

*A Publication
for the Radio-Amateur
Especially Covering VHF,
UHF and Microwaves*

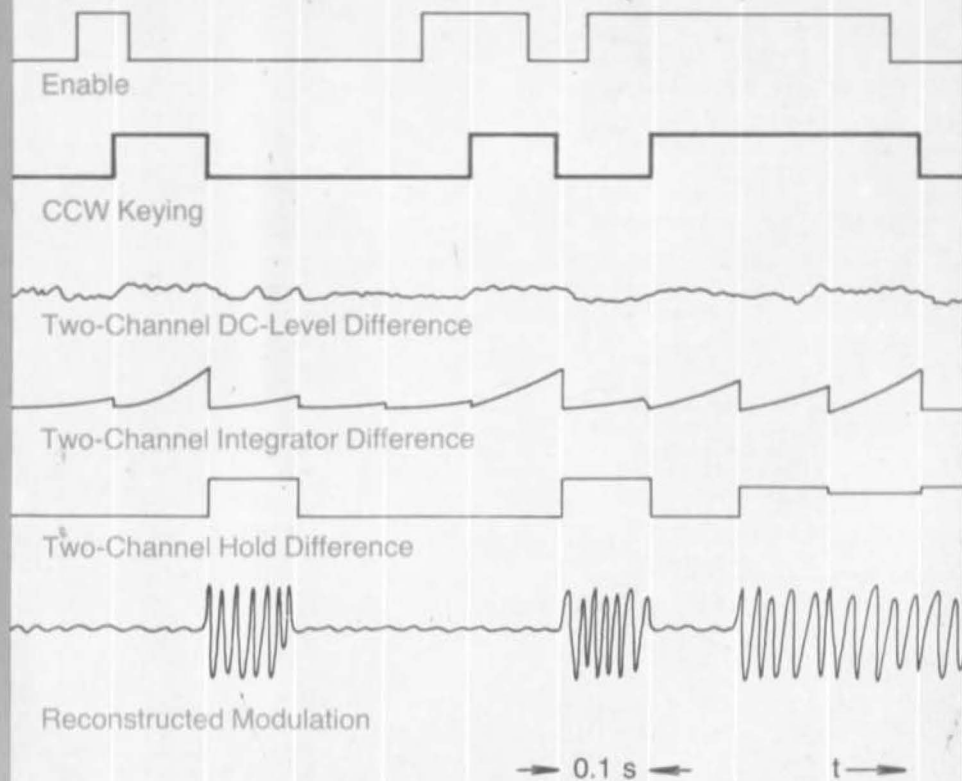


VHF

communications

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The Elements of Coherent CW Transmission (CCW)





VHF communications

A Publication for the Radio Amateur
Especially Covering VHF, UHF, and Microwaves

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In the Focus

Coherent CW transmissions (CCW) has been used by a dedicated group of amateurs in the United States for some years now. Especially Ray Petit, W 7 GHM, gave so many basic ideas that it is impossible to discuss CCW without mentioning him.

A two-part article by Charles Woodson, W 6 NEY, gives both, an excellent explanation of the principles and the hardware required for starting in this mode. This is the reason why I decided to introduce this new technique in VHF COMMUNICATIONS with a reprint of his article. The reactions to the German-language article (which appeared earlier this year) were enthusiastic – many amateurs have already started working on this new technique, and some are developing new hardware. This will enable us to publish further articles on Coherent Transmission Techniques, which are capable of saving both transmitter power and bandwidth – simultaneously enhancing the telecommunication quality – this opens fantastic possibilities!

In my opinion however, the breakthrough of this technique will come with coherent RTTY (in the PSK mode?) rather than CCW, although the possible reduction of bandwidth is not that spectacular, but CRTTY can easily be matched with computer techniques!

Finally I would like to make you the same offer as I did to the readers of my German magazine: If you are interested in Coherent Transmission Techniques (CTT), just send me a postcard – I will then distribute a list of all readers interested.

With kind 73s
your's
DL 3 WR



Ulf-Dietmar Ernst, DK 9 KR

To Be Discussed: Coherent Communications Technology

Coherent communications is a very interesting form of communications where the transmitter and receiver are phase and frequency locked to another using optimized filters. This technology has been used for some time with advantage for space, satellite, and military communications. It is somewhat surprising that this mode has not had any effect on amateur radio communications until now.

The author wishes to mention a few fundamentals of coherent communications to readers of VHF COMMUNICATIONS, which he hopes will create an interest in experimenting with this mode.

Since there are various methods of coherent communications, it is felt that these should be discussed, so that one can determine the best method for amateur applications.

The values given in different articles for the improvement of the signal-to-noise ratio of 10 to 20 dB are a little misleading at first. However, these improvements can really be obtained, and comprise two different components. The main improvement of the signal-to-noise ratio

is obtained by correctly selecting the lowest bandwidth possible for this transmission channel.

The required bandwidth for Morse communications is only a few Hz, whereas one order of magnitude more is required for radio teletype transmissions. However, the advantage of these low bandwidths can, at present, not be utilized to the full without locking frequency and phase between transmitter and receiver, and without solving the main problem of low bandwidths: Construction of a narrow-band filter that does not ring! Coherent methods using optimized computer-controlled, narrow-band filters only allow us to utilize the required bandwidth for the transmission of the signal.

The use of the coherent transmission technology on its own only improves the signal-to-noise ratio by a few dB (in the order of 3 to 6 dB). However, there are considerable differences, which means that it is important to select the most favourable coherent transmission method. We differentiate between:

Orthogonal and
Antipodal signals.



	Gain of Signal-to-noise ratio		Error-Probability Rate per 1000 signs
Non-coherent, orthogonal	0 dB	0 dB	100
Coherent orthogonal	2 dB	3 dB	30
Differential coherent antipodal	4 dB	3 dB	10
Coherent antipodal	5 dB	6 dB	3
Error per 1000	10	100	

Table 1

In the case of a digital, binary signal such as RTTY, this means that:

In the orthogonal mode, »mark« and »space« are transmitted on two different frequencies, which must not appear at the same time. This is the principle of the orthogonal mode, which is guaranteed in the RTTY-mode due to the code used.

With antipodal RTTY-signals, on the other hand, »mark« and »space« signals are at the same frequency but exhibit a phase difference of 180°.

The time-dependent »mark« and »space« only differ in their sign [$\sin x = -\sin(x + 180^\circ)$] and are thus opposite. They are therefore designated antipodal.

Instead of transmitting the actual »mark« and »space« signals, it would be possible to only transmit the difference, in other words the switchover between these two states. Assuming that the last state was »space«, the sequence »mark-mark-space-mark« would be transmitted as »mark-space-space-mark«. This method will increase the reliability of the transmission.

The **table** gives gain and error-rate values for the various modes, after ensuring that all measures have been carried out to improve the si-

gnal-to-noise ratio by limiting the bandwidth (1). It will be seen that a coherent antipodal transmission offers the highest gain over a conventional transmission.

Also we can see that the advantages of the coherent principle are greater when the signal path is poor (high error rate). Since the transmission paths are usually poor for amateur applications and a certain readability should still be present at the limits, this clearly shows the advantage of coherence.

This means that a coherent antipodal transmission technology provides the best method of improving the reliability of amateur radio transmissions. The phase transmission mode P 1 is required for binary transmissions (RTTY). It is necessary for the receiver to know the exact phase and frequency of the transmitter so that its computer-controlled filters can filter out the transmitted signals and reconstruct them.

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McGraw-Hill, New York, 1966



Charles Woodson, W 6 NEY

Coherent Telegraphy Transmissions

Part 1: The Concept

This two-part article is published with the kind permission of the ARRL and is based on an article in the May and June 1981 editions of QST.

Would you think it is possible for you to decrease your output power by a factor of ten and increase readability by the same amount — simultaneously ? It's really possible.

The more we know about something we seek, the easier it is to find. This principle applied to Morse CW communications is called coherent CW or CCW. On-the-air trials of this technique have shown it will provide an improvement of more than 20 dB in communications effectiveness over ordinary CW methods. This same principle can be used with RTTY, ASCII and FSK signals, but this discussion will focus on CW keying.

CW signals may be analyzed as a series of digital units, all of which have (at least approximately) a unit of time in common. For convenience, I'll call this time unit a »frame«. Each frame contains either a »mark« (key down) or a »space« (key up). **Figure 1** illustrates this concept.

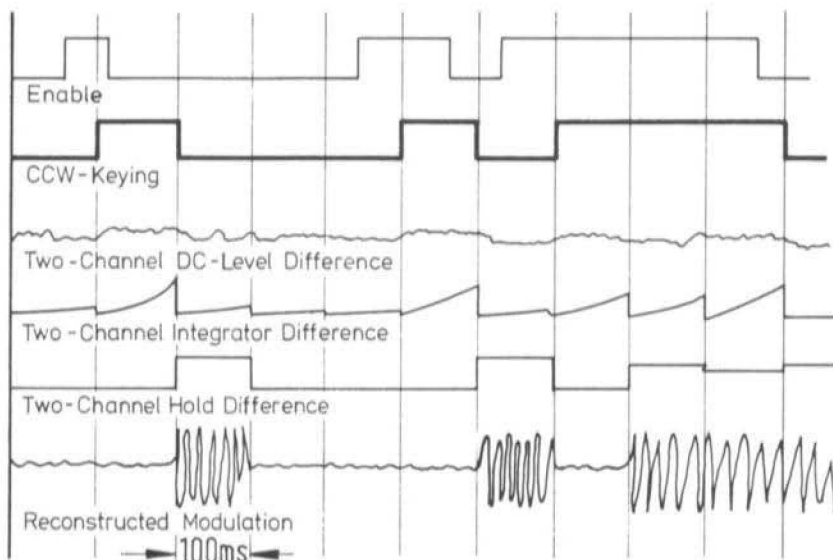
Ordinary CW dots, dashes and spaces begin at somewhat arbitrary times, depending on when the operator happens to press the key. Thus, the frame length varies to a considerable degree, and you can't predict when each frame starts and ends. With CCW, all dots,

dashes and spaces are exact multiples of the basic time unit and occur within predictable time frames. This includes any pauses during transmission. When received, CCW signals sound like any other CW signal except that they are being sent very precisely, as with a perfect »fist«. This characteristic is utilized to permit the use of very narrow bandwidth filters.

CW FILTERS

In general, receiver filters with bandwidths much wider than that of the desired signal are less effective because they allow reception of additional noise and undesired signals. At 12 wpm a CW signal occupies about 10 Hz of the spectrum, yet 500- or 2300-Hz-wide filters are frequently used for CW reception. With a 500-Hz filter, one hears the 10-Hz-wide desired signal and 490 Hz of noise and QRM ! By analogy, an SSB operator using a similar approach would listen to 100 kHz of the band at one time !

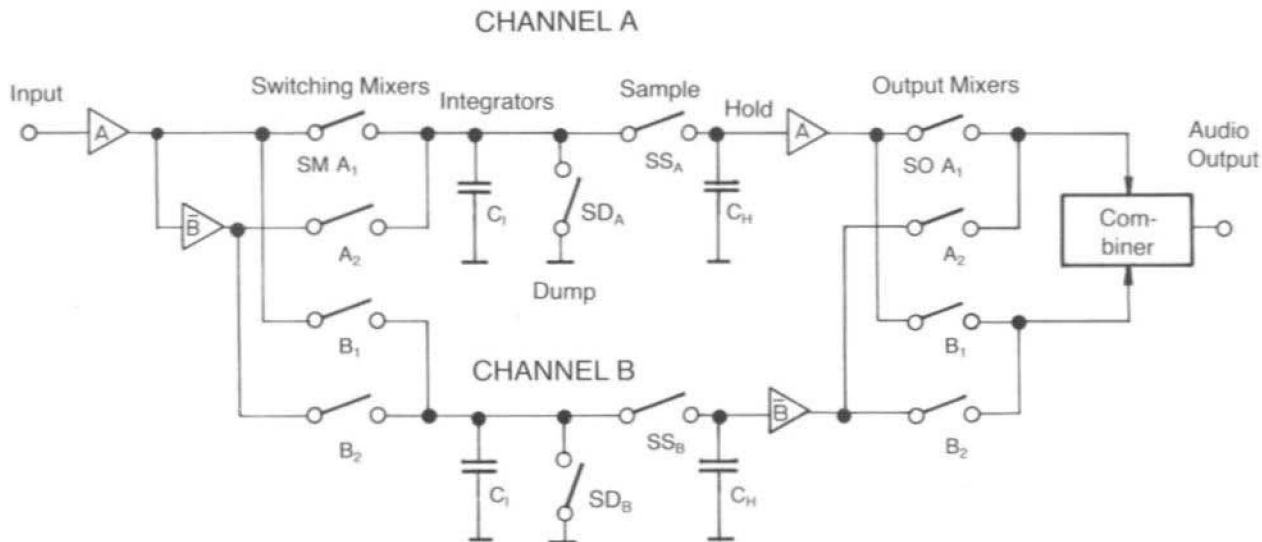
High-Q analog CW filters are not useful at the narrow bandwidths approaching the bandwidth of a 12-wpm CW signal. Such filters, with bandwidths less than 500 Hz, tend to »ring« or produce an output after the signal ceases. The human ear is confused by such ringing. Also, the receiver stability and resetability required in conjunction with the filter, on the order of a few hertz, is difficult to achieve.



The elements of CCW communication. Frames, in 0.1-second units, are shown on the horizontal axis. The enable (top waveform) shows the closure of a manual key by the operator. When referenced to the precise frame times, it can be seen that the dots, dashes and spaces of the enable are not accurate in length. Note that with the CCW-keyer waveform a mark or space is begun only at the beginning of the frame period and continues for the full period(s).

As received, the signal is mixed with QRM and QRN. The difference between the DC voltages from the switching mixers of the two channels (third waveform) is a function of the desired, but weak signal. An integrator sums the power (voltage) received over the frame period. This sum is sampled at the end of the period and held until the beginning of the next period. The recovered modulation is used to key an audio signal for detection by ear.

Fig. 1: Elements of a CCW-transmission



W 6 NEY

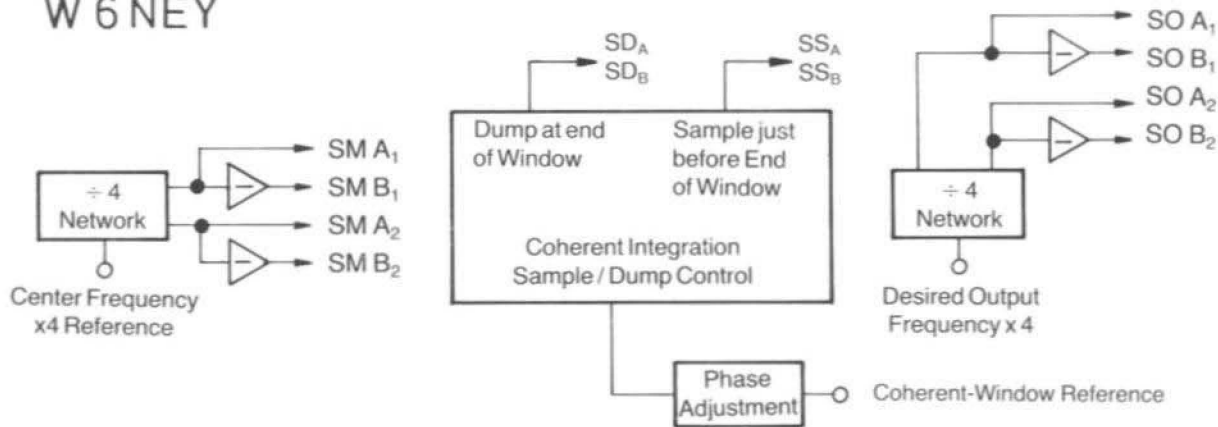


Fig. 2: Block diagram of a CCW-Filter



Phase-locked loop (PLL) filters with time constants long enough to produce bandwidths of only a few hertz unfortunately take tens of seconds to achieve lock. PLLs also tend to lock on the strongest signal in the passband and are, therefore, sensitive to QRM. PLL filters have their place of importance, but not with the bandwidths required here.

