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ESPECIALLY COVERING VHF, UHF AND MICROWAVES

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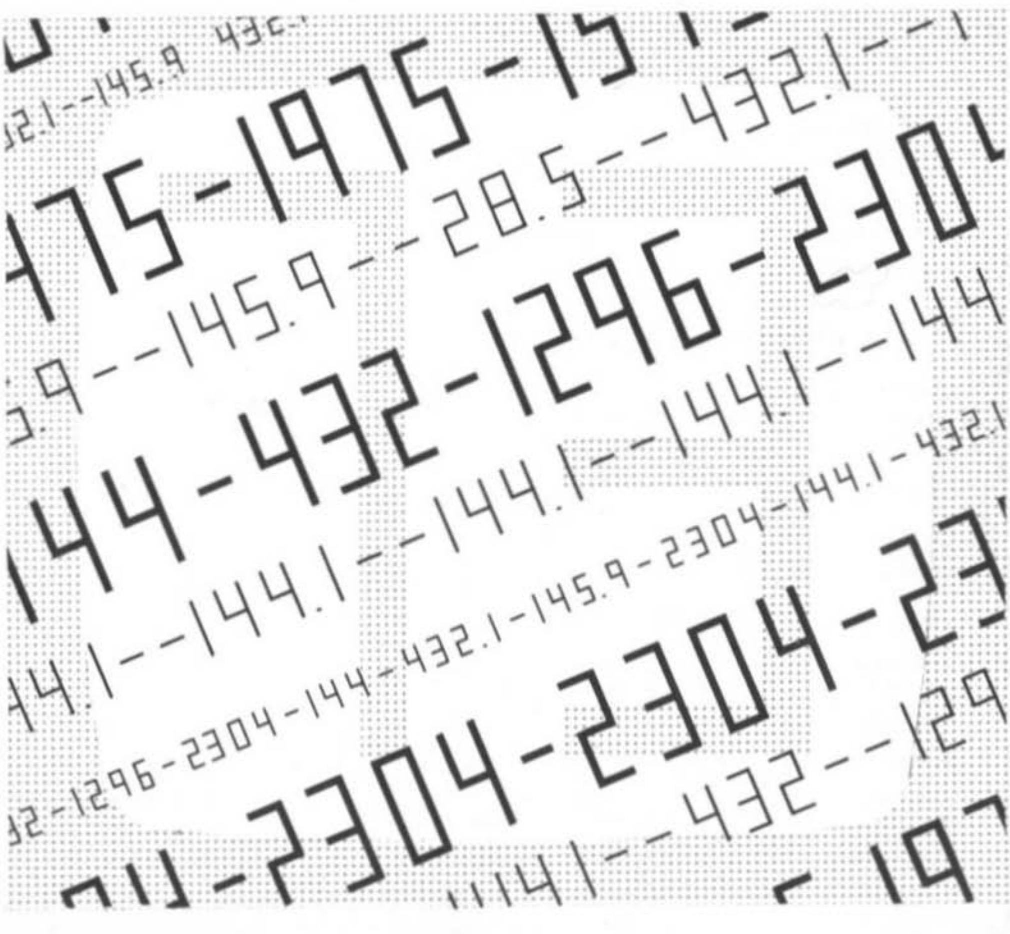

VHF COMMUNICATIONS

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Another year draws to a close, and also another volume of VHF COMMUNICATIONS. We hope that you have enjoyed reading the magazine, and have found it informative.

Fortunately, we have been able to keep the subscription price at DM 16,— for 1976. Since the DM has fallen with respect to a number of currencies since the beginning of 1975, a number of the national subscription prices will have also been reduced to a price lower than for 1975. We always attempt to keep our prices stable in spite of rising costs.

VHF COMMUNICATIONS will look even better next year when we go over to a new photoset system for preparing the texts. Of course, we will be bringing lots of outstanding designs and theoretical articles in 1976.

The publishers DJ 3 QC and G 3 JVQ/DJ 0 BQ and their representatives would like to take this opportunity of wishing you a very prosperous, Happy New Year 1976.

A TRANSMIT CONVERTER FOR 144 MHz WITH SCHOTTKY RING MIXER

by F.Weingärtner, DJ 6 ZZ

The idea of using an inexpensive Schottky ring mixer IE-500 in a transmit converter came from DK 1 OF's article (1) describing his 144/9 MHz converter. Even the very first prototype was found to be uncritical and reliable. The author was easily able to obtain an output power of 10 W (single tone) at an operating voltage of 12 V using a four-stage amplifier chain (BF 224/C 1-12/B 3-12/ B 12-12) subsequent to the mixer.

Following the construction of the prototype, a small, universal module was developed that was to provide an output power of 3 W at an operating voltage of 12 V. This module is especially suitable for the construction of fixed and portable stations, and can be connected to a linear amplifier when required. The photograph given in Figure 1 shows the author's prototype of this module, which possesses the same dimensions as the matching receive converter DK 1 OF 016.

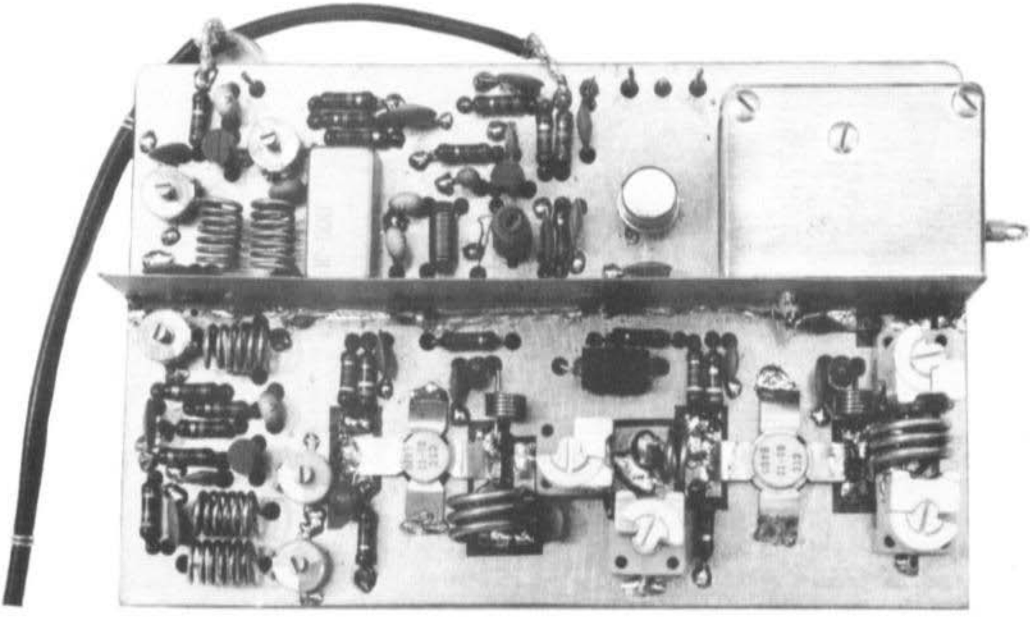

Fig.1: Author's prototype of the transmit converter DJ 6 ZZ 005

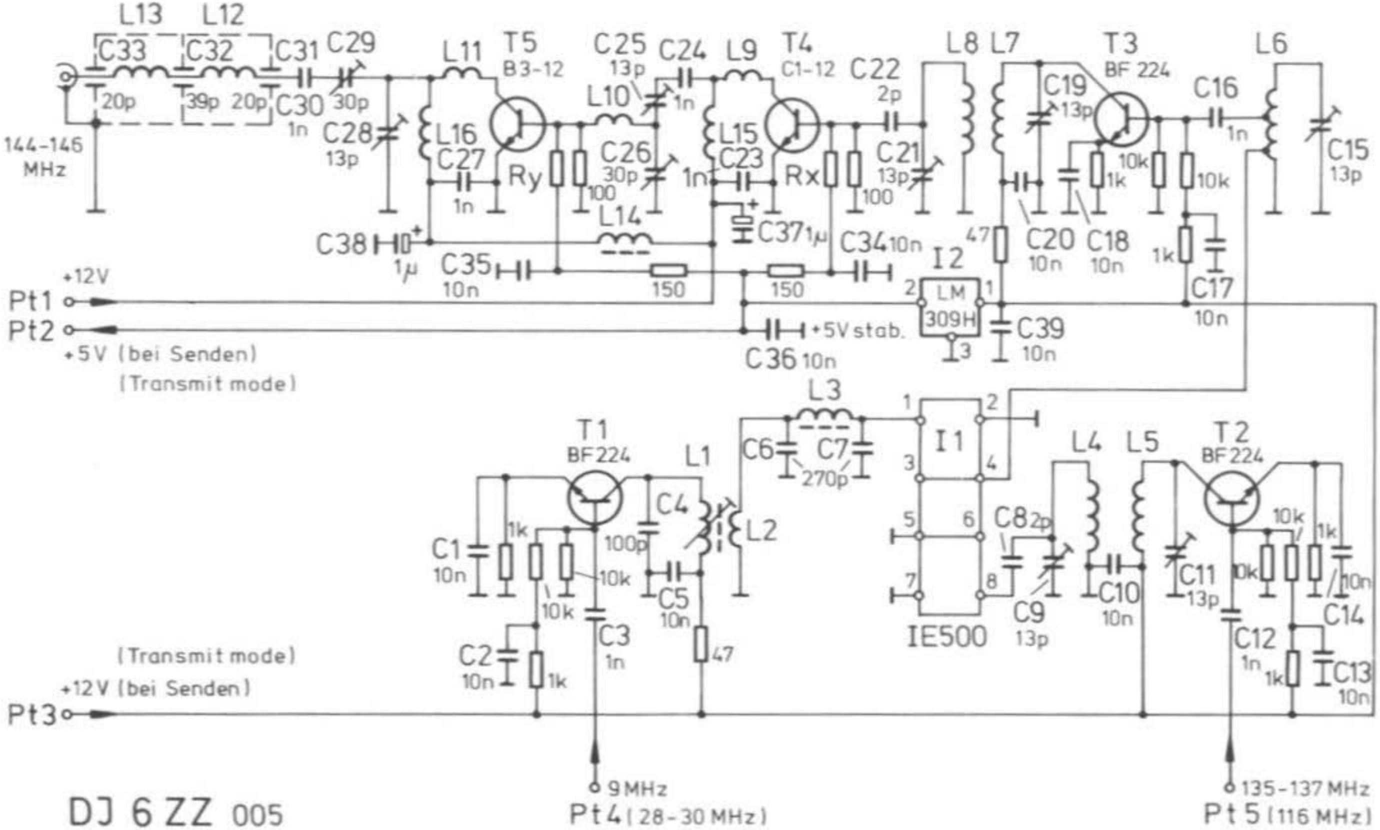

Fig.2: Circuit diagram of the transmit converter using a Schottky ring mixer

1. CIRCUIT DESCRIPTION

As will be seen in the circuit diagram given in Figure 2, the transmit converter does not possess its own local oscillator. This means that the transmit converter can be used to convert SSB signals either from a 9 MHz exciter (e. g. DK 1 OF 018/019), or from a shortwave transmitter in the range of 28-30 MHz. In the first case, a local oscillator signal is required that is variable in the range of 135 MHz and 137 MHz (e. g. DK 1 OF 011/014); in the second case, a fixed local oscillator frequency of 116 MHz is required. In order to ensure exact transceive operation, it is advisable for this local oscillator signal to be taken from the receive converter (e. g. Microwave Modules converter MMC 144/28 LO).

The SSB-signal is fed to connection Pt 4, and the local oscillator signal to Pt 5. The level of both signals should be in the order of 200 mV (RMS). Both signals are then amplified, filtered, and fed to the Schottky ring mixer. The required conversion product is filtered out in the first resonant circuit comprising inductance L 6. Since this circuit is relatively wide, the signal is passed to a further bandpass filter after being amplified in the first linear amplifier comprising transistor T 3 before being amplified to an output power of 3 W in the two power transistors T 4 and T 5. The lowpass filter is provided to filter out any harmonics.

An integrated 5 V voltage stabilizer is used for stabilizing the operating points of the driver and final output transistors (T 4, T 5), and feed the base voltage divider.

The collectors of transistors T 3 and T 5 are always connected to the operating voltage; they are blocked in the receive mode with the aid of their base resistors. In the transmit mode, connection Pt 3 is connected to + 12 V. This is so that all stages are provided with the required voltages. The stabilized voltage of 5 V can be monitored at Pt 2 or used for further stages. The fixed voltage stabilizer (TO-5-case) can be loaded up to 200 mA.

2. CONSTRUCTION

All components are accommodated on a double-coated PC-board designated DJ 6 ZZ 005. The dimensions of this board are 125 mm x 80 mm. The component locations and the conductor lanes are given in Figure 3. A 20 mm high screening panel is required which is directly soldered onto the PC-board. If the output, lowpass filter is to be constructed separately and then soldered into place, it will be necessary for the screening panel to be provided with a hole for the feedthrough capacitor C 31. The photograph of the author's prototype given in Figure 1, indicates how the various parts are put together. It will be seen that the filter forms an integral part of the module.

Attention should be paid that a trimmer should be used for C 21 that possesses a low temperature coefficient, since the resonant circuit comprising L 8/C 21 is relatively narrow band. Plastic foil trimmers have been found to be suitable.

It is necessary to connect bridge Br and solder it to both sides of the PC-board before soldering the integrated ring mixer into place.

All components that are to be grounded are directly soldered to the ground surface. The two connections of the plastic foil trimmers are shortened to 1 mm in length and bent in a suitable manner so that the trimmers are directly mounted onto the board. The resistors R_x and R_y comprise two parallel-connected resistors which are soldered into place during the alignment process after their values have been determined experimentally.

All fixed capacitors are disc-types and their connections should be as short as possible without causing a short to the ground surface. Capacitor C 30 (1 n) is used for DC-blocking and is connected between the output stage and the filter.

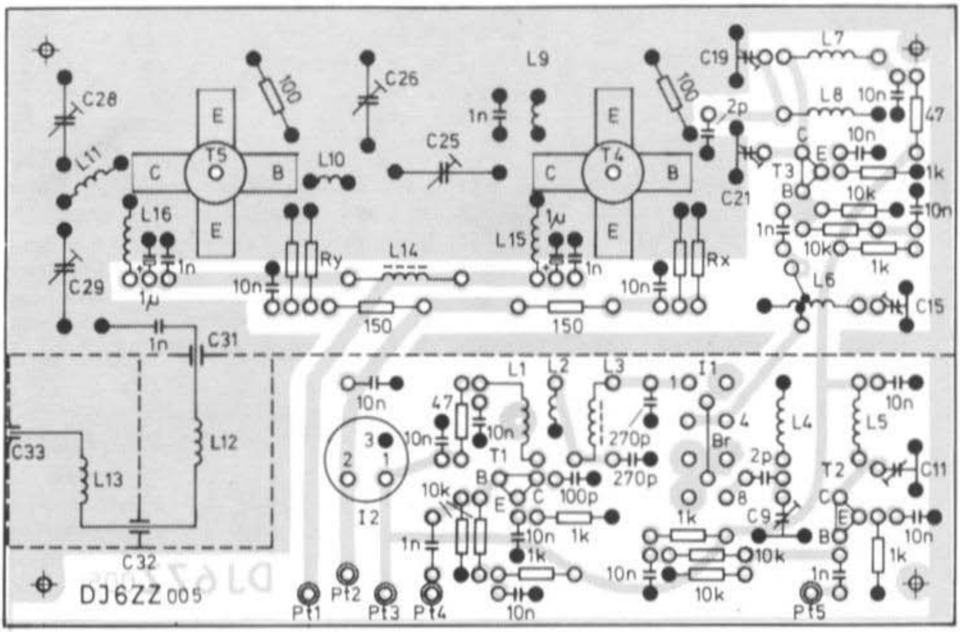

Fig.3: Component locations and conductor lanes on the lower side of DJ 6 ZZ 005

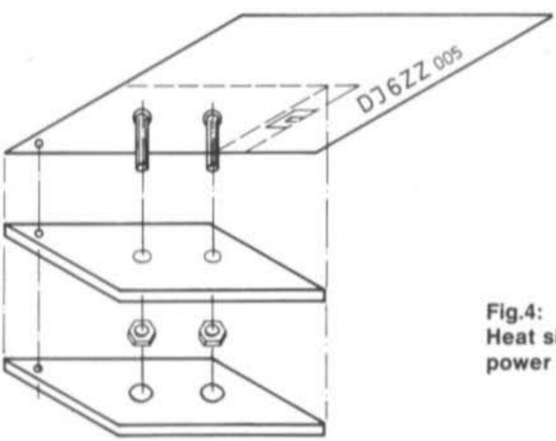

Fig.4: Heat sink for the power transistors

2.1. COOLING OF THE POWER TRANSISTORS

The driver transistor T 4 (C 1-12) is soldered to the head of a 4 mm brass screw. It is now possible for transistor T 4 and the output transistor T 5 to be screwed to a common brass or aluminium block for cooling. The PC-board is screwed to the heat sink as shown in Figure 4.

3. ALIGNMENT AND PREPARATIONS FOR OPERATION

a) All air-spaced trimmers of the power transistor stages should be adjusted for minimum capacitance.

b) Connect the operating voltage of +12 V to Pt 3. A voltage of +5 V should be present at Pt 2.

c) Inject a carrier signal to Pt 4 and a local oscillator signal to Pt 5, both at 200 mV. Whilst monitoring the required signal in a 2 m receiver, align L 1, C 9, C 11, C 15 and C 19 one after the other several times for maximum signal.

d) Connect the operating voltage of + 12 V via a mA-meter also to Pt 1 and select the values of resistors R_x and R_y so that a quiescent current of 20 mA is obtained for T 4 and 50 mA for T 5. Since the required resistance values are not always available, the PC-board provides holes for two parallel resistors (orientation values for R_x and R_y are in the order of 470 Ω). CAUTION! During the selection process it could possible for the stabilized voltage to be shorted to the screening panel which would overload the 5 V stabilizer and even damage it.

   If the amplifier stages commence oscillation during the alignment procedure (mA-meter indicates too high a value), switch the operating voltage off and on again.

e) Connect a terminating resistor of 50 Ω or 60 Ω in conjunction with a power meter or reflectometer. Inject a carrier and local oscillator signal. Align C 21, C 25, C 26, C 28 and C 29 several times one after the other for maximum output signal. If the amplifier chain does not operate in a stable manner, which is not to be expected, it will be necessary for the value of the coupling capacitor C 22 to be reduced. Finally, all circuits should be checked to see whether they are aligned for maximum output power.

f) If a carrier frequency.of 28-30 MHz and a local oscillator signal of 116 MHz is to be used, it is necessary for C 4 to be reduced to 47 pF, L 3 to be self-supporting without core, and for trimmers C 9 and C 11 to be possibly exchanged for larger values or for inductances L 4 and L 5 to be increased by one turn.

g) The alignment of the output filter should be made as described by DC 6 HL in (2).

4. SPECIAL COMPONENTS

T 1 - T 3:	BF 224 (TI) or BF 199 (Siemens)
T 4:	C 1 - 12 (CTC)
T 5:	B 3 - 12 (CTC)
I 1:	IE 500 or SRA-1
I 2:	LM 309 H, SG 309, or TBA 625

L 1: 20 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound on a 5 mm dia. coilformer with SW-core (red)
L 2: 4 turns of 0.3 mm dia. (29 AWG) enamelled copper wire wound on L 1 (on the same side as the core)
L 3: 12 turns of 0.4 mm dia. (26 AWG) enamelled copper wire wound in the threading of a SW core (red), 3 mm dia., 9 mm long
L 4, L 5: 7 turns of 0.8 mm dia. (29 AWG) silver-plated copper wire wound on a 5 mm former, self-supporting
L 6: 6 turns as L 4, coil tap at 2 and 3.5 turns from the cold end
L 7, L 8: As L 6 but without coil taps
L 9, L 11: 3 turns of 1.2 mm dia. (17 AWG) silver-plated copper wire wound on a 9 mm former, self-supporting, rotated by 90^{o} with respect to L 10
L 10: 2 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound on a 5 mm former, self-supporting
L 12, L 13: 4 turns of 1 mm dia. (18 AWG) silver-plated copper wire wound a 7 mm former, self-supporting, spaced approx. 1.5 mm between turns
L 14, L 15: 5 turns of 0.5 mm (24 AWG) enamelled copper wire close wound on a 5 mm former, self-supporting
L 16: Ferroxcube 6-hole core with 2.5 turns (Philips)

C 26, C 29: 3-30 pF air-spaced trimmer
C 25, C 29: 3-13 pF air-spaced trimmer
C 9, C 11, C 15, C 19, C 21: 3-13 pF plastic-foil trimmer, 7 mm dia.
C 31, C 33: 20 pF; C 32: 39 pF feedthrough capacitors

5. REFERENCES

(1) J. Kestler: A Receive Converter for 144 MHz/9 MHz with Schottky-Diode Ring Mixer
VHF COMMUNICATIONS 6, Edition 4/1974, Pages 204-214

(2) G. Otto: A 144 MHz Linear Amplifier with 25 W Output at 12 V to 14 V
VHF COMMUNICATIONS 5, Edition 2/1973, Pages 81-90.

