 (1973), Edition 1, Pages 47-53.

New AM/FM 2 m Transmitter Module AT 222
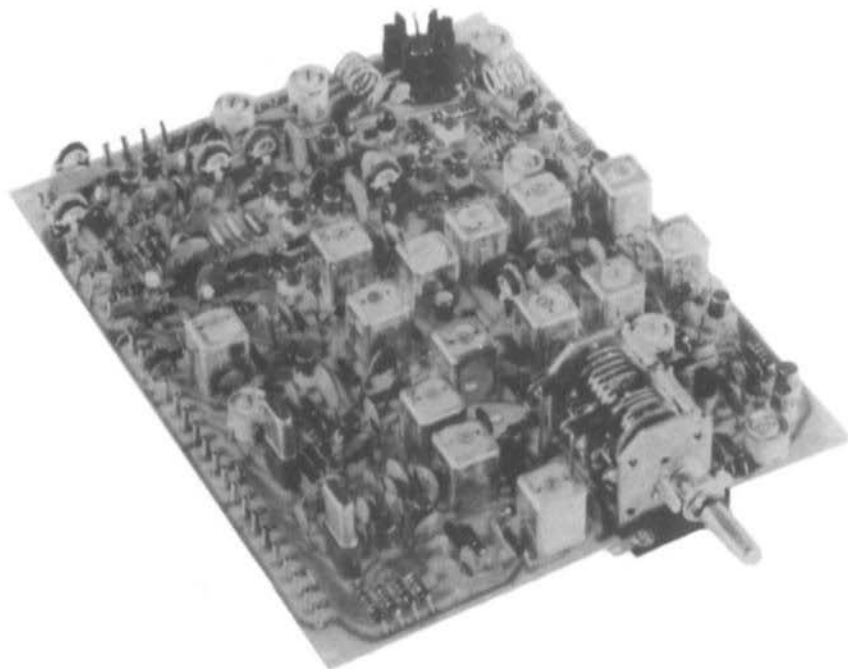
- Switchable to AM and FM
- Built-in synthesis VFO (stability better than 100 Hz/h at 144 MHz) or crystal control
- Built-in speech processor

- Output 1 W FM; 1 W PEP AM
- 12 VDC operation
- Matching linear amplifier for 8 W output available

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

Part III: The Carrier Oscillator

by H. J. Franke, DK 1 PN

The first module of this integrated receiver system (the SSB/CW portion) was described in (1). The following article is to describe the carrier oscillator which is just as suitable for transmit and receive applications.

1. CIRCUIT DESCRIPTION

Figure 1 gives the circuit diagram of the carrier oscillator. Two completely independent crystal oscillators are used whose outputs are connected together. The required carrier oscillator is switched on with the aid of a switch which grounds the emitter resistor of the transistor in question. The construction is very simple since no inductances are required. The output signal is not taken from the collector as is usually the case but from the base. The low harmonic voltage of the crystal is tapped off at this position and is virtually not affected by the rest of the circuit. The amplitude of the output voltage is adjusted with the aid of the feedback resistors R 205 or R 208 and by the voltage division ratio of capacitors C 202 and C 208. The given values are suitable for operation from a stabilized operating voltage of 6 V which is also required for the Plessey ICs. In this case, a low-harmonic RF voltage of 70 mV to 100 mV is available at an impedance of 50 Ω to 60 Ω. The voltage division ratio can be altered for other voltage values. However, the effective capacitance of the voltage divider should always amount to approximately 300 pF.

Fig. 1 : Circuit diagram
of the carrier oscillator

2. CONSTRUCTION

The carrier oscillator is also designed for installation in a TEKO box 2A. The PC-board DK 1 PN 002 has dimensions of 65 mm x 45 mm and can be provided with a 13-pin connector. Figure 2 gives the component location plan, and Fig. 3 gives a photograph of the author's prototype.

Fig. 2 : Component locations on PC-board DK 1 PN 002

Fig. 3 : Prototype of the carrier oscillator

2.1. SPECIAL COMPONENTS

T 201, T 202: BF 224, BF 173 or similar silicon NPN RF transistors

C 201, C 207, C 211: Approx. 10 nF ceramic capacitors
C 202: 560 pF - 680 pF styroflex capacitors
C 203: 680 pF styroflex capacitor
C 206, C 210: 220 pF styroflex capacitors
C 204, C 208: 22 pF ceramic capacitors
C 205, C 209: 4 - 20 pF ceramic or plastic foil trimmer 7 mm in dia.

Crystal: HC-25 U, fundamental mode, parallel resonance 13-pole connector. TEKO box 2A

3. REFERENCES

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

SECOND FRENCH EDITION OF VHF COMMUNICATIONS

The great popularity of the first omnibus edition of VHF COMMUNICATIONS lead not only to a reprint of same but also to the publication of a second omnibus edition. This edition is packed full of most of the latest designs published in VHF COMMUNICATIONS. Please inform your French speaking friends of the availability of both omnibus editions. Further details can be obtained from our French representative under the following address:

Mlle. Christiane Michel, F 5 SM, Les Pillès, F-89117 PARLY, France

OSCAR 6 OPERATIONS SUMMARY

by Perry I. Klein, K 3 JTE

May 25, 1973

OSCAR 6, first in the series of AMSAT-OSCAR-B missions, continues to operate satisfactorily after seven months in orbit. During the first week in May, the operating schedule was modified making the translator available for communications on Thursdays, Saturdays and Mondays, Greenwich Mean Time, and OFF on other days. The purpose of this change was to subject the nickel-cadmium battery to shorter, more frequent charge-discharge cycles. This procedure appears to be working well, and may well extend the useful lifetime of the spacecraft.

The temperature of the battery, which had risen to as high as 47 $^{\circ}$C (117 $^{\circ}$F) in early February and had been a cause for concern, has now dropped to a more comfortable value. There is now no reason to believe that we will not achieve the one-year planned lifetime, and possibly even exceed it, although we may find it necessary to further modify the operating schedule from time to time in an attempt to extend OSCAR's operating life to the maximum possible.

As many of the users of OSCAR 6 have noticed, we have inituated AMSAT official bulletin transmissions through the satellite translator, and these are generally given on the reference orbits (the first orbit of each Greenwich day, the same orbit during which the satellite is turned on briefly for telemetry recording on the OFF days). VE 2 BYG, K 1 HTV, W 3 TMZ and K 7 BBO have been serving as AMSAT Official Bulletin Stations for these transmissions, which include reports of special experiments and any changes in the operating schedule.

We can now identify the calls of some 1,100 stations who have made one or more contacts via OSCAR 6, very nearly half of these stations being outside the USA. These include stations from 59 countries, including CT2, DL, DU, EA, EI, F, FC, FP8, G, GI, GM, GW, HA, HB, HG, HK, I, JA, KP4, KX6, LA, LX, LU, LZ, OE, OH, OHØ, OK, ON, OX, OZ, PA, PY, SM, SP, SV, TF, UA1, UB5, UC2, UG6, UJ8, UR2, UW6, VE, VK, VKØ, VP2V, VP9, VU, W, XE, YU, ZE, ZL, ZS, 4X, 6Y, and 8P.

In the United States, all 50 states have been on at one time or another. The number of stations on from each of the US call areas is W1: 38; W2: 64; W3: 47; W4: 64; W5: 34; W6: 91; W7: 53; W8: 30; W9: 61; WØ: 52; KH6: 2; KL7: 5.

The operator apparently leading with the most reported satellite contacts is K 7 BBO with over 3,300 QSO's (Dave is averaging about 500 satellite QSO's a month), and several stations now have over 45 states confirmed through the satellite.

We have one report of operation from VE8, and activity in Africa, South America and the Far East seems to be very sparse. We urge persons in these regions to equip for satellite operation during the coming months.
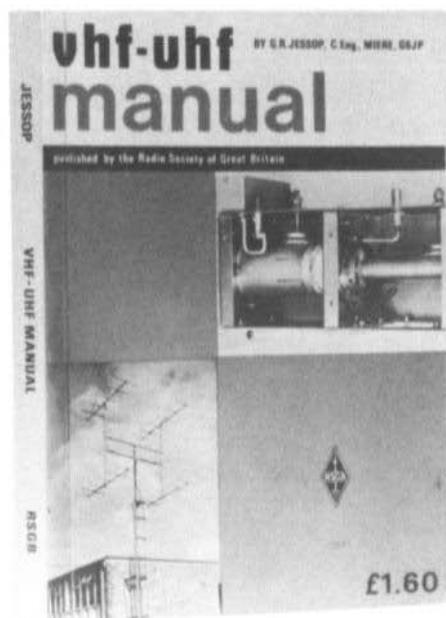
A special message commemorating World Telecommunication Day was transmitted over OSCAR 6's Codestore message storage system on May 17 using the 29.45 MHz beacon.