The filter we need will provide a bandwidth of only a few hertz without ringing and without a tendency to lock on the QRM. Such a filter improves the signal-to-noise ratio dramatically. A 1-W signal copied through a 10-Hz bandwidth filter is comparable to a 50-W signal heard through a 500-Hz filter or a 230-W signal heard through a 2300-Hz filter.

THE CCW STATION

Typically, CCW stations agree on an operating frequency (e.g., 14,049,000 Hz \pm 2 Hz) and a frame length (usually 0.1 second, the speed of 12 wpm), and acquire the «framing» – when each frame starts and ends – as part of the signal-tuning process. Thus, the frequency, frame length and frame phase are all known at the receiving end and are used to advantage in the detection process.

To achieve the necessary frame-length accuracy and to get on the operating frequency within the narrow tolerance of the filter, all frequency-determining oscillators in both the transmitter and receiver of the CCW station must be highly stable and accurate. The stability and accuracy requirements are obtainable by using carefully built crystal oscillators which are compared to a reference frequency such as WWV. Time discipline for the transmitted signal is determined by a reference oscillator which is divided to provide a 10-Hz synchronizing signal for the transmitter keyer. The CCW filter at the receiving station uses timing signals derived from the station reference oscillator. These timing signals tell the receiver filter when to expect a frame to begin and end.

THE COHERENT INTEGRATING FILTER

Figure 2 shows a block diagram of the filter which makes possible the efficient reception of a CCW signal. The major blocks of each of the two filter chains are: input mixers, integrators, sample-and-hold circuits, output mixers, and the timing and control circuitry. The reason for the two chains will be examined later; for now, we'll follow the signal through one chain.

The Mixer: The first part of each filter chain is a switching mixer where the desired signal (along with adjacent QRN and QRM) is mixed with a reference signal of the same frequency as the desired signal. (Solid-state switching is performed in the actual circuit, but for simplicity, mechanical contacts are shown in Fig.2). The reference signal is obtained from a stable source such as the timing and control circuitry, and it determines the center point of the CCW filter. A signal at the desired frequency comes out of the mixer as a DC voltage – the stronger the signal, the larger the voltage. An off-frequency signal, however, comes out of the mixer as a low-frequency AC voltage: We mix the incoming signal right down to zero beat. Undesired signals will be distinguished from the desired signal because they are not exactly zero beat.

The Integrator: An op-amp integrator comprises the second part of each filter chain. We use the integrator to distinguish the desired signal (the zero-beat DC voltage) from the undesired signals (low-frequency AC voltages) coming from the mixer. The integrator may be thought of as a moderately large capacitor. A synchronizing «dump» signal from the timing and control circuitry shorts out this capacitor at the start of each time frame. Any desired signal (DC voltage) during the time frame causes the capacitor to charge. The resulting voltage at the end of the time frame is a function of the strength of the desired signal received during that frame.

QRM and QRN, being off frequency, appear as AC signals to the integrator capacitor.



These charge the capacitor for part of the time frame, but discharge it for other parts of the same period. Consequently, signals off frequency do not have as great an effect on the integrator output as do signals exactly on the desired frequency. That is how the CCW filter achieves its selectivity.

As an example, consider an interfering carrier which is 10 Hz above or below the desired signal. Following the switching mixer, this QRM appears as a 10-Hz AC voltage. If the filter is set to the CCW standard frame length of 0.1 second, then the 10-Hz interfering signal goes through one complete cycle during the integrating period. At the end of the time frame, the QRM-produced voltage at the integrator output is zero. Thus, the CCW filter has a null just 10 Hz above and below its center frequency. There are also similar nulls at other 10-Hz multiples.

Sample-and-Hold and Integrator Reset:

At the end of each time frame, a «sample» signal from the timing and control circuit transfers the voltage at the integrator output to the sample-and-hold circuit. That circuit «remembers» that voltage for the following interval. Once the sample-and-hold has acquired the integrator output voltage, a dump signal from the timing and control circuitry shorts out the integrator capacitor. It does this by means of a CMOS analog switch connected across the capacitor. This allows the integrator to start over again with zero voltage at the start of the next time frame.

Resetting the integrator at the end of each time frame lets the CCW filter avoid the ringing (or intersymbol interference) common to other narrow-bandwidth filters. Note that this is possible only because the CCW filter «knows» when each time frame begins and ends. It is here that the time discipline of the transmitted signal is used to advantage in the detection process.

Output Mixer: This last block of the filter chain is much like the input mixer: it functions as an amplitude modulator, using the sample-and-hold output voltage to control the amplitude of a sidetone. The purpose of this mixer is to con-

struct a sidetone for the human operator to hear.

WHY TWO CHANNELS ?

If the incoming signal is in phase with the center reference, then the mixer output is always positive. The integrator which follows will see a positive DC voltage. If the signal is out of phase with the reference, then the mixer output is always negative. The integrator will see a negative DC voltage. The positive or negative DC voltage charges the integrator capacitor, the sample-and-hold «remembers» that charge during the next time frame, and the output mixer generates a sidetone whose amplitude is proportional to the voltage on the sample-and-hold capacitor. But if the signal is 90° out of phase with the reference frames, then the mixer output will at times be positive and at other times be negative during a given input cycle. This output will be averaged to zero by the integrator. The result is no filter output from this channel.

The situation is different for each channel because the A channel input mixer is operated by a reference which is 90° out of phase with the B channel reference. Thus, if a signal is 90° out of phase with the A channel, it will be in phase (or 180° out of phase) with the B channel. At all phase differences between the two channels, the product of the two channels is always the desired signal despite the phase relationship between the center frequency reference and the incoming signal.

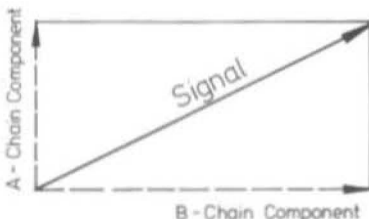


Fig. 3:
Required signal as a vector



If the desired signal is graphed as a phasor (as in **Figure 3**) one might say that the B channel picks up the X component of that phasor, and the A channel picks up the Y component of the phasor. The two-channel output mixers are also driven with signals 90° out of phase. That way, the output tones combine vectorially. The result is that the combined output is a tone whose amplitude reflects the amplitude of the desired signal, regardless of the signal phase. The phase of the output tone also reflects the phase of the desired signal.

The theoretical response curve of the filter may be developed. We won't go into the mathematical details except to say that the amplitude response is a $\sin x/x$ curve, like that in **Figure 4**. For a 0.1-second frame length, the nulls in the filter response occur every 10 Hz either side of the center frequency. The 3-dB points on this curve are 9 Hz apart; the 6-dB points are 12 Hz apart.

Figure 5 compares the CCW filter (0.1 second frames) with an ordinary 500-Hz CW filter and a 2700-Hz SSB filter. On this scale it is impractical to show the numerous nulls in the CCW-filter response; shown instead is the envelope of the primary response.

HOW MUCH IMPROVEMENT ?

One way of comparing CCW with the ordinary CW method is to consider the filter noise bandwidth. This is the bandwidth of an ideal steep-sided filter which would pass the same amount of random noise as the filter being considered. For 0.1-second frame length CCW, the filter noise bandwidth is 10 Hz. This equates to an approximate superiority of 17 dB over a 500-Hz CW filter and about 24 dB over a 2300 Hz filter.

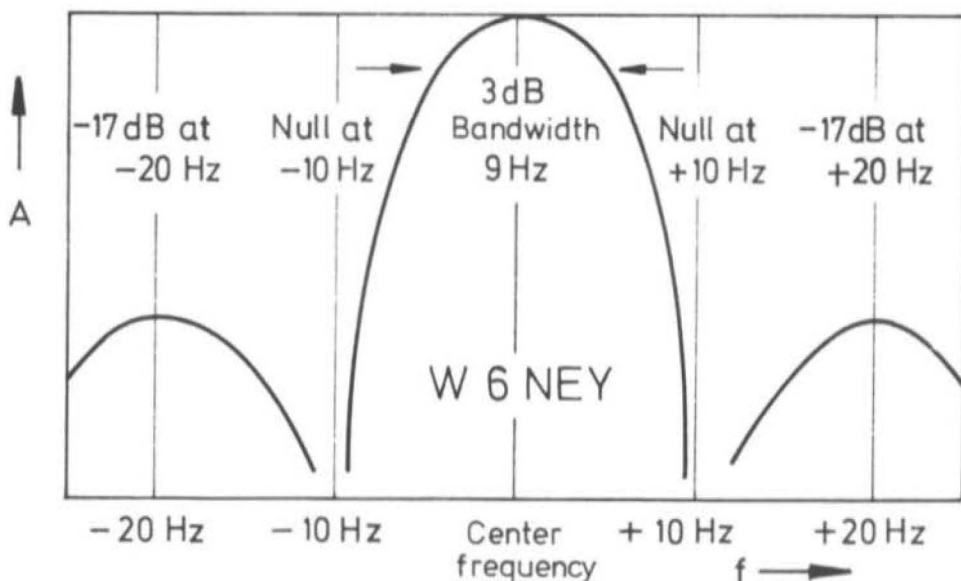


Fig. 4: Amplitude response of a 10 Hz bandwidth CCW filter

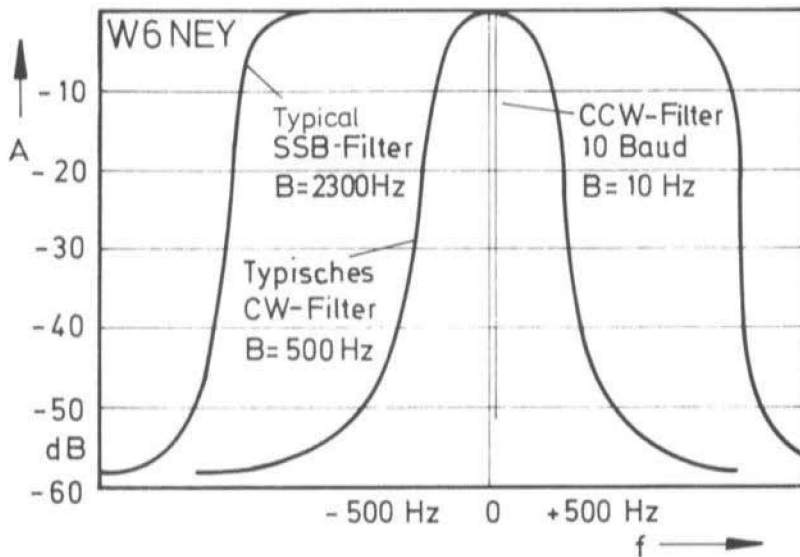


Fig. 5: Comparison of three filters

Outer curve: Typical 2300 Hz SSB-filter
Center curve: Typical 500 Hz CW-filter
Inner curve: A CCW-filter for 10 Baud, bandwidth is 10 Hz

Such estimates should be reasonably accurate with respect to noise, but when QRM is present, the CCW filter probably does even better. Using a CCW system of 0.1-second frames with ground wave in the presence of natural noise, and adjusting power for matching readability, I have measured an approximate 16-dB improvement over a 470-Hz crystal filter; this is close to the theoretically expected value.

Narrowing the CCW bandwidth by using longer frame times provides an additional signal-to-noise advantage at the price of slower information transmission rates. A 0.1-second integration period gives about 24 dB improvement over a 2300-Hz crystal filter; a 1-second integration period (1.2 wpm), 34 dB; a 10-second

period, (0.12 wpm), about 44 dB. These speeds are slow, but the improvement in effective communication with lower power is quite fascinating.

The improvement gained by long-frame CCW is limited by phase modulation introduced by the propagation path. For 14-MHz signals, motion in the F layer typically produces 2 or 3 Hz of phase (or frequency) modulation for a JA to W6 path. (We have also observed what appears to be propagation time delays under poor band conditions.) When the filter passband becomes so narrow that this modulation exceeds the filter bandwidth, further improvement in signal-to-noise ratio cannot be obtained by narrowing the filter passband.



In evaluating filter effectiveness, noise bandwidth does not tell the whole story; there are psychological considerations, too. The human ear is frequency sensitive, and the human brain can focus on particular CW signal frequencies amid the noise and QRM. Skillful CW operators use this capability well. My observations have led me to conclude that this skill is worth at least a 6-dB margin when using a 2300-Hz filter. QRM, however, is often a confusion factor and therefore causes more degradation of copy than an equivalent amount of random noise. These psychological factors are difficult to quantify, but probably reduce the advantage of CCW over ordinary CW.

Figure 6 shows graphically the results of on-the-air comparisons between CW and CCW made in 1975. Transmissions were made on 14,049,000 Hz from JR 1 ZZR at power levels of 10 watts, 1 watt and 0.1 watt using CCW and a vertical groundplane antenna on a four-story building. A three-element beam was used for reception at W 6 BB. The CCW signals were received simultaneously as CW and CCW signals, and were recorded on separate channels of a stereo cassette recorder. We selected sample periods from the cassette recording and played back the signals to four moderately experienced CW operators. The average proportion of copy shown on the graphs is based upon words considered copied. The copy content was taken from radio journals. Extrapolation of these data indicate an estimated 25-W CW signal as equivalent to a 0.1-W CCW signal in communications effectiveness, or a 24-dB superiority for CCW.

CONCLUDING REMARKS

The CCW technique appears to be most promising, especially where signals are weak compared to the noise and QRM. Under high absorption and QRN conditions (as often experienced on 80 and 160 meters) the additional selectivity of CCW would be helpful; we don't have data on that yet however.

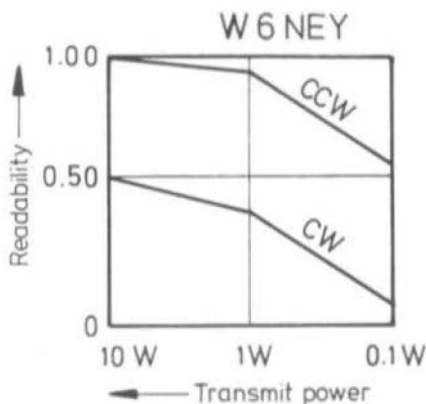


Fig. 6: Average readability of simultaneously transmitted CW and CCW transmissions when received by four different operators. Three different output power levels were used

CCW might be used for EME communication, but the problem is complicated because of lunar-motion Doppler effects. One might need a computer to calculate the frequency at which the signal is expected to return. Also, achieving the necessary frequency stability of 1 or 2 Hz is more difficult at the higher frequencies used for EME.

Some of the simplest rigs are the easiest to convert for CCW operation. To obtain the full advantage of the CCW mode, however, receiver quality should be high. In Part 2, I will describe the equipment and methods used for communicating by CCW.