A STEREO VHF-FM RECEIVER WITH FREQUENCY SYNTHESIZER
PART 3: POWER SUPPLY AND NOTES REGARDING PART 1 AND 2

by J.Kestler, DK 1 OF

1. POWER SUPPLY

A power supply has been developed to provide the three required voltages of +5 V, +15 V and -6 V. This power supply requires a power transformer with two 12 V windings. Such transformers are readily available. All three voltages are stabilized, and fixed voltage stabilizers are used for the two voltages requiring the higher current values. This results in relatively simple circuits with a low number of components and with distinct advantages with respect to short-circuit, current limiting, and thermal overload.

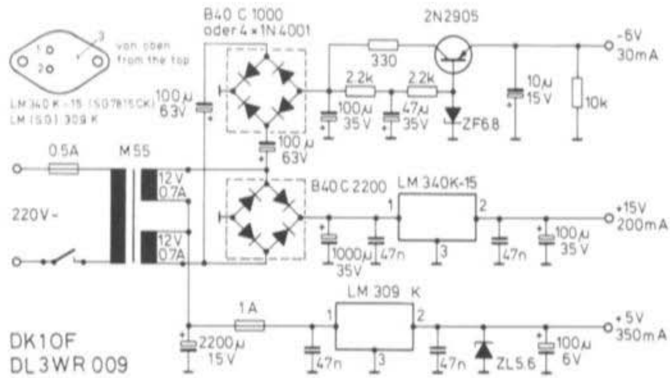

Fig.1: Circuit diagram of the power supply for the VHF-FM receiver

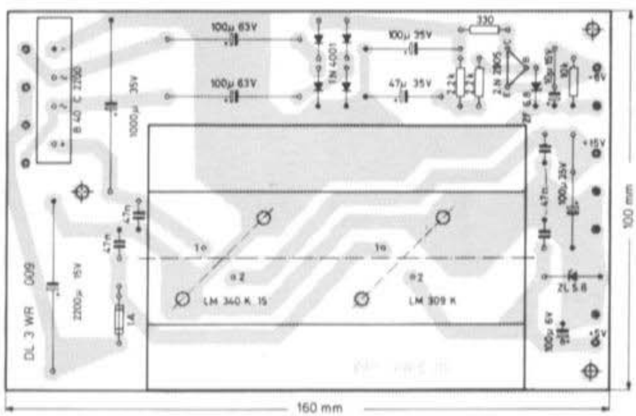

Fig.2: PC-board DL 3 WR 009 for the power supply DK 1 OF

The single-coated PC-board für this power supply is designated DL 3 WR 009 and its dimensions are 160 x 100 mm. The two fixed -voltage stabilizers are mounted on a common heat sink of 70 mm in width and 100 mm in length (possessing four fins on each side), which is fixed to the PC-board by the integrated circuits. The heat sink and the large electrolytic capacitors with working voltages of up to 70 V determine the size of this PC-board (Fig. 2).

2. MOBILE OPERATION

One prototype of the described receiver is used by the author as car radio. No difficulties have been found up till now in the temperature range of -10 to +50 °C.

2.1. POWER SUPPLY

The power supply voltage of + 15 V given in the circuit diagrams of the modules is not critical. The receiver will also satisfacturely be operating with an operating voltage of 10 V, which means that it can be run from a car battery (12 V, negative ground) without difficulties. If an AC-alternator is in use, it will be necessary for a filter link as shown in Figure 3 to be used. A transformer core having approximately 120 turns of 1 mm dia. wire has been found to be successful. This filter can be loaded up to approximately 4 A, which means that a mobile station can also be connected in addition to the broadcast receiver. The operating voltage for the frequency synthesizer (+5 V) can be obtained easily using an integrated stabilizer LM 309 K (see Fig. 1). If the module DK 1 OF 023 (AGC, squelch and tuning indicator) is also to be used, it will be necessary for a voltage of -6 V to be provided.

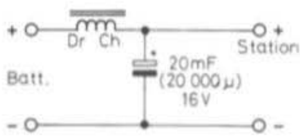

Fig.3: Supressing the AC-alternator

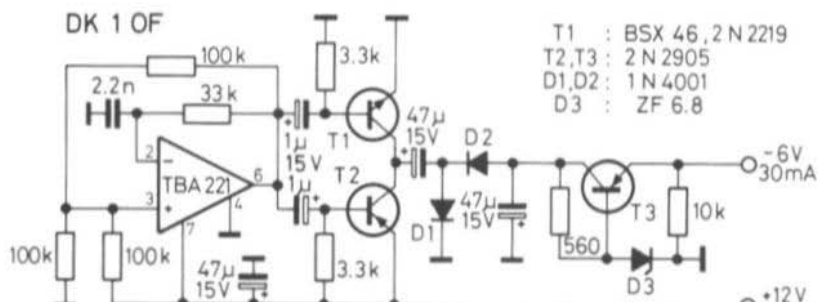

Fig.4: DC-DC converter +12V/-6V for operating the VHF-FM receiver in a car

Figure 4 gives the circuit diagram of a DC-DC converter suitable for this. The integrated circuit TBA 221 operates as an astable multivibrator and generates a square-wave voltage of approximately 7 kHz which in turn drives the push-pull output stage comprising transistors T 1 and T 2. The signal is then rectified in diodes D 1 and D 2 and fed to a simple stabilizer circuit comprising T 3/D 3. This circuit can be loaded up to a maximum of 40 mA.

2.2. TRAFFIC-NEWS DECODER

As was mentioned in Part 1 of this article a number of FM transmitters in Germany radiate a 57 kHz pilot-tone for the traffic-news network. This information can easily be used to switch on a pilot lamp or LED-indicator. A suitable decoder for this is the integrated circuit CA 3090, which is also used as stereo decoder.

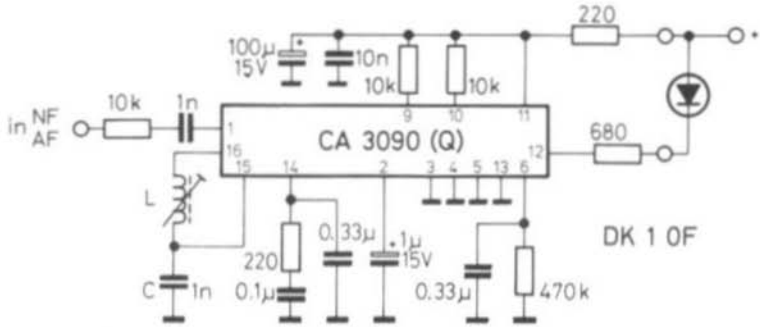

Fig.5: Circuit diagram of a traffic-news decoder

The circuit diagram of this decoder is given in Figure 5. The series-resonant circuit L/C should have a resonant frequency of 228 kHz (57 kHz x 4); L is a potted core of 14 x 8, A_L = 160 (B 65541-K0160-A022) and possesses 50 turns of 0.3 mm dia. (29 AWG) enamelled copper wire (approx. 400 µH). The PC-board DK 1 OF 024 can be used for construction.

CONSTANT-AMPLITUDE SSB - ADVANTAGEOUS OR NOT ?

by R.Lentz, DL 3 WR

The advantages of the modulation modes whose amplitude remains constant are well known: Narrow-band frequency modulation (NBFM) or frequency-shift keying (FSK), for example, cause considerably less interference in broadcast and television receivers and AF-amplifiers than single sideband (SSB), double sideband (AM), or keyed CW signals. Since SSB and CW are far superior modes for DX-communications, a modulation mode was required that solved the problems of envelope demodulation of the amplitude variations, and still provided the DX capability of CW and SSB. The ideal case would therefore be a type of SSB having a constant amplitude.

1. PHASE-LOCKED SSB

The first practical development that is known was described by PA Ø EPS in 1972 (1). In this description, the SSB-signal was theoretically infinitely limited and fed to a phase-locked loop (TTL). This mode was designated phase-locked single sideband (PL-SSB), and operated according to the following principle:

A (9 MHz) SSB-signal is limited infinitely so that a constant amplitude results, even during the intermediate spaces between words. As is to be expected, a very wide frequency spectrum is generated. According to standard RF-clipper technology (5), a second SSB filter would now be required to suppress the undesirable harmonics outside of the passband. However, this would lead to a certain degree of amplitude fluctuation which was to be avoided.

Instead of this, a second oscillator is phase-locked to the clipped SSB-signal. The oscillator generates a signal of constant amplitude, but the problem is now to keep the spectrum of the phase-locked oscillator so narrow as possible. Values were found experimentally by PA Ø EPS for the PLL-filter which result in a usable compromise between bandwidth and readability of the signal. Figure 1 gives a block diagram for generation of PL-SSB.

Whereas PA Ø EPS built up his original circuit with individual transistors and discrete components, PA Ø LQ described a more advanced version using the inexpensive integrated circuit TBA 120 (Fig. 2). The integrated circuit operates as a limiter-amplifier and phase detector. The limiter is effective at input voltages of more than approximately 200 μV; an AM-suppression of more than 40 dB is achieved at approximately 5 mV. This means that the input of this circuit can be directly connected to the SSB filter.

Connection 8 of this IC supplies a control voltage which varies between 3 and 9 V at an operating voltage of 12 V. It is connected via the phase-correcting network, which must be found experimentally, to the varactor diode of the voltage-controlled oscillator (VCO). Inductance L 1 is tuned to the VCO-frequency with the aid of the two 1 nF capacitors. When using this circuit, this frequency can be in the order of 3 to 12 MHz. The value of capacitor C 1 should be selected so that a RF-voltage of approximately 200 - 400 mV peak-to-peak is present at inductance L 1.

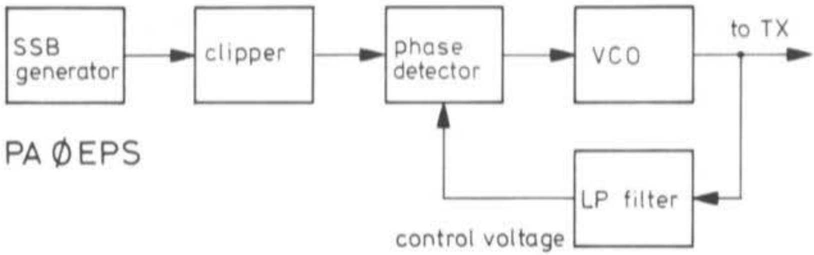

Fig.1: Block diagram of a PL-SSB exciter

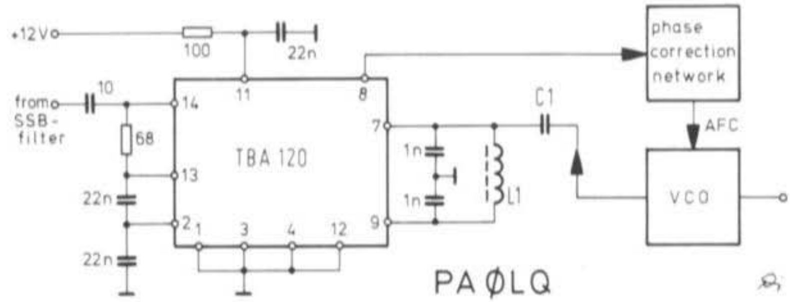

Fig.2: Circuit for converting a SSB-signal into a PL-SSB-signal

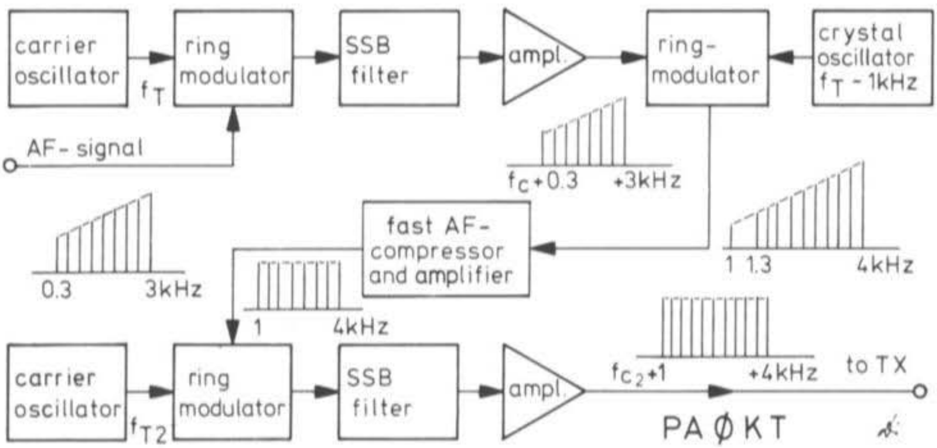

Fig.3: Block diagram of a CA-SSB exciter with fast AF-compressor

Careful selection of the time constants in the PLL-filter allow a relatively clean, pleasant signal to be generated that provides good readability at weak signal strengths and does not exhibit a wide signal spectrum. The signal generated in the circuit given in Figure 2 can then be converted to the required shortwave or VHF band where it is then amplified to the required output level. In contrast to conventional SSB, PL-SSB, can be fed to class C amplifiers.

2. CONSTANT AMPLITUDE SSB

In a recent three-part article (3), PA Ø KT writes that the spectrum of a PL-SSB signal is still wider than desirable even when carefully optimized. This especially undesirable, when several PL-SSB stations are active in a VHF/UHF contest. The cause of this is probably the large and abrupt phase variations in the zero passes of the limited RF-signal. PA Ø KT then goes on to describe two other methods of generating a SSB signal of constant amplitude (CA-SSB) that does not cause any increase in spectrum bandwidth.

The first system described by PA Ø KT is shown in Figure 3. A fast AF-compressor is used which is able to control the amplitude variations, but is, of course, not able to amplify the level of the suppressed carrier up to the peak level during the spacing between words. For this reason, the AF-signal is firstly converted into a SSB-signal in the conventional manner and the balanced mixer adjusted so that a residual carrier of approximately -30 dB is obtained. The SSB-signal is now fed to a second ring mixer and re-converted into the AF-level. The oscillator used for the second conversion, however, oscillates at a frequency that is 1 kHz lower than the original carrier oscillator. This allows the original, residual carrier to be heard as a 1 kHz audio tone, and the whole AF voice spectrum is 1 kHz higher. This signal is now fed to an AF-compressor having very short-time constants where a constant amplitude signal results. The compressor should have a dynamic range of at least 30 dB in order to increase the value of the carrier signal to the peak level during pauses between words. Figure 4 shows a circuit in which the gain of an integrated operational amplifier type 741 is controlled with the aid of two field-effect transistors. The output voltage of this circuit is a constant 4 V at input voltages between 5 mV and 2 V.

This constant-amplitude signal (with 1 kHz audio tone in the pauses between words) is now fed to a conventional SSB exciter. The distant station will not notice that the transmitted signal is shifted by 1 kHz, since it will tune in to zero beat in order to obtain the original AF-spectrum. It is, however, important that the second ring mixer possesses a very high carrier suppression since a 1 kHz audio tone would be audible (beat between the suppressed carrier of the SSB exciter and the 1 kHz tone).

This system was used by PA Ø KT for several months and better reports were received than with PL-SSB. The weak point of the system is still the AF-compressor. The time constants in the control circuit must be so short that they are in the same order as the duration of one cycle of the lower AF-frequencies. If the time constants are too short, the lower frequencies will be controlled and jumps in the signal amplitude will be exhibited. For this reason, a second system was developed using a RF-compressor. Not only is the operation better, but it is also easier to realize.

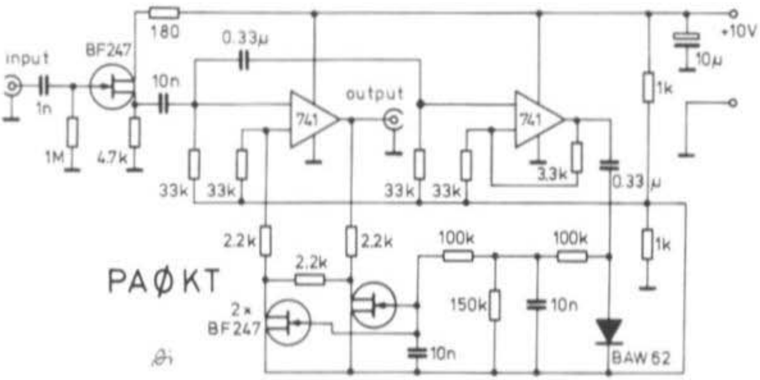

Fig.4: Circuit diagram of a fast AF-compressor for a CA-SSB system

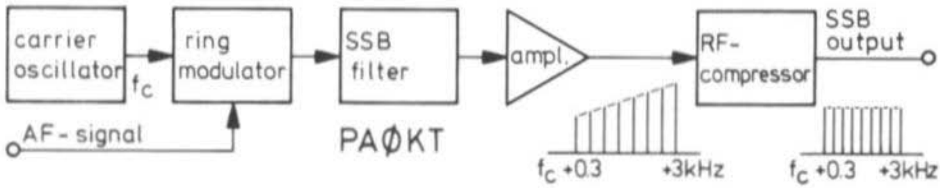

Fig.5: Block diagram of a CA-SSB exciter using an RF-compressor

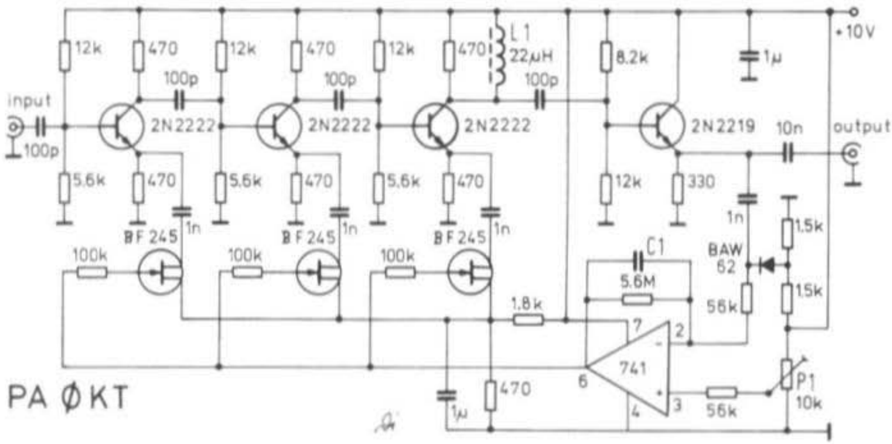

Fig.6: Circuit diagram of an RF-compressor

2.1. CA-SSB WITH RF-COMPRESSOR

A block diagram is given in Figure 5 to show how a SSB-signal of constant amplitude is obtained using a RF-compressor (3), (4). Firstly a conventional SSB-signal is generated having a partially suppressed carrier, which is then fed to a compressor which directly processes the RF-signal. With a time constant of approximately 100 µs, the compressor is easily able to follow all amplitude variations of the envelope and to bring these to a constant level. However, compared to the duration of one cycle of the carrier frequency (0.1 µs at 10 MHz), the time constant is infinitely great.

The circuit diagram of such a RF-compressor is given in Figure 6. It will be seen that three field-effect transistors are used as variable resistors in the emitter circuits of three amplifier stages. The reference voltage of the control circuit and thus the output level can be adjusted with the aid of potentiometer P 1. The FETs control an operational amplifier having a voltage gain of 100. The time constant is determined by the circuit comprising capacitor C 1 and the parallel resistor of 5.6 MΩ. Values between 10 and 100 pF are suitable for C 1. Although the circuit is principally wideband up to approximately 10 MHz, it is favourable for a resonance to be provided for the required output voltage using inductance L 1 together with the circuit capacitance. Since the characteristics of field effect transistors vary considerably, it may be necessary for these transistors to be selected. Furthermore, the FETs used can be matched individually by varying the source voltage divider (470 Ω/1.8 kΩ).

In practice, the control range of this circuit amounts to approximately 30 dB. The maximum output voltage is in the order of 4 V, at input voltages of between 10 mV and 1 V peak-to-peak. The input RF-voltage is adjusted so that the residual carrier is present with approximately 30 mV. In this case, the modulation peaks of the SSB-signal should not be in excess of 1 V.

This type of constant amplitude SSB is very effective. Several Dutch radio amateurs have, however, mentioned that the signal must fail when the input level is less than -40 dB. This can be proved using a two-tone test signal having a very pure spectrum, and equal amplitude. However, these conditions cannot be expected to take place during voice transmissions.

It should, however, be mentioned that this signal still possesses small amplitude variations. According to experiments made by PA Ø KT using this form of CA-SSB, virtually all cases of interference in (Hi-Fi) audio equipment disappeared. As is well known, this interference is caused by an envelope demodulation in audio equipment. In a few cases, the interference was still present, and the original type of PL-SSB seems to be the only answer.

3. NOTES

PA Ø SE mentions that a RF-compressor is also a good means of increasing the mean voice level in SSB-transmitters since it provides virtually the same advantages as a RF-clipper (5), but does not require a second sideband filter (even when a certain increase in bandwidth cannot be avoided).

TJ 1 EZ (ex. PA Ø EZ) states that there is no difference in principle between compressors and clippers since they only differ in the length of the time constants. In his opinion, compressors and clippers can be assumed to be amplitude modulators with the voice signal at one input and the control signal at the other. As in the case with every type of modulator sum and difference frequencies are generated that are combined with the original signal. The faster the control voltage is varied, the wider will be the frequency spectrum, and the greater will be the distortion products that increase the bandwidth of the original signal.