The first reported aeronautical mobile communication via OSCAR 6 was by W 6 OAL, who reported working K 7 BBO April 27th on Orbit 2431 over a distance of approximately 5,000 miles. The transmitter aboard the aircraft was a Gonset Sidewinder operating on SSB with less than 5 watts PEP to a simple whip antenna.

W 2 GN for the past several months has been quite successful with an automobile OSCAR terminal, and has been very popular with his special state DX-peditions to Vermont and Kentucky to put these rare states on the air. AMSAT is encouraging more mobile terminal operation with OSCAR 6. In particular, operation from small private aircraft, small boats and automobiles (especially on SSB) would provide a very effective demonstration of the usefulness of amateur satellites for small-terminal communication. In addition, operation using totally hand-held equipment or operation from a bicycle or motorcycle would be impressive "firsts", and we urge anyone interested in these activites to give it a try. If possible, make tape recordings of some of the mobile contacts and send them to AMSAT, P.O. Box 27, Washington, D.C. 20044 U.S.A.

VHF — UHF MANUAL

MINIATURE RECEIVE CONVERTER FOR 432 MHz/144 MHz FOR PORTABLE OPERATION AND DF-HUNTS

by G. Hoffschildt, DL 9 FX

A miniature receive converter is to be described which allows adequate sensitivity and stability to be achieved with a minimum of expense. It represents a very cheap manner of getting started on the 70 cm band. The author designed this converter as a plug-on unit for his 2 m receiver for 70 cm DF-hunts (fox hunts). The converter is enclosed in a small case (54 mm x 80 mm x 35 mm) and sufficient room is provided to accommodate a 9 V battery. This means that it is completely independent and only needs to be connected between the 70 cm antenna and the 2 m receiver. The weight including case and battery is only 150 g which means that it is very suitable for portable applications.

The author has two special 2 m receivers for DF-fox hunts under development which will be described in later editions of VHF COMMUNICATIONS. However, the described converter can be used with any 2 m receiver.

1. CIRCUIT DETAILS

The simple circuit diagram of this 70 cm converter is shown in Figure 1. However, due to the simplicity of the circuit, it is necessary for special UHF transistors to be used. The transistor type used in the input and mixer stages is suitable for use up to frequencies of 1 GHz. This transistor exhibits a power gain of 12.5 dB at 800 MHz and a noise figure of 5 dB (typical). An especially fast switching transistor is used in the oscillator/multiplier stage so that harmonics of sufficient strength are generated. The transistor type 2 N 709 exhibits a storage time t_S of only 6 ns.
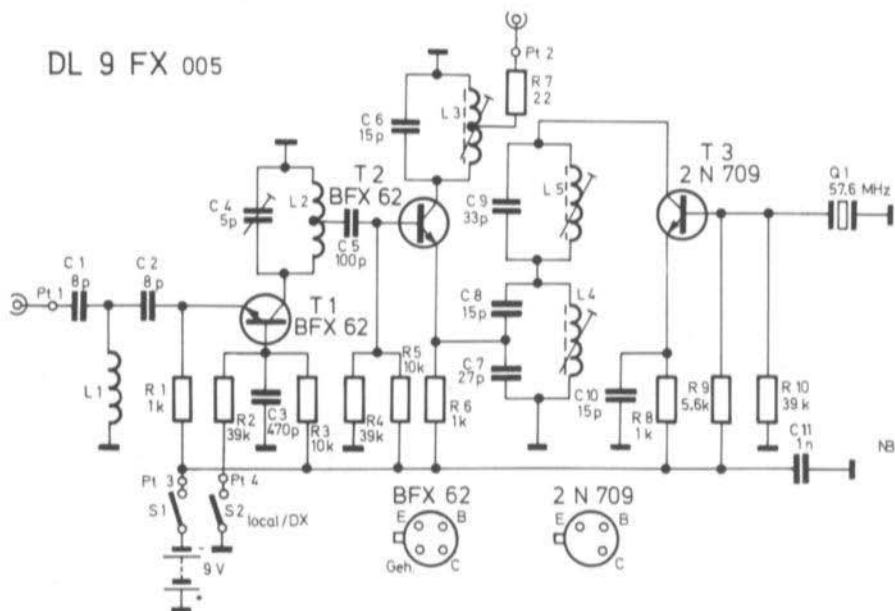

Fig. 1: Circuit diagram of a simple 432 MHz/144 MHz receive converter

The preamplifier transistor T 1 operates in a non-critical common-base circuit. The emitter is fed with the input signal via a high-pass link comprising capacitors C 1 and C 2 as well as inductance L 1. The cut-off frequency of this high-pass filter is in the order of 400 MHz so that interference signals, especially from 2 m transmitters and VHF broadcast transmitters are suppressed. A high-pass filter is not so critical with respect to neutralization of the preamplifier stage in this frequency range as a resonant circuit. In addition to this, the construction is simpler and no alignment is required.

Further selectivity is provided between the preamplifier and mixer stage in a circuit comprising a printed inductance. The alignment is made with trimmer capacitor C 4. The mixer transistor T 2 is fed with the 288 MHz oscillator voltage from a capacitive voltage divider connected to the emitter. Capacitor C 7 of the voltage divider bypasses the emitter for the required frequency of 432 MHz. The 144 MHz intermediate frequency circuit is connected to the collector. Resistor R 7 simplifies the matching to the subsequent 2 m receiver and ensures a constant gain over the whole 2 m band.

The oscillator operates at 57.6 MHz. The collector circuit comprising inductance L 5 must be tuned to this frequency whereas the other resonant circuit comprising inductance L 4 is aligned to the 5th harmonic (288 MHz). This circuit filters out the required frequency and feeds it to the emitter of the mixer.

The ground surface of the PC-board is connected to the positive pole of the operating voltage so that all "cold" ends of the coils can be directly grounded. When the full gain of the converter is not required (when near to the transmitter during DF-fox hunts or with very strong local stations) it is possible to switch off the preamplifier stage. In this case, the base resistor R 2 (+U_B) is disconnected from ground by switch S 2.

Fig. 2: Printed circuit board DL 9 FX 005

2. CONSTRUCTION

The described 70 cm converter is accommodated on a single-coated PC-board of 75 mm x 50 mm. It has been designated DL 9 FX 005. It possesses a cut-out for the 9 V battery. Figure 2 shows the component locations and conductor lanes. An 11 mm hole should be drilled at the top and bottom of the case for mounting the input BNC connector and output BNC plug. If a PL 259 connector is used, a 10.4 mm hole will be required. The converter is usually directly plugged on to the 2 m receiver.

The two switches (sliding or toggle types) are either mounted on one of the sides, or adjacent to the input socket. The author's prototype shown in Fig. 3 gives some idea as to the construction of the unit. The completed PC-board is mounted with the aid of soldering tags fixed below the input and output socket. These solder tags are soldered to the ground surface on the input side and glued into place at the IF output.
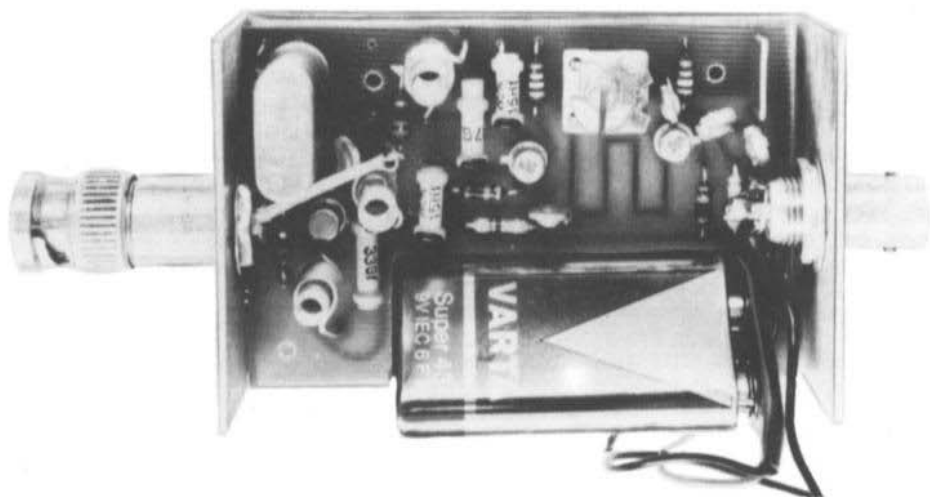

Fig. 3 : Author's prototype of the simple 70 cm converter

2.1. COMPONENT DETAILS

T 1: BFX 62 (Siemens) or BF 180 (Philips)
T 2: BFX 62 (Siemens) or BF 200 (AEG-Tfk, Philips)
T 3: 2 N 709 (Texas Instruments, Philips)

L 1: Silver-plated copper wire of 1 mm dia. (18 AWG) bent in a square with 10 mm spacing between the ends and mounted 8 mm above the board.
L 2: Printed inductance on the board.
L 3: 3.5 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 4.3 mm diameter coilformer with VHF-core; output tap: 0.25 turns from the cold end.
L 4: 0.75 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on coilformer and core as for L 3.
L 5: 7.5 turns of 1 mm dia. (18 AWG) wound on a 4.3 mm coilformer with SW-core (pink).