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letter 75:7. Note: Back copies of volumes of the Coherent CW Newsletter (CCWN) are available from CCWN, 2301 Oak St., Berkeley, CA 94708: 1975, \$5; 1976, \$5; 1977, \$10; 1978, \$10. Volumes 75 and 76 are well summarized in the Weiss article in CQ. Most of volumes 77 and 78 are summarized in this article. Further volumes of the CCWN are not planned, but a book on CCW is being assembled by Petit. This article has benefited from suggestions by: Jim Maynard, K 7 KK; Ray Petit, W 7 GHM; Keitaro Sekine, JA 1 BLV; and Ed Johnson, W 2 ZWA/JA 1 YVW.

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Michael Martin, DJ 7 VY

A Wideband Driver for the Shortwave Bands

The following article is to describe a driver amplifier for the shortwave bands, which is designed to match the HF-VHF transverter described by DK 1 OF in (1). This low-intermodulation amplifier provides an output power of 4 W PEP between 1 and 30 MHz, and can either be used for qrp-communications, or for driving a tube amplifier equipped with two 6JB6 or 6146 types. Two such amplifiers will also allow low-reactive two-tone signals to be generated for the measurement of intercept point values of > 40 dBm.

1. TECHNICAL SPECIFICATIONS OF THE AMPLIFIER

Input impedance: $Z_{in} = 50 \Omega$
 Load impedance: $Z_{out} = 50 \Omega$

(For values $\neq 50 \Omega$, the frequency response and intermodulation values will be different).

Noise figure: NF = 3 dB at 30 MHz

Output power:

$P_{out} = 4 \text{ W PEP} \triangleq 2 \times 30 \text{ dBm} \triangleq 40 \text{ V}_{pp}$ into 50Ω

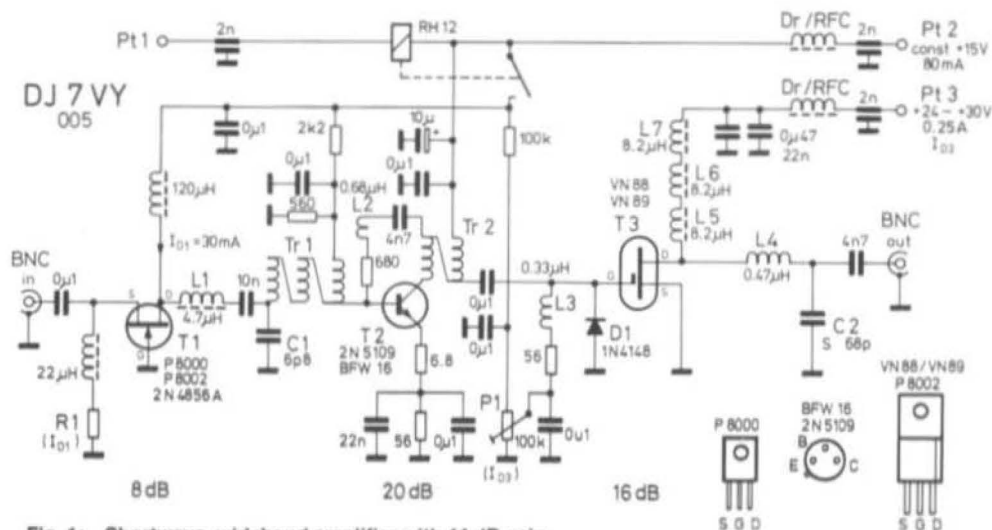


Fig. 1: Shortwave, wideband amplifier with 44 dB gain between 1 and 30 MHz; $P_{out} = 4 \text{ W PEP}$



Fig. 2:
IM-rejection at full drive,
measured at 3.6 MHz
with $U_D = 30$ V;
Hor.: 20 kHz/div.
Vert.: 10 dB/div.

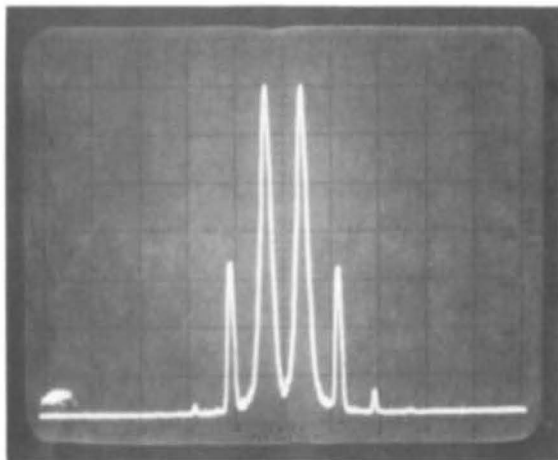


Fig. 3:
Conditions as for Fig.1,
but $U_D = 15$ V !

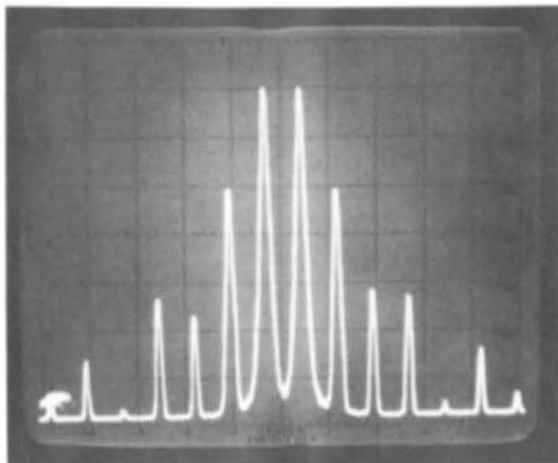
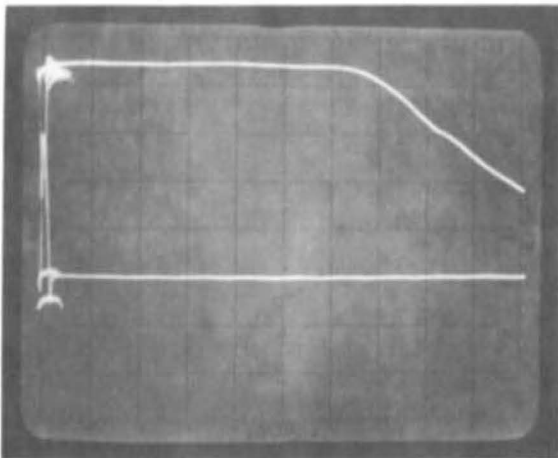


Fig. 4
Frequency response from
0 to 50 MHz;
Hor.: 5 MHz/div.
Vert.: 10 dB/div.





Gain:

$G_p = 44 \text{ dB} \pm 0.3 \text{ dB}$, 1 to 30 MHz

Input signal for full drive:

$P_{in} = 2 \times -14 \text{ dBm} \triangleq 2 \times 44 \text{ mW}$

Harmonic suppression at $P_{out} = 30 \text{ dBm} = 1 \text{ W}$
(measured at 9 MHz): $2 f_1 = 22 \text{ dB}$,
 $3 f_1 = 36 \text{ dB}$, $4 f_1 = 56 \text{ dB}$, $5 f_1 = 76 \text{ dB}$

Third order intercept point:

$IP_{in} = 4 \text{ dBm}$ $IP_{out} = 48 \text{ dBm}$

Intermodulation rejection at full drive:

$IM = 36 \text{ dB}$ (42 dB ref. PEP)

(no 7th, 9th and 11th order products as with bipolar power transistors).

Intermodulation values for other output power levels can be calculated according to the following equation:

$$IM = 2 (IP_{out} - P_{out})$$

e.g. $36 = 2 (48 - 30)$.

Power requirement:

15 V / 80 mA, 30 V / 250 mA.

Dimensions: 72 mm x 110 mm x 45 mm
(including heat sink).

2. CIRCUIT DESCRIPTION

As can be seen in **Figure 1**, the input signal is capacitively coupled to the high-current FET T 1 (P 8002), which possesses an input impedance of 50Ω in the common-gate circuit used. The subsequent stage T 2 equipped with transistor 2 N 5109 in a feedback configuration also possesses an input impedance of 50Ω which is transformed up in transformer Tr 1 to provide an output load of 450Ω for the first stage. Inductance L 1 and capacitor C 1 are provided for compensating the frequency response at 30 MHz. This is done by extending the gate-drain capacitance of T 1 to form a π -link. Inductances L 2 and L 3 are provided for improving the frequency response at the upper frequency limit. The second stage is terminated with 56Ω , which ensures a good neutrali-

zation of the V-MOS-FET output stage even in the case of complex load impedances.

V-MOS-FETs are transistors in which the current does not flow horizontally (as usual) but vertically within the chip geometry. They exhibit a very high cut-off frequency, good reliability, and the absence of secondary breakthroughs. However, they are somewhat sensitive to overvoltages, which means that the operating voltage must always be less than half of the maximum voltage given in the data sheet.

A 90 V / 2 A transistor type VN 89 was selected for T 3, which possesses a power dissipation of 12 W. The transistor operates in class A, and diode D 1 ensures that the gate bias voltage is increased correspondingly with a slight delay so that even greater output powers than 4 W can be achieved, of course, with a higher current drain. The output capacitance of T 3 is compensated for with the aid of L 4 and C 2.

At an output impedance of 50Ω , the amplifier provides intermodulation values that are far in excess of those of bipolar transistor drivers. **Figure 2** shows the intermodulation rejection at full drive, where the power of a single tone is 30 dBm $\triangleq 1 \text{ W}$, corresponding to 4 W PEP. The IM-rejection is -36 dB with respect to a single tone, or 42 dB when referred to PEP. This corresponds to a third order intercept point IP_{out} of 48 dBm. With a gain of 44 dB, this amounts to an input IP-value of $+4 \text{ dBm}$.

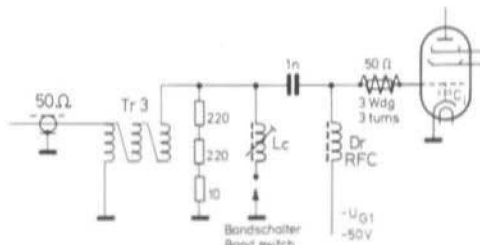


Fig. 5:
A tube amplifier in a grounded cathode circuit can be driven via a wideband transformer: Tr 3 (as Tr 1)



Figure 3 shows the amplifier with the same input signal after the operating voltage of T 3 has been reduced to 15 V. The driver should then only be driven up to approximately 2 W PEP.

Figure 4 shows the frequency response of the amplifier between 0 and 50 MHz.

The current drain can be switched to zero in the standby mode during VOX-operation.

Figure 5 shows a wideband transformer (50 Ω : 450 Ω) for driving a tube power amplifier. The input capacitance C_{in} of the tube must be compensated for with L_c :

$$\left(L_c = \frac{1}{\omega^2 C_{in}} \right)$$

A voltage of 120 V (peak-to-peak) present across 450 Ω is more than sufficient for all popular power amplifier tubes. Since the resonant circuit comprising L_c and the tube, and connection capacitances is dampened with 450 Ω , it will not be necessary to carry out any tuning of the driver within any band. Any possible grid current will not cause any limiting as encountered with high-impedance tube drivers such as the 12 BY 7.

2.1. Components

- T 1: P 8000 or P 8002 (TI) with heat-sink, see text
 T 2: 2 N 5109 or BFW 16 (Siemens) with large cooling fins.
 T 3: VN 88 or VN 89 (Siliconix) with mica disk
 D 1: 1N 4148 or 1N 4151
 Tr 1: Toroid core R 10-N 30 (Siemens) with 3 x 8 turns of 0.5 mm dia. enamelled copper wire
 Tr 2: Toroid core R 10-N 30 with 2 x 9 turns of 0.5 mm dia. enamelled copper wire

All cores and chokes: Fixed inductances for 12.5 mm spacing

- | | | | |
|------|--------------|------------|-----------------|
| L 1: | 4.7 μ H | L 5 - L 7: | 8.2 μ H |
| L 2: | 0.68 μ H | | 1 x 22 μ H |
| L 3: | 0.33 μ H | | 1 x 120 μ H |
| L 4: | 0.47 μ H | | |

2 x 6-hole ferrite core chokes (Philips) for filtering the operating voltage.

- C 1: 6.8 pF ceramic disk capacitor, spacing 5 mm
 C 2: 68 pF styroflex capacitor, spacing 12.5 mm

Capacitors up to 470 nF: ceramic disk, tubular, or multilayer capacitors for 5 mm spacing, also 7.5 mm spacing for the larger values;

- 1 tantalum electrolytic: 10 μ F
 3 feedthrough capacitors of approx. 2 nF for solder mounting
 1 trimmer potentiometer, 100 k Ω , horizontal mounting, spacing 10/5 mm
 1 relay RH-12 V (National)
 2 BNC-connectors for single hole mounting with double solder tags
 1 metal case with heat sink

3. CONSTRUCTION

The amplifier can be constructed using the 70 mm x 90 mm double-coated PC-board DJ 7 VY 005 (see **Figure 6**), and can be enclosed in a metal case of 72 x 110 x 30 mm complete with heat sink. **Figure 7** shows the author's prototype built up on a 2.5 mm Vero board. Construction does not present any problems due to the 50 Ω interstage technology.

Transistor T 1 is screwed to the cover of the metal case below the through-contacts after increasing the length of its connections to allow them to be placed through the board. The cooling tag of T 3 is bent by 90°, placed through the cover and screwed to the heat sink. The

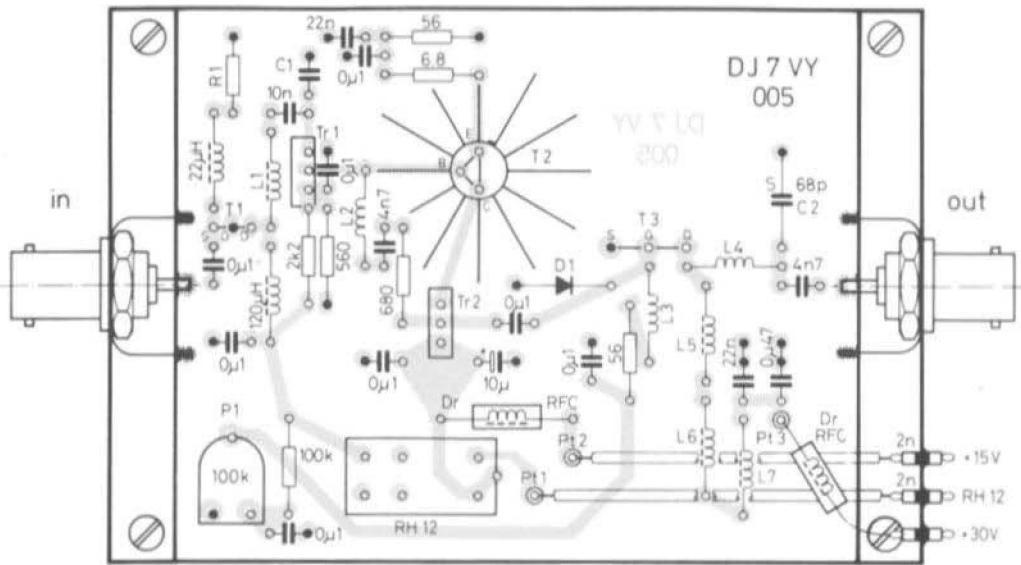


Fig. 6: The double-coated PC-board is only 90 mm x 70 mm and can be mounted in a suitable metal case

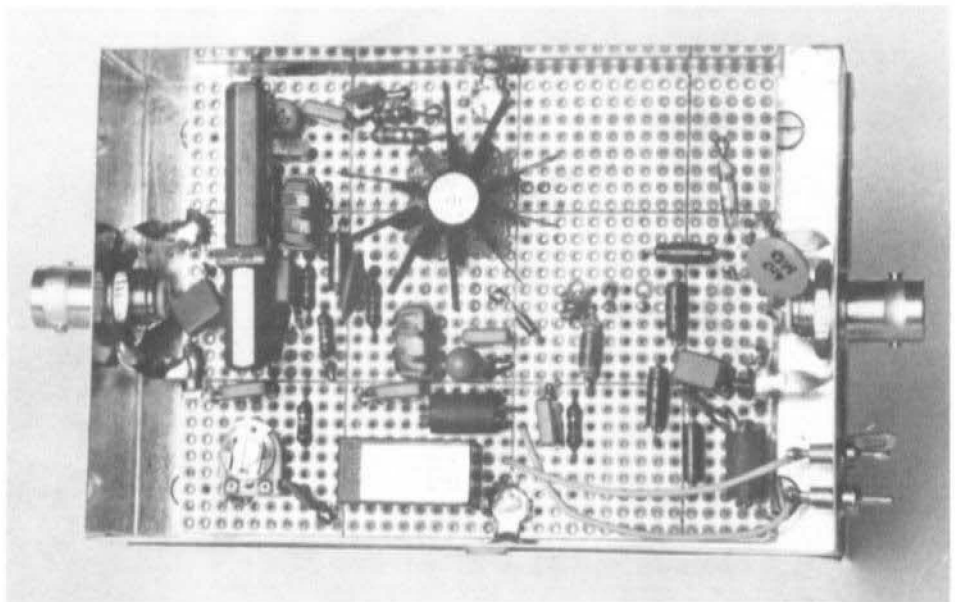


Fig. 7: In the author's prototype using a Vero board, T 1 (left) was cooled with two hexagonal spacers; T 3 is below the board



mica disk for insulating the drain should be provided with heat-conductive paste or with two drops of oil.