A clipper represents the extreme case since the "control signal" is very complex and extremely wideband. This is the spectrum that is present directly after the clipper, and before the filter. On the other hand, the increase in bandwidth in the case of a RF-compressor such as that designed by PA Ø KT is relatively low and therefore acceptable. It should be mentioned that the circuits of the RF-compressor are not too critical and that other circuits can be used. The criterium of the system being described by PA Ø KT is that a compressor is used at RF.

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A FOUR-DIGIT FREQUENCY COUNTER FOR 250 MHz USING A 7-SEGMENT LED-READOUT

by E.Zimmermann, HB 9 MIN

This article is to describe a frequency counter which is so small that it can be used as the digital readout in VHF transmitters and receivers. The complete counter comprises four modules (Fig. 1):

1. Divide by 10 prescaler up to typically 250 MHz by DJ 6 PI (1)
2. Counter module HB 9 MIN 001
3. Indicator module HB 9 MIN 002
4. Time base (2) with an output frequency of 1 kHz (TTL level)

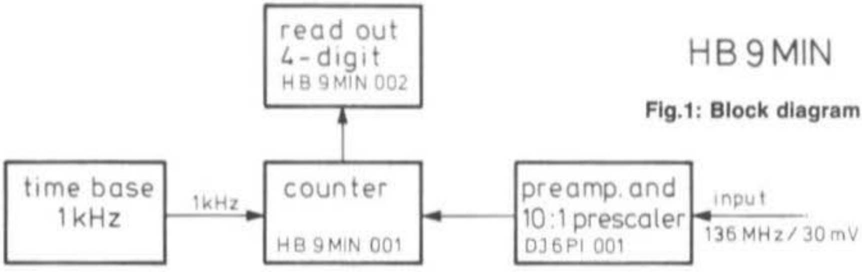

Fig.1: Block diagram

Since the prescaler (1) and time base (2) have already been described in this magazine, the following article is to describe the modules HB 9 MIN 001 and 002 (see Fig. 2 and Fig. 3). The counter module is accommodated on the double-coated PC-board HB 9 MIN 001 (100 mm x 60 mm), and is equipped with nine TTL integrated circuits of the SN 74 series. Integrated MOS-circuits were not used due to their relatively low frequency limit, and for price reasons (in 1974).

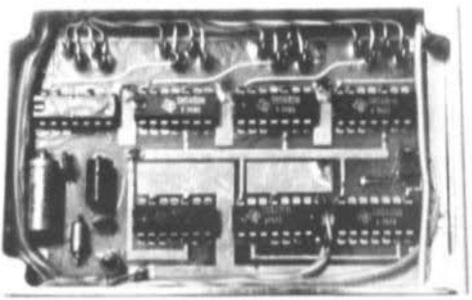

Fig.2:
The counter module HB 9 MIN 001

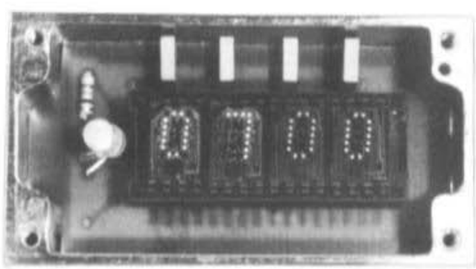

Fig.3:
The indicator module HB 9 MIN 002

The PC-board of the indicator module HB 9 MIN 002 is single-coated and its dimensions are only 68 mm x 39 mm. This module is equipped with four integrated 4 x 7 point LED readouts type hp 5082-7300. These readouts comprise a BCD-storage, BCD-decimal decoder and the driver which means that they can be directly connected to the counting decades SN 7490, see Figure 4. The small dimensions of the counter are only possible using this type of readout which saves two integrated circuits. Unfortunately, they are also the most expensive components in the counter.

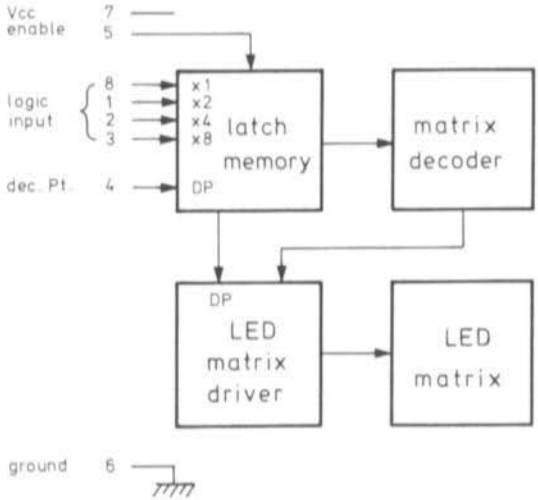

Fig.4: Block diagram of an indicator unit hp 5082-7300

1. SPECIFICATIONS

Frequency range without prescaler:

with SN 7490 N as first counter:	typ. 20 MHz
with SN 7490 A or SN 7490 N - S 1:	typ. 35 MHz
with SN 74LS90N (low power Schottky):	typ. 42 MHz

Resolution: 100 Hz

Accuracy: With a time base stability of 1×10^{-7} (2) and at a measuring frequency of 136 MHz: $\pm$ 14 Hz $\pm$ 1 digit

Current drain at U_B = 5 V and T = 25 °C: counter: typ. 250 mA
indicator: typ. 380 mA

Working temperature range of the readout: -25 to +85 °C

Working temperature range of the counter module: dependent on the IC:

SN 74...N and SN 74L...N:	0 to +70 °C
SN 84...N and SN 49...N:	-25 to +85 °C

2. OPERATION (Fig. 5)

The integrated circuits I 1 and I 2 (SN 7490) divide the timebase frequency of 1 kHz by five which results in a total division factor of 25. The resulting frequency is 40 Hz. A favourable solution of this problem is to use the integrated circuit type SN 49710 P, which comprises two divide by 5 and one divide by 2 dividers. This solution was used in the author's prototype, which is the reason why Figure 2 differs from the described version. In order to ease construction, the circuit has been modified so that the more popular type SN 7490 N could be used.

The integrated circuit I 3 (SN 7492 N) is connected as divide by 12 divider and generates the clock pulses that are evaluated in two NAND-gates having three inputs each (I 4: SN 7410 N). The storage strobe and reset impulses are available at the outputs (Fig. 6). The third gate in I 4 and two gates in I 5 (SN 7400 N) are connected as inverters; the third gate of I 5 represents the counting gate (pins 8, 9 and 10).

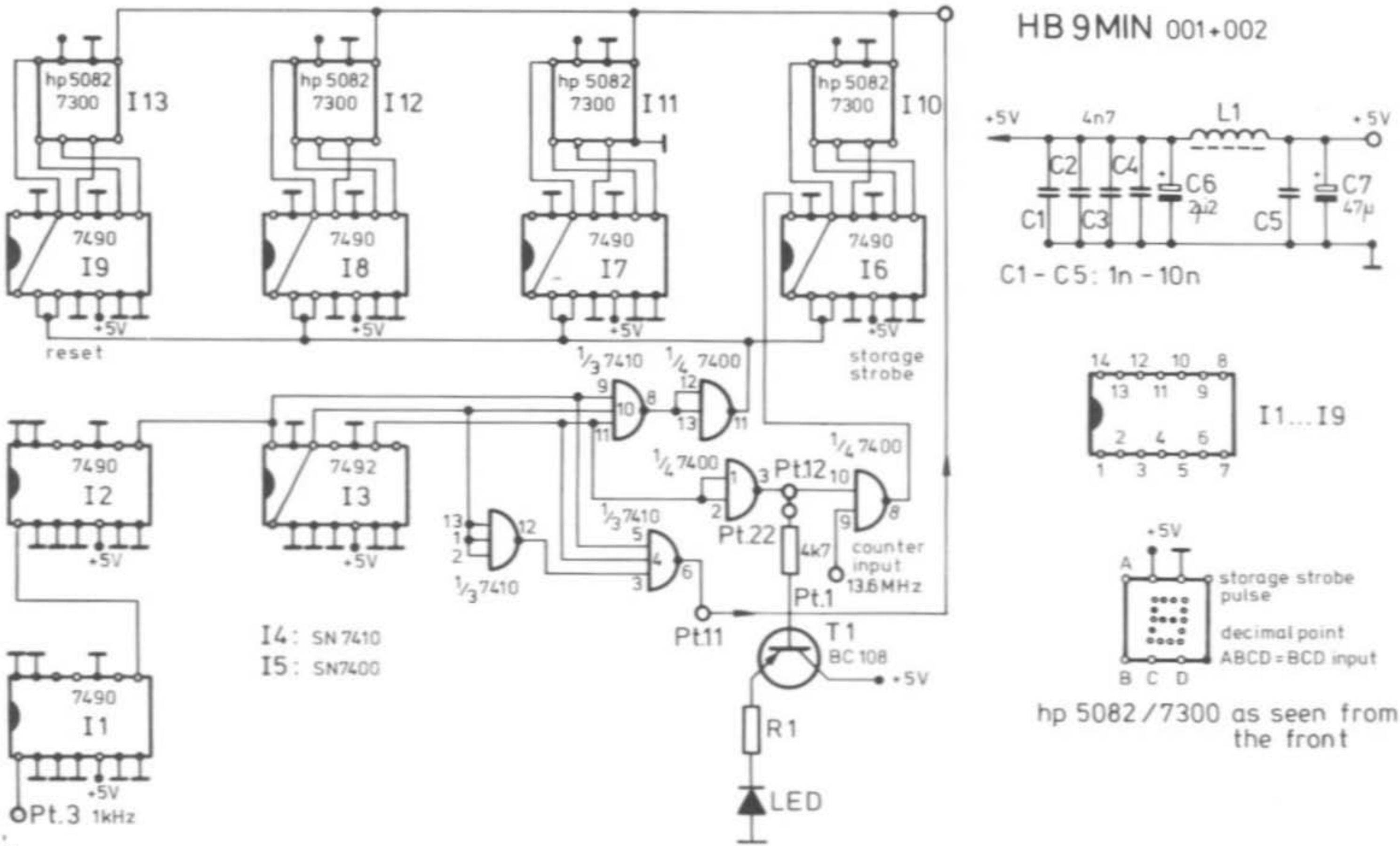

Fig.5: Circuit diagram of a 4-digit frequency counter with 7-segment LED-readout with integrated storage/decoder

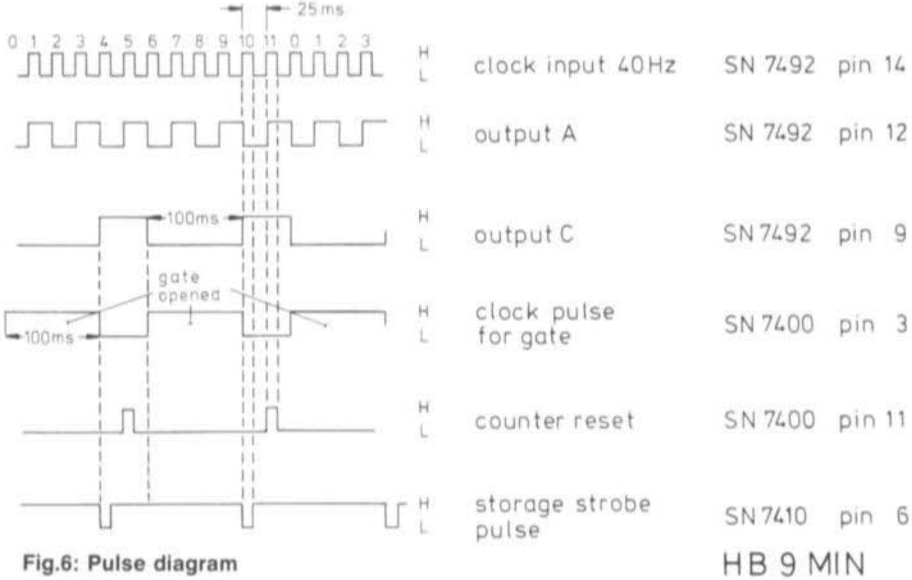

Fig.6: Pulse diagram

The four decade counters I 6 to I 9 (SN 7490 N) count the number of pulses passed through the gate during the 100 ms measuring period and pass them to the storage in BCD-code from where they are fed to the readout. A 12.5 ms L-pulse at the end of the measuring period resets the storage to zero and accepts the new count from the decades; this value is then stored and indicated until a new storage strobe pulse actuates the reset and transfer of the new result. 12.5 ms after the storage strobe pulse, a 12.5 ms H-pulse is generated which resets the counting decades to zero. After a further period of 12.5 ms, the counting gate will again open and the new storage strobe pulse will be generated.

The decimal points provided in the readouts HP 5082 - 7300 are illuminated when pin 4 is grounded. This is made in the kHz-decade so that an indication ···.· kHz results.

3. CONSTRUCTION

It will be seen in Figure 7 and 8 that the construction is very uncritical in the case of both modules. Attention should be paid in the case of the counter module that the double-coated PC-board HB 9 MIN 001 should be soldered on both sides of the board if a board with through-contacts is not available. Two bridges must be made on PC-board HB 9 MIN 002. This allows a single-coated PC-board to be used. The four readout units are plugged into a 32 pin dual-in line socket. The connection leads are directly soldered to the PC-board.

The interconnection of the counter module to the indicator module (which should be mounted in the most favourable position) is made with 20 leads; 4 x 4 are required for the BCD-transfer, one for the storage strobe, one for the monitoring of the gate time and one each for +5 V and ground. These interconnection cables should be screened so that no harmonics of the TTL pulses are injected into the receiver or transmitter. The screening can easily be made with the aid of the outer conductor of a coaxial cable into which all twenty wires can be inserted.

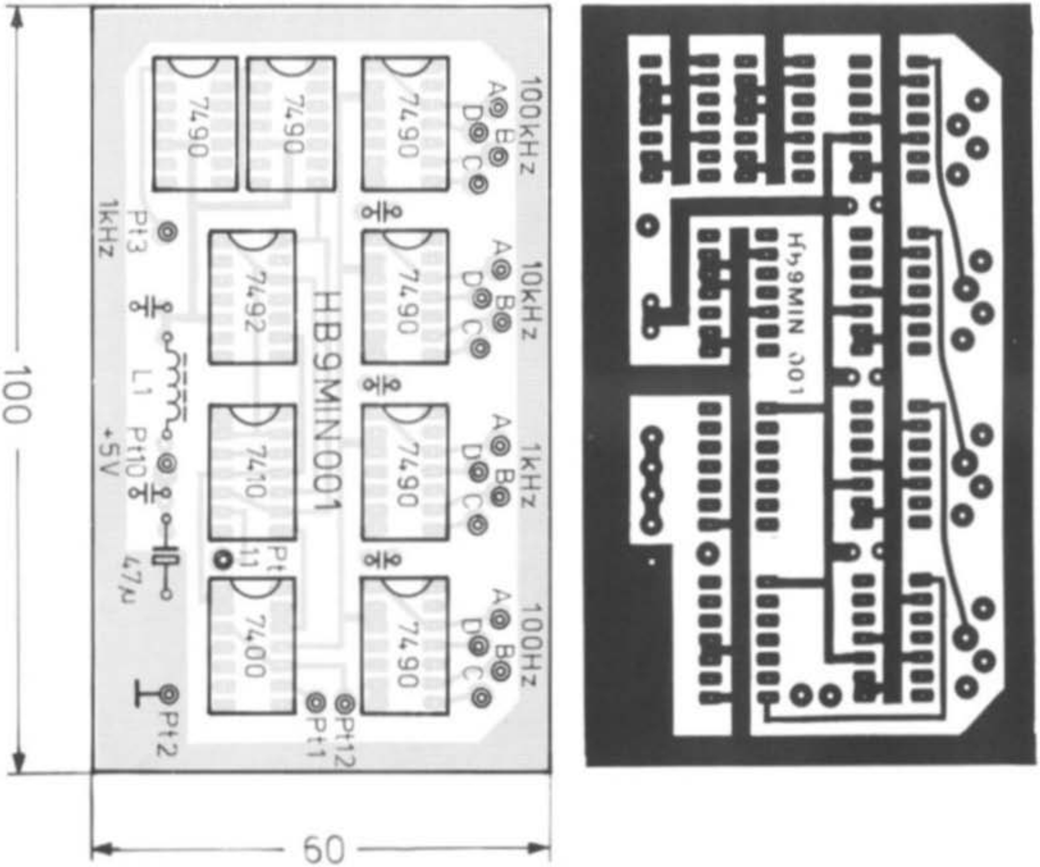

Fig.7: Component locations on the counter board HB 9 MIN 001

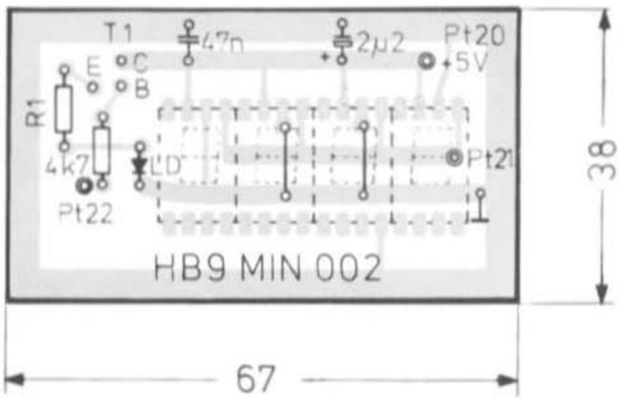

Fig.8: The indicator board HB 9 MIN 002

In order to avoid the mentioned interference, the following additional points should be observed:

Bypass the 5 V line several times; preferably with various capacitance values.

All interconnection leads should be screened.

All modules should be screened sufficiently according to RF practice (good multi-pin connectors for the 20 connections, coaxial sockets for the measuring frequency and time-base input).

If interference is injected into the receiver via the preamplifier 9582 of module DJ 6 PI 001, this can be avoided using a low-reactive buffer stage (e. g. a dual-gate MOSFET tuned to the measuring frequency).

4. COMPONENTS

I 1, I 2, I 6 - I 9: SN 7490 N (for I 6 possibly SN 74LS90N)
I 3: SN 7492 N
I 4: SN 7410 N
I 5: SN 7400 N
I 10 - 13: hp 5082-7300 (Hewlett-Packard);
if possible select for equal brightness

T 1: BC 108 or similar silicon NPN AF-transistor

LED: Any light emitting diode

R 1: Approx. 100 Ω, dependent on the LED used.

C 1 - C 5: Ceramic disc capacitors of between 1 nF and 10 nF
C 6: Approx. 2.2 μF tantalium electrolytic
C 7: 47 μF tantalium electrolytic

L 1: Ferrite choke approx. 22 μH

5. NOTES

No alignment is required on the described modules. Before putting the modules into service, the circuit should be checked for wiring faults and incorrect polarity of the integrated circuits. The counter can be extended to more than four digits, it is only important, that the fan-out of ten is not exceeded with respect to the gates of the storage strobe and reset pulses. If necessary, higher power gates can be used.

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AMSAT PHASE III PROGRAM

by David Hull, VK 3 ZDH

A conference was held at the Goddard Space Flight Center, Greenbelt, Maryland, U.S.A. over the period 20th to 24th March, 1975. It was convened to define the next satellite(s) in the OSCAR series and to decide the responsibilities of the national groups involved towards developing these satellites.

Those who attended included Larry Kayser, VE 3 QB and Bob Pepper, VE 2 AO from AMSAT Canada, Karl Meinzer, DJ 4 ZC from AMSAT Deutschland, Chuck Swedblom, WA 6 EXV and Dick Kolbly, K 6 HIJ from the San Bernardino Microwave Society, Jan King, W 3 GEY and Perry Klein, K 3 JTE from AMSAT HQ and Dave Hull, VK 3 ZDH from the WIA Project Australis.

The principal area of discussion was OSCAR 8 and the possible launch vehicle/ orbit opportunities for this project. Without going too much into the alternative possibilities, which included a joint VK/VE satellite in an OSCAR 6/7 orbit, it can be stated that the conference decided to go ahead on development of an AMSAT Phase III advanced spacecraft for launch in mid-1978 and to concentrate all effort to that end.

The development is constrained by the launch date of the last ITOS launch on the Delta 2910, a call-up mission with a mid-1978 target. Failing this launch the Titan 3C/377 Military launch could be considered as could the Space Shuttle scheduled for an expected first launch in June 1979. The orbit possibilities of these launches are 900 mile, Sun Synchronous (as per OSCAR's 6 and 7) for the Delta, Geostationary Synchronous for the Titan, and low altitude low inclination for the Shuttle. None of these orbits was considered entirely satisfactory for the Amateur Satellite service worldwide at our present state of development.

An optimum location for the Geostationary satellite was impossible to find; it would serve only one area for long periods at a time. The 900 mile orbit had been fully explored with OSCARs 6 and 7 and there seemed little point to a lower shuttle height orbit. The only alternative seemed to be an initial launch into a 900 mile orbit with a subsequent in-flight manoeuvre to raise the apogee of the satellite to such a height that a considerable radio range would result for much of the orbit.

What the conference had in mind was to provide a viable alternative to the 20 metre band without any of the propagation problems of the HF bands. This in-flight manoeuvre would require the spacecraft to be fitted with an Apogee Kick Motor (an AKM, a small internal rocket motor) and this would be a completely new development for the OSCAR Series. This motor would be fired by ground control some orbits after launch at a time determined by the orbit mechanics. It is anticipated that the AKM will push the satellite into an initial apogee over the North Pole of 7.2 earth radii. About 1000 watts EIRP would be required for effective communication at apogee.