C 4: 1.7 - 6 pF airspaced trimmer with four connections.

All other capacitors: Ceramic disc types for 5 mm spacing,
C 1, C 2 and C 10: TC = -750 x 10^{-6}/°C (violet dot).

All resistors should be capless, carbon resistors.

1 crystal 57.600 MHz HC-6/U, plugged into hollow rivets.
1 case 54 x 35 x 80 mm
2 miniature sliding or toggle switches.
1 BNC connector, 1 BNC plug.

3. ALIGNMENT

The oscillator is firstly brought to oscillation by aligning the core of inductance
L 5. The frequency can be checked with the aid of a dipmeter. The core is
rotated into the coil until oscillation is commenced reliably on switching on the
operating voltage. The 70 cm converter is then connected to a 2 m receiver
and the IF circuit aligned for maximum noise at the centre of the band with
the core of inductance L 3.

A weak 70 cm signal should now be fed to the input. If no beacon, transmitter,
calibration spectrum generator (1) or 2 m transmitter with VFO is available,
a dipmeter can be tuned to 108 MHz and brought to the vicinity of the converter.
The intermediate circuit comprising trimmer C 4 as well as the 288 MHz circuit
with inductance L 4 can now be aligned for maximum S-meter reading. Finally,
inductance L 5 (oscillator circuit) is now carefully aligned for highest S-meter
reading. However, attention should be paid that no position is selected where
the noise in the receiver increases in jumps (self-oscillation of the oscillator).

4. REFERENCES

(1) K. Eichel: Simple VHF-UHF Calibration Spectrum Generator
VHF COMMUNICATIONS 2 (1970), Edition 4, Pages 240-243.

AF Amplifier AA1

Output power:	1.5 W (at 12 V)
Distortion factor:	1 % (12 V, 1 kHz, 1 W)
Frequency response:	100 Hz to 15 kHz (-3dB)
Sensitivity:	12 mV (for 1 W, 12 V, 1 kHz)
Operating voltage:	12 V—15 V
Current drain:	3—230 mA

Small integrated AF amplifier with an output power of
1.5 W that is suitable for multiple applications.
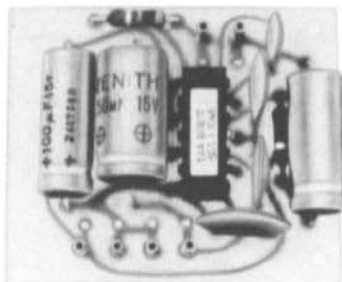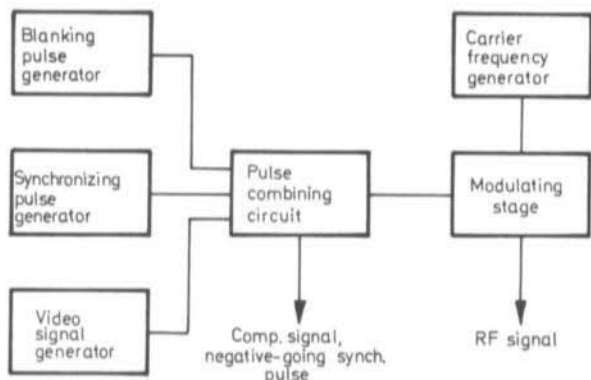
TV PATTERN GENERATOR

by K. Wilk, DC 6 YF

A relatively simple television pattern generator is to be described that satisfies all requirements required by the CCIR television standard described in (1). A pattern generator is very useful to the ATV enthusiast, or TV repairman of which there are a great number amongst the readers of this magazine. A pattern generator does not only simplify matters in the construction, alignment and repair of television systems, but also offers the ATV amateur a welcome composite video signal for transmission purposes.

The complete unit comprises a power transformer and the following four modules:

> Pulse centre DC 6 YF 001
> Pattern generator DC 6 YF 002
> Power supply DC 6 YF 003
> Grid and dot generator DC 6 YF 004

Fig. 1:
Block diagram of
the pattern generator
module

1. PATTERN GENERATOR DC 6 YF 002

The block diagram of the individual circuits accommodated on this printed circuit board are given in Figure 1. The three input stages used for generating the blanking and synchronizing pulses as well as video signal also provide circuits for combining the following signals:

Horizontal and vertical blanking pulses; horizontal and vertical synchronizing pulses, as well as video information.

These pulses are obtained from the pulse centre and are processed in a similar manner in integrated digital TTL-circuits. The basic video information is horizontal and vertical bars. They are in fact square-wave pulses whose frequency is an even multiple of the deflection frequency. These pulses are taken from the frequency dividers of the pulse centre DC 6 YF 001.

The subsequent "pulse combining stage" combines the output signals of the input stages, and provides a composite video signal for driving the modulator. This module is also equipped with a carrier frequency generator for the video car-

rier. The output signal of the modulator stage is an amplitude-modulated double-sideband signal by which the maximum amplitude corresponds to that of the synchronizing pulses.

1.1. THE MULTIPLE DRIVER SN 7406 N

The integrated TTL-circuit type SN 7406 N contains 6 driver stages. It is thought that several details are of interest. Figure 2 shows the connection diagram of the dual-in-line case. The integrated circuit is equipped with 6 independent inverters with TTL inputs. The power capabilities of the outputs, which are in the form of "open collectors", is considerable: The maximum output voltage corresponds to 30 V and an output, collector current of 40 mA is permissible. One application for this circuit is to obtain a transition from TTL levels to higher output voltages.
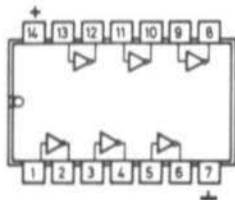
Fig. 2 :
Connections of the integrated
circuit SN 7406 N

1.2. GENERATION OF THE COMPOSITE VIDEO SIGNAL

The composite video signal, e.g. the complete video signal with all synchronizing pulses, should be directly available from the pattern generator. This signal is also required in a somewhat different form for driving the modulator. Two stages are used for this that are the three stages for generation of the blanking and synchronizing pulses and video signal, together with the pulse combining stage. The circuit diagram of these circuits is given in Figure 3.

The inverters I 203 a - f (SN 7406 N) with open collectors at the output provide the blanking, synchronizing and video pulses for the mixer. Their amplitudes must be in a certain relationship to another so that the outputs of the circuits provide impulses approximately equal to those laid down in the TV-standard. The input stages for the blanking and synchronizing pulses are built up in the same manner. The NAND-gates I 202 a and I 202 b (SN 7400 N) operate as OR-gate and mix the horizontal and vertical pulses in such a manner so that both vertical and horizontal impulses are present in combined form at the input of inverter I 203 a and I 203 b (SN 7406 N).

The integrated circuit I 203 f is provided with video impulses from the video processing stage with gates I 201 a, b, c and d (SN 7400 N), I 202 c and d and the inverter I 203 c. These signals are fed via a selector switch so that the required video signal can be selected.