The position of the connectors can be seen in Figure 7; however, the input and output can also be on another side of the case. If they are to be mounted beside each other, it is recommended that an RF-tight screening of the input connector is provided and that a choke should be used for suppressing any sheath-current on the outside of the cable between connector and PC-board (ground the outer conductor at both sides of the cable!). RF-currents on the outside of the cable can be suppressed by placing several turns of the coaxial cable through a R 10/N 30 toroid core.

After winding the transformer, it will now be possible for the components to be mounted on the board. The value of R 1 is determined by measuring the drain current at various values of R 1. After passing the connection leads of T 1 and T 3 through the through-contacts, the board is firstly fixed in the case by soldering at several points around the edge. This is followed by mounting the connectors, the operating voltage feed-throughs, and T 1 and T 3. The operating voltage is connected after placing potentiometer P 1 to zero.

4. CONNECTION AND ALIGNMENT

A signal generator with a variable output voltage of maximum 100 mV is sufficient for driving the amplifier; two 100 Ω / 2 W carbon resistors can be connected together as output load.

Potentiometer P 1 is aligned for a quiescent current of 250 mA via T 3, after switching on the amplifier via Pt 1.

CAUTION: It is necessary for the 15 V operating voltage to be stabilized, since any slight voltage fluctuations will cause large variations of the quiescent current!

An output voltage of 40 V (peak-to-peak) should be measured with an HF-probe connected to the output when tuning the signal generator between 1 and 30 MHz. If a sweep-frequency measuring system is available, it is possible for the frequency response to be displayed as given in Figure 4. The amplifier has been constructed several times even using different types of construction, and has worked immediately under all conditions.

When using the amplifier as QRP-transmitter, a miniature matchbox should be used in order to ensure a good 50 Ω termination, which then also provides a further harmonic suppression.

If the gain is still not high enough, a further P 8002 preamplifier can be provided with a 2:1 transformer in the drain circuit to provide a further gain of 6 dB, or 8 dB, if a 3:1 transformer is used. In this case, a frequency response compensation similar to L 1 and C 1 must be found experimentally.

REFERENCES

- (1) J. Kestler, DK 1 OF:
A Universal Converter for HF and VHF
VHF COMMUNICATIONS 8,
Edition 3/1976, pages 159-174

Murphy's Law – as experienced by DL 6 WU:

»The more often one measures the gain of an antenna, the smaller it becomes.«



Leif Åsbrink, SM 5 BSZ

The Optimum 6 Element Yagi-Antenna

In an article in the IEEE Transactions on Antennas and Propagation (Vol. AP 23, 1975, page 8 ff.), two authors C.A.Chen and D.K.Cheng described a method of designing optimized Yagi antennas using computer calculations. The article also included a numerical example for an optimum 6-element Yagi antenna. In this case, optimum means the highest possible gain, independent of boom length and bandwidth.

This article is now to give the data for constructing a 70 cm and 2 m version of this antenna, however, it is not meant to be a fool-proof method of constructing antennas, but is to describe a method of constructing an antenna that really works.

THE CHEN/CHENG ANTENNA

Over the years, many radio amateurs have tried to construct an antenna according to the theoretical calculations of Chen/Cheng. The results were usually unsatisfactory. Such an antenna has been described in Ham Radio Magazine.

The reason that this antenna does not usually operate correctly is partly due to the fact that it is very critical, and partly due to the calculation method. It is necessary for the antenna to be measured and optimized after construction.

A signal generator and receiver are required for this measurement that can be tuned over a frequency range of $\pm 5\%$ of the required center frequency. Furthermore a site must be

found and be prepared where the radiation pattern can be measured in this frequency range. The angular resolution should be in the order of 0.5° . For the level measurements, either a precision attenuator will be required, or the receiver must be linear (AGC off, maximum AF gain, CW or SSB mode, and decrease the RF-gain as required), so that the level can be measured at AF-level using a normal AF-voltmeter.

The radiation pattern can be measured by rotating the antenna in an axis that runs parallel to the elements. A suitable arrangement is shown in **Figure 1**. It is also possible to measure between the roofs of two houses having a suitable spacing from another. The most important point is that one avoids ground reflections and any other reflections from neighbouring objects.

If the measuring path is satisfactory, the radiation diagram is now measured at various frequencies, and the frequency is selected at which the pattern corresponds as close as possible to the theoretical values given in **Table 1**.

This is followed by calculating how far this frequency is from the design center frequency in per cent and by changing the lengths of the elements by twice this percentage. The positions of the elements remain unchanged.

If the radiation patterns are now measured again at various frequencies, it will be found that a pattern will be present which is much more similar to the tabular values, and far nearer to the design frequency than before.

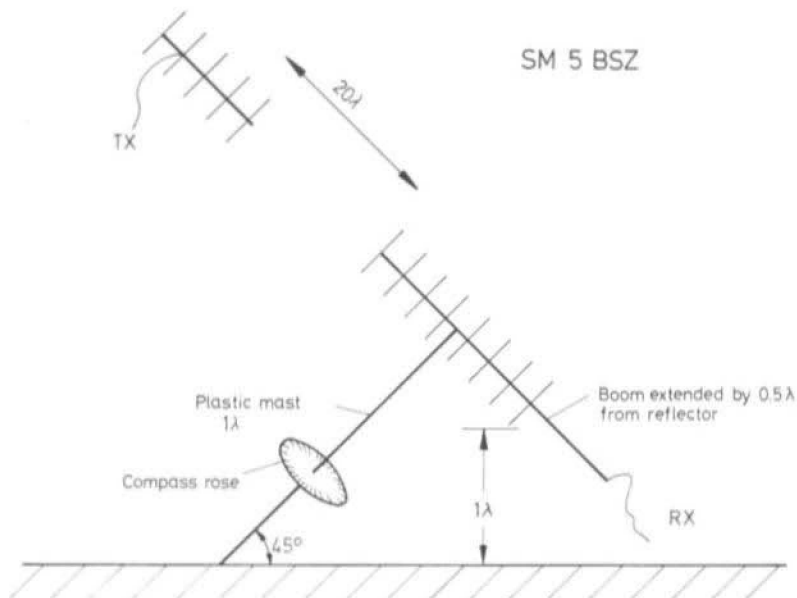


Fig. 1: An antenna measuring set-up that avoids ground reflections

Angle (deg.)	Level (dB)	Note
0	0	Main beam
19	- 3	
25.5	- 6	
30	- 9	
38.5	- 20	First dip
53	- 11	First sidelobe
71.5	< - 30	Second dip
83	- 17.5	Second sidelobe
96	< - 30	Third dip
108	- 18.5	Third sidelobe
120	- 28	Fourth dip
138	- 13	Fourth sidelobe
156	- 25	Fifth dip
180	- 10	Back lobe

Table 1: The radiation pattern values of the Chen/Cheng antenna

After carrying out one or two further corrections according to the given example, one will

obtain the final (= theoretical) radiation pattern at the design frequency. The gain will then amount to 11.5 dB over a dipole, which is approximately 1 dB more than can be obtained with other antennas having a comparable length.

ANTENNA MEASUREMENTS MADE AT ANNABODA/SWEDEN

An antenna optimized according to the described procedure participated at the Annaboda antenna measurement contest in 1980. The gain measured was 11.8 dB, however, it seems that the 70 cm measurements indicated gain values that are too high by 2 to 3 tenths of a dB. A direct integration of the radiation pattern will give a gain of 11.6 dB. Chen and Cheng give a theoretical gain value of 11.25 dB, which is probably incorrect for some unknown reasons.



However, the exact gain value down to a tenth of a dB is only of theoretical interest; it is more important that this antenna is able to provide noticeably more gain than any other antenna of similar length. The measured values at Ännaboda, which are very accurate for the 2 m antennas, can be used for comparison. These were published in the Swedish Amateur Radio Magazine QTC. The results for 144 MHz were given in edition 4/81, and the 432 MHz results in edition 5/81.

PROVED CONSTRUCTIONS

For those readers that do not wish to optimize the antennas according to the described measuring and optimizing procedure, a description is now to be made of an optimized Chen/Cheng antenna for the 144 MHz and 432 MHz band.

The 432 MHz antenna is constructed using a tubular PVC boom of 19 mm outer diameter and 1.5 mm wall thickness. All elements are constructed from 5.0 mm diameter solid aluminium rods. They are pressed into holes in the boom.

The dimensions are given in **Table 2**.

Element length (mm)	Spacing from reflector (mm)
324.5	0
308.5	173.5
297.0	374.5
293.0	656.0
296.0	880.5
293.0	1173.5

Table 2: Dimensions of the 70 cm antenna

The 144 MHz antenna is designed for a center frequency of 144.4 MHz. The boom is constructed from aluminium tubing with an outer diameter of 25 mm. All elements are constructed from aluminium tubing with an outer dia-

boom using special element clamps which will possibly have an effect on the resonant length of the elements. The element clamps can be obtained from SM 5 ERW, PI 8820, S-64300 VINGÅKER/Sweden.

If the elements are to be mounted onto the boom in a different manner, or when the boom diameter is changed considerably, it will be necessary for the antenna to be measured and realigned. When using the dimensions given in **Table 3**, the antenna will operate satisfactorily between 144.0 and 144.6 MHz.

Element length (mm)	Spacing from reflector (mm)
1004	0
954	519
922	1120
907	1963
917	2634
907	3511

Table 3: Dimensions of the 144 MHz antenna

DESIGN OF A SUITABLE GAMMA MATCH

Although the »gamma match« should be part of the fundamental knowledge of any person interested in antennas, a high-quality version of such a match is to be briefly described. The main feature is that the capacitor is made in a coaxial manner using a PTFE tube as dielectric. This construction ensures that it is impossible for water to run to a position where strong electrical fields are present. As can be seen in **Figure 2**, the PTFE-tube also protrudes by approximately 10 mm on both sides of the capacitor, which means that the path is very long for leakage currents when the antenna is dirty and wet. Of course, the surfaces of the two copper parts, and the end of the coaxial cable must be protected against corrosion. The result is a gamma match that is suitable for the high effi-



The author hopes that radio amateurs will forget «the resonant frequency of an antenna», since such a concept is confusing! The only thing that is in resonance is the match, and any antenna can be matched impedance-wise at any frequency, — even when the frequency of optimum gain is far away. Radio amateurs should understand that

Antenna gain = Good radiation pattern !

Is a single dB important ?

Good, normal Yagi antennas such as the Tonna 9-element, HyGain 8-element, Cush Craft 11-element, Wisi 10-element, or Jay-beam 10-element, are all on a straight line in a diagram giving the gain as a function of boom length (such diagrams should be drawn in a linear scale, and not in dB). This line crosses a gain value of 11.5 dB at a boom length that is 25% greater than that of the Chen/Cheng antenna. Furthermore, normal Yagi antennas require at least 50 % more elements.

This means, that the described antenna is thus 25% better than conventional Yagi antennas — corresponding to 0.97 dB. The gain differences between conventional Yagis of equal length and the Chen/Cheng antenna amounts to approximately 1 dB.

Normally, it is assumed that a signal difference of 1 dB will not be audible. This is, however, only (approximately) true when the signal is well out of the noise. This is also valid for a very weak continuous carrier. However, it is not commonly known that the «probability of detection» Morse signals is affected by small differences of the signal-to-noise ratio.

Experiments have shown that a good telegraphist can usually copy Morse signals at a certain speed without error, however, the errors will increase considerably when the signal approaches the noise level. If the signal-to-noise ratio is selected so that 85% of the letters can be copied correctly, a deterioration of the signal-to-noise ratio by only 1 dB will decrease the number of correctly copied letters to 60%!

For those radio amateurs that wish to carry out communication at the technological limits (e.g. EME), this represents a considerable difference. The price of this 1 dB more antenna gain is a reduced bandwidth, however, such low-signal communication is always made within the first 500 kHz of the amateur band.

Portable Antennas

operating according to log-periodic principles

LP 2/2 m — LP 3/2 m — LP 5/2 m

Two, three, and five element portable antennas for 144 MHz.

Gain 4.2 dBd, 7 dBd, and 8 dBd respectively.

Prices:	LP 2/2 m:	DM 29.80
	LP 3/2 m:	DM 49.00
	LP 5/2 m:	DM 79.00

LP 3/70 cm and LP 5/70 cm

Three and five element portable antennas for 70 cm.

Gain 4.5 dBd and 8 dBd respectively.

Prices:	LP 3/70 cm:	DM 39.00
	LP 5/70 cm:	DM 49.00



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A Receive Converter for the Geostationary Weather Satellites METEOSAT – GOES – GMS

Part 2: The Local Oscillator Module

Part 1 of this article (VHF COMMUNICATIONS 4/1981) described the interdigital filter-mixer module with IF-preamplifier. Part 2 is to describe the matching local oscillator module.

METEOSAT transmits WEFAX (APT) transmissions on two frequencies: 1691.0 and 1694.5 MHz. These two frequencies are to be converted to a first intermediate frequency of 137.5 MHz. This intermediate frequency is also used in professional systems. It may seem reasonable to use a first intermediate frequency in the 144 MHz band, however, this is not advisable since there are no amateur receivers that possess a suitable bandwidth at IF and audio level for the automatic picture transmission (APT). Furthermore, there is more danger of breakthrough of strong 2 m stations. This means that a special receiver is required and this can be at the standard first IF, which is also covered by the DC3NT receiver. A further advantage is that one of the orbiting weather satellites (NOAA 6) also operates on this frequency.