To this end, and to further advance our command techniques, it was decided to fly, also for the first time, an onboard computer. This unit would integrate the Command, Telemetry and general housekeeping of the whole spacecraft. The Computer would interface directly with Ground Station Equipment (GSE) computers in the worldwide chain of command stations. The Spacecraft computer would also arrange the transmission of telemetry in any format (RTTY, CW, BCD et al) as decided by the software fed from the command stations. Commands and operating schedules would also be decided in like manner by ground loaded software.

All this is an interesting technical exercise from the participant point of view, but what about the OSCAR users?

The principal transponder would be a linear unit of 150 kHz bandwidth with reception either in the 2 m or 70 cm band and transmission in the alternative (70 cm or 2 m band). The exact choice of uplink, 2 m or 70 cm, and thus downlink, was not decided and the conference chose to refer this choice to a poll of interested parties.

In general, VE and VK with some of the W's favoured 2 m up and 70 cm down; the DL and AMSAT HQ representatives were in favour of the alternative (as in OSCAR 7).

Two or three Beacons will be flown. There will be a beacon at each end of the passband and, possibly, a 2304 MHz beacon if the present problem with the FCC on this question can be overcome.

The responsibilities of the groups involved in building the spacecraft were laid down as follows:

AMSAT Deutschland:	Design major units of spacecraft, i.e., transponder, integrated housekeeping unit including computer. Build prototype spacecraft.
AMSAT Canada:	Build spacecraft, both prototype and flight units.
Project Australis:	Design and build GSE equipment with ground computer etc., provide prototype for test use and 5 - 6 integrated units for world command stations before launch. Provide software for both spacecraft and GSE computers.
San Bernardino Microwave Society:	Design and build 2304 MHz beacon.
AMSAT HQ:	Provide overall system management, procure components, arrange launch, provide operations management once spacecraft is in orbit.

As will be seen this is an ambitious program and is, of course, subject to future changes and modifications as circumstances may demand. The planned spacecraft is, however, a logical expansion of the AMSAT-OSCAR program and we believe within the capabilities of the international participants given reasonable fortune and support.

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NOISE IN RECEIVE SYSTEMS

by R.Lentz, DL3WR

One often finds both in professional and amateur circles that the various definitions of sensitivity are mixed together in technical discussions. If the relationships between the various definitions are not known, this often leads to considerable confusion. For instance, those who work with equipment for FM communications speak of the sensitivity of their equipment in μV (sometimes without mentioning or even knowing the appropriate signal-to-noise ratio on what it is based), whereas those involved with SSB communications will talk of noise figures in dB or kT_0. Specialists in satellite communications and radio astronomy finally will give their sensitivity in noise temperature. This shows how difficult it is even for engineers to discuss sensitivity on a common basis, let alone radio amateurs.

The first part of this article is to try to find a clear manner of conversion from one definition to another. A number of diagrams are to be given to ease the conversion for practical use. Firstly, however, NOISE itself is to be studied to find the various sources, and the various terms are to be briefly defined. This is designed as a refresher for the more experienced readers, and as noise fundamentals for those that have not studied the subject in detail.

The second part considers the calculation of the sensitivity of a receive system, (antenna, preamplifier, feeder, receiver), as well as how such a system can be optimized. It will also be shown how the signal-to-noise ratio can be calculated from the signal-plus-noise-to-noise ratio. This conversion is necessary, for instance, when measuring the gain or polar diagram of an antenna under very weak signal conditions (e.g. with the aid of solar noise), since the indicated level in dB is not always the true dB-level.

1. NOISE SOURCES

To limit the scope of this article, noise resulting from, for instance, distortion products, lightning, ignition interference, power-line carried interference, TV line-frequency oscillators etc. is not to be discussed but only the continuous noise audible as a "hissing" noise in the loudspeaker under no-signal or weak-signal conditions. This noise is a statistically random voltage which originally comprises all frequencies (white noise). Noise is generated in resistors, semiconductors and tubes, whereas the antenna picks up galactic noise.

1.1. THERMAL NOISE

This noise category comprises noise generated in all types of resistances: as a circuit component, dissipation resistance, connection or surface resistance of resonant circuits, feeders, waveguides, antennas, vacuum tubes, semiconductors. Reactive resistances (impedances), however, do not generate noise.

The atoms in a conductor are never stationary, but are in a continuous irregular oscillation around their quiescent positions (Brownian motion); the intensity of this movement increases with the temperature of the conductor (thermal agitation). This agitation puts the free electrons also into movement and causes a continuous shift of charge carriers. In this way a voltage is generated across the resistor, whose magnitude and polarity are continuously changing. The mean value of this noise voltage U_n increases with the absolute temperature T, the resistance R of the conductor and the bandwidth B of the amplifier connected to the resistor. The noise power P_n results from the noise voltage and resistance.

In the frequency range B a thermal noise source (resistor) will provide an available noise power to a connected consumer under matched conditions:

$$\boxed{P_n = k \times T \times B} \qquad (1)$$

Where: k = Boltzmann's constant 1.38×10^{-23} Ws/°K
T = Absolute temperature of the resistance in °Kelvin (°K)
B = Bandwidth of the amplifier in Hz

The information given in this section shows that:

The thermal noise of resistances of all types can be reduced by cooling.

The bandwidth should only be as wide as absolutely necessary to pass the required modulation mode.

1.2. CURRENT NOISE

The direct current flowing through active components such as vacuum tubes, semiconductors, diodes, generates a noise power according to various mechanics (e.g. shot-effect noise, flicker-effect noise, current distribution noise, recombination noise). The magnitude of this noise depends on the frequency, technology, circuit and current flow. For instance, we know from experience that some transistor types produce a lower noise than others, that mixer circuits are noisier than amplifier circuits, and that AF-input transistors are operated at very low current levels.

All resistors also generate a current noise in addition to the thermal noise when a current flows through them. This noise is generated by inhomogenities of the resistance material. Small resistors are noisier than larger ones, metal-film resistors are not so noisy as carbon resistors. High noise levels are produced by poor contacts when current flows through them.

1.3. NOISE RECEIVED BY THE ANTENNA

Electromagnetic energy is picked up by the antenna from earth, from the atmosphere and from space with various scattered and discrete sources such as the sun. It is indicated in the receiver as noise. It is important to know that different noise sources are dominant in the different frequency ranges. Figure 1 shows average values of the various noise components in the frequency range from 10 MHz to 10 GHz, as received using an omnidirectional antenna (1). The values given for cities and suburban areas (USA) seem however to be higher than to be expected in Europe. The noise figure of a good receiver (Section 2.3.) is also given for comparison.

In the shortwave range upto about 20 MHz, the noise picked up by the antenna is higher than the "electronic" noise generated in a good receiver. At higher frequencies, the receiver noise becomes more and more the factor determining the sensitivity threshold. As can be seen in Figure 1, it is not worthwhile trying to obtain an extremely low noise figure with a shortwave receiver for use upto 20 MHz, and probably reduce its large-signal capabilities. A noise figure of 10 dB is usually sufficient except for the two higher shortwave bands (21 and 28 MHz). If a VHF or UHF converter is to be used, it will be advisable to calculate the system sensitivity as given in Section 4.

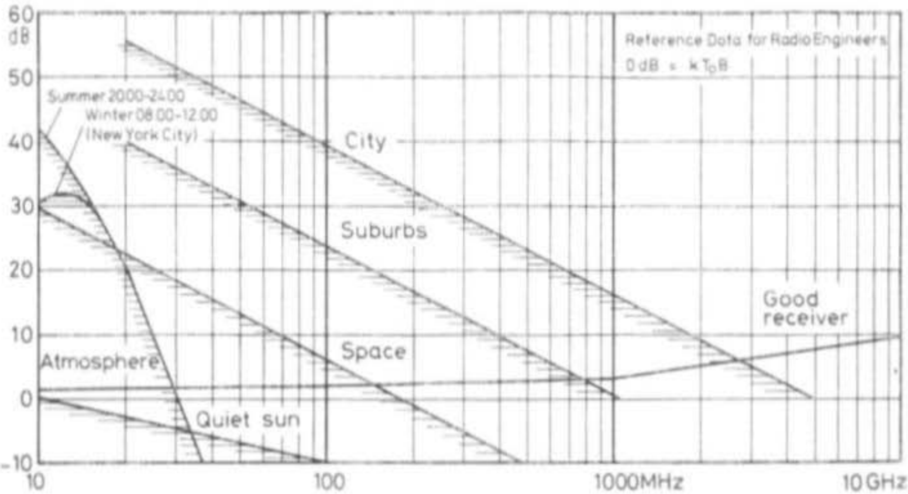

Fig.1: Interference and noise power level picked up by an omnidirectional antenna

2. VARIOUS DEFINITIONS FOR SENSITIVITY

2.1. NOISE FACTOR F

The noise factor is defined as follows when the required signal $P_{r\ in}$ and noise power $P_{n\ in}$ are present at the input of a fourpole such as an amplifier, mixer, receiver, attenuator, etc. and the required output signal $P_{r\ out}$ and noise signal $P_{n\ out}$ are present at the output:

$$F = \frac{P_{r\ in}/P_{n\ in}}{P_{r\ out}/P_{n\ out}} = \frac{(P_r/P_n)_{in}}{(P_r/P_n)_{out}} \qquad (2)$$

or with $P_r/P_n = S/N$:

$$\boxed{F = \frac{(S/N)\ in}{(S/N)\ out}} \qquad (2a)$$

It will be seen that the noise factor F indicates the factor by which the signal-to-noise ratio is deteriorated on passing through a four-pole (receiver, amplifier etc.).

2.2. ADDITIONAL NOISE FACTOR F_a

The noise factor F has practically always a value of greater than 1. In the ideal case where the four-pole does not generate any intrinsic noise, the noise factor will be F = 1. The intrinsic noise component of the four-pole itself is defined as the additional noise factor F_a.

The noise factor F therefore comprises:

$$F = 1 + F_a \tag{3}$$

This means that the additional noise factor F_a is:

$$\boxed{F_a = F - 1} \tag{3a}$$

Noise factor F and additional noise factor F_a are often used and exchanged randomly (especially on data sheets and advertising material). As long as F is far greater than 1, the errors will be small but as soon as the noise factor F of low-noise amplifiers approaches 1, the error can be extremely great.

2.3. NOISE FIGURE $F_{(dB)}$

The noise figure is also defined with F as the noise factor. It is only that the noise factor F has been converted in to a logarithmic scale and is given in dB (decibel). This is as follows:

$$\boxed{F_{(dB)} = 10 \log F} \tag{4}$$

2.4. THE kT_o FACTOR

If one assumes an ambient temperature of 17 °C (T_o = 273 + 17 = 290 °K) and inserts this into equation (1), the following value will be obtained for kT_o:

$$\begin{aligned} kT_o &= 1.38 \times 10^{-23}\ Ws/^{o}K \times 290\ ^{o}K \\ &= 4 \times 10^{-21}\ Ws = 4 \times 10^{-21}\ W/Hz \end{aligned}$$

An ideal, noise-less receiver would have a sensitivity threshold of 1. Under practical conditions, one measures how much higher the signal power must be than for the ideal receiver in order to obtain the intrinsic noise threshold of the receiver $(S/N)_{out} = 1$.

The measuring value is $F = \frac{n \times kT_o}{kT_o} = n$. This is a dimensionless figure corresponding to the noise factor F.

This kT_o value has established itself for sensitivity measurements in a number of countries (2). It is not to be found in English or American literature. Even in those countries that use kT_o, this definition is being replaced more and more by the terms noise figure $F_{(dB)}$ and noise temperature T.

2.5. SENSITIVITY THRESHOLD IN μV

As already mentioned, the sensitivity threshold is the lowest power necessary to obtain a ratio of required power P_r to noise power P_n of 1 at the output.

If a noise-less receiver with a bandwidth B (Hz) is connected to a noise source of temperature T (°K), the receiver will receive the following noise power P_n (W) according to equation (1):

$$P_n = kTB$$

If the noise source has an impedance Z_{out} and the receiver has an input impedance Z_{in} and if matched conditions are present ($Z_{out} = Z_{in} = Z$), it is possible to calculate the effective noise voltage U_n from the available noise power P_n:

$$\boxed{\frac{U_n^2}{Z} = kTB \quad \text{or} \quad U_n = \sqrt{Z \times kTB}} \qquad (5)$$

Assuming, for instance, that the bandwidth of a receiver is 300 kHz, the ambient temperature is 290 °K (17 °C), and the input impedance is 50 Ω, the sensitivity threshold of this receiver will be as follows (intrinsic noise = 0 or F = 1 or F = 0 dB):

$$U_n = \sqrt{50\,\Omega \times 1.38 \times 10^{-23}\,\mathrm{Ws/^oK} \times 290\,^oK \times 3 \times 10^5\,\mathrm{Hz}} = 0.25\,\mu V$$

If the bandwidth of the receiver is reduced from 300 kHz to 3 kHz, in other words by factor 100, the sensitivity threshold will improve by factor 100 = 10. This amounts to 0.025 μV if the other conditions (F = 0 dB; $T = T_o$ = 290 °K; Z = 50 Ω) are still valid.

Under practical conditions, the intrinsic noise of a non-ideal receiver must be added to the sensitivity threshold. For calculation, the measured noise factor (not noise figure in dB) must be inserted into equation 5 as follows:

$$\boxed{U_{n\ threshold} = \sqrt{F \times Z \times kTB}} \qquad (5a)$$

At F = 6 dB or F = 4, the threshold voltage will be:
$U_{n\ threshold}$ = 0.025 μV x 2 = 0.05 μV in the last example.

2.6. SENSITIVITY IN μV

Usually it is not the sensitivity threshold that is given for a receiver but the input voltage level for a certain signal-to-noise ratio. As long as the RF-signal-to-noise ratio $(S/N)_{RF}$ prior to demodulation is meant, the calculation is made as follows:

Example 2.6.1.
Given is a sensitivity threshold of 0.5 μV
($T = T_o$; B = 300 kHz; F = 6 dB; Z = 50 Ω).

Required is the input voltage for a signal-to-noise ratio of 26 dB (power ratio of 400):

$$U_{(26\ dB)} = U_{n\ threshold} \times \sqrt{400} = 0.5\,\mu V \times 20 = 10\,\mu V$$

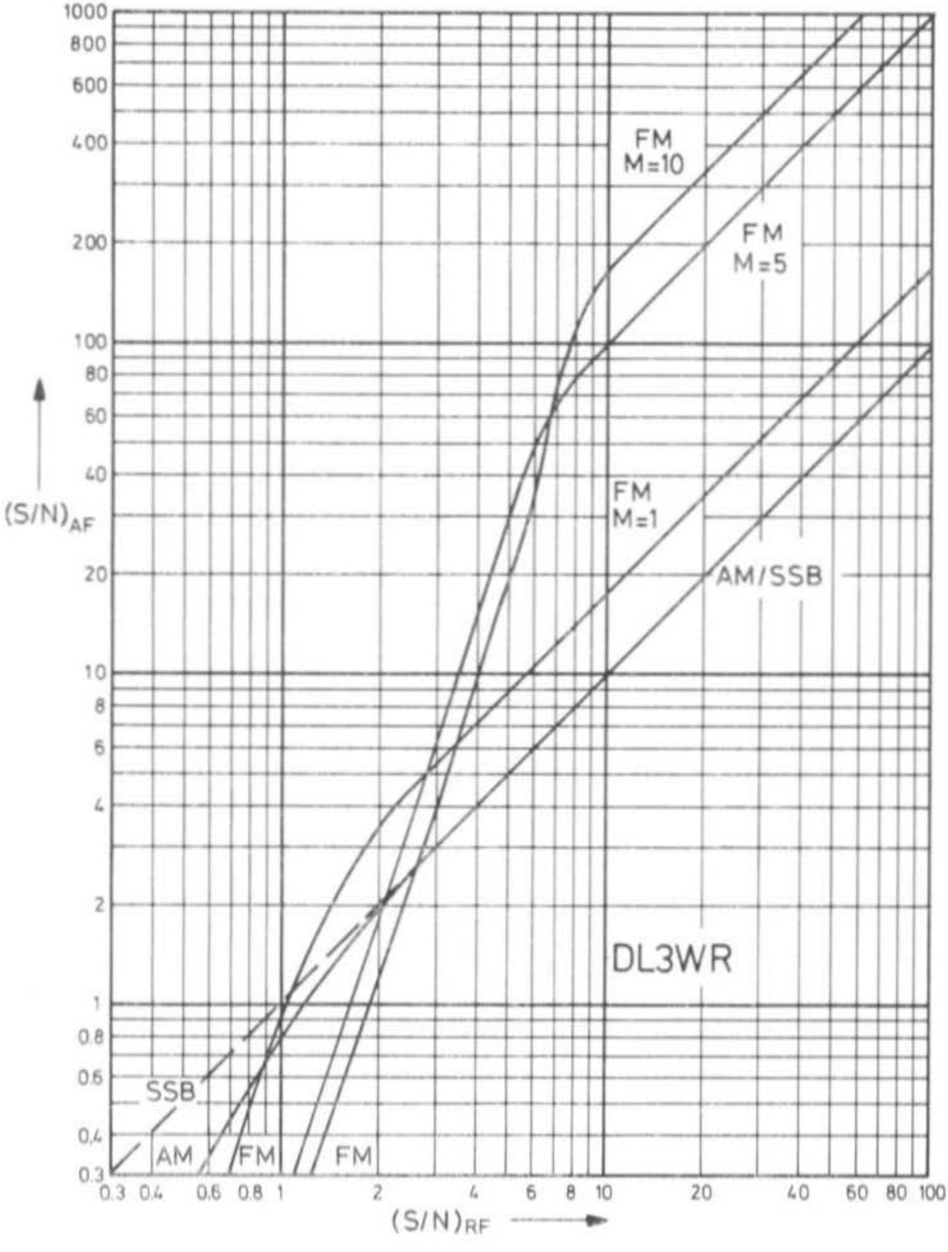

Fig.2: Audio S/N ratio as a function of the RF S/N ratio with SSB, AM and FM with various modulation indexes

However, it is usually the audio signal-to-noise ratio $(S/N)_{AF}$ that is given for FM-receivers. This means that the modulation mode characteristics must also be considered. These are compared in Figure 2, which are taken from References (3) and (4). It will be seen that the AF signal-to-noise ratio of frequency modulated signals increases far faster on increasing the RF signal-to-noise ratio than AM signals, and the greater the modulation index M (M = deviation/modulation frequency), the greater will be the absolute values. However, a certain threshold must be exceeded, and SSB is clearly superior below this threshold.

Example 2.6.2.
Given is again a sensitivity threshold of 0.5 µV; required is the input voltage for an AF signal-to-noise ratio of 26 dB, at a modulation index of M = 10 (e.g. deviation 5 kHz; mod. frequency 500 Hz).

As can be seen in Figure 2, the following is valid for FM with M = 10 with $(S/N)_{AF}$ = 400 (= 26 dB) : $(S/N)_{RF}$ = 24. According to the previous example we can now calculate the required input voltage as follows:

$$U_{(26\ dB)\ AF} = U_{n\ threshold} \times \sqrt{24} = 0.5\ \mu V \times 4.9 = 2.45\ \mu V$$

Example 2.6.3.
Given are values for amateur repeater operation:
$Z = 50\ \Omega$; $T = T_o$; $B = 12$ kHz; deviation = 4 kHz; $f_{mod} = 1.75$ kHz;
F = 3 dB.

Required are: Sensitivity for an AF signal-to-noise ratio of 20 dB:

$$U_{n\ threshold} = \sqrt{2 \times 50\ \Omega \times 1.38 \times 10^{-23}\ Ws/^{o}K \times 290\ ^{o}K \times 12 \times 10^{3}\ Hz} = 0.07\ \mu V$$

$$M = \frac{4\ kHz}{1.75\ kHz} = 2.3$$

According to Figure 2: $(S/N)_{RF} \approx 16$ at $(S/N)_{AF} = 100$ and M = 2.3

This means: $U_{(20\ dB)\ AF} = U_{n\ threshold} \times \sqrt{16} = 0.07\ \mu V \times 4 = 0.28\ \mu V$

2.7. NOISE TEMPERATURE

A resistor with a temperature 0 °K (= -273 °C) produces zero noise power: $P_n = kTB = 0$, thus F = 1 or $F_a = \emptyset$. If the temperature of the resistor is increased to ambient temperature of $T_o = 290\ ^{o}K$, the resulting noise power will be $1\ kT_oB$. The noise temperature of a four-pole is then the temperature by which a resistor produces the same noise power as the four-pole under test (amplifier, receiver, cable, antenna).

A directional antenna with negligible intrinsic noise pointed towards the earth's surface or horizon will have a noise temperature of T = 290 °K (ambient temperature 17 °C). If on the other hand, the antenna is orientated towards a quiet (cool) part of the sky, values of less than 10 °K can be obtained at microwave frequencies.