The pulse-combining stage combines the individual signals to form a composite video signal. This stage is equipped with diode D 201 to D 204 and transistor T 201. The diode having the highest voltage at the anode will conduct. The voltage divider 5.6 kΩ/1.2 kΩ with diode D 204 ensures that the required white level is present at the RF output if no pulses are present. The combined pulses are triggered with transistor T 201. The video signal for the modulator can be taken from the emitter of transistor T 202. A video signal with negative going synchronizing pulses is available at the collector which is passed through the buffer stage equipped with transistor T 203 and provides an output signal of approximately 1.4 V_{pp} across 75 Ω at the video output Pt 217.
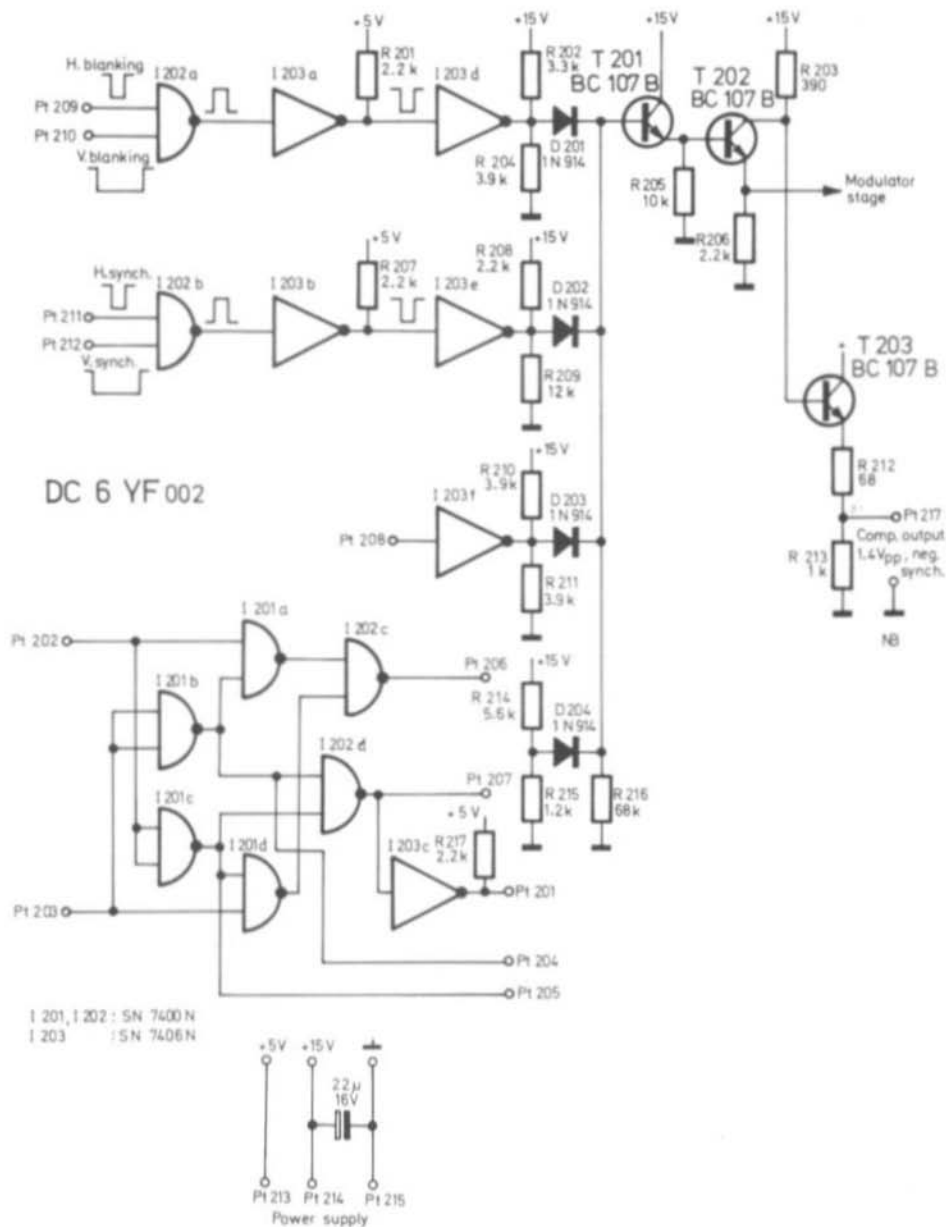
Fig. 3: Generation of the composite video signal, circuit diagram

The input pulses for the video information are, as previously mentioned, taken from the frequency dividers of the pulse centre. They will appear as a grid subsequent to a OR-processing or as a chessboard after an EXCLUSIVE-OR-processing. In this case, the horizontal and vertical pulses are alternately blanked. The following signals will be available with the aid of an inverter stage and the inverter stages already available as buffers for the vertical and horizontal bars:

Horizontal bars	Negative grid	Chessboard
Vertical bars	Positive grid	

The required pattern is selected with the aid of a switch and fed to the input of I 203 f.

In addition to this, it is possible for various frequencies to be selected for the bars, which means that the number of bars can be selected. This is valid both for the vertical and horizontal bars.
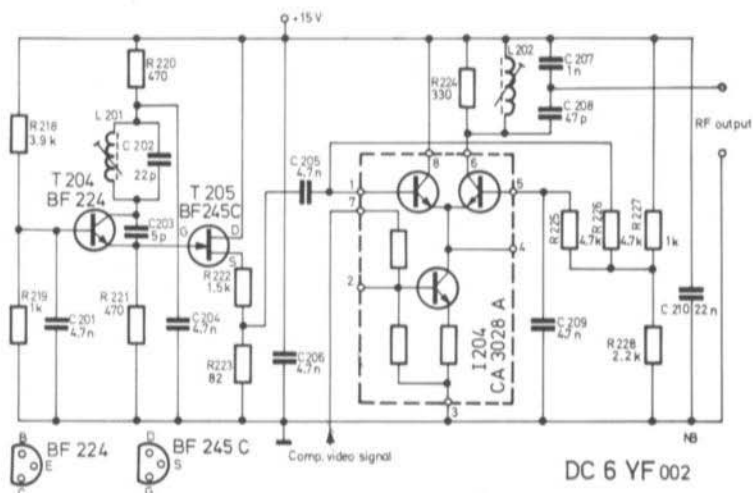

Fig. 4: Circuit diagram of the carrier frequency oscillator and modulator

1.3. CARRIER FREQUENCY GENERATOR AND MODULATOR

Figure 4 shows the circuit diagram of these stages. The carrier frequency generator is equipped with the oscillator transistor T 204 and the buffer T 205 and offers no special features. The RF-signal is fed to the RF input of the modulator at pin 1 of the analog integrated circuit CA 3028 A (I 204). The circuit roughly corresponds to the recommendations of the manufacturers. The composite video signal is present at pin 7. The RF amplitudes of this signal are as follows:

White level	10%	Synchronizing pulse level	75%
Video signal	65%	Blanking pulse level	100%

The RF-signal can be taken from the collector circuit and fed via a coaxial cable to the TV-receiver or TV-system. In the author's prototype, both RF stages are aligned for a video carrier of 48.25 MHz which corresponds to CCIR VHF channel 2 (TV band I). If required, the frequency can be modified for any required channel or modified to the IF frequency of 38.9 MHz.
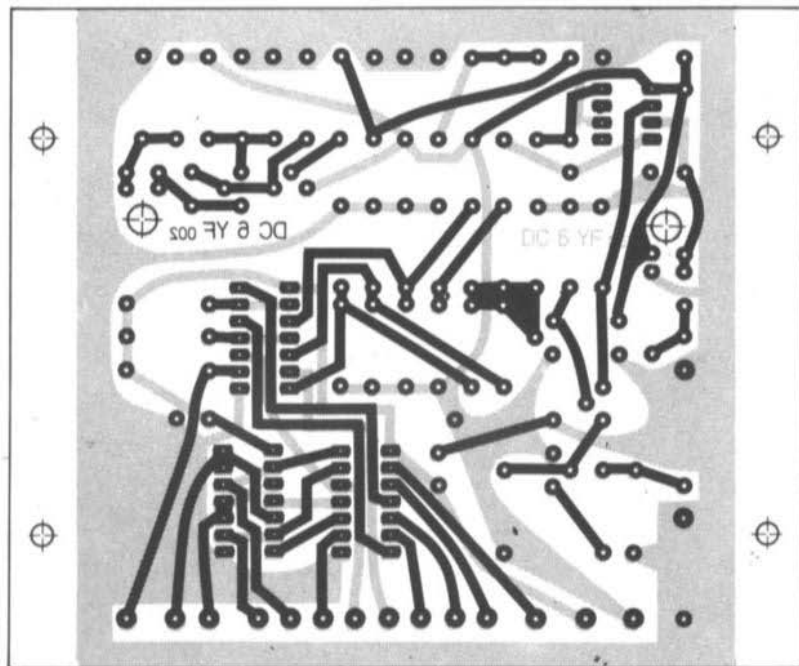
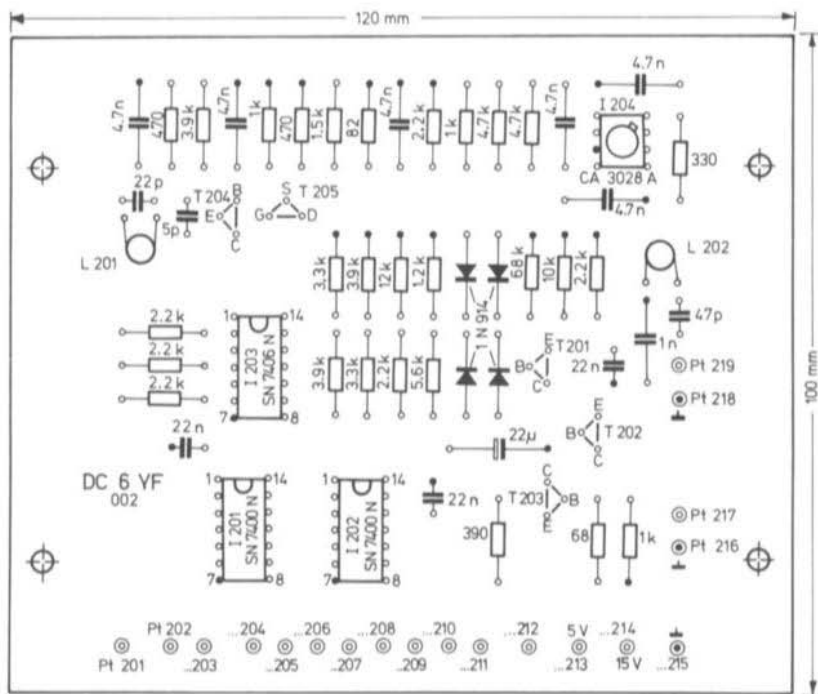Fig. 5 : Printed circuit board DC 6 YF 002, double-coated, through-contacts