For those readers outside of Europe, Middle East and Africa, it should be noted that the GOES and GMS-series of satellites only transmit on a single WEFAX frequency of 1691.0 MHz.

CIRCUIT DESCRIPTION

In the case of METEOSAT, two local oscillator frequencies are required: 1557.0 and 1553.5 MHz. These frequencies are derived from crystals in the 97 MHz range, which are multiplied by 16. The circuit diagram of the local oscillator module is given in **Figure 1**.

Each crystal is provided with its own oscillator circuit and subsequent buffer stage. Frequency selection is made using a small relay by remotely switching the 8.5 V stabilized operating voltage. This rather extensive circuitry has the advantage that each crystal can be adjusted to the required frequency.

The two buffer stages drive a common 97 MHz resonant circuit which comprises inductance L 5 and the two capacitors of 18 pF and 27 pF respectively. The first of two transistor doubler stages is connected to this capacitive tap. The resulting frequencies are given in the circuit diagram. This is followed by a straight-through amplifier for 389 MHz and a varactor multiplier with a $\lambda/2$ circuit at the output.

At the lower frequencies, the coupling from stage-to-stage is made with the aid of band-pass filters in order to mainly suppress spec-

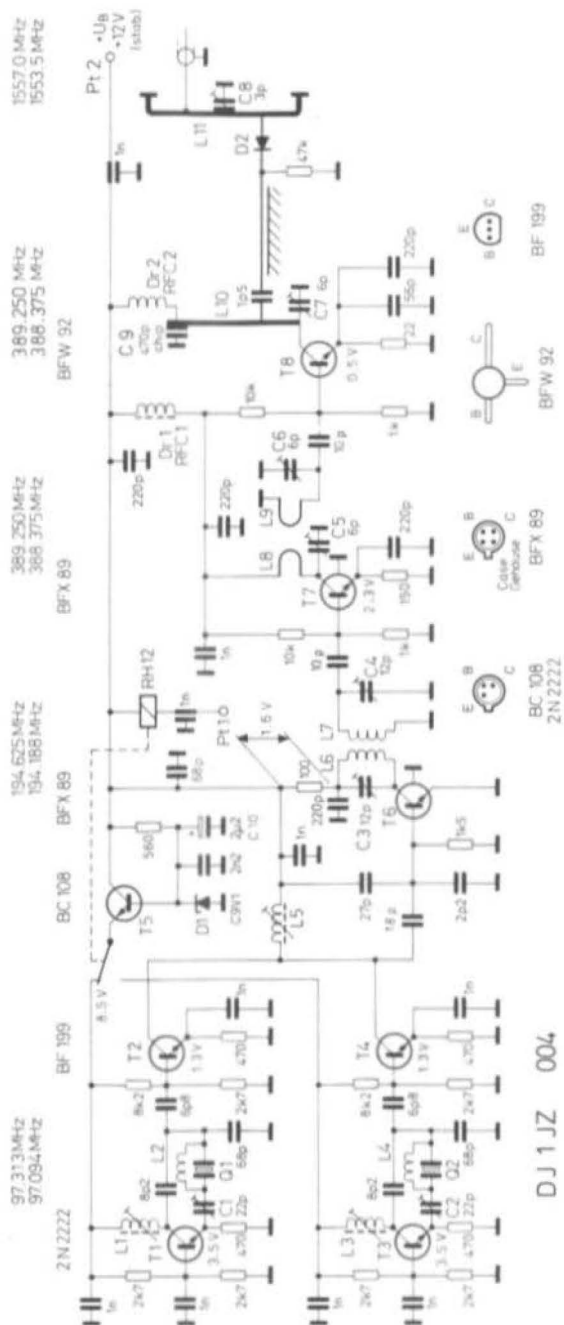


Fig. 1:
The 6-stage local oscillator
uses a separate circuit for
each crystal to provide a stable
behaviour at the exact frequency



tral lines that are in the vicinity of the output frequency, (1557 \pm 97 MHz). Interfering lines spaced further from the frequency are suppressed in the interdigital filter of the mixer module.

COMPONENTS

- T 1, T 3: 2N 918, 2N 2222, or BFX 89, BFY 90
- T 2, T 4: BF 199, BF 224
- T 5: BC 108, BC 182, or similar silicon NPN-transistor
- T 6, T 7: BFX 89 or BFY 90 or 2N 5179
- T 8: BFW 92 or BFR 34 A
- D 1: Zener diode 9.1 V
- D 2: Silicon switching diode 1N 914, 1N 4148, or 1N 4151 or similar
- L 1, L 3: 4.5 turns of 0.8 mm dia. silver-plated copper wire wound on a 6 mm ripped coil former with VHF-core
- L 2, L 4: 20 turns of enamelled copper wire of approx. 0.5 mm dia., wound on a 3 mm former, self-supporting, soldered to the conductor side of the board.
- L 5: 5.5 turns, otherwise as L 1
- L 6, L 7: 2.75 turns of 0.8 mm dia. silver-plated copper wire, wound on a 6 mm former, pulled out to a spacing = wire diameter, and soldered into place with 3 mm spacing to the board.
- L 8, L 9: U-shaped bracket made from 0.8 mm dia. silver-plated copper wire bent around a 6 mm former, and soldered into place so that the top is 10 mm over the board.
- L 10: 40 mm long piece of 1.5 mm dia. silver-plated copper wire; one end bent 7 mm and the other 3 mm at right angles; solder short end to chip capacitor.
- L 11: 56 mm piece of 2 mm dia. silver-plated copper wire bent at right angles at 3 mm from each end and soldered to the board with a spacing of 3 mm at both ends. Trimmer C 8 is at the center. Coil tap for D 2 is 8 mm from the bent, tap for the output is 6 mm from the other bent.
- RFC 1: Ferrite bead with 3.5 turns of max. 0.4 mm dia. enamelled copper wire
- RFC 2: 15 turns of approx. 0.5 mm dia. enamelled copper wire wound on a 3 mm former, self-supporting
- C 1, C 2: 22 pF plastic foil trimmer 7.5 mm dia. (Philips: green)
- C 3, C 4: 12 pF plastic foil trimmer, 7.5 mm dia. (Philips: yellow)
- C 5 - C 7: 6 pF plastic foil trimmer, 7.5 mm dia. (Philips: grey)
- C 8: 3 pF ceramic tubular trimmer, miniature type for single-hole mounting (Philips: type 801)
- C 9: Approx. 470 pF ceramic disk capacitor without connection leads
- C 10: Approx. 2.2 μ F tantalum electrolytics, drop type
- 2 pieces ceramic feedthrough capacitors for solder mounting; value approx. 1 nF.
- All other capacitors: ceramic disk capacitors for a spacing of 5 mm.
- All resistors: Carbon resistors for a spacing of 10 mm
- Relay: RH-12V or RS-12V (National)
- Crystals in a welded metal case HC-47/U, fifth overtone; for higher demands on the temperature response: type XS 2106 (KVG), otherwise: type XS 2101 (KVG)
- Q 1: 97.3125 MHz
- Q 2: 97.09375 MHz
- 1 piece sheet-metal case with the inner dimensions: 145 x 70 x 28 mm.

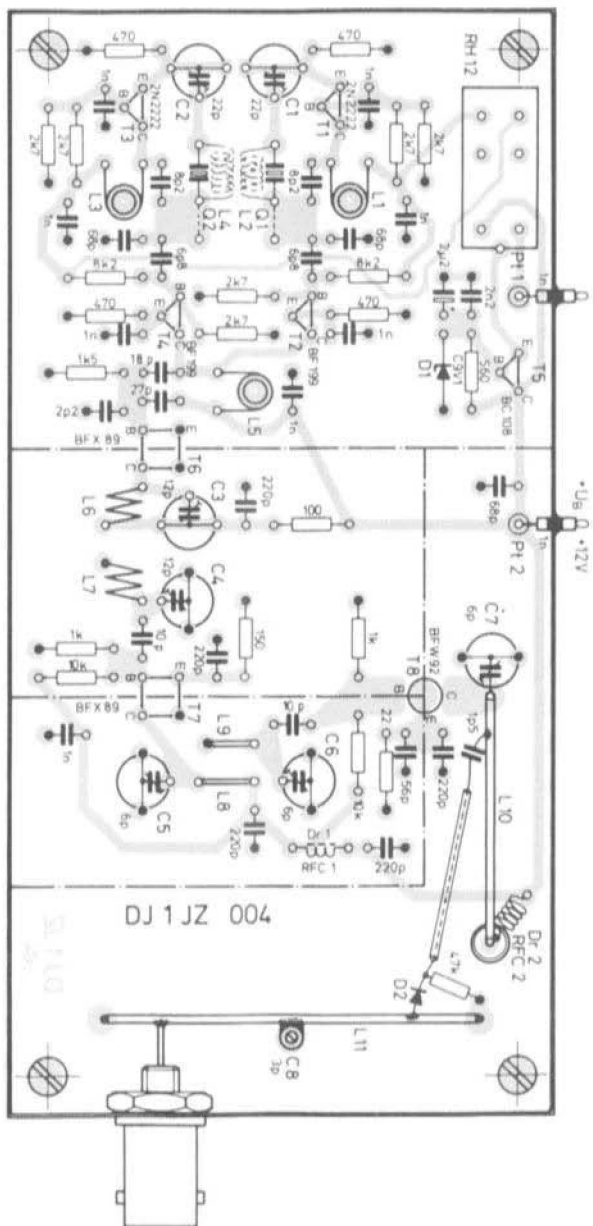


Fig. 2: Inductances L 2 and L 4, and transistor T 8 are mounted on the conductor side of the board

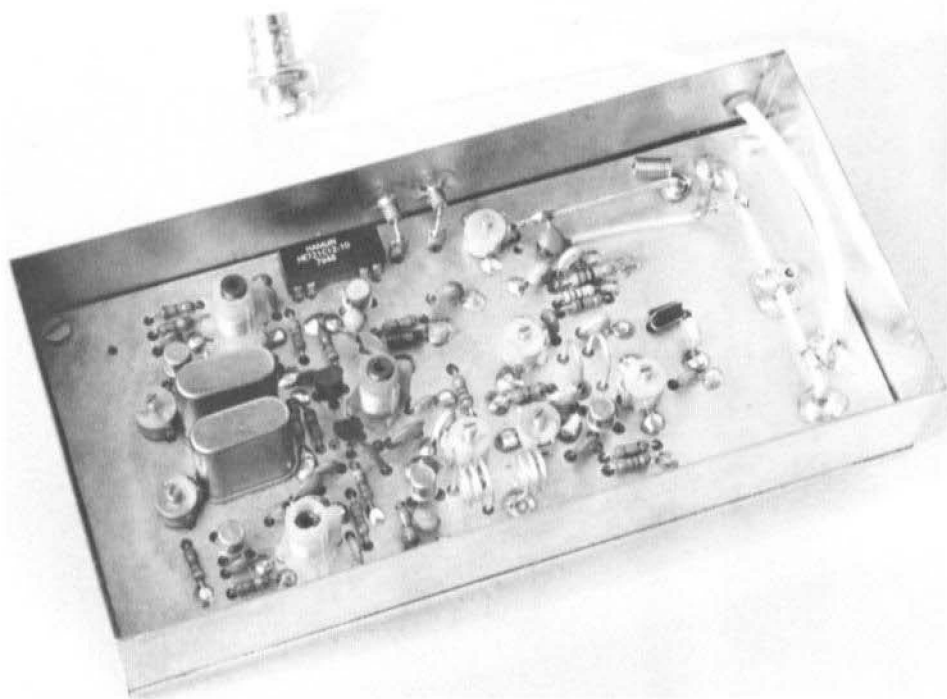


Fig. 3: The author's prototype was photographed without screening panels so that the components can be seen more easily

CONSTRUCTION

The circuit given in Figure 1 can be constructed using the PC-board shown in Figure 2. The dimensions of this board are 145 mm x 70 mm. The PC-board is double-coated, but only etched on one side. The ground surface is continuous on the component side and is only removed with the aid of a 3 mm drill at those positions where the component connections are passed through the board. All ground connections are made on the component side. Only three components are mounted on the conductor side: the two inductances for compensating the crystal holder capacitance (L 2 and L 3), and transistor T 8.

The mounting position of most components is given by the holes in the board. It is only necessary to note the positions of the components of the varactor multiplier.

The 1.5 pF capacitor from L 10 to diode D 2 keeps one connection lead of approximately 25 mm in length. Whereas its other connection (approx. 3 mm long) is connected to L 10 (approximately 5 mm from trimmer C 7), an insulating tube is passed over the long connection and the end is soldered to diode D 2 and the 47 k Ω resistor. These two components are soldered into place normally with very short connection leads. The wire provided with insulating tube is depressed onto the board in its full length so that it forms a type of stripline. The



described construction can be seen in the prototype shown in **Figure 3**.

The two feedthrough capacitors are soldered into the case in the vicinity of the appropriate connection points Pt 1 and Pt 2. Either a coaxial socket (as shown in Figure 2) or a thin coaxial cable (PTFE-cable or semi-rigid cable) can be soldered to the output tap on L 11.

The case is soldered all around the edge of the PC-board so that the conductor side is spaced approximately 3 to 5 mm from the lower cover.

Several dashed lines are shown in the components location plan and indicate the position where screening panels should be soldered. These prevent a capacitive breakthrough of the lower frequencies to the output, which means that the spectrum is cleaner. This measure should always be made when using the oscillator module in conjunction with a wideband mixer. In our case, the interdigital filter of the mixer module provides sufficient selectivity.

ALIGNMENT

The following equipment is required:

A high-impedance DC-voltmeter, a frequency counter for up to 500 MHz (e.g. DK 1 OF 044/045 + DJ 6 PI 005), a wavemeter up to approx. 2000 MHz (VHF COMMUNICATIONS 2/1975, p.90-92), as well as a power meter (VHF COMMUNICATIONS 4/1981), or at least an indicator.

Firstly check the stabilized voltage at the emitter of T 5 and the operation of the relay switching, as well as the DC-voltages at the emitter resistors of oscillator and buffer stages.

The power meter and frequency counter are now connected to L 5 and the selected crystal frequency should be indicated. The oscillator should cease operation above and below the crystal frequency when rotating the core of L 1 (L 3). If this is not the case this will indicate that

the compensating inductance L 2 or L 4 is incorrect. In this case, the oscillators should be switched off and the resonant frequency of these coils should be measured together with the associated crystal using a dipmeter. The number of turns is then varied until the resonant frequency corresponds to that of the crystal.

This problem should not be encountered with the given crystals, and it is usually only necessary for L 1 (L 3) to be aligned for maximum power (with a good commencement of oscillation) and to align C 1 (C 2) for the exact frequency. Due to the interaction, this should be repeated several times until frequency, power, and oscillation behaviour are as required.

The frequency doubler stages equipped with T 6 and T 7 can be aligned for maximum collector current; the final values measured in conjunction with the prototype are given in the circuit diagram.

It is very possible that the actual values differ from those of the prototype, and this is not important as long as a sufficient output power is provided at the output frequency (minimum: 1 mW). The frequencies should be checked at each stage, in order to ensure that an unwanted tripling does not take place.

The 389 MHz amplifier equipped with transistor T 8 should be aligned in conjunction with a power meter coupled to the stripline circuit L 10. The frequency of this stage should also be checked. If no power meter is available, one can disconnect the 47 k Ω resistor of the varactor quadrupler from ground, and bypass the free end with 220 pF. It is then possible for the DC-current to be measured. This at least allows a preliminary alignment to be made, even when C 7 must be aligned afterwards.