It is easy to convert the noise factor F to noise temperature T, since it has been seen that T = 0 °K corresponds to F_a = 0, and the additional noise factor F_a = 1 corresponds to T = T_o:

$$\boxed{T = (F - 1) \times T_o = (F - 1) \times 290\ °K} \qquad (6)$$

Assuming an amplifier possesses a noise factor of F = 3 (or F_a = 2), its noise temperature will be:

$$T = (3 - 1) \times 290\ °K = 580\ °K$$

3. RELATIONSHIP BETWEEN THE DEFINITIONS

This section is to bring a number of diagrams to ease practical conversion now that the calculation is known from the previous sections:

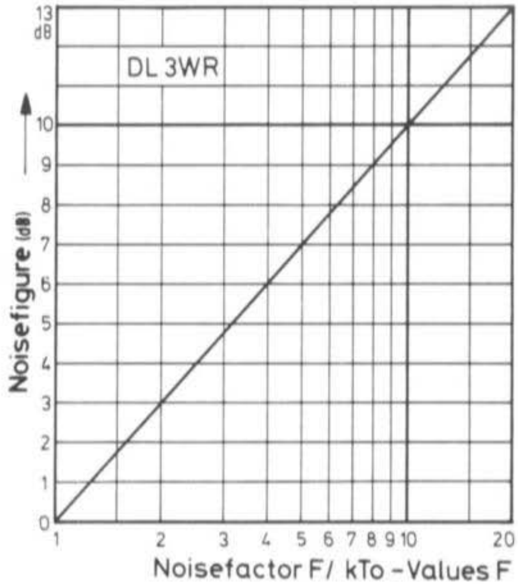

Fig.3: Diagram for converting noise factor F into noise figure F_{dB}

The first diagram, given in Figure 3, shows the relationship between the noise figure $F_{(dB)}$ and the noise factor F. It can also be used to convert kT_o-values into noise figure values. For this diagram equation 4 is valid which is transposed as follows:

$$F = 10 \exp \frac{F_{dB}}{10} \qquad (4a)$$

Figure 4 allows the conversion from sensitivity threshold in μV to noise factor or noise figure. This diagram has been calculated according to equation 5. If the example given in section 2.5. is used, where F = 4 dB was measured, the sensitivity threshold at a bandwidth of B = 3 kHz and Z = 50 Ω will be 0.04 μV.

In order to keep the diagram as clear as possible, only the lines corresponding to 60 Ω have been drawn except for a bandwidth of 1 MHz where 50 Ω, 60 Ω and 75 Ω are given. However, the 50 Ω and 75 Ω lines can be easily found by parallel shift of the lines.

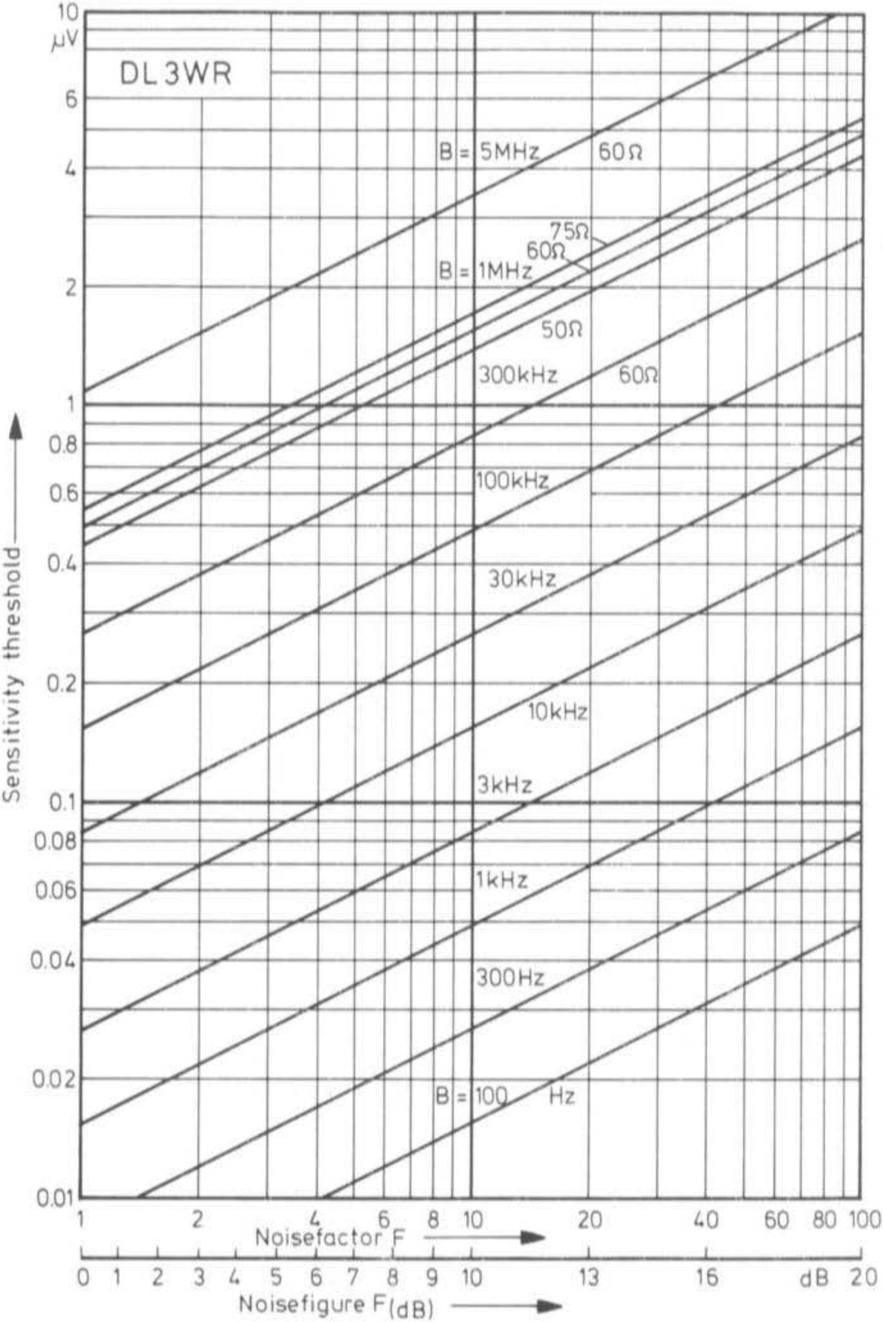

Fig.4: Diagram for converting sensitivity threshold into noise figure or noise factor

Equation 6 has been used to calculate the diagram given in Figure 5. Scales for both noise factor and noise figure are given for ease of use.

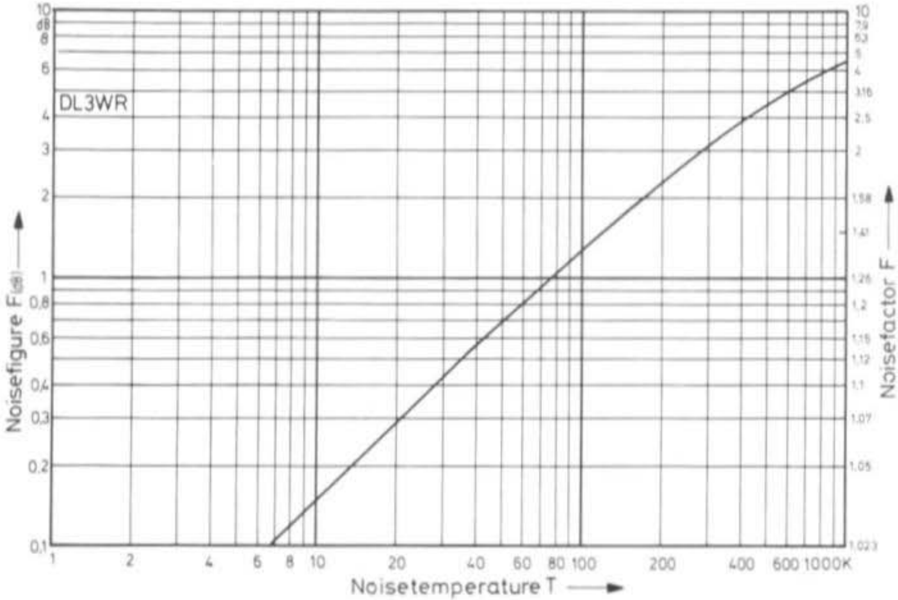

Fig.5: Diagram for converting noise figure and noise factor into noise temperature

Figure 6 finally gives all the sensitivity definitions in the form of a nomogram to allow quick conversions even though it does not have the fine resolution of the previous diagrams.

The noise power level in dBm (dB referred to 1 mW) is also given which can be calculated from equation 1 as follows, assuming $T = T_o = 290\ ^oK$ and $B = 1$ Hz:

$$P_n = kT_oB = 4 \times 10^{-21}\ W$$

This noise power is now referred to 1 mW:

$$4 \times 10^{-21}\ W \ \hat{=}\ -204\ dBW = -174\ dBm$$

If the bandwidth is 1 kHz instead of 1 Hz (Factor of 1000 wider = 30 dB), the noise power level will be -144 dBm.

Use of the nomogram:
The measured noise figure (or noise factor or noise temperature) is joined to the bandwidth of the IF amplifier of the receiver by a straight line. This straight line will cross the noise level scale, and this intersection will show which signal power level is required to produce a signal-to-noise ratio of 1 (0 dB). The corresponding voltage across 50 Ω is given to the right of this scale.

Example: The sensitivity threshold at a noise figure of F = 3 dB (F = 2 or T = 290 °K) and an IF-bandwidth of 500 Hz will be - 144 dBM or 0.014 μV across 50 Ω.

The noise power level of a receive system is very useful for calculation of the expected signal-to-noise ratio when the transmit power, cable loss, antenna gain and path loss are known.

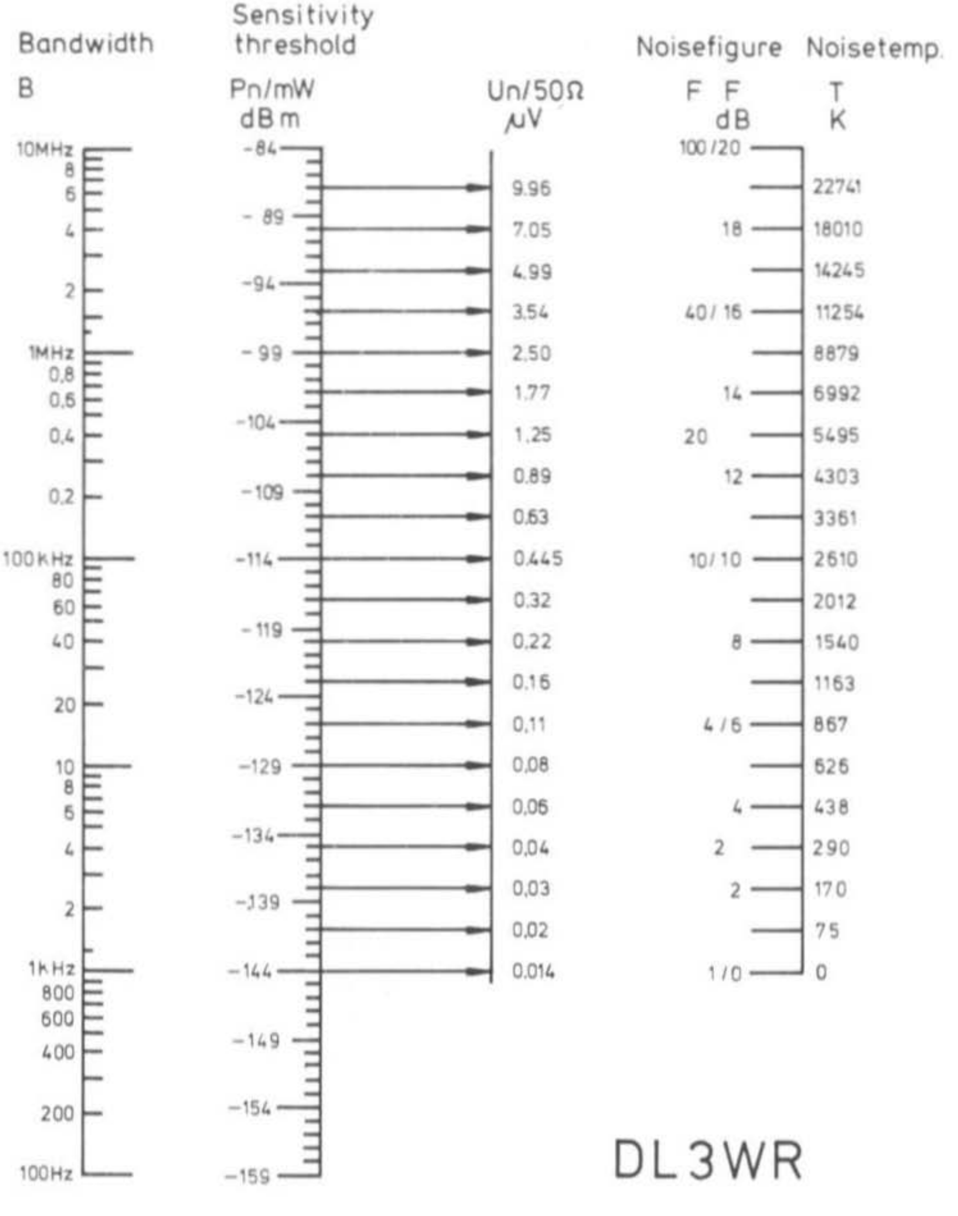

Fig.6: Nomogram for conversion of the various noise definitions

4. SENSITIVITY OF A RECEIVE SYSTEM

A receive system comprises various sections having different noise levels: IF-amplifier and mixer stages usually produce considerably more noise than the RF-amplifier stages. The feeder between the antenna and the first RF amplifier stage is a very important factor determining the overall noise figure of a receive system. As can be seen in Figure 1, all noise contributions of a receive system become more and more important with increasing frequency.

A receive system therefore comprises antenna, feeder, possibly a preamplifier with subsequent cable, as well as the actual stages of the receiver (RF-amplifier, mixer, IF-amplifier etc.). It is possible to determine the noise figures and gain figures of the individual sections or at least those of the receiver and the preamplifier as well as the loss of the feeder cable. Of interest to us are:

The overall noise figure of the receive system;
How the system can be optimized.

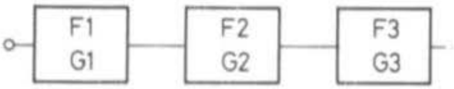

Fig.7:
Series connection of several sections of a receive system

Figure 7 shows a series circuit of several sections of a receive system. Such a configuration can comprise, for instance, RF-amplifier, mixer, and IF amplifier. The first section can, of course, also be the feeder cable, and section two the complete receiver. In this case, only two sections are considered. The noise factors F of each section of the receive system, as well as the gain (or loss) of all sections except the last must be known in order to establish the overall noise factor according to equation 7.

$$\boxed{F_{tot} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_2 \times G_1} + \frac{F_4 - 1}{G_3 \times G_2 \times G_1}} \qquad (7)$$

corresponding to:

$$F_{tot} = F_1 + \frac{F_{a2}}{G_1} + \frac{F_{a3}}{G_2 \times G_1} + \ldots \qquad (7a)$$

Attention: Noise and gain factors must be used in equation 7 and not dB values. Conversion can be made with the aid of Figure 3.

4.1. ALL SECTIONS HAVE A POSITIVE GAIN

The practical use of equation 7 can be best shown with the aid of an example:

Example 4.1. Required is the overall noise figure of a receiver having the following sections:

1. RF-amplifier with F_1 = 3 dB (F_1 = 2) and G_1 = 10 dB (G_1 = 10)
2. Mixer with F_2 = 10 dB (F_2 = 10) and G_2 = 6 dB (G_2 = 4)
3. IF amplifier with F_3 = 8 dB (F_3 = 6.3) and G_3 = 60 dB (G_3 = 10^6)
4. Rest of receiver: negligible even at high noise figures F_4

$$F_{tot} = 2 + \frac{10-1}{10} + \frac{6.3-1}{4 \times 10} + \frac{F_4 - 1}{10^6 \times 4 \times 10}$$

$= 2 + 0.9 = 0.133$ The 4th term can be neglected

$= 3.03 \mathrel{\hat{=}} 4.8$ dB due to the large nominator.

Equation 7 and the example show that it is mainly the first section that determines the overall noise figure. As long as all sections possess positive gain values (not losses) and reasonable noise figures, it will be seen that the noise contribution of the subsequent sections becomes rapidly smaller so that the third, or at least the fourth section can be neglected.

It would be ideal for the first section to have the lowest possible noise figure and exhibit a high gain of say 100 (20 dB). In this case, the second term in our example would only be 0.09! However, since both these characteristics can hardly be realized simultaneously, and since the large signal handling capabilities must also be considered, a certain compromise must be made between noise figure and gain. Very low noise figures are usually possible at gain factors of 10 to 20 (10 - 13 dB).

4.2. THE FIRST SECTION HAS A LOSS

A practical case is now to be considered where the first section of our receive system is a lossy feeder (3 dB loss) between antenna and receiver, whereas all other sections remain as before. When considering the fact that the additional noise figure of passive components corresponds to the loss factor, the following results:

<u>Example 4.2.1.</u>

$F_1 = 2;$	$G_1 = 1/2 = 0.5$	(feeder)
$F_2 = 2;$	$G_2 = 10$	(RF-amplifier)
$F_3 = 10;$	$G_3 = 4$	(Mixer)
$F_4 = 6.3;$	$G_4 = 10^6$	(IF-amplifier)

$$F_{tot} = 2 + \frac{2-1}{0.5} + \frac{10-1}{10 \times 0.5} + \frac{6.3-1}{4 \times 10 \times 0.5}$$

$$= 2 + 2 + 1.8 + 0.27$$

$$= 6.07 \mathrel{\hat{=}} 7.8 \text{ dB}$$

The above calculation is also valid when the first section of a receiver is a passive mixer (e.g. Schottky ring mixer). With, for example, $F = 6$ dB and $G = -7$ dB $= 1/5$ which is directly followed by the IF-amplifier with $F = 6.3$ and $G = 10^6$, the following total noise factor results:

<u>Example 4.2.2.</u>

$F_1 = 4;$ $G_1 = 1/5 = 0.2$
$F_2 = 6.3;$ $G_2 = 10^6$

$$F_{tot} = 4 + \frac{6.3-1}{0.2} = 4 + 26.5 \mathrel{\hat{=}} 30.5 = 14.8 \text{ dB (!)}$$

It will be seen from these examples that:

The noise figure of a receive system is always higher than the attenuation of the first section.

All subsequent sections have a greater effect on the overall noise figure than would be the case when the first section possessed a gain instead of a loss.

A low-noise preamplifier at the receiver end of a lossy cable will not provide very much improvement. However, if this low-noise preamplifier is placed at the antenna end of the lossy cable then a considerable improvement of the overall noise figure will be obtained. This is indicated clearly in the following example:

Example 4.2.3.

$F_1 = 2$; $G_1 = 10$ (preamplifier)
$F_2 = 2$; $G_2 = 0.5$ (feeder)
$F_3 = 2$; $G_3 = 10$ (RF-amplifier)
$F_4 = 10$; $G_4 = 4$ (mixer)

$$F_{tot} = 2 + \frac{2-1}{10} + \frac{2-1}{0.5 \times 10} + \frac{10-1}{10 \times 0.5 \times 10}$$

$$= 2 + 0.1 + 0.2 + 0.18 = 2.48 \triangleq 3.9 \text{ dB}$$

It is possible to calculate how much improvement a low-noise preamplifier could provide in front of the passive mixer used in example 4.2.2. A slight improvement of the overall noise figure in example 4.2.2. could be obtained when a low-noise first IF amplifier was used: with $F_2 = 2$ dB $= 1.6$: $F_{tot} = 7 = 8.5$ dB.

The improvement of the overall noise figure provided by placing a low-noise preamplifier at the antenna end of the feeder increases with the loss of the cable. The aim is, to maintain the signal-to-noise ratio at the antenna without too much deterioration. However, there are problems such as weather proofing, stability, relay switching etc. that must be solved.

4.3. THE ANTENNA AS FIRST SECTION OF A RECEIVE SYSTEM

If the antenna is also to be considered in the calculation of the overall sensitivity, it will be necessary to work with noise temperatures. An antenna pointing towards the horizon (not to the sun) will attain a noise temperature of 300 °K as indicated in section 2.7., assuming that it does not possess any noticeable losses. It is only necessary to convert the values given in the various examples for the total noise figure to noise temperatures (Fig. 5) and then establish which section of the system provides the highest noise contribution and is therefore worthwhile to be improved. Of course, the information given in Figure 1 must be considered which shows that this is only worthwhile at frequencies of 430 MHz and higher.

Let us now calculate the improvement obtained when exchanging a very low-noise, two-stage preamplifier at 23 cm (F = 3 dB; G = 20 dB) by an uncooled parametric amplifier as given in reference (8) having T = 100 °K, G = 20 dB. Both are practical at this frequency. Calculations are to be made with the antenna pointed towards the horizon, and also when pointed towards the moon at a high angle.