Fig. 6 : Component locations of PC-board DC 6 YF 002

- 181 -

1.4. POWER SUPPLY

The pattern generator and pulse centre modules require a positive voltage of 5 V at approximately 450 mA for the TTL-circuits, and +15 V at 50 mA for the other stages. A suitable power supply has been designed and has been designated DC 6 YF 003 (see section 2.).

1.5. CONSTRUCTION OF THE PATTERN GENERATOR

The pattern generator is accommodated on a double-coated PC-board with through-contacts. This PC-board has the dimensions 120 mm x 100 mm. The PC-board, which has been designated DC 6 YF 002 is shown in Figure 5; the component locations in Figure 6. A photograph of the author's prototype is given in Figure 7.

The following connection points are as follows (H= horizontal; V= vertical):

Pt 201: Output, grid pattern
Pt 202: Input, horizontal bars
Pt 203: Input, vertical bars
Pt 204: Output, V. bar pattern
Pt 205: Output, H. bar pattern
Pt 206: Output, chessboard pattern
Pt 207: Output, white grid pattern
Pt 208: Input, video signal
Pt 209: Input, H. blanking pulses
Pt 210: Input, V. blanking pulses

Pt 211: Input, H. synch. pulses
Pt 212: Input, V. synch. pulses
Pt 213: + 5 V
Pt 214: + 15 V
Pt 215: 0 V
Pt 216: Composite video output (0)
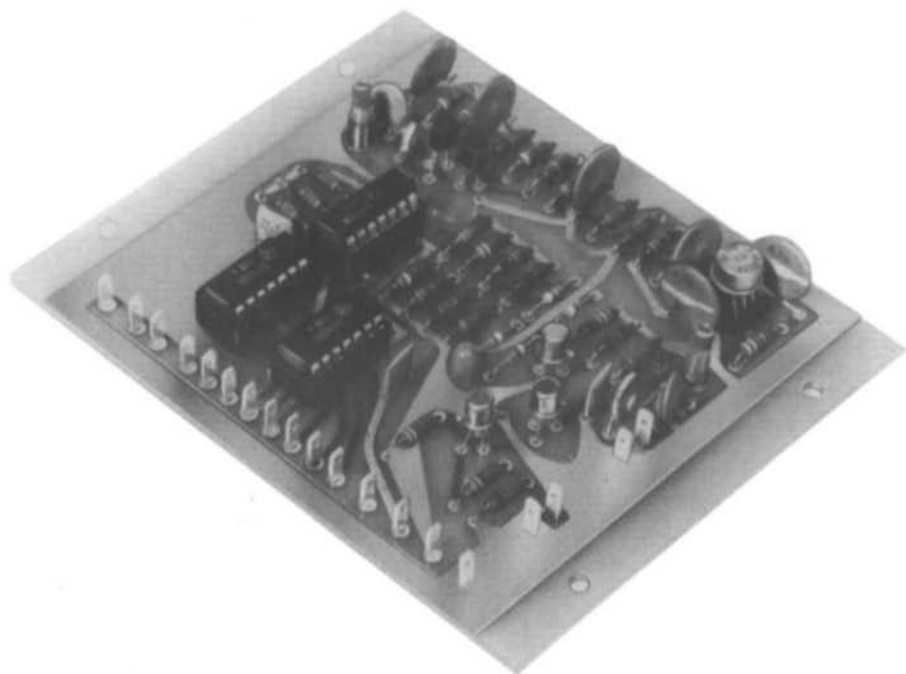Pt 217: Composite video output
Pt 218: RF-output (0)
Pt 219: RF-output

Fig. 7: Photograph of the author's prototype

1.6. COMPONENT DETAILS

I 201, I 202: SN 7400 N (Texas Instruments)   T 201, T 202, T 203: BC 107 B
I 203: SN 7406 N (TI)   T 204: BF 224
I 204: CA 3028 A (RCA)   T 205: BF 245 C, W 245 C

D 201 - D 204: 1 N 914 or similar

Ceramic capacitors: 6 x 4.7 nF; 3 x 22 nF

1 tantalium drop-type electrolytic 22 μF/16 V

L 201: 10 turns of 0.5 mm dia. (24 AWG) enamelled copper wire
      wound on a 4.3 mm coilformer with HF core (red).
L 202: 5 turns as for L 201

1.7. ALIGNMENT OF THE INDUCTANCES

For the alignment of the two resonant circuits, the pattern generator module
is connected on its own. The TTL-circuits should preferably be removed or
soldered into place later. After this, the two resonant circuits are aligned to
the required frequency. The coil data given refers to a frequency of 48.25 MHz.
It will be necessary to increase the number of turns to obtain an output fre-
quency of 38.9 MHz. The adjustment of inductance L 2 is relatively uncritical.

2. POWER SUPPLY DC 6 YF 003

The power supply module DC 6 YF 003 has been developed for the pattern ge-
nerator but is also suitable for universal use. The power supply module com-
prises two completely independent parts, whose circuit and construction are
identical. Each circuit corresponds to a conventional stabilizer circuit. The
PC-board can be equipped as required to satisfy the various requirements as
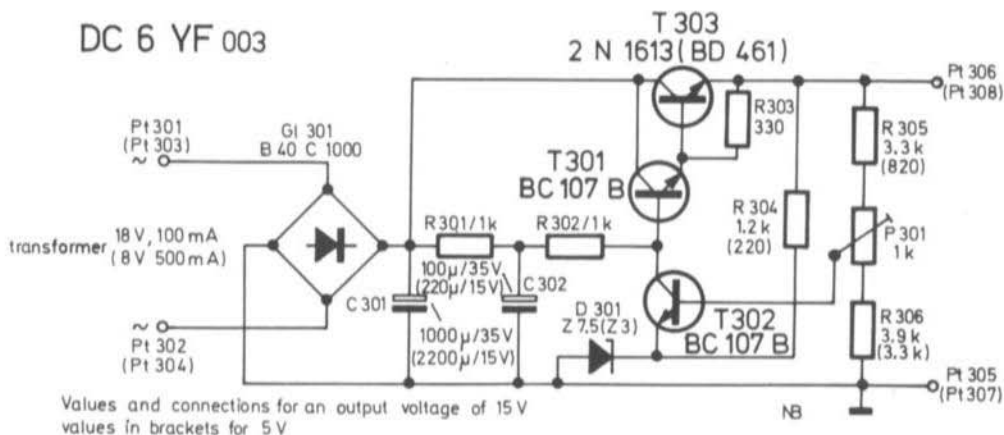to output voltage and current.