The varactor quadrupler must finally be aligned for maximum power at the output frequencies. The interdigital mixer can then be used as selective indicator by measuring the mixer diode current.

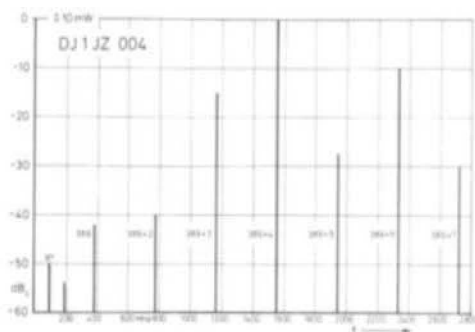


Fig. 4:
Only 389 MHz and harmonics of this are present in the spectrum of the author's prototype, now equipped with screening panels. These spurious lines are virtually eliminated in the interdigital filter of the mixer module

Since a switching diode is used in the multiplier in an application not planned by the manufacturer, the stability and output power differ greatly from diode to diode. It may therefore be advisable for a number of these diodes to be tried out to select the one with the best characteristics.

The following values were measured on the author's prototype by the publisher:

U_B :	12 V
I_B :	82 mA
P_{out} :	10 mW

In the prototype, a BFR 34 A was used for T 8, and an 1N 4148 was used for D 2. **Figure 4** shows the spectrum at the output with screening panels provided, but without passing through the interdigital filter.

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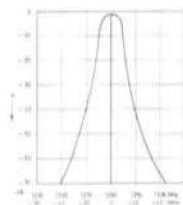
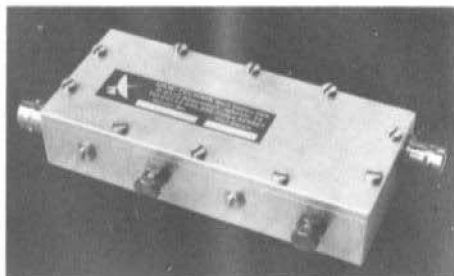
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Rolf Heidemann, DC 3 QS

A Gunn Oscillator for the 24 GHz Band

A simple, reliable Gunn oscillator for the 24 GHz amateur band is to be described to help achieve increased activity on this band. During the design, special attention has been paid to obtain a wideband, good impedance matching of the Gunn diode to the load in order to ensure that the sensitive semiconductor will not be brought into a critical, oscillation range.

For this reason, the height of the waveguide is reduced in the resonance range, although this reduces the Q of the resonator and the noise behaviour of the oscillator. However, stability has priority, especially when high performan-

ce, and thus expensive diodes are to be used. Figure 1 shows the cross-section of the Gunn oscillator; the body of the choke is shown separately in Figure 1a. The photographs given in Figures 2 and 3 show an author's prototype which was milled out of two pieces of aluminium.

An alternative type of construction can be made by using a standard waveguide material R 220 (WG 20) together with a wedge which can then be soldered into place. This wedge must be made from copper in order to ensure a good heat dissipation. Such a construction is shown in Figure 4.

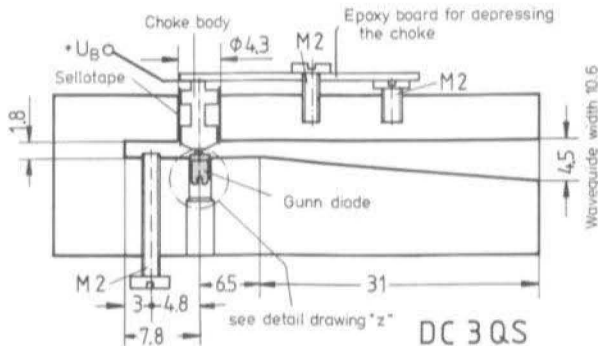


Fig. 1: Cross section of the power-Gunn oscillator for the 24 GHz band

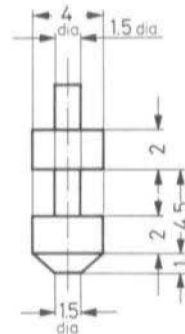


Fig. 1a: The choke of the oscillator is lathed from brass

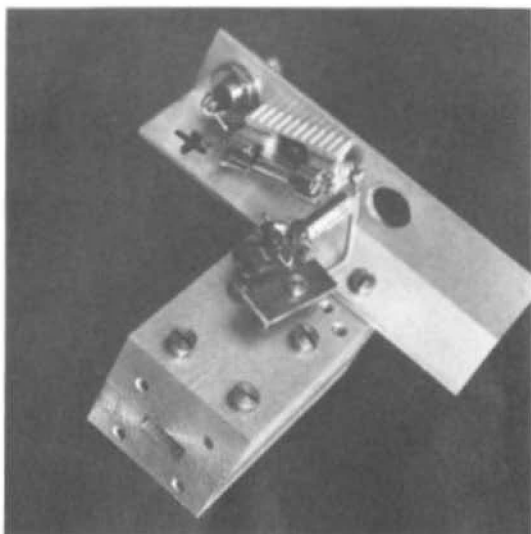


Fig. 2:
External view of the
Gunn oscillator

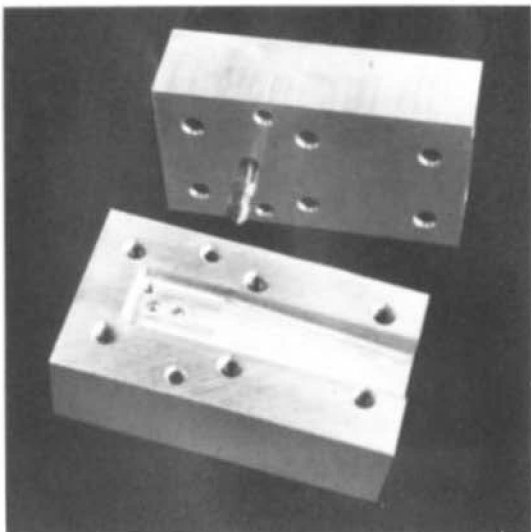
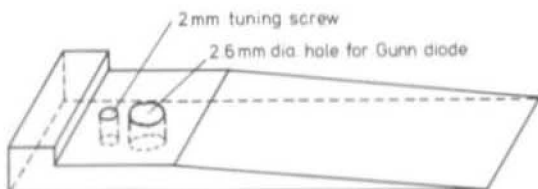


Fig. 3:
Two-piece construction
and choke constructed
as shown in Fig.1a



Solder the copper wedge into a WG 20 waveguide

Fig. 4:
Copper wedge for constructing the oscillator
shown in Fig. 1 in conjunction with a R 220
waveguide

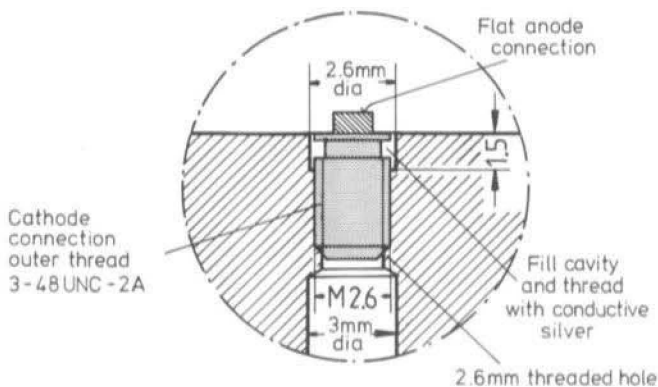


Fig. 5:
The Gunn diode can be mounted in this manner into a 2.6 mm thread, which provides a favorable solution

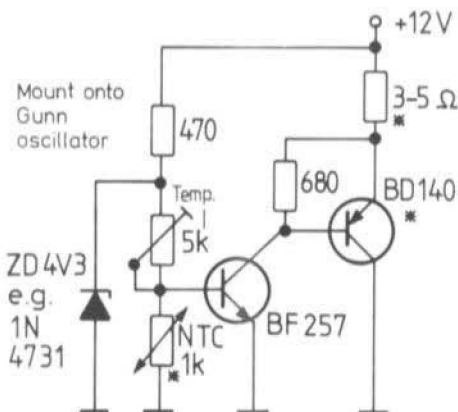


Fig. 6
Circuit for temperature-stabilization of the oscillator

As was mentioned previously, special attention has been paid to a good cooling of the Gunn diode. In the case of 100 mW diodes, the cathode connection to be cooled is provided with a thread which is designated 3-48 UNC-2A. Since such threads can usually not be cut in Europe, the author has used the method shown in **Figure 5** with success. In this case, a M 2.6 thread is cut after which the thread and all cavities up to the cathode connection are filled with a fluid conductive silver that will harden.

In order to improve the frequency stability of the oscillator, the temperature is stabilized using the circuit shown in **Figure 6**. With the

aid of trimmer potentiometer P the temperature is brought to a value above the highest ambient temperature to be expected plus the temperature increase due to the power dissipation of the Gunn diode. Due to the thermally favourable installation of the diode shown in **Figure 5**, a 100 mW Gunn oscillator has been operating at the author's location for a long period without problems both as an oscillator of a transceiver, and as a pump oscillator for a parametric preamplifier for the 5.7 GHz band.

K-band diodes of all types can be used in the described cavity oscillator. The author has tried types TEO 132 (Plessey) and GC-5610 (Frequency Sources/GHz-Division).



Michael Ulbricht, DB 2 GM

A Noise Generator for VHF and UHF

It is very interesting to know how sensitive one's home-made converter, preamplifier or receiver really is. In order to measure this, one requires a noise generator. A very useful, and reproducible noise source is to be described in this article.

There are various methods of generating a noise voltage. The most well-known is to use a saturation diode. A high-vacuum diode with a tungsten cathode generates a noise power that is directly proportional to the diode current. This results in an easy calibration. Such a source was described in (4).

Noise generators can be constructed according to the thermal method using carbon or metal-layer resistors. In this case, one utilizes the

fact that the noise power is directly proportional to the temperature. If a resistor is subjected to various temperatures, it will generate an exactly defined noise power, however, these noise levels are very low.

A simple and cheap method of generating a noise power with the aid of semiconductors was shown in (1). This had the disadvantage that the noise power was not reproducible, which meant that the generator must be calibrated before making measurements. However, the great advantage was the simple construction. In many cases, one does not require an exact noise figure, but only wishes to carry out comparisons to other modules, or to a previous construction. Such a semiconductor noise generator is now to be described.

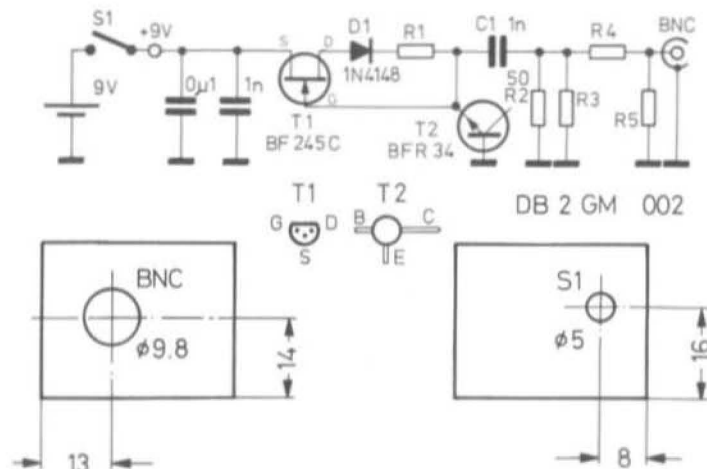


Fig. 1:
The emitter-base diode of a UHF-transistor is the actual noise diode, which is fed from a constant-current source equipped with T 1.
Below: Holes in the case

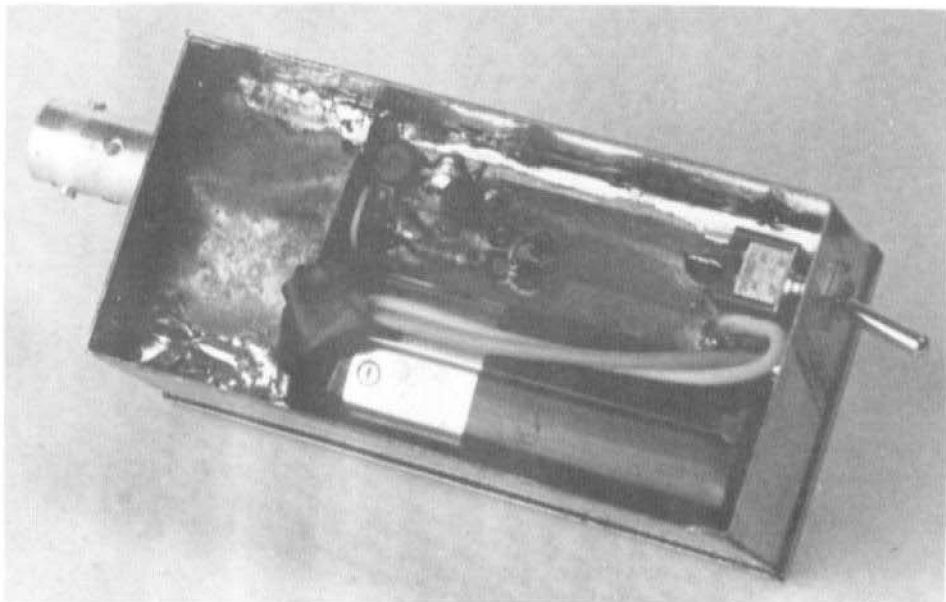


Fig. 2: The photograph shows that this noise source is a real weekend project.

CIRCUIT DESCRIPTION

The base-emitter diode of an RF-transistor (T 2) in **Figure 1** is operated in its blocked direction. The transistor is provided with its operating current via a constant-current source, in order to ensure that the noise power remains constant. This current source is constructed in a conventional manner using a field effect transistor (T 1). The constant current is adjusted with the aid of R 1, and diode D 1 is used for temperature compensation.

Resistor R 1 has a value of 2.2 k Ω in the case of the author's prototype. The noise power did not increase at lower values, and the noise power was correspondingly less at greater values. Since a DC-voltage of approximately 4.8 V is present at the noise diode T 2, a chip capacitor C 1 was provided for DC-isolation.

Attenuator (R 3, R 4, R 5) is provided after the terminating resistor R 2. This allows the noise power to be reduced. However, the main purpose of this attenuator is to improve the matching, since the return loss is improved by the attenuation factor of this attenuator. Of course, an attenuator in a PC-board construction has its limits. The return loss deteriorates at higher frequencies, which means that the matching also deteriorates. For this reason, the author's prototype (see **Figure 2**) is only provided with a 6 dB attenuator using metal-layer resistors. Information given in (2) can be used for calculating other attenuation values. Any further attenuation can be made using a good external attenuator.