An antenna temperature of 50 °K is assumed for the second case. Subsequent cable loss and receiver noise figure are not to be considered since the preamplifier is located directly at the antenna, and the high gain of the preamplifier is sufficient to blanket all succeeding contributions.

$$\boxed{T_{tot} = T_{ant} + T_{Rx}} \qquad (8)$$

Example 4.3.1. (Horizon)

T_{ant} = 300 °K

$T_{preamp.1}$ = 290 °K (3 dB) $\qquad$ $T_{tot\ 1}$ = 590 °K ≙ $F_{tot\ 1}$ = 4.9 dB

$T_{preamp.2}$ = 100 °K $\qquad$ $T_{tot\ 2}$ = 400 °K ≙ $F_{tot\ 2}$ = 3.9 dB

Example 4.3.2. (Moon)

T_{ant} = 50 °K

$T_{preamp.\ 1}$ = 290 °K $\qquad$ $T_{tot\ 1}$ = 340 °K ≙ $F_{tot\ 1}$ = 3.4 dB

$T_{preamp.\ 2}$ = 100 °K $\qquad$ $T_{tot\ 2}$ = 150 °K ≙ $F_{tot\ 2}$ = 1.8 dB

If the preamplifier is not to be located directly at the antenna, it is possible to use a diagram given in reference (9) in order to convert the noise temperature at the antenna to the noise temperature at the end of the cable.

$$\boxed{T_{ant + cable} = \frac{290\,^{\circ}K\,(L - 1) + T_{ant}}{L}} \qquad (9)$$

Where: L is the loss of the cable as factor and not as dB value.

For tropospheric communication along the surface of the earth (antenna pointed towards the horizon), an antenna temperature of 300 °K or more must always be assumed. In the case of EME (Moonbounce) or satellite communications, the noise temperature of an unsuitable antenna can remain high inspite of the fact that it is pointing towards quiet direction in space. In the latter case, even an expensive parametric amplifier will not be able to improve the system sensitivity to any degree.

If a parametric amplifier is available, it will be worthwhile to study the individual noise contributions of the antenna. The total noise produced by an antenna comprises cosmic noise T_c, atmospheric noise T_a, ground noise T_g and intrinsic noise of the antenna T_i:

$$\boxed{T_{ant} = T_c + T_a + T_g + T_i} \qquad (10)$$

The cosmic and atmospheric noise temperatures are dependent on the frequency and on the direction in which the antenna is pointing. They are therefore magnitudes that cannot be affected by the antenna designer.

The intrinsic noise temperature T_i of the antenna, however, can be reduced by use of low-loss materials (highly conductive material and large, smooth surface areas), as well as the use of suitable antenna designs such as horn reflectors and cassegrain antennas. The noise temperature T_g contributed by heat radiation of the earth can be kept at a minimum by maintaining a good front-to-back ratio and clean polar diagram.

Two equal-gain antennas can possess different noise temperatures, and can therefore noticeably differ in satellite and EME communications. The author would, however, like to point out that these considerations are only valid in the GHz range when using preamplifiers with noise figures of 3 dB or less at the antenna.

5. SIGNAL-TO-NOISE RATIOS AT WEAK SIGNAL LEVELS

When are measured dB not true dB? This does not refer to measuring errors in equipment (non-linear amplifiers, demodulators, meters, etc.), but refers to gain measurements at very low signal-to-noise ratios. In this case the gain indicated in dB can be far lower than the actual gain. This can be more clearly seen in the following example:

Example 5.1.

It is assumed that two 70 cm stations are comparing their receive systems by measuring solar noise. Station A receives the solar noise 5 dB above the background noise (antenna facing away from the sun), and station B 3 dB above noise. The assumption that system A is 2 dB more sensitive than system B is, however, false! A conversion must be made from signal-plus-noise-to-noise to signal-to-noise ratio:

$$\boxed{\frac{S}{N} = \frac{S+N}{N} - 1} \qquad (11)$$

Once again the numerical values and not dB values should be inserted!

Station A: $\frac{S+N}{N} = 5 \text{ dB} \mathrel{\hat{=}} 3.162$ Station B: $\frac{S+N}{N} = 3 \text{ dB} \mathrel{\hat{=}} 2$

$$\frac{S}{N} = 3.162 - 1 = 2.162 \mathrel{\hat{=}} 3.35 \text{ dB} \qquad \frac{S}{N} = 2 - 1 = 1 \mathrel{\hat{=}} 0 \text{ dB}$$

The difference between the two systems is therefore 3.35 dB and not 2 dB.

Since the conversion from (S + N)/N to S/N and vice versa is somewhat complicated due to the intermediate conversion into the linear scale, a diagram was calculated for this conversion which is given in Figure 8. It allows the corresponding values to be read off on the appropriate scale. At higher signal strengths of (S + N)/N = 10 dB, the differences become smaller and smaller so that the error due to the included background noise can be neglected. For this reason, the signal strengths during gain measurements should be at least 10 dB to 20 dB or more above noise. Sometimes this is not possible as can be seen in the following example:

Example 5.2.

The beamwidths of a high performance antenna are to be measured with the aid of solar noise. This is achieved by pointing the antenna at a point on the sun's path so that the sun passes through the beamwidth of the antenna. The resulting antenna diagram could be similar to the curve given in Figure 9.

Since the angular speed of the sun of 15^{o}/hour is known, it is possible to convert the time scale into degrees of arc. Of interest are the 3 dB beamwidths and the suppression of minor lobes.

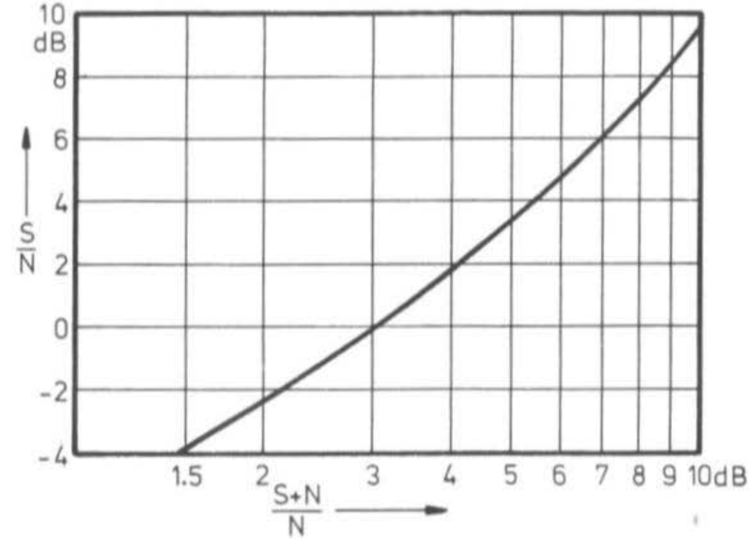

Fig.8: Diagram for converting (S+N)/N to S/N

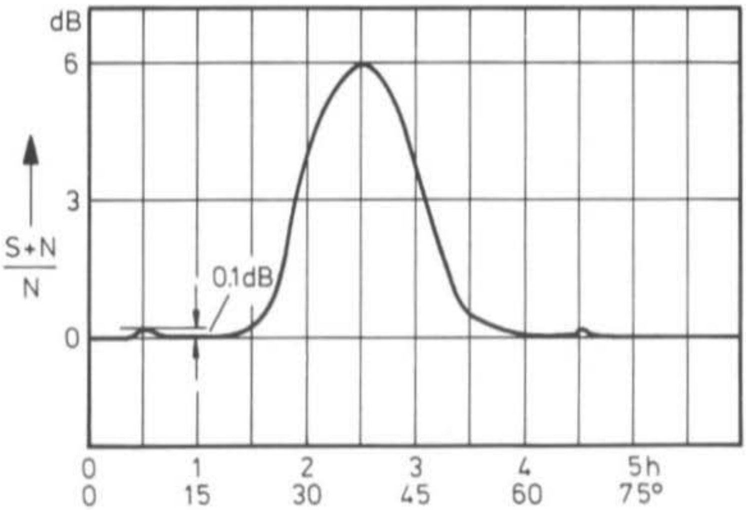

Fig.9: Antenna xy-diagram recorded with the aid of solar noise

If we take a value of 3 dB less than the maximum value, the resulting value would be, for example, 17°. On the other hand, if we convert the (S + N)/N = 6 dB into S/N = 4.77 dB and subtract 3 dB from this value, we will obtain S/N = 1.77 dB and (S + N)/N = 4 dB. The required beamwidth is therefore at the 4 dB points or 2 dB down on the maximum value. This corresponds to about 14° in Figure 9.

The suppression of the minor lobes seems to be only 5.9 dB, however, when this is converted we find:

	(S + N)/N	S/N
Maximum	6 dB	4.77 dB
Minor lobes	0.1 dB	-16.33 dB
Difference		21.1 dB

This means that the minor lobes are more than 20 dB down on the main lobe, which represents a considerable difference to the "measured" 5.9 dB given in Figure 9!

If the solar noise is strong enough, the following simple method of determining the -3 dB beamwidths can be used:

At the maximum of the main lobe a 3 dB attenuator can be inserted into the signal path between antenna and preamplifier (not between preamplifier and receiver) and the indication is noted. The sun then passes through the whole beamwidth of the antenna and the time taken for the sun to pass between these two points corresponding to a value 3 dB down on the maximum is determined.

After this it is only necessary for the time to be converted into degrees of arc (1° for every 4 minutes). If it is possible to rotate the antenna by 90° in its longitudal axis, the measurement can be repeated to establish the beamwidth in the other plane. The antenna gain can then be established from these two beamwidths (11).

When both beamwidths are known, it is in principle possible to establish the sensitivity threshold with the aid of solar noise. Instructions on this were given in (12), however, this author incorrectly calculated the noise temperature (our equation 6) from the noise factor instead of the additional noise factor. Since the measurement requires a constant noise source which is not provided by the sun due to variations of sunspot activity and other effects, the calculation can only be classed as an approximation. For this reason, it is not to be discussed in further detail here.

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A COLINEAR ANTENNA FOR THE 13 cm BAND

by K.Hupfer, DJ 1 EE

The described colinear antenna for the 13 cm band (2304-2306 MHz) was firstly described at the Munich UHF-Convention in 1973. Due to the increasing interest for this frequency band, the antenna is now to be described in detail.

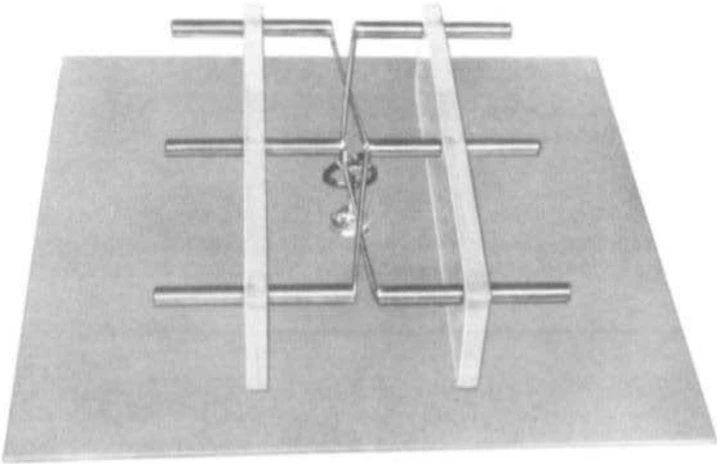

Fig.1: Colinear antenna for the 13 cm band

1. DESIGN

As can be seen in Figure 1, the antenna consists of six $\lambda/2$ elements in front of a common reflector panel. The spacing of the elements from this panel has been selected to be $0.22\ \lambda$ so that the natural free-space feedpoint impedance of a λ-radiator is only slightly affected. The overall length L_{mech} of a λ-radiator is then only mainly dependent on the relationship of radiator length to thickness, and from the spacings between the two $\lambda/2$ elements (see Fig. 2).

When using the same material as the author: 5 mm diameter silverplated brass tube, a shortening factor of approximately 20% will result according to (1). This means that the mechanical length is:

$$L_{mech} = \lambda \times 0.8 \text{ corresponding to } 130\text{ mm} \times 0.8 = 105\text{ mm}$$

The spacing s should amount to 5 mm, which means that each individual $\lambda/2$ radiator possesses a mechanical length of 50 mm.

According to (1), the feedpoint impedance of a centre-fed λ/dipole with a length-to-diameter ratio of 130 : 5 = 26 amounts to a feed-point impedance of Z_{λ}-dipole = 550 Ω. If three such λ-dipoles are connected together in the wellknown manner by crossing the feeder cables, a feedpoint impedance Z_{fp} will result for the colinear antenna which is as follows:

$$Z_{fp} = \frac{Z_{\lambda}\text{-dipole}}{\text{number of }\lambda\text{-dipoles}} = \frac{550}{3} = 183\ \Omega$$

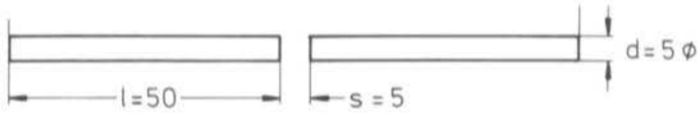

Fig.2: λ-dipole

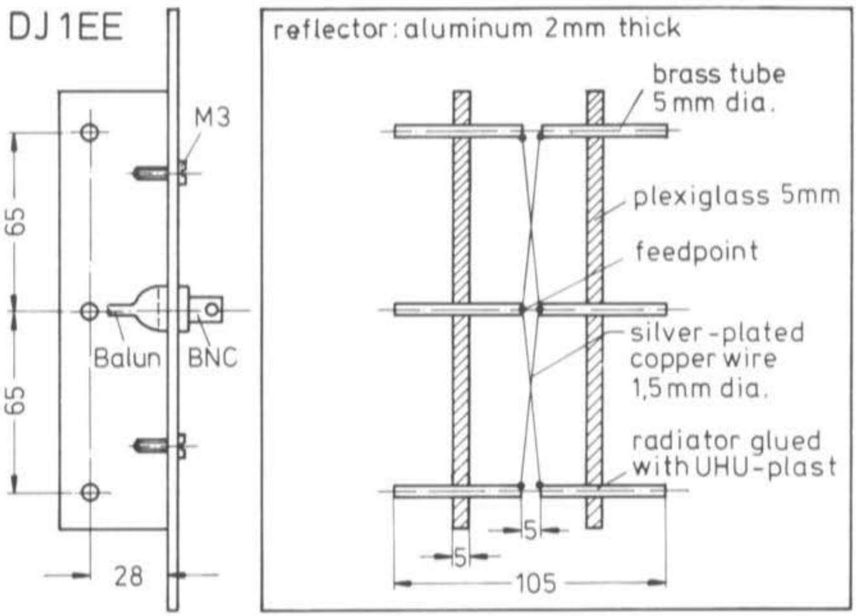

Fig.3: Construction details of the colinear group for 2304 MHz

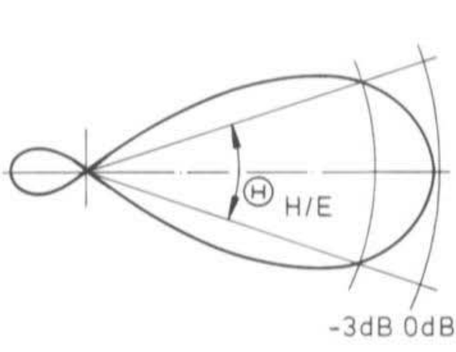

Fig.4: Determining the -3 dB beamwidth

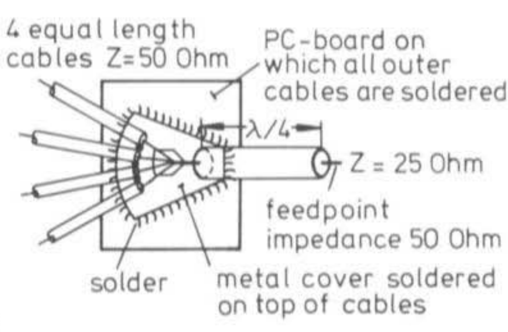

Fig.5: Interconnection and matching of four individual groups

By using a $\lambda/2$ balun, this impedance can be transformed at a ratio of 4 : 1 to a feedpoint impedance of 183/4 = 46 Ω unbalanced. This means that it provides a suitable match for a 50 Ω coaxial cable. Teflon-coaxial cable with a solid copper outer conductor should be used for the $\lambda/2$ balun. Suitable cables are RG 141 or RG 142/U with a 3.6 mm outer diameter and a velocity factor of 69.5%. Such cables possess a loss of 13.5 dB/100 ft at 1 GHz or 27 dB/100 ft at 3 GHz (corresponding to 44 dB or 88 dB/100 m).

The reflector panel should not be too small in order to ensure a good front-to-back ratio. The dimensions of the reflector plate, dipoles and mounting parts are given in Figure 3. It is important that the outer conductor of the coaxial cable from the balun has contact around the whole circumference of the BNC-socket.

2. ANTENNA GAIN

The gain of this antenna was determined using two different methods: The first measurement was made by comparing the field strength with an antenna having a known gain. It is not recommendable for a dipole to be used as reference antenna since the result can be affected by the mount and feeder as well as by ground and other reflections. The author used a simple dipole in front of a reflector panel whose gain was known to be 3 dB. This ensured that no inaccuracies can be caused by reflections from behind the dipole.

Furthermore, the gain was calculated from the measured -3 dB beamwidths of the main lobe (Fig. 4). According to (2) the antenna gain referred to an isotropic source can be calculated according to the following formulas:

$$G = \frac{41253}{\Theta_H \times \Theta_E} \quad \text{with } \Theta \text{ in degrees.}$$

The results of both measurements were of sufficient accuracy and indicated a gain of approximately 10 dB referred to a dipole. In addition to this, the matching was measured. The SWR-value was 1.22 at 2.3 GHz and resonant frequency was 2.34 GHz (SWR = 1).

3. A GROUP OF FOUR SUCH COLINEAR GROUPS

Four of the described colinear groups were then mounted by the author on a common reflector panel. The spacing between each group was the same as if four individual plates were joined together. Each of the individual groups was provided with its own balun transformer and the four equal-length 50 Ω cables from the individual antennas are connected in parallel. The resulting impedance at this point of 50/4 = 12.5 Ω is transformed in a $\lambda/4$ transformer having an impedance of 25 Ω to the required 50 Ω (see Fig. 5). If the $\lambda/4$ length of 25 Ω impedance cannot be manufactured, two $\lambda/4$ lengths of 50 Ω cable can be connected in parallel (Editors: If a $\lambda/4$ length is too short for mechanical reasons, a multiple of $\lambda/4$ can be used). The gain of this group of four colinear arrays was determined to be 6 dB more than an individual colinear group. This means that the gain of the whole array is in the order of 16 dB.

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A MINIATURE RECEIVER FOR THE 2 m BAND

by G.Rühr, OH 2 KT

The described receiver was developed to obtain a miniature receiver equipped with integrated circuits for applications in the aircraft bands of 118 to 136 MHz in the A3-mode as well as on the 2 m amateur band in the F3 or A3-modes. Required was also a receiver that covers the whole range with the aid of a VFO, that could, however, be crystal-controlled. The current drain should be as low as possible, however, the audio output power should be sufficient both for head-set and loudspeaker use.

Although the author did not use any special miniature components with exception of the ceramic capacitors, the described receiver fits into a cast aluminum case having the dimensions of 110 mm x 60 mm x 30 mm. Figure 1 and Figure 2 show the author's prototype. It will be seen that the crystal filter (an industrial surplus type) is by far the largest component. If a smaller filter is available, it is possible to decrease the size of the receiver still further.

This description does not represent a complete construction article, but is brought more for information since the mechanical construction depends greatly on such available parts as variable capacitor and crystal filter. For this reason, no PC-board design is to be published.