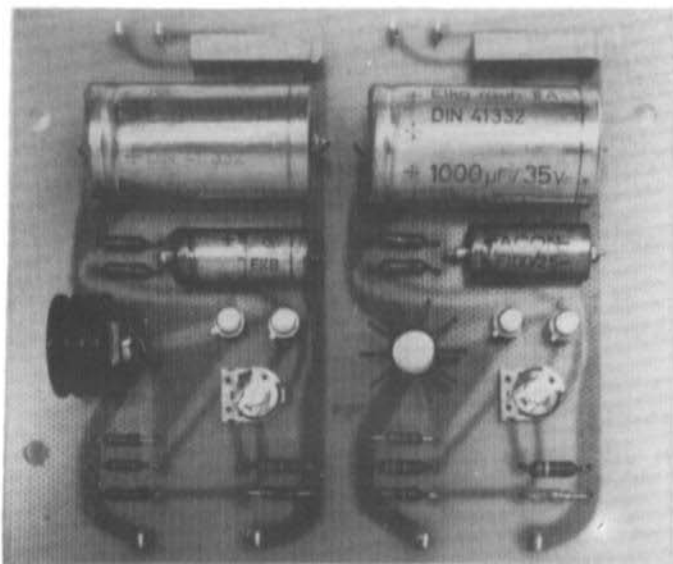Fig. 8: Circuit diagram of the power supply

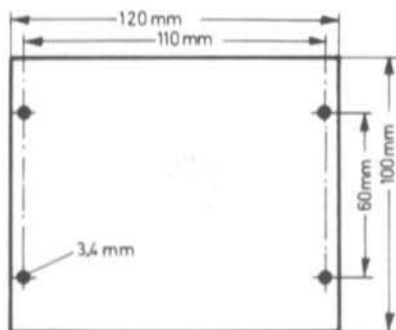Fig. 9 : Photograph of the author's prototype

Fig. 10 :
Dimensions of the heat sink

The circuit diagram given in Figure 8 gives the values for output voltages of 5 V at a current of approximately 450 mA and 15 V with a current of approximately 50 mA. These values are suitable for operating the pattern generator. It is necessary for the pass-transistor to be provided with cooling fins or a heat sink in both cases. In the author's prototype (Figure 9), a special heat sink was used for the pass-transistor of the 5 V circuit which is not readily available. It is therefore recommended that a heat sink be used that has roughly the same dimensions as the PC-board and fits into the mounting holes on the PC-board. The power transistor is then mounted into the heat sink in an insulated manner. The provision of a generous heat sink is also advisable when

the pattern generator is to be extended later. The dimensions of a 2 mm thick aluminum plate suitable for use as a heat sink is given in Figure 10. The holes for the transistor have not been marked, since various types of transistors can be used. In the author's prototype, a power transistor type BD 461 manufactured by Texas Instruments was used as pass-transistor in the 5 V supply. This is a plastic transistor in a flat SOT 32 case with mounting hole.

2.1. CONSTRUCTION

The printed circuit board DC 6 YF 003 is given with component location plan in Figure 11. The board is single-coated and the dimensions are 120 mm x 100 mm which are the same as those of PC-board DC 6 YF 001 and DC 6 YF 002.
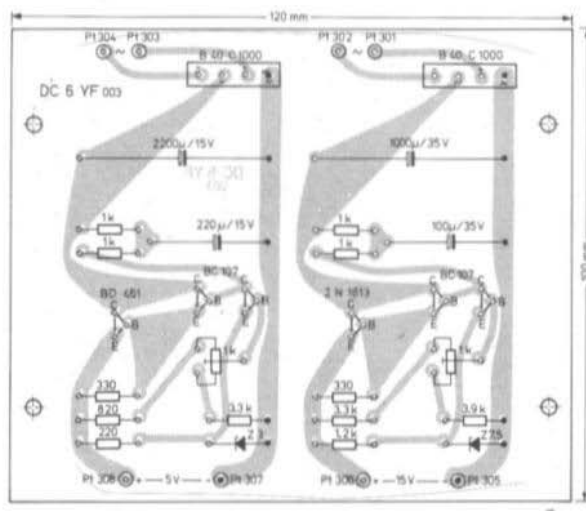

Fig. 11: PC-board DC 6 YF 003 and component locations

The connection points on the board are:

Pt 301 }
Pt 302 }   18 V AC (from power transformer)

Pt 303 }
Pt 304 }   8 V AC (from power transformer)

Pt 305 (0)
Pt 306 (+)   15 V output

Pt 307 (0)
Pt 308 (+)   5 V output

2.2. COMPONENT DETAILS

RECT 301: Silicon bridge rectifier B 40 C 1000
T 301, T 302: BC 107 B or similar silicon transistor with high current gain
T 303: 2 N 1613 or BD 461

P 301: 1 kΩ trimmer potentiometer

D 301: Zener diode ZF 7.5 or ZF 3 or BZY 85/C7V5 or BZY 85/C3

C 301: 1000 μF/35 V or 2200 μF/15 V
C 302: 10 μF/35 V or 220 μF/15 V

2.3. ADJUSTMENT

The 5 V power supply is adjusted to the nominal voltage with a load resistor of approximately 18 Ω (2 W), and the 15 V supply with a resistor of 270 Ω (1 W) (trimmer resistor P 301).

3. THE COMPLETE PATTERN GENERATOR

Figure 12 gives the interconnections between the various modules.
The following components are also required:
Power transformer; a selector switch for horizontal and vertical bars, as well as for the type of pattern.
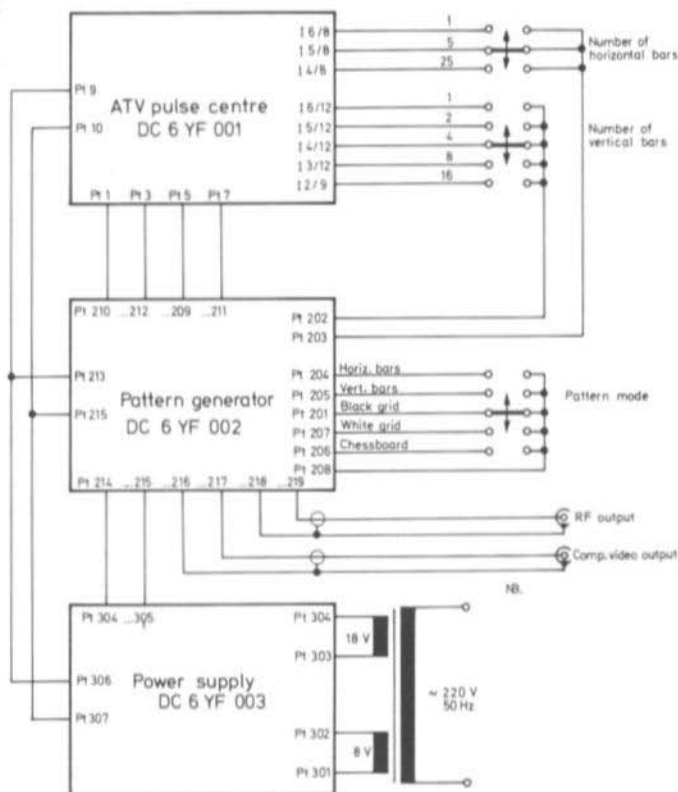

Fig. 12: Block diagram of the complete pattern generator
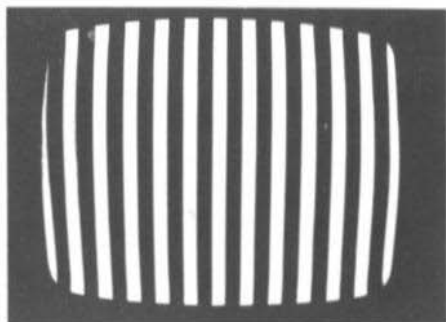
Fig. 13 : Horizontal bar pattern
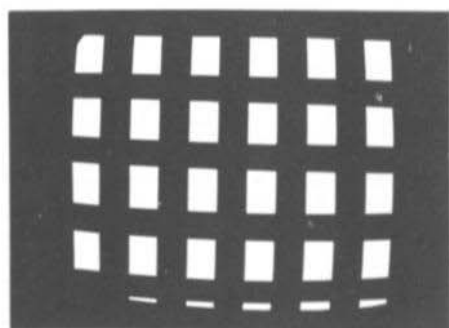

Fig. 14 : Vertical bar pattern
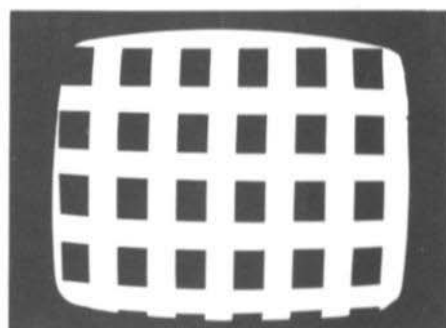

Fig. 15 : Black grid pattern
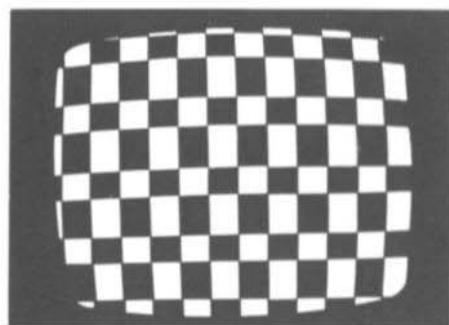

Fig. 16 : White grid pattern
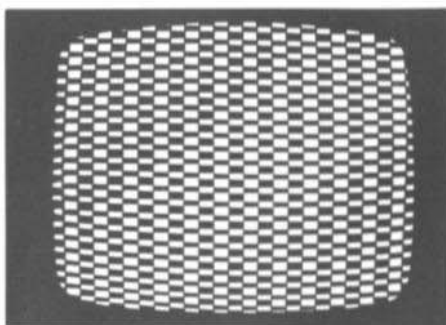

Fig. 17 : Chessboard pattern

Fig. 18 : Chessboard pattern

The vertical bar signals are available in a ratio of 1 : 1 due to the double divider. They are taken directly from circuits I 2 to I 6 of the .TV-pulse centre:

15.625 kHz (1 bar)   at point I 6/12   125 kHz (8 bars) at point I 3/12
31.25   kHz (2 bars)   at point I 5/12   250 kHz (16 bars) at point I 2/9
62.5   kHz (4 bars)   at point I 4/12

The number of visible bars is approximately 20% less since a number of bars are present in the blanked portion of the image.