An improvement of the return loss resulted using a thin brass or copper plate pressed around the metal-layer resistors of the attenuator. It is possible for this to be soldered to

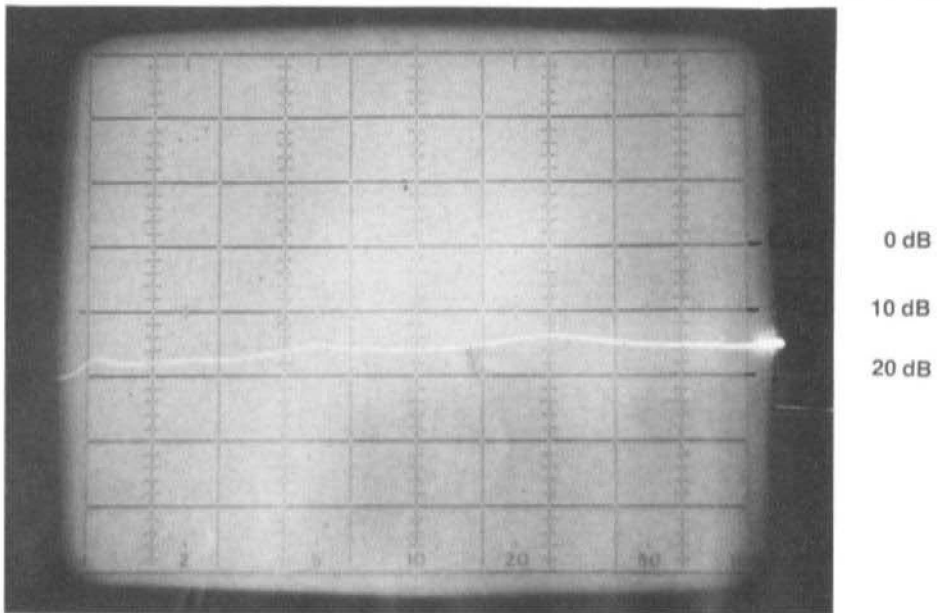


Fig. 3: Return loss between 150 MHz (left edge) and 1.3 GHz measured after the 6 dB attenuator. Value between 18 and 14 dB

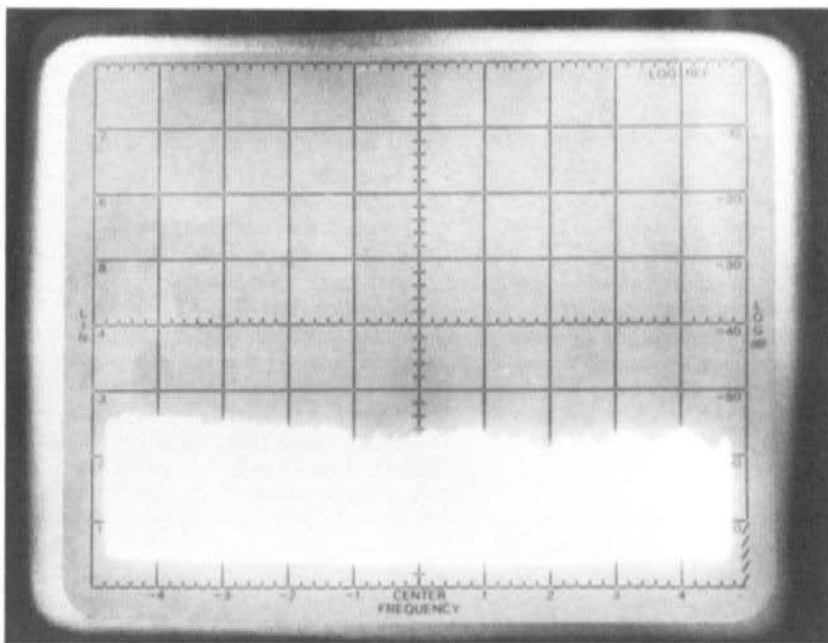


Fig. 4: The ENR-value measured between 50 and approximately 1300 MHz decreases by approximately 3 dB over this frequency range



ground on both sides. The mounting of the plate can be seen in the photograph.

A return loss of 15 dB was obtained using this construction and a 6 dB attenuator (R 3, R 4, R 5) at 1.3 GHz. The matching is better at 145 MHz and 432 MHz, and for those readers only requiring a noise generator in this frequency range, the plate can be deleted.

Figure 3 shows the return loss at the output of the noise generator. The noise spectrum as a function of frequency is given in **Figure 4**.

CONSTRUCTION

Figure 5 shows a small PC-board for construction of the noise generator. It is designated DB 2 GM 002, and its dimensions are 72 mm x 35 mm. It can be mounted in a 30 mm high metal case. The free part of the board should be cut out for accommodating the 9 V battery, and for the protruding part of the BNC-connector.

The components are directly soldered to the conductor side of the board as shown in **Fig. 6**. There are no holes in the board, and the connector leads of the components are directly soldered to the conductor lanes. The outside panel of the case is provided with two holes for the BNC-connector and switch as shown in **Figure 1**.

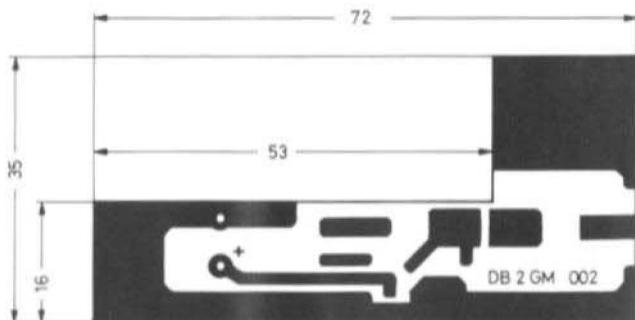


Fig. 5: PC-board DB 2 GM 002 is single-coated and the components are mounted on the conductor side

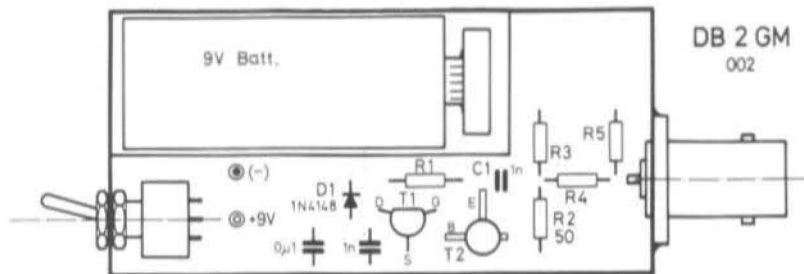


Fig. 6: Component locations showing positions of the switch, BNC-connector, and battery in the case



The BNC-connector is now placed in one half of the case and soldered into position. The PC-board can now be fitted into this half of the case so that the connection pin of the BNC-connector is soldered onto the board in the vicinity of the attenuator. After this, the edge of the PC-board is soldered to the metal case. The covers of the case can be fitted tightly, which means that they need not be soldered into place. This is followed by the rest of the wiring. A 9 V battery is used as voltage supply.

COMPONENTS

T 1:	BF 245 C
T 2:	BFR 34 A
R 1:	see text
R 2:	50 Ω or other impedance

Metal-layer resistors for the attenuator:

3 dB:	R 3, R 5: 294 Ω	R 4: 17.5 Ω
6 dB:	R 3, R 5: 150 Ω	R 4: 37.5 Ω
10 dB:	R 3, R 5: 71 Ω	R 4: 96.0 Ω

C 1: 1 nF chip capacitor
(value uncritical)

Sheet-metal case:
Cover dimensions: 37 x 74 x 30 mm

BNC connector

OPERATION

The noise generator can now be connected to a receiver, and should operate immediately, if wired correctly. The required noise power can be selected with the aid of resistor R 1. It is possible, of course, for the noise source to be calibrated by comparing it to a professional noise generator.

MEASURING THE NOISE FIGURE

The noise power P_n is calculated as follows:

$$P_n = k \times T \times B$$

where: k is the Boltzmann konstant k
= 1.38×10^{-23} Ws/K,

T the absolute temperature in Kelvin,
and

B the bandwidth of the system.

The noise generator produces a noise power of $P = (ENR + 1) \times k \times T \times B$. ENR is the **Excessive Noise Ratio**, in other words the factor that must be determined for the noise generator during the calibration process.

A simple measuring system for noise measurements is shown in **Figure 7**.

The module to be measured is provided with an attenuator at the output. A receiver is used as indicator and is connected to the output of the attenuator. At first, the noise generator remains switched off. The attenuator is now adjusted for a certain indication on the S-meter of the receiver. The noise generator is now switched on, which will increase the indication on the receiver. Using the attenuator, the indicated value is reduced to the same value as was indicated before switching on the noise source. The difference in attenuation that can be read off on the attenuator allows the noise figure of the module to be calculated in the following manner:

$$NF = \frac{ENR}{a - 1} \quad (\text{noise figure}) \quad \text{or}$$

$$T = \left(\frac{ENR}{a - 1} - 1 \right) \times 290 \text{ K}$$

(noise temperature)

where 'a' is the attenuation factor.



Fig. 7: Measuring system for noise-figure measurements



This must be placed as factor into the equation and not in dB. This can be calculated as $a = 10^{a/10}$.

An example is now to be given to clarify the measurement and calculation.

With the noise generator switched off, a readable indication is obtained on the receiver with an attenuation value of 4 dB. After switching on the noise generator, one will require a further attenuation of 8 dB in order to obtain the same indication. This means that the attenuation difference is $a = 8 \text{ dB} - 4 \text{ dB} = 4 \text{ dB}$. This corresponds to an attenuation factor $a = 10^{4/10} = 2.51$. If the noise generator possesses an ENR of 10, the noise figure of the module under test can be calculated as:

$$NF = \frac{10}{2.51 - 1} = 6.6 \triangleq 8.2 \text{ dB.}$$

Even if the ENR-value of the noise generator is unknown, it will still be possible to make comparisons and to establish whether a modification to the module results in an improvement or not.

The larger the difference in attenuation to achieve the same indication, the lower will be the noise figure of the module.

A very good description of all terms in noise technology is to be found in (3).

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A New Generation of Transverters for 1296 MHz

ST 1296/144 A:	2 m IF, output 1 W	DM 699,—
ST 1296/144 B:	2 m IF, output 3 W	DM 775,—
ST 1296/28:	10 m IF, output 1 W	DM 855,—



We would like to introduce our second generation of 23 cm transverters for operation in conjunction with either 10 m or 2 m transceivers.

- Double-conversion both on transmit and receive with the 10 m version to obtain the extremely high image and spurious rejection and clean spectrum.
- Overall noise figure of the receive converter typically 3.9 dB
- Transverters are available in the following versions:





Julian Gannaway, G 3 YGF and
D. Holmes, G 4 FZZ

Some Pitfalls in Noise Figure Measurements

In 1981, G 3 YGF examined virtually all 144 MHz preamplifiers available on the UK market with respect to noise figure, gain, selectivity, large-signal capabilities and matching. These measurements were made with the aid of professional measuring equipment. The results of these measurements were published in RADIO COMMUNICATION (November and December 1981) as »The Effects of Preamplifiers on Receiver Performance, and a Review of Some Currently Available 144 MHz Preamplifiers«. Enquiries and discussions following this article led the authors to write down the experience gained on making these noise-figure measurements, which are now to be discussed in VHF COMMUNICATIONS.

Unfortunately, using expensive measuring equipment does not always mean that one will automatically receive correct measuring results; there are many pitfalls. Users of home-made noise generators must pay even more attention to the sources of measuring errors. This article therefore is of

great assistance to VHF COMMUNICATIONS readers, since a home-made noise measuring system is to be described in a later edition of this magazine.

1. MEASURING SYSTEM

The noise figures of preamplifiers or converters are usually measured in an automatic noise-figure measuring system in conjunction with a calibrated wideband noise source. The leading manufacturers of such equipment are HP and AIL. The latter called its system »Precision Automatic Noise Figure Indicator« (PANFI), and this designation is to be used in this article. Older versions of the PANFI only have a single input frequency of 30 MHz, which means that one or more converters must be used to carry out noise-figure measurements. Such a measuring system is shown in **Figure 1**.

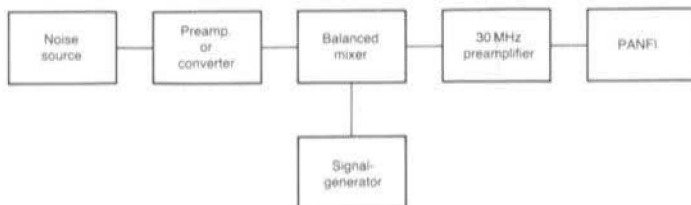


Fig. 1: Most measuring set-ups contain several amplifiers and mixers in front of the PANFI



2. SOURCES OF ERROR

Non-reproducible, unclear or incorrect measured values can be caused by the following:

2.1.

The source impedance (VSWR) of the noise source can change between that present when switched on and switched off. This means that the output noise power of the amplifier to be measured will also change, independently of the noise level from the noise source.

This source of error can be avoided by using an attenuator between noise source and test object. A value of 10 to 15 dB should be sufficient. Of course, the noise-figure value obtained must be corrected for by deducting the value of the attenuation. For this reason, it is necessary to know the value of the attenuation very accurately and for the attenuator to possess a VSWR of better than 1.05 : 1. If differing results are obtained when measuring with and without the attenuator, this will indicate that the VSWR of the noise source is changing. The measured values obtained with the attenuator should therefore be more accurate.

2.2

The ENR-value of even professional noise sources can possess quite a large tolerance in its specifications; a typical value is 0.1 to 0.5 dB. This error limits the overall measuring accuracy of the system and is unavoidable. If an attenuator is used on the noise source as mentioned above, the tolerance of the attenuation value must be added to the tolerance of the ENR-value.

2.3.

The noise components of the second, and any subsequent stages must also be taken into consideration. To do this, the system noise figure at the input of the second stage, and the

gain of the first stage must be known. This noise figure should be fairly low – maybe 3 to 5 dB – and the input matching must be good so that no instability is present or induced within the test object. It may be necessary to provide attenuators with values of between 3 and 10 dB in some cases in order to neutralize them. It should be noted that the noise figure of the second stage may be considerably higher than the value measured in a 50 Ω system if the output impedance of the first stage is not 50 Ω resistive.

2.4.

Any of the preamplifiers or converters may be unstable, or oscillating at a frequency outside of the frequency band of operation. GaAs-FETs and bipolar transistors in VHF/UHF circuits can cause oscillations at several GHz. If a spectrum analyzer or at least a detector is not available for checking this over a very wide frequency range, (between DC and light !) hysteresis effects, fluctuations of the DC-currents or of the noise figure are sure signs of problems. These unwanted effects can be provoked by touching various parts of the circuit.

2.5.

The overall gain of the measuring system may be high enough to cause one or more stages at the end of the chain to be driven into their non-linear range, or saturation. In particular, when one is operating with bandwidths of maybe 10 MHz or more, the wideband noise power can very soon reach considerable values.

Since low-signal amplifiers become non-linear even at an input power level of – 20 or – 30 dBm, and since the thermal noise at a bandwidth of 1 MHz is in the order of – 110 dBm, this results in a relatively small gain range in which the measuring system can operate correctly. The overall gain should be high enough so that the noise level with the noise source switched off is approximately 20 dB over the threshold at which the automatic gain control of the noise measuring system begins to operate. In the case of the PANFI, this is at – 70 dBm, which means that the gain between the



input of the test object and the input of the PANFI may amount to approximately 60 dB. (Editorial note: Even the short-term peak values of the noise power, which are far in excess of the ENR-value should not drive any part of the amplifier chain into the non-linear range. For this reason, approximately 20 dB should be added to the ENR-value when determining the linear operating range).

Any excess gain is better reduced by removing amplifiers than by using attenuators. It is not advisable to use a large number of amplifiers and converters together and hope for the best!

2.6.