Fig.1: Author's prototype of the 2 m receiver **Fig.2: The FM version for the 2 m band**

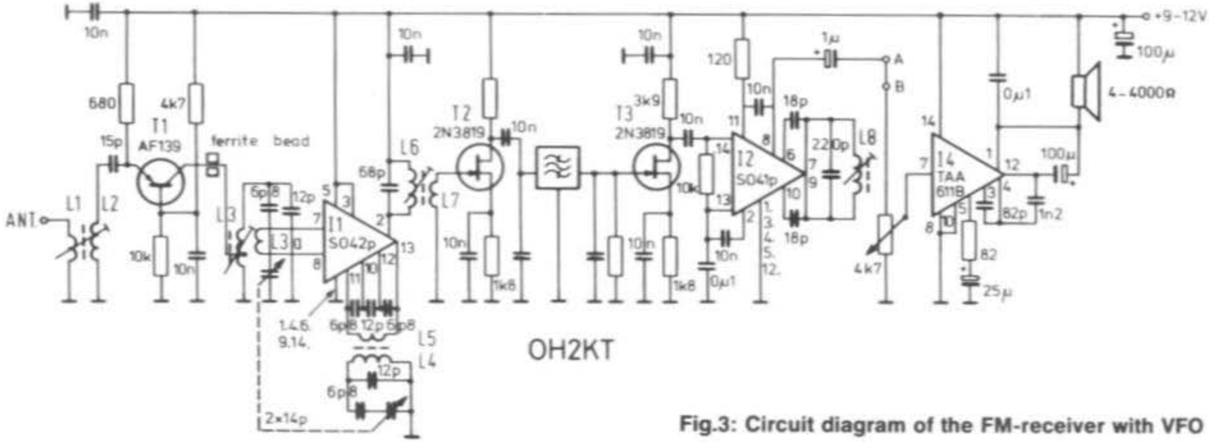

Fig.3: Circuit diagram of the FM-receiver with VFO

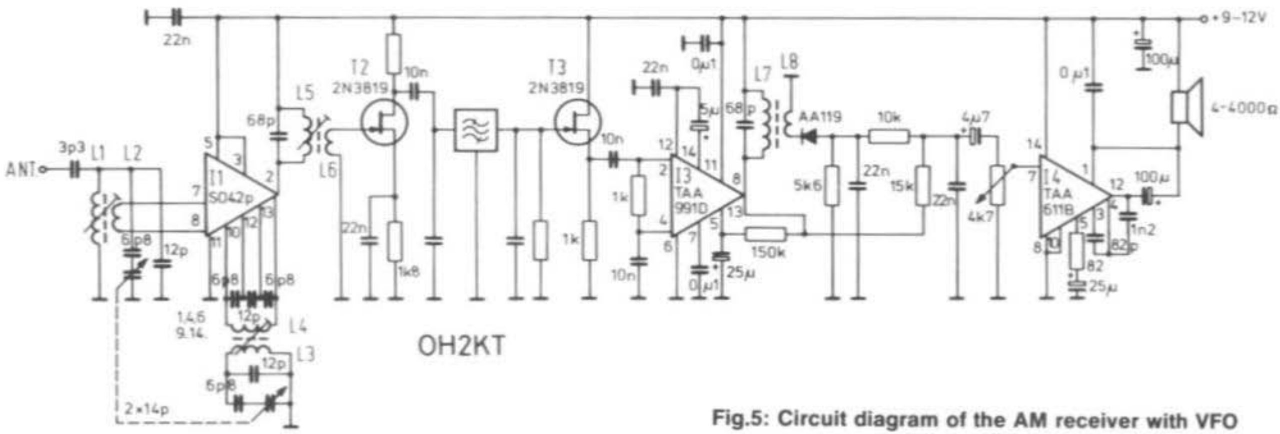

Fig.5: Circuit diagram of the AM receiver with VFO

1. CIRCUIT DESCRIPTION

The circuit diagram of the 144 MHz version with VFO is given in Figure 3. A transistor AF 139 (or BF 324) is used in the RF-amplifier which then drives the integrated circuit SO 42 P which is used as mixer and local oscillator. The intermediate and oscillator circuits are tuned using a small VHF double-variable capacitor with reduction gearing. A variable tuning of the input stage is not required since it is sufficiently wideband. If only one frequency is to be used, or where a very narrow band of frequencies is required, it is also possible for the intermediate stage to have a fixed tuning. The local oscillator operates below the receive frequency in order that no interference is received from FM transmitters in the 160 MHz band due to the relatively low image rejection of this simple receiver.

The output circuit of the mixer is followed by a field effect transistor for matching to the crystal filter (10.7 MHz). A further FET is also to be found at the output of the filter for matching. The values of the capacitors at the input and output of the filter, as well as the resistors in the drain circuit of the first FET and in the gate circuit of the second FET are dependent on the specifications of the crystal filter used.

The second FET is followed by the integrated circuit SO 41 P which comprises a limiting IF-amplifier and coincidence demodulator. This integrated circuit is a low-current version of the well known type TBA 120. The drain resistor of the matching FET between crystal filter and IF-IC can be replaced by a resonant circuit if enough room is available in order to obtain a higher overall gain of the IF amplifier. The whole AF-gain of the receiver is generated in the integrated circuit TAA 611 B, which provides enough output even for loudspeaker operation.

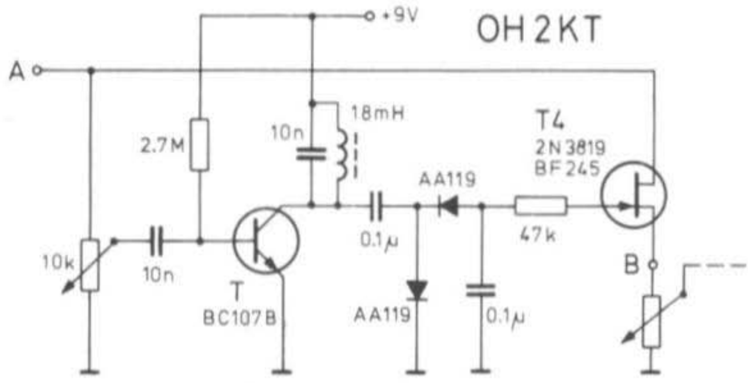

Fig.4: Circuit diagram of a squelch circuit

The receiver was designed for an operating voltage of 9 V, however, any voltage up to 12 V can be used. At voltages of less than 9 V, the gain will fall off rapidly. The connection leading to the volume potentiometer can be disconnected at point A and B for installation of a squelch circuit. A circuit of a suitable squelch is given in Figure 4. It will be seen that it uses an amplifier tuned for a frequency in the upper portion of the audible range, e.g. 12 kHz, which feeds a rectifier circuit, which in turn feeds a FET connected in the form of a resistor in front of the volume potentiometer (1). Under non-signal conditions, the noise in the order of 12 kHz is filtered out, amplified and rectified and used to block the FET. When a signal is received, the noise voltage will fall as will the rectified voltage and the FET will open.

The AM-version of the receiver shown in Figure 5 naturally differs from the FM version in the intermediate amplifier and demodulator. In this case, an integrated circuit TAA 991 is used as IF amplifier. This integrated circuit possesses a built-in automatic gain control (2). No RF amplifier transistor was used in front of the mixer circuit of the AM receiver in order to avoid having to connect the RF-stage to the automatic gain control. The AM-receiver also only possesses one input circuit, and has a correspondingly low image rejection. In addition to this, no squelch was provided because it could not be used in the described manner.

Both receiver versions can also be equipped with a crystal-controlled oscillator when the built-in oscillator circuit of the IC SO 42 P is disabled. The crystal oscillator can, for instance, operate at one third of the local oscillator frequency (overtone crystal for approximately 45 MHz). The oscillator can be followed by a frequency tripler using a separate transistor. Figure 6 shows a suitable crystal oscillator and tripler as well as its connection to the mixer IC. If the polarity is changed, a transistor type BF 224 can be used.

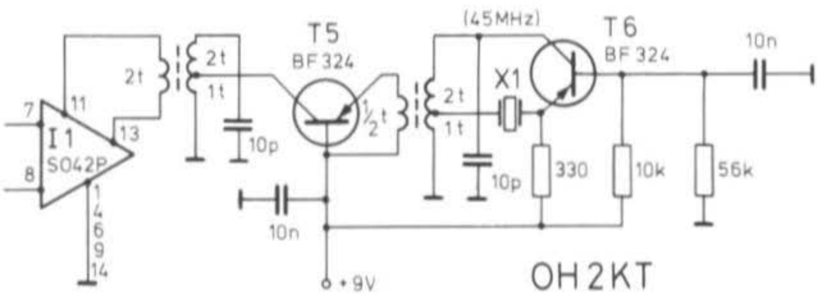

Fig.6: Crystal oscillator and coupling

The coupling of the crystal oscillator to the tripler should be made so that the collector current of the very steep transistor does not increase in excess of 2 to 3 mA. One coupling turn is already too much.

2. CONSTRUCTION

The construction of the receiver should be made on a double-coated PC-board. The other side is in the form of a continuous ground surface. All holes for components that are not grounded should be countersunk on the other side in order to avoid short circuits to ground.

2.1. SPECIAL COMPONENTS

I 1: SO 42 P (Siemens)
I 2: SO 41 P (Siemens)
I 3: TAA 991 D (Siemens)
I 4: TAA 611 B (Siemens)

T 1: AF 139 (various manufacturers)
T 2 - T 4: 2 N 3819, BF 245 (Texas Instruments)
T 5 - T 6: BF 324 (Siemens)

Crystal filter for 10.7 MHz for a channel spacing of 25 kHz, or narrower for AM (ITT, KVG)
Crystal filters for 9 MHz can also be used.
IF inductances: IF-filter 12 mm x 12 mm for required IF.

All RF-coils are wound on 5 mm dia coilformers with VHF core.

Coil data for Figure 3:

L 1: 2 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 2: 4 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 3: 3 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire, coil tap: 1 turn from cold end
L 3a: 2 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 4: 3 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire
L 5: 2 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 6: 18 turns of 0.3 mm dia. (29 AWG) enamelled copper wire, 3 μH
L 7: 10 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 8: 10 turns of 0.3 mm dia. (29 AWG) enamelled copper wire, 1 μH

Coil data for Figure 5:

L 1: 3 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire
L 2: 2 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 3: 3 turns of 0.8 mm dia. (20 AWG) silver-plated copper wire
L 4: 2 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 5: 18 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 6: 10 turns of 0.3 mm dia. (29 AWG) enamalled copper wire
L 7: 18 turns of 0.3 mm dia. (29 AWG) enamelled copper wire
L 8: 12 turns of 0.3 mm dia. (29 AWG) enamelled copper wire

3. MEASURING VALUES

FM receiver for 145 MHz with 10.7 MHz crystal filter, bandwidth 15 kHz:
Current drain without signal at 9 V: 16 mA
Current drain without signal at 12 V: 18 mA

All subsequent values measured at 9 V:
Sensitivity at f_{mod} = 1 kHz, deviation = ± 5 kHz for 12 dB SINAD:
1 μV (2 μV RMS)

SINAD-ratio at 50 μV (100 μV RMS) : 20 dB
SINAD-ratio at 500 μV (1 mV RMS) : 34 dB

Image rejection 20 dB
AM-suppression at 0.5 mV: 25 dB
Overload threshold: 10 mV
Limiter threshold (drop in AF-voltage by 3 dB ref. to 1 mV input voltage):
2 V (4 V RMS)

$$\text{SINAD} = \frac{\text{signal + noise + distortion}}{\text{noise + distortion}};$$

bandwidth of the measuring system: 20 kHz.

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A SIMPLE BANDPASS FILTER FOR THE 2 m BAND

by H.J.Brandt, DJ 1 ZB

This bandpass filter was a result of the requirements of the author and several other amateurs and was mainly developed to suppress the harmonics of transistor power amplifiers (1). The filter is easily able to handle output powers up to at least 50 W. The construction is extremely simple and the bandpass filter is easy to align. The suppression of harmonics and sub-harmonics of the 2 m band amounts to more than 30 dB. A further selectivity of approximately 20 dB can be expected from the PA-stages of the transmitter. The antenna itself will also provide a further attenuation of out-of-band signals.

The simple, single-chamber construction using two circuits has been criticized by a few experts. Of course, it would have been possible to obtain better results with more extensive circuitry, intermediate panels and more resonant circuits. However, most amateurs do not have the required measuring equipment to align them, and they are therefore of little use. This filter has been received with great enthusiasm in the Munich area of Germany and a large number have been built.

1. CONSTRUCTION

The filter is enclosed in a silver-plated box of 94 x 50 x 25 mm. Where such a box is not available, it can be obtained from the Publishers or their representatives. The coaxial sockets for input and output can be mounted on panel A or B. Since a large number of various types of plugs will be used when constructing this filter, only the centre of the socket is given. In the case of position A, BNC-sockets for single-hole mounting should not be used, since they protrude too far into the inner chamber of the filter.

The internal construction of the filter is given in Figure 2. Two line circuits made from 2 mm diameter silver-plated copper wire are tuned with the aid of air-spaced trimmers. They are spaced 15 mm from another and are mounted in anti-phase so that neither the hot or cold ends can couple. This spacing has been found to be most favourable using this type of construction, both with respect to the insertion loss and ripple.

The input and output coupling is made inductively and galvanically using lines of 1 mm diameter, silver-plated copper wire that are soldered to the line circuit at a point where an impedance of 50 Ω is present. Each of these coupling lines is formed into a small inductance (3 turns of 4 mm inner diameter, length approx. 7 mm) just before the coaxial socket. These inductances are not effective within the passband range of the filter, but represent "chokes" for higher frequencies in the blocking range. They are provided to ensure that the filter is relatively free of spurious resonances. The first deteriorations of the stop-band attenuation are to be observed above 1100 MHz.

The cold ends of the trimmers and resonant lines are grounded to the filter box with the aid of a silver-plated double solder tag as can be seen in Fig. 2. Conventional, solder-coated or tin-plated solder tags will cause a considerable increase of the insertion loss, since the RF-current is especially high at this point. The spacing of the resonant-line circuits from the base of the box is far smaller than from the cover. This means that the RF-currents are mainly concentrated to the bottom of the box. For this reason, it is permissible for the

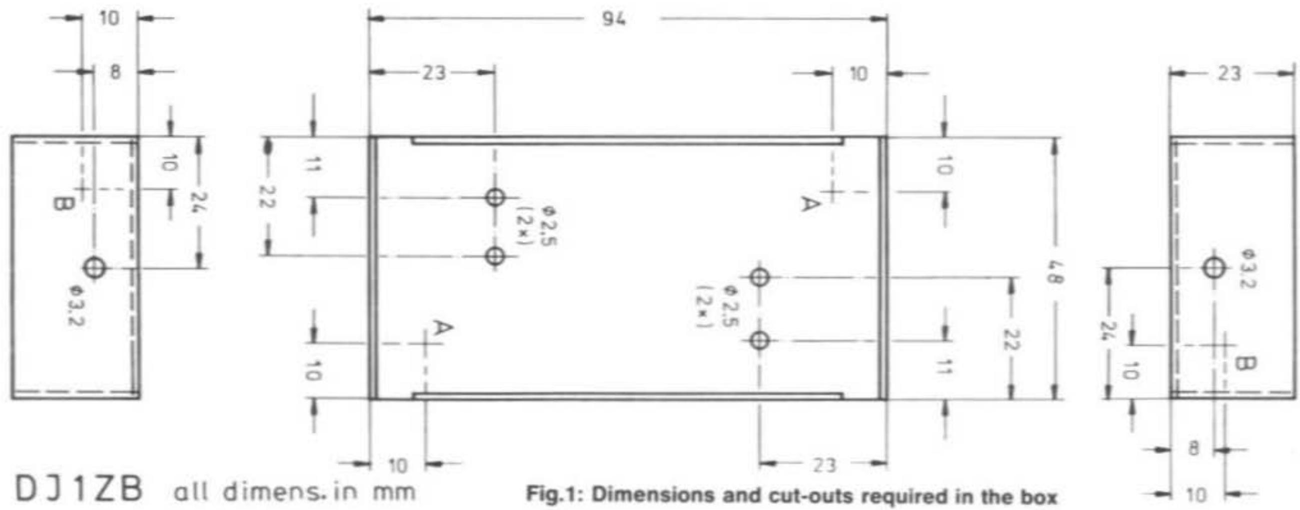

Fig.1: Dimensions and cut-outs required in the box

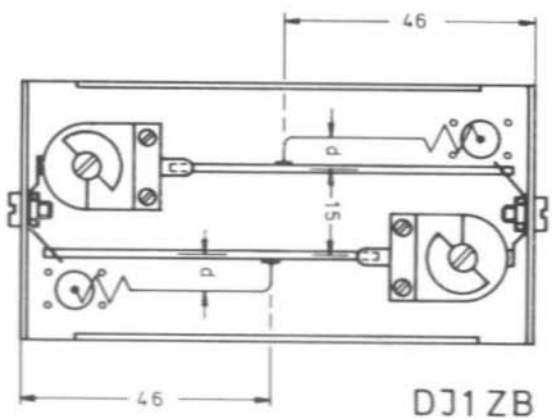

Fig.2: Internal construction of the filter

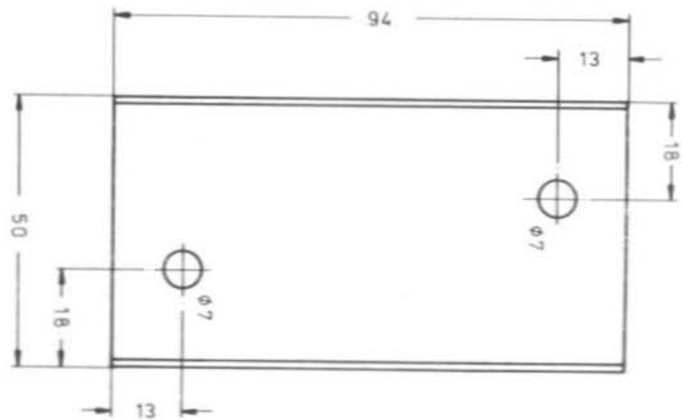

Fig.3: Alignment holes in the cover

cover to be only grounded with the aid of the four self-tapping screws provided. The position of the alignment holes in the cover are given in Figure 3.

1.1. SPECIAL COMPONENTS

One silver-plated box of 94 mm x 50 mm x 25 mm.
Two silver-plated double solder tags.
Two air-spaced trimmers of 3 - 45 pF for chassis mounting, complete with solder tags.
Two coaxial sockets (BNC or smaller)
2 mm dia. (12 AWG) and 1 mm dia. (18 AWG) silver-plated copper wire.

2. ALIGNMENT

The alignment can be made in a simplified, or more complex manner according to which measuring aids are available to the individual amateur. The transmitter to be used for the alignment should be insensitive to mis-match. If this is not the case, it should be protected by using low drive, reducing the operating voltage, or inserting an attenuator into the signal path directly after the transmitter (at x). If possible, a dummy-load of the correct impedance should be used for the termination Z. Of course, it is also possible for your own antenna to be used if nothing else is available. However, the standing-wave ratio of the antenna will tend to degrade the measurement. It goes without saying that this should only be done on inactive frequencies and that monitoring pauses should be made during transmissions after repeating the call-sign at regular intervals.

2.1. ALIGNMENT FOR THE MOST FAVOURABLE PASSBAND BEHAVIOUR

The measuring arrangement given in Figure 4 is used. The transmitter is firstly set up near the centre of the band, and the SWR-meter switched to forward power. Both circuits of the filter are now aligned for maximum forward power. The alignment frequencies are now tuned to the upper and lower band limits and the filter circuits tuned until a suitable bandpass characteristic is obtained over the whole band.

This alignment is not the most favourable, but the required filtering will be provided. The input impedance of the filter can exhibit a considerable deviation from the required impedance, and it may be necessary for the output stage of the transmitter (without attenuator at position x) to be corrected for maximum forward power.

2.2. ALIGNMENT FOR MINIMUM REFLECTED POWER

The filter is firstly aligned as was explained in Section 2.1. to obtain the most favourable bandpass characteristics. The SWR-meter is now inserted into the signal path in front of the filter (Fig. 5) and switched to reflected power indication. The filter is then aligned for minimum reflected power, after which the filter is turned around (reverse the input and output) and the alignment repeated. If the results are not satisfactory, a more favourable minimum of the reflected power should be tried by varying the spacing d (usually 6 mm, see Fig. 2), or by shifting the coupling taps on the resonant lines. In order to ensure that no unwanted transformation takes place, the filter is always reversed and the measurement repeated. An adjustment will always be found where the reflected power is very low over the whole 2m-band and does not increase until the band limits are exceeded.

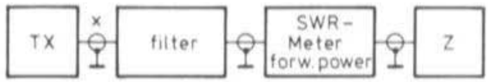

Fig.4: Measuring set-up for best passband behaviour

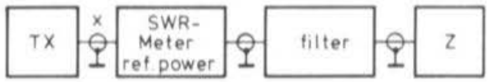

Fig.5: Measuring set-up for alignment of minimum reflected power

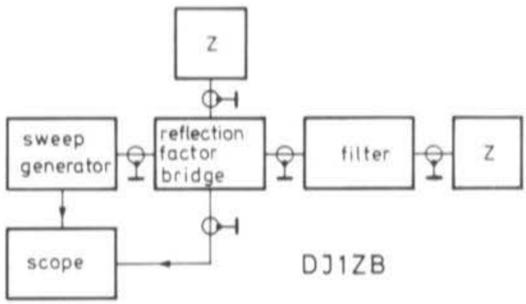

Fig.6: Measuring set-up for alignment of minimum reflected power using a swipped-frequency measuring system and reflection-factor bridge

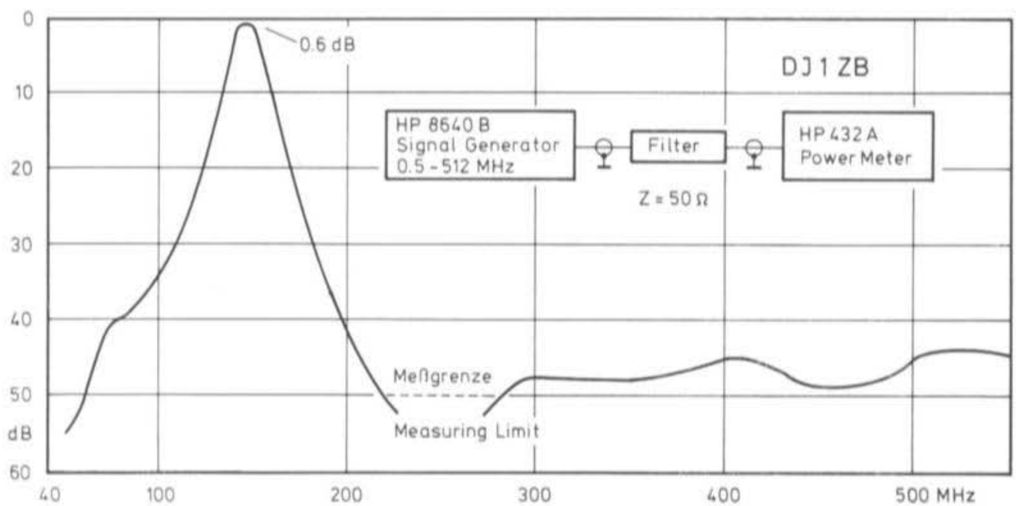

Fig.7: Filter characteristics

If a swept-frequency measuring system (Polyskop) and a reflection-factor measuring bridge (e.g. Telonic RHO-Tector Model TRB-50) can be used, the alignment is, of course, far easier and quicker (Fig. 6). The selectivity curve of a filter aligned on such a measuring system can be seen in Figure 7. The insertion loss of 0.6 dB is the only power-limiting factor of this filter since this loss will be converted into heat. It has been found, that an even lower insertion loss can be obtained if the filter is aligned for a somewhat larger bandwidth.