The horizontal bar signals can be taken from the TV-pulse centre in a ratio of 4 : 6 as given below and, if required, in a ratio of 4 : 1 (connection points given in brackets):

50 kHz (1 bar) at point I 6/11 (I 6/8)
250 kHz (5 bars) at point I 5/11 (I 5/8)
1250 kHz (25 bars) at point I 4/11 (I 4/8)

The number of vertical and horizontal bars can be switched separately with the aid of switches, as can the pattern mode.

The method used to obtain the video image is especially simple due to the use of the TV-pulse centre and its dividers and is sufficient for most applications. A disadvantage is that the signal for the horizontal bars is only available in a ratio of 4 : 6 and not 1 : 1 and that the squares with the grid and chessboard patterns are not square. An additional PC-board (DC 6 YF 004) is to be described in one of the next editions of VHF COMMUNICATIONS which can be used together with the pattern generator in order to obtain a grid with square fields and a dot pattern which is required for the adjustment of colour television receivers. This additional board does not replace any of the previously described boards but provides extra possibilities.

3.1. OUTPUT SIGNALS

The composite video signal with negative going synchronizing pulses can be taken from connections Pt 16 (0) and 17 with an amplitude of 1.4 V (peak-to-peak). The RF-output at Pt 18 (0) and Pt 19 can be directly connected to the coaxial input socket of the TV-receiver or via a 60 Ω/240 Ω transformer to the balanced 240 Ω input.

3.2. CONSTRUCTION

Each of the previously aligned and tested modules can be combined with the additional parts to form a complete pattern generator. Due to the fact that all boards have the same dimensions, they can be mounted one above the other using long screws and spacer bushings. The switches should be mounted so that the leads are as short as possible.

3.3. APPLICATIONS

The pattern generator is, of course, especially suitable for providing test signals for alignment of all stages of a TV-receiver with the exception of UHF-tuner and sound circuit. Example: The vertical deflection, picture height and linearity can be adjusted easily when the narrowest horizontal bar pattern of 25 bars is used. However, the pattern generator is not only suitable for providing signals for a TV-receiver but also for feeding an ATV transmitter during test transmissions, thus saving the life of the camera tube.
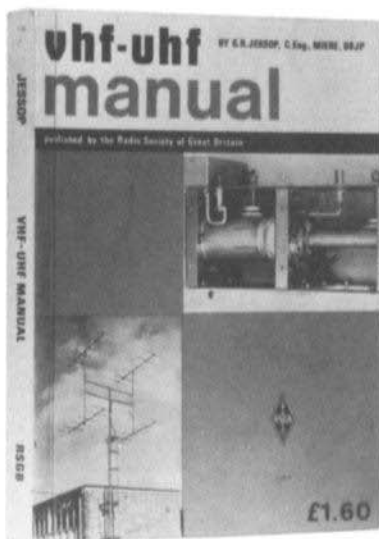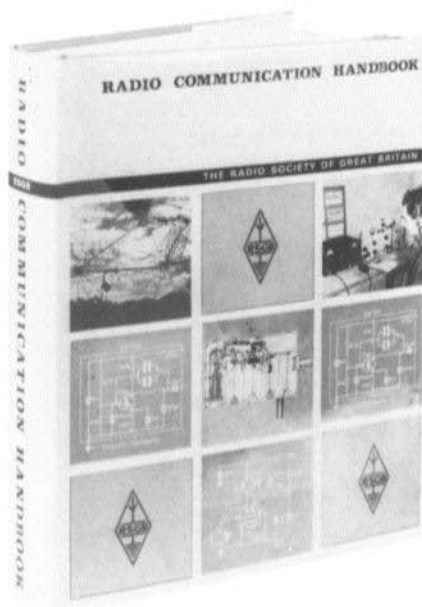
3.4. OPERATION

Of course, it is necessary for the individual modules and the interconnection wiring between the boards to be checked before commencing operation. This is especially valid with respect to the voltages from the power supply, which especially endanger the integrated circuits.

3.5. PHOTOGRAPHED IMAGES

Figures 13 to 18 show photographs of the various patterns that can be generated with the described pattern generator.

4. REFERENCES

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

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

(3) G. Sattler: A Modular ATV Transmitter
    VHF COMMUNICATIONS 6 (1973), Edition 1, Pages 2-15

DK 1 OF 001	V C O		Ed.	3/73
PC-Board	DK 1 OF 001	(with printed plan)	DM	8.00
Semiconductors	DK 1 OF 001	(3 transistors, 3 diodes, 1 IC)...	DM	26.30
Minikit 1	DK 1 OF 001	(1 potted core, 1 ferrite choke, 1 ceramic trimmer)..........	DM	10.00
Minikit 2	DK 1 OF 001	(9 capacitors, 8 resistors, coil wire, 2 solder pins)..........	DM	31.80
Kit, complete	DK 1 OF 001	with all above parts............	DM	75.00

DK 1 OF 002	VARIABLE FREQUENCY DIVIDER		Ed.	3/73
PC-Board	DK 1 OF 002	(Double-coated, through contacts)	DM	28.00
Semiconductors	DK 1 OF 002	(8 ICs)	DM	93.60
Minikit	DK 1 OF 002	(3 capacitors, 25 solder pins) ...	DM	6.50
Kit, complete	DK 1 OF 002	with all above parts............	DM	126.00

DK 1 OF 003	PHASE COMPARITOR		Ed.	3/73
PC-Board	DK 1 OF 003	(with printed plan)	DM	10.00
Minikit 1	DK 1 OF 003	(3 tranistors, 1 diode, 1 IC, 2 potted cores)	DM	26.40
Minikit 2	DK 1 OF 003	(10 capacitors, 14 resistors, coil wire, 8 solder pins)	DM	26.80
Kit, complete	DK 1 OF 003	with all above parts............	DM	62.00

DK 1 OF 004	REFERENCE OSCILLATOR		Ed.	3/73
PC-Board	DK 1 OF 004	(with printed plan)	DM	7.50
Minikit 1	DK 1 OF 004	(2 transistors, 1 ceramic trimmer, 2 styroflex capacitors)	DM	6.60
Minikit 2	DK 1 OF 004	(2 capacitors, 5 resistors, 4 solder pins)	DM	4.20
Crystal	1.000 MHz	XS 6002 with horizontal holder ...	DM	24.50
Kit, complete	DK 1 OF 004	with all above parts	DM	42.00

DK 1 OF 005	960 : 1 FREQUENCY DIVIDER		Ed.	3/73
PC-Board	DK 1 OF 005	(with printed plan)	DM	7.50
Semiconductors	DK 1 OF 005	(3 ICs)	DM	26.00
Minikit	DK 1 OF 005	(1 capacitor, 8 solder pins)	DM	3.30
Kit, complete	DK 1 OF 005	with all above parts	DM	36.00

DK 1 OF 006	FREQUENCY MULTIPLIER (6 times)		Ed.	3/73
PC-Board	DK 1 OF 006	(with printed plan)	DM	8.00
Minikit 1	DK 1 OF 006	(3 transistors, 3 coil sets, 2 ceramic trimmers)	DM	22.20
Minikit 2	DK 1 OF 006	(14 capacitors, 11 resistors, coil wire, 6 solder pins).......	DM	19.80
Kit, complete	DK 1 OF 006	with all above parts............	DM	50.00

DK 1 OF A	ADDITIONAL PARTS FOR THE SYNTHESIZER		Ed.	3/73
Semiconductors	DK 1 OF A	(16 diodes)	DM	11.20
Preselector switches	DK 1 OF A	(4 switches with accessories) ...	DM	112.00
Minikit	DK 1 OF A	(12 feedthrough capacitors)	DM	47.70
Kit, complete	DK 1 OF A	with all above parts...........	DM	170.00
All kits	DK 1 OF 001 to A with all parts		DM	520.00
All PC-Boards	DK 1 OF 001 to 006		DM	68.00

DJ 6 PI 001	PRESCALER for upto 250 MHz		Ed.	3/73
PC-Board	DJ 6 PI 001	(with printed plan)	DM	7.00
Minikit 1	DJ 6 PI 001	(3 ICs, 1 transistor, 2 diodes, 2 tantalium, 6 ceramic capacitors)	DM	149.20
Minikit 2	DJ 6 PI 001	(14 resistors)	DM	5.60
Kit, complete	DJ 6 PI 001	with all above parts...........	DM	160.00

DC Ø DA 001	RECEIVE CONVERTER 70 cm/10 m		Ed.	3/73
PC-Board	DC Ø DA 001	(Double-coated)	DM	20.00
Minikit	DC Ø DA 001	(3 transistors, 4 trimmers, 2 coilformers with cores)	DM	30.60
Kit	DC Ø DA 001	with all above parts...........	DM	50.00

Matching transmit converter DJ 6 ZZ 002 (10 m/70 cm) in 2/71

DK 1 PN 002	CARRIER OSCILLATOR for SSB TRANSMITTERS AND RECEIVERS		Ed.	3/73
PC-Board	DK 1 PN 002	(with printed plan)	DM	8.50
Minikit	DK 1 PN 002	(2 transistors, 2 trimmers, 4 styro-flex caps., 3 ceram. caps., 13 pole connector, 1 TEKO case 2A	DM	24.40
Kit	DK 1 PN 002	with all above parts...........	DM	32.00

DL 9 FX 005	MINIATURE 432 MHz/144 MHz CONVERTER		Ed.	3/73
PC-Board	DL 9 FX 005	(with printed plan)	DM	8.00
Minikit 1	DL 9 FX 005	(3 transistors, 1 trimmer, 3 coil-formers with cores, 2 sliding switches, 1 battery connector, 1 BNC socket, 1 two-coloured case)	DM	62.75
Minikit 2	DL 9 FX 005	(10 ceramic capacitors, 10 resistors, coil wire)	DM	9.50
Crystal	57.6000 MHz	(HC-25/U)	DM	28.00
Kit, complete	DL 9 FX 005	with all above parts...........	DM	105.00

DC 6 YF 002	TV PATTERN GENERATOR		Ed.	3/73
PC-Board	DC 6 YF 002	(Double-coated, through contacts)	DM	28.50
Semiconductors	DC 6 YF 002	(4 ICs, 5 transistors, 4 diodes)	DM	39.10
Minikit	DC 6 YF 002	(1 tantalium cap., 6 ceram. caps., 2 coilformers with cores)	DM	6.80
Kit	DC 6 YF 002	with all above parts...........	DM	73.00

DC 6 YF 003	DOUBLE POWER SUPPLY		Ed.	3/73
PC-Board	DC 6 YF 003	(with printed plan)	DM	14.00

CRYSTALS and CRYSTAL FILTERS
for equipment described in VHF COMMUNICATIONS

Crystal filter	XF-9A (for SSB) with both sideband crystals		DM	106. --
Crystal filter	XF-9B (for SSB) with both sideband crystals		DM	137. --
Crystal filter	XF-9C (for AM; 3.75 kHz)		DM	137. --
Crystal filter	XF-9D (for AM; 5.00 kHz)		DM	137. --
Crystal filter	XF-9E (for FM; 12.00 kHz)		DM	137. --
Crystal filter	XF-9M (for CW; 0.50 kHz) with carrier crystal		DM	106. --
Crystal discriminator XD-09-08 matching XF-9E			DM	78. --

Crystal	96.0000 MHz	(HC-6/U) for 70 cm converters (DL 9 GU, DL 9 JU)	DM	21.50
Crystal	96.0000 MHz	(HC-25/U) for 70 cm converters (DL 9 GU, DL 9 JU)	DM	28. --
Crystal	95.8333 MHz	(HC-25/U) for 70 cm converters (DC 6 HY)	DM	28. --
Crystal	84.5333 MHz	(HC-6/U) for 24 cm converters (DL 3 WR)	DM	21.50
Crystal	67.3333 MHz	(HC-6/U) for 70 cm / 10 m converters	DM	18. --
Crystal	66.5000 MHz	(HC-6/U) for synthesis VFO (DJ 5 HD)	DM	18. --
Crystal	65.7500 MHz	(HC-6/U)) for TX + RX converters	DM	21.50
Crystal	65.5000 MHz	(HC-6/U)) 130 / 130,5 /	DM	16.50
Crystal	65.2500 MHz	(HC-6/U)) 131 / 131,5 MHz	DM	21.50
Crystal	65.0000 MHz	(HC-6/U)	DM	16.50
Crystal	62.0000 MHz	(HC-6/U) for synthesis VFO (DJ 5 HD)	DM	18. --
Crystal	45.4780 MHz	(HC-25/U) for VXO (DJ 9 ZR)	DM	24.50
Crystal	42.0000 MHz	(HC-6/U) for 70 MHz converters (G 3 JHM)	DM	15.60
Crystal	38.9000 MHz	(HC-6/U) for DJ 4 LB 001 ATV-TX	DM	20. --
Crystal	38.8170 MHz	(HC-6/U) for OSCAR 6 converter	DM	20. --
Crystal	38.6667 MHz	(HC-6/U) for 2-m-converters	DM	13.70
Crystal	27.8000 MHz	(HC-25/U) for 24 MHz synthesis VFO		
		(DL 3 WR 007)	DM	25. --
Crystal	27.0000 MHz	(HC-25/U) (bandlimit 432 MHz)	DM	14. --

STANDARD FREQUENCY CRYSTALS

5.0000 MHz	(HC-6/U)	for calibration spectrum generators (DC 6 HY)	DM	25. --
1.0000 MHz	(HC-6/U)	for calibration spectrum generators		
		(DJ 4 BG, DC 6 HY)	DM	20.50
1.0000 MHz	(HC-6/U)	for 75°C crystal oven	DM	41. --
100 kHz	(HC-13/U)	for calibration spectrum generators (DC 6 HY)	DM	28. --
Crystals	72.... MHz	(HC-6/U) for 2 metre transmitters	DM	21.50
	Please give frequency - delivery approx. 8 weeks.			
Crystals	72.... MHz	(HC-25/U) for 2 m transmitters, with crystal holder	DM	33. --
	Please give frequency - delivery approx. 8 weeks.			
Crystals	other frequencies are available on request.			
	Please give frequency and type - delivery approx. 8 weeks.			
	HC-6/U	....	DM	26. --
	HC-25/U	....	DM	34. --
Crystal oven	XT-1	(6 V)	DM	82. --
Crystal oven	XT-2	(12 V)	DM	82. --
Ceramic filter CFS-455 D	(DC 6 HL 007 FM strip)		DM	47. --

TERMS OF DELIVERY and BANK ACCOUNTS

THE NEW J-BEAM
MOONBOUNCERS

Introducing the new generation of J-Beam MOONBOUNCER antennas. The coopera-tion between J-BEAM and VHF COMMUNICATIONS has now led to the introduction of a completely new range of the versatile MOONBOUNCER antennas. These an-tennas have been developed according to the very latest technologies. This has re-duced the length of the 10 element crossed Yagi, for instance, from 4.92 m to 3.65 without any reduction of the performance whatsoever. The quality of the new range has also been improved considerably and is now virtually of full professional stan-dard. Especially non-corrosive aluminium is used throughout, with cast aluminium mounts for the elements, and a heavy-galvanized steel mast clamp. All in all, an an-tenna that will not rot away and fall down every few years, but an antenna that is constructed for a long life.

All of the MOONBOUNCER antennas can be either connected for circular polarisa-tion at the antenna with one feeder to the shack, or if two feeders are fed down to the shack, it is possible to select vertical, horizontal, as well as clockwise and anti-clockwise circular polarization.

Circular polarisation is most certainly the polarisation of the future. The advantages of this form of polarisation were discussed in a recent article by G 3 JVQ/DJ ⌀ BQ in VHF COMMUNICATIONS. The possibility of switching to any required polarisation to find the momentary most favourable polarisation is a great advantage of the MOONBOUNCE antennas.

The following three types are available, which can be stacked and bayed to form arrays suitable for extreme DX modes such as MS and EME:

Type	Elements	lstr. Gain (dipole)	Hor. Beamwidth	Bom length
5XY/2 m	2 x 5	11 dB (8.8 dB)	45°	1.672 m
8XY/2 m	2 x 8	12.2 dB (10.0 dB)	40°	2.600 m
10XY/2 m	2 x 10	15.2 dB (13.0 dB)	33°	3.650 m