3. APPLICATION

As has been already mentioned, the main application of this filter is to suppress harmonics and sub-harmonics from 2 m-transmitters. Several commercially available transmitters possess a good harmonic filter, whereas the suppression of the local oscillator and other lower frequencies is sometimes relatively poor. If this filter is connected into the antenna feeder, it will also provide an additional suppression of out-of-band signals during reception which can be of great advantage to those amateurs living in the vicinity of strong transmitters.

3.1. ALIGNMENT OF THE POWER OUTPUT STAGES OF TRANSMITTERS

As could be noticed in the case of a 2 N 3375 power amplifier, it was difficult to establish whether the output stage was tuned to the fundamental or first harmonic. In such cases, it is advisable for such a bandpass filter for the required band to be connected between transmitter and dummy-load so that the alignment of the output stage is guaranteed to be at the required frequency.

It should also be mentioned in the same context that if an insertion loss of more than 0.6 dB is determined when inserting a correctly-aligned filter into the signal path, this is probably (with the exception of slight differences in the matching) due to the fact that the harmonic content falsified the original measurement. The output stage should finally be aligned for maximum forward power in a measuring arrangement as shown in Figure 4.

3.2. EXPERIMENTS WITH TEST CIRCUITS

In the VHF/UHF range, preamplifier stages, multiplier or output stages are often tested on a dummy load. Such a dummy load provides, however, a termination for all frequencies, even for those which would cause it to oscillate. For this reason, it is more favourable for a bandpass filter to be connected in front of the dummy load so that the termination is only made for the required frequency (possibly more so than when using the actual antenna), so that any tendency to oscillation can be seen immediately and avoided at an early stage of development.

4. NOTES

The versatility of such a filter for experimental and normal operation have now led the author to develop a similar filter in the same case for the 70 cm-band. The first filter prototypes have shown that smaller trimmers, line circuits with a lower impedance and greater spacings between the filters are required. Exact measurements and dimensions are to be given in a later article.

Editors: Printed, bandpass filters for 2 m (3) and 70 cm (4) have already been described in this magazine.

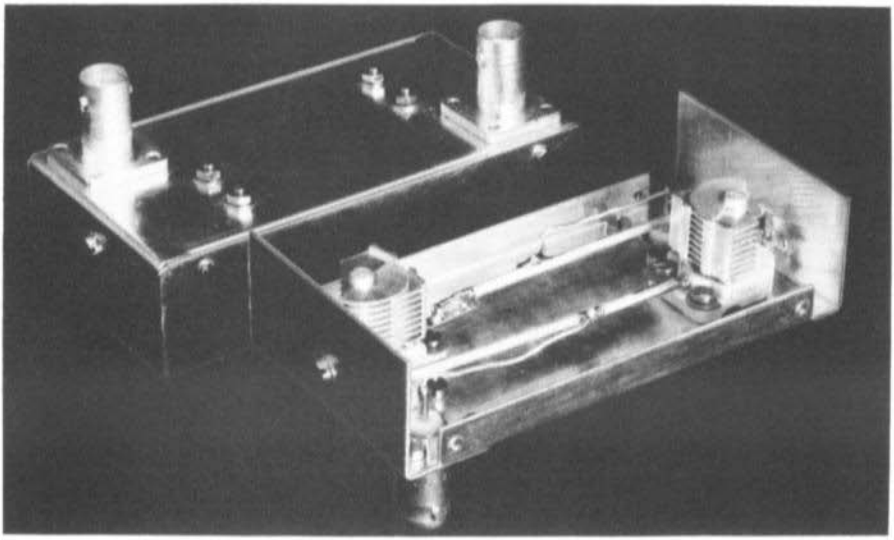

Fig.8: A prototype filter

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A NUMERICAL INDICATION SYSTEM

by K.Wilk, DC 6 YF

An indicator system is to be described that allows large numerals to be indicated on a television screen, indicating, for instance, a measuring result. The readout has been limited to four digits since this is sufficient for indication of:

Measured values such as voltage, current, resistance
Time (hours and minutes)
Counter results of frequency and occurancies

A larger number of digits is seldom necessary due to the limits of the attainable accuracy of the measuring system. In addition to this, three or four digits represent a display with the most favourable dimensions for indication on the screen of a TV-tube.

The video system generates a video signal that can be fed directly to a TV-monitor or modified TV-receiver (video circuit) (1). A decimal point can be given to the right of each digit.

1. THE SYSTEM

The numerical indicating system comprises:

1. Pulse generator	(DC 6 YF 008)
2. Numerical generator	(DC 6 YF 009)
3. Multiplexer	(DC 6 YF 010)
4. Video signal synthesizer	(DC 6 YF 008)

The four circuits given above are accommodated on three separate PC-boards having the designations given in brackets. The dimensions of the boards conform to the European standard of 100 mm by 160 mm. They are designed as plug-in modules using 31 pin connectors. The complete system operates from a 5 V supply.

2. PRINCIPLE OF OPERATION

The operation of the circuits is to be described in conjunction with the block diagram given in Figure 1: The numerical signs, and the decimal points are stored and read by feeding appropriate signals to the data inputs. Five lines are required for each data input: BCD-coding for the digit and on/off coding for the decimal point. This means that 20 data lines are required for the four digits. The digits 0 to 9 are formed from seven segments of vertical and horizontal bars.

2.1. THE PULSE GENERATOR

The pulse generator represents the clock of the system and provides the following pulses:

1. Clock pulses (1 MHz)
2. Line synchronizing pulses
3. Vertical synchronizing pulses

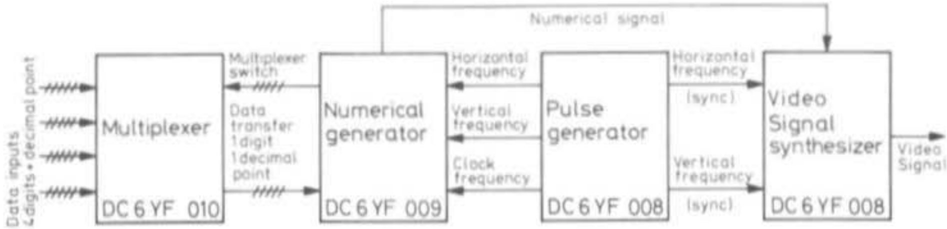

Fig.1: Block diagram of a numerical indicator system

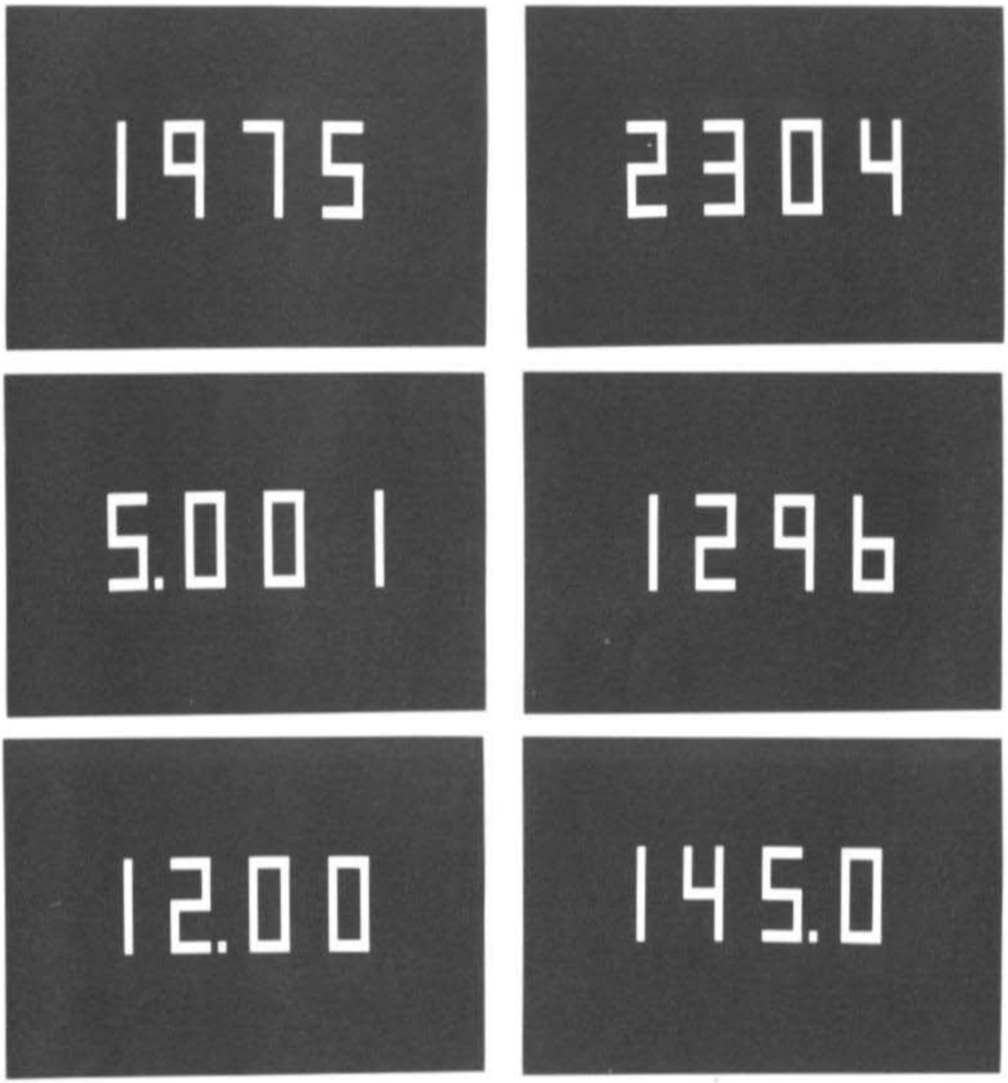

Fig.2: Types of digits that can be indicated

2.2. THE NUMERICAL GENERATOR

This module is fed with clock pulses that are an even multiple of the line frequency (1 MHz). When displayed on a TV-screen, the clock pulses would result in 50 vertical bars. The horizontal pulses are then divided by ten so that 20 horizontal bars result. It is possible in this manner to divide the screen into a virtually square grid, if the positive slope of the bars is used as limit. The square fields are now combined to form segments of the digit. The decimal point only comprises a single field. Each of the segments is driven individually and can be displayed or suppressed. An indication of 8. 8. 8. 8. results when all segments are energized. The individual segments are driven by a BCD seven segment decoder. The required digit must be fed to the decoder in BCD-code until the electron beam is in that portion of the TV-screen.

2.3. MULTIPLEXER

Since the numerical generator can only process one digit at a time the multiplexer is required to ensure that the required digits are fed to the data lines at the correct time.

The multiplexer is fed from the numerical generator.

2.4. VIDEO SIGNAL SYNTHESIZER

A video signal is formed in this module from the vertical and horizontal synchronizing pulses of the pulse generator and the signal from the numerical generator. The video signal with a peak amplitude of 1.2 V into 75 Ω (negative going synchronizing pulses) is fed via a 75 Ω cable to the video monitor.

3. NOTES

The extent of circuitry is the same to indicate either dark digits on a white field or white figures on a dark field. In practice, however, it has been found that white digits are more easily readable. For this reason, this mode is used in the practical circuit.

Full construction details are to be given in one of the next editions of VHF-COMMUNICATIONS.

3. REFERENCES

(1) K. Wilk: A Domestic TV-Receiver as Video Monitor
VHF COMMUNICATIONS 6, Edition 3/1974, Pages 186-190.

INDEX TO VOLUME 7 (1975) OF VHF COMMUNICATIONS

PRICE LIST OF MATERIAL
described in Edition 4/1975 of VHF COMMUNICATIONS

DJ 6 ZZ 005	**TRANSMIT CONVERTER FOR 144 MHz**		**Ed.4/1975**
PC-board	DJ 6 ZZ 005	(double-coated, without through-contacts)	DM 15.—
Semiconductors	DJ 6 ZZ 005	(1 IC, 5 transistors)	DM 82.—
Minikit 1	DJ 6 ZZ 005	(9 trimmer capacitors, 3 feedthrough capacitors)	DM 26.—
Minikit 2	DJ 6 ZZ 005	(1 Schottky ring mixer, 1 coilformer with core, 1 core, 1 ferrite choke, 20 ceramic caps., 2 tantalum caps.)	DM 52.—
Kit	**DJ 6 ZZ 005**	**with the above parts**	**DM 175.—**
Matching receive converter		DK 1 OF 016/017	DM 108.—
DL 3 WR 009	**POWER SUPPLY FOR VHF-FM RECEIVER**		**Ed.4/1975**
PC-board	DL 3 WR 009	(with printed plan)	DM 18.—
Semiconductors	DL 3 WR 009	(1 bridge rectifier, 6 diodes, 2 integrated stabilizers)	DM 58.—
Minikit	DL 3 WR 009	(9 electrolytics, 4 ceramic capacitors, 4 resistors, 1 fuse holder, 1 heatsink)	DM 26.—
Kit	**DL 3 WR 009**	**with the above parts**	**DM 98.—**
Set of metal screening panels for DK 1 OF 020-022			DM 15.—
HB 9 MIN 001/2	**FOUR-DIGIT FREQUENCY COUNTER**		**Ed.4/1975**
PC-board	HB 9 MIN 001	(double-coated, through contacts)	DM 16.—
PC-board	HB 9 MIN 002	(with printed plan)	DM 7.—
Semiconductors	HB 9 MIN 001	(9 ICs, 1 transistor, 1 LED)	DM 54.—
LED-readouts	HB 9 MIN 002	(4 hp 5082-7300)	DM 224.—
Minikit	HB 9 MIN	(1 choke, 2 electrolytics, 5 ceramic capacitors)	DM 12.—
Kit	**HB9MIN001/2**	**with the above parts**	**DM 313.—**
DJ 1 ZB	**BANDPASS FILTER**		**Ed.4/1975**
Kit	DJ 1 ZB	(1 silver-plated case, 2 silver-plated lugs, 2 trimmer capacitors)	DM 42.—

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CRYSTALS and CRYSTAL FILTERS
for equipment described in VHF COMMUNICATIONS

CRYSTALS and CRYSTAL FILTERS

Crystal filter	XF-9A	(for SSB) with both sideband crystals	DM 110.—
Crystal filter	XF-9B	(for SSB) with both sideband crystals	DM 148.—
Crystal filter	XF-9C	(for AM; 3.75 kHz)	DM 150.—
Crystal filter	XF-9D	(for AM; 5.00 kHz)	DM 150.—
Crystal filter	XF-9E	(for FM; 12.00 kHz)	DM 150.—
Crystal filter	XF-9M	(for CW; 0.50 kHz) with carrier crystal	DM 110.—
Crystal filter	XF-9NB	(CW filter 8 pole; 400 Hz; with carrier crystal)	DM 172.—
Crystal filter	QF-9FO	(for FM; 15 kHz)	DM 165.—
Sideband crystal	XF-901	(8.9985 MHz)	DM 15.—
Sideband crystal	XF-902	(9.0015 MHz)	DM 15.—
Carrier crystal	XF-900	(9.0000 MHz)	DM 15.—
Crystal	96.0000 MHz	(HC- 6/U) for 70 cm converters	DM 26.—
Crystal	96.0000 MHz	(HC-25/U) for 70 cm converters	DM 34.—
Crystal	95.8333 MHz	(HC-25/U) for 70 cm converters	DM 34.—
Crystal	78.8580 MHz	(HC- 6/U) for ATV TX (DJ 4 LB)	DM 26.—
Crystal	67.3333 MHz	(HC- 6/U) for 70 cm / 10 m converters	DM 22.—
Crystal	66.5000 MHz	(HC- 6/U) for synthesis VFO (DJ 5 HD)	DM 22.—
Crystal	65.7500 MHz	(HC- 6/U)	DM 22.—
Crystal	65.5000 MHz	(HC- 6/U)	DM 22.—
Crystal	65.2500 MHz	(HC- 6/U)	DM 22.—
Crystal	65.0000 MHz	(HC- 6/U)	DM 22.—
Crystal	64.3333 MHz	(HC- 6/U) for ATV converter (DJ 5 XA)	DM 22.—
Crystal	62.0000 MHz	(HC- 6/U) for synthesis VFO (DJ 5 HD)	DM 22.—
Crystal	57.6000 MHz	(HC-25/U)	DM 33.50
Crystal	57.6000 MHz	(HC- 6/U)	DM 22.—
Crystal	38.9000 MHz	(HC- 6/U) for DJ 4 LB 001 ATV-TX	DM 25.—
Crystal	38.6667 MHz	(HC- 6/U) for 2-m-converters	DM 17.—
Crystal	1.4400 MHz	(HC- 6/U) for synthesizer DK 1 OF 012	DM 22.50

STANDARD FREQUENCY CRYSTALS

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Crystal	1.0000 MHz	(XS 0605) for 75° ovens	DM 50.—
Crystal oven	XT-2 (12 V)	75°C	DM 82.—
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Crystal socket	for HC-25/U	horizontal mounting	DM 5.—
Crystal socket	for HC-25/U	vertical mounting	DM 1.50
Crystals	72...... MHz	(HC-25/U)	DM 23.—

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Ceramic filter	CFS-455 D	for FM IF-strip DC 6 HL 007	DM 70.—

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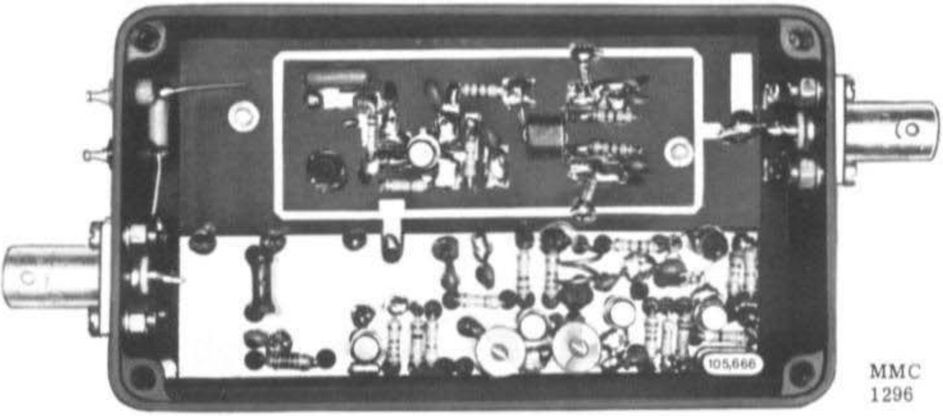

MMC
1296

1296 MHz CONVERTER Microstripline, Schottky diode mixer
IF: 28-30 MHz or 144-146 MHz / Noise figure: typ. 8.5 dB / Overall gain 25 dB

432 MHz CONVERTER 2 silicon preamplifier stages.
MOSFET mixer. All UHF circuits in microstrip technology.
Noise figure: typ. 3.8 dB / Overall gain: typ. 30 dB
IF: 28-30 MHz or 144-146 MHz 9-15 V / 30 mA

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Overall gain: typ. 30 dB / IF: 28-30 MHz, others on request.
9-15 V / 20 mA

VARACTOR TRIPLER 144/432 MHz Max.input at 144 MHz: 20 W
(FM,CW) - 10 W (AM). Max. output at 432 MHz: 14 W

VARACTOR TRIPLER 432/1296 MHz Max.input at 432 MHz: 24 W
(FM, CW) - 12 W (AM) / Max.output at 1296 MHz: 14 W

ATV CONVERTER Input frequency range: 430-440 MHz
IF range: CCIR Channel 3 (48 MHz) / Noise figure: typ. 3.8 dB
Bandwidth: 10 MHz / Gain: typ. 25 dB / 9-15 V / 30 mA

All modules are enclosed in black cast-aluminium cases of 13 cm by 6 cm by 3 cm and are fitted with BNC connectors. Input and output impedance is 50 Ohms. Completely professional technology, manufacture, and alignment. Extremely suitable for operation via OSCAR 7 or for normal VHF/-UHF communications.

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Type TA   Type TA-S   Type TA 4

Type U 3   Type U 4   Type U 5

Model	Type	Frequency	Gain	Weight	Features
TA	5/8 λ	144-175 MHz	3 dB	275 g	Glass-fibre whip
TA-S	5/8 λ	144-175 MHz	3 dB	275 g	Glass-fibre with 5 m cable
TA 4	1/4 λ	144-175 MHz	0 dB	130 g	Stainless steel (PH 17-7)
U 3	5/8 λ	400-470 MHz	3 dB	100 g	Silver-plated, epoxy coated
U 4	Colinear	420-470 MHz	4 dB	150 g	Stacked $\lambda/4$ and 5/8 λ
U 5	Colinear	420-470 MHz	5 dB	175 g	Stacked 5/8 λ and 5/8 λ

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