The noise figure indicated on the PANFI can be affected by strong local transmissions on any of the intermediate frequencies used, as well as on their image frequencies and spurious responses. Since the PANFI possesses a bandwidth of 5 to 10 MHz, and since most converters have bandwidths of the order of 1 to 10 MHz, interference is possible in a very wide frequency range. One source of interference is the oscillator chains of other converters! One will require a spectrum analyzer with a resolution of several kHz in order to check whether the intermediate frequency is free of other signals. The required examination would be much easier to carry out if the IF bandwidth was only a few kHz, since the frequency band used could be monitored aurally.

The measuring system is especially sensitive to interference at the input frequency of the test object, since the signal level is in the vicinity of the thermal noise level. Any unwanted signal must be approximately 20 dB below this level in order to ensure that the measurement is not falsified. Unfortunately, one is able to receive a surprisingly large number of local signals on most receivers, even with no antenna connected. By comparing two measurements made at slightly different frequencies within the bandwidth of the system, it is possible for such problems to be seen by indicating unexpected differences in the measurement.

2.7.

If insufficient filtering is provided, the mixer can also mix signals with the harmonics of the local oscillator. This possibility is worsened due to the harmonic relationships between the frequencies of many amateur bands, and their use as intermediate frequencies.

The excellent characteristics of ring mixers are often not obtained because they are not terminated with a real 50Ω at all connections and for all frequencies involved. A correct termination is only to be found in very few cases.

In practice, one actually finds converters with very strange spurious responses. For instance, there is a 432 MHz to 144 MHz converter with a spurious response that is only 13 dB down on the required frequency! This is caused by the local oscillator frequency of 288 MHz, whose second harmonic also mixes itself with the signal: $432 - 288 = 144$; $576 - 432 = 144$!

2.8.

The individual mixers, but also some of the tested converters, are just as sensitive in the image frequency range as in the required frequency range. Since noise sources generate wideband noise, a noise figure is measured that is somewhere between the single-sideband and the double-sideband value, according to the filtering. It is therefore essential to use effective bandpass filters for the various required frequency ranges in the measuring system.

2.9.

Most converters produce considerable spurious signal levels at the input and output connectors (in the order of -20 dBm), usually from the local oscillator. If one of these signals at the output of the converter falls within the bandwidth of an IF-preamplifier, it may cause overload effects and reduce the gain for the noise signals.



If the input or the output of a converter is connected to a wideband mixer, one or more of these spurious outputs can compete with the true local-oscillator and result in a large number of spurious reception products and effects.

3. DISCUSSION

The main reason for these difficulties is the fact that the signal levels and frequencies are relatively unknown, especially when a measuring system is to be built up for a special reason such as a VHF/UHF convention where the components are not known, and are being operated in an unknown RF-environment. The behaviour of the test objects with respect to stability, gain, frequency response, spurious reception points, and interference lines also cause a number of unknown variables.

Problems can be avoided by using some additional measuring equipment.

3.1.

With the aid of a spectrum analyzer, one can examine the signal levels of the frequencies in question at the various stages in the chain and discover any presence of instability or RF-breakthrough. At least one should have some form of wideband detector in order to discover the presence of interference frequencies.

3.2.

Gain values and frequency responses can be measured with the aid of a network analyzer or a signal generator and power meter.

3.3.

Effective filtering is absolutely essential in order to avoid problems caused by overload and spurious-reception points. An effective bandpass filter is necessary between every mixer and the subsequent (pre)amplifier, and a further filter at the input of the mixer to attenuate the image frequency.

3.4.

Fluctuating measured values on the PANFI indicate RF-breakthrough or instabilities. Sometimes, these fluctuations are dependent on the movement of persons or equipment varying the RF-breakthrough. Such effects can only be avoided by thoroughly screening all equipment, or by switching off the interfering signal sources.

3.5.

An overload condition of amplifiers is shown by a large increase of the measured noise figure, for instance, when a preamplifier is connected in front of a converter, or a further preamplifier placed in front of the existing system.

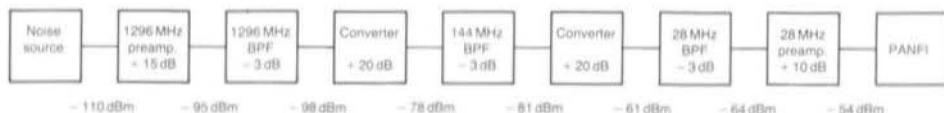


Fig. 2: Noise-figure measuring set-up for the 23 cm band including the necessary filtering, with typical gain values and practical power levels

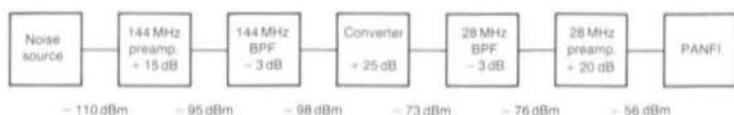


Fig. 3: Noise-figure measuring set-up for the 2 m band;
the indicated power levels are valid for an IF bandwidth of 1 MHz

3.6.

Interfering signals in the required frequency range are indicated when the measured noise figure considerably varies when one varies the frequency of a local oscillator or the intermediate frequency of the PANFI. High-order oscillator harmonics can run all over the whole IF-bandwidth when the fundamental signal is only varied by a few hundred kHz. If everything is operating correctly, the noise-figure values should not vary noticeably over the bandwidth of the IF, even when this is several MHz.

4. FINAL CONCLUSIONS

It is recommended to use a special converter designed for each required frequency band in order to avoid problems encountered with wideband mixers. Even then, sufficient selectivity should still be provided before and after

each mixer; this can be built in, or added in the form of external filters. Suitable set-ups are shown in **Fig. 2** for measurements on 1296 MHz, and in **Figure 3** for 144 MHz.

It was previously mentioned that the PANFI requires an input level of at least -70 dBm, (AGC actuation) and a maximum input level of approximately 0 dBm. The amplifier/converter chain should provide a noise level that is not much more than 10 to 20 dB in excess of this AGC-threshold when the noise source is switched off. This means that a maximum range of 50 to 60 dB remains for the ENR value (approximately 5 - 15 dB according to the attenuator value used), and the peak values of the noise power.

Noise-figure values can only be trusted when the previously mentioned checks have been successful and when the measuring system provides constant, reproducible results with a number of different preamplifiers. Any unusual or unexpected behaviour should make one suspicious and lead to a careful, systematic examination.



Leif Åsbrink, SM 5 BSZ

Dynamic Range of 2 m Transceivers

Part 1: Introduction

The satisfaction of DX communications on the 144 MHz band is often spoiled nowadays due to interference from neighbouring stations. The cause is usually an inadequate dynamic range of transmitters and receivers! The greatest problems with modern equipment are caused by the noise sidebands of the local oscillators. They decrease the dynamic range in both receivers and transmitters; often to a very great extent. These problems were discussed in an article published by DJ 7 VY (1), which also included a description of a good, clean VFO.

The following article represents an introduction into a series of articles in which

simple modifications are to be described for popular 144 MHz transceivers available on the market. These modifications will result in a considerable improvement of their dynamic range.

Basic Considerations

Table 1 gives three typical situations to indicate how high the dynamic range, and how low the noise sidebands must be if 2 stations are to operate simultaneously. The worst case is assumed where the two antennas are pointing directly to another.

Output power (W)	Antenna (Element/gain)	Distance (km)	Dynamic range (BW = 3 kHz)	Noise sideband suppression (dB/Hz)
300	40/17.5 dB _D	2	149 dB	- 184
50	10/11 dB _D	5	120 dB	- 155
10	Halo/- 1 dB _D	5	89 dB	- 124

Table 1:
Required dynamic range and noise sideband suppressions for three typical situations



The values given in Table 1 show us that a very large dynamic range is very useful, and the characteristics of the equipment should be optimized as much as possible within reasonable cost limits. The situation with modern equipment is usually opposite and the dynamic range is usually limited by lack of care during the design of the circuit.

In this article, and the following descriptions, the dynamic range will always be referred to a SSB bandwidth of approximately 3 kHz, i.e. the values in Table 1, fourth column, can be directly compared to the diagrams. The importance of the numerical values can be easily understood in the case of the transmit spectrum; however, in the case of the receivers we differentiate between the two-tone dynamic range, and the three-tone dynamic range.

The two-tone dynamic range is measured as the level of an interfering signal that reduces the signal-to-noise ratio of a required weak signal by 3 dB. In this case, the interference signal level is measured in dB over the noise floor (at a bandwidth of 3 kHz). In modern receivers, the two-tone dynamic range is usually limited by the noise sidebands of the local oscillator. It can, however, be determined by the overload characteristics of any amplifier or mixer, and will, in this case, be in the vicinity of the 1 dB compression point (in dB over the noise floor).

The three-tone dynamic range – or intermodulation-free dynamic range – is related to the third-order intercept point, and is measured as the level (in dB above the noise floor) of one of two equally strong signals that produce intermodulation products of the same level as the noise floor.

In the above discussion, all signals are assumed to be within the 144 MHz band. Image-frequency response, second-order intermodulation (e.g. 48 MHz TV-signals plus 96 MHz FM carriers), and other problems caused by out-of-band signals, are not to be discussed here, since they can be easily suppressed using suitable filters.

The two-tone dynamic range may be very low at certain frequencies due to spurious reception points, however, this is a minor problem that is not usually present in modern 144 MHz transceivers, and is therefore not to be discussed here.

Finally, the three-tone dynamic range is not to be discussed in detail here, for the following reasons: In practice, the interference caused by intermodulation is not so severe since at least two, very strong signals must be present simultaneously; the interference caused in this manner only appears at a limited number of frequencies, which means that a skilled operator can avoid them.

A receiver having a good two-tone behaviour, will automatically possess a good three-tone range, whereas the opposite may not be true due to the noise sidebands.

The intercept point and its importance has already been discussed in detail in amateur radio publications.

Request specifications regarding the dynamic range before purchasing your next transceiver!

The aim of this series of articles is to improve the very unsatisfactory situation. This information has been partly published in the Swedish Magazine »Radio och Television«. The author wishes to make radio amateurs aware of the fact that there are very great differences in the dynamic range between different transceivers. If more amateurs request dynamic range specifications, the manufacturers will be forced to put more effort into these problems if their equipment is to remain popular on the market. The manufacturer that is able to quote a transmit signal purity of – 100 dB in 25 kHz spacing, and – 110 dB in 100 kHz spacing (bandwidth 3 kHz), and to quote the same numerical values for the two-tone dynamic range of the receiver, would have a considerable advantage over other manufacturers on the market.



Measured values indicate the present standard of transceivers

The costs involved in achieving the previously mentioned dynamic range are negligible. This will be proved in the following articles of this series. In order to illustrate the present situation, the author would like to show the results of measurements that were mainly made at the VHF-meeting held in Annaboda in 1981. The illustrations show the signal purity of transmitters in the CW-mode.

Figures 1 to 4 show typical measured values for types FT 225 RD, FT 221, IC 211/245, and TS 700 in that order, both in their original condition (upper curve), and after modification (lower curve). The modifications will be described in subsequent articles of this series.

In the case of transceivers IC 260 E/IC 251 E, the measured values spread considerably from model to model; **Figure 5** shows four of these. The reason for this spread has not been examined. Furthermore, no experiments were made, to improve this type of transceiver. The run of the two upper curves already indicates a poor filtering in the phase-locked loop.



Fig. 1:
Vertical scale = power difference to the main carrier;
Horizontal scale = frequency difference to the carrier.
Noise sidebands from the FT 225 before (above) and after (below) modification



Fig. 2:
As in Fig.1 – in this case for the FT 221

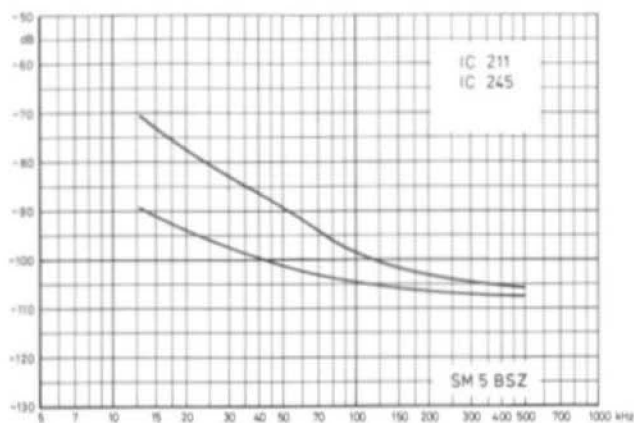


Fig. 3:
As in Fig.1 – now for the
IC 211/IC 245

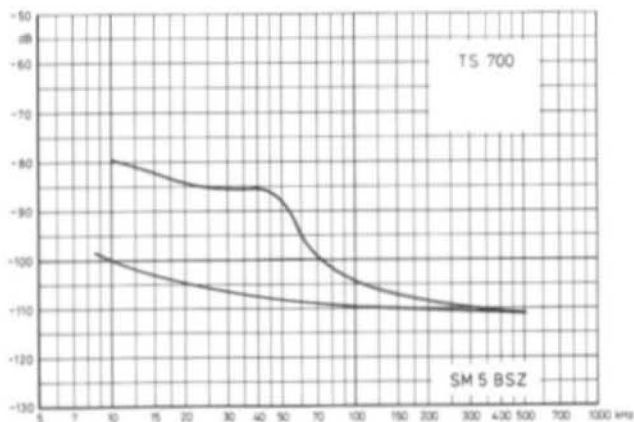


Fig. 4:
As in Fig.1 – this is for
the TS 700 (see part 2
of this series)

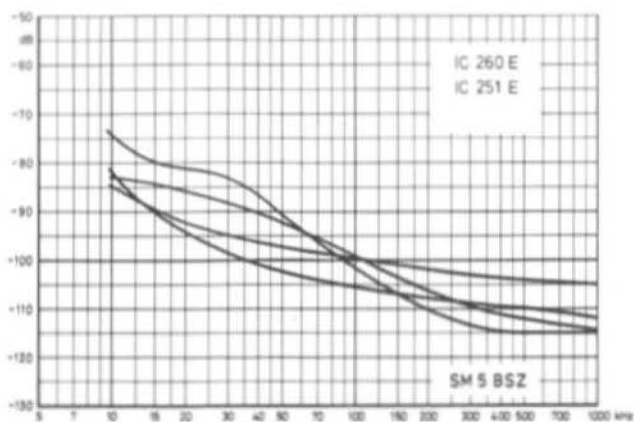


Fig. 5:
As in Fig.1 – measured
values of 4 different
IC 260 E/IC 251 E
exhibit a spreading
of up to 20 dB



Figure 6 shows the measured values for two different transceivers type FT 480 R. In the case of these two transceivers, the noise is caused by too low a signal level at the output of the crystal filter. In principle, this should be easy to modify, but due to the compact construction, this may be difficult in practice.

Figure 7 shows measured values of two different TR 2400. This transceiver is clearly poorly designed and constructed, and the manufacturer is well advised to improve this unit as soon as possible.

Typical measured curves are given in **Figure 8** for transceivers type IC 202 and IC 2 E. A large number of these transceivers were measured,

and all showed values very near to these curves. **Figure 8** also includes values for transceiver type IC 22; however, only one unit was available for the measurement.

Figures 9, 10, and 11 are also only based on one transceiver per type. It is therefore unknown to what extent the measured values spread from unit to unit.

All measurements were carried out with a home-made transceiver having a very large dynamic range. **Figure 12** shows the performance of the measuring system, that is obtained using a IC 202 followed by a 144 MHz crystal filter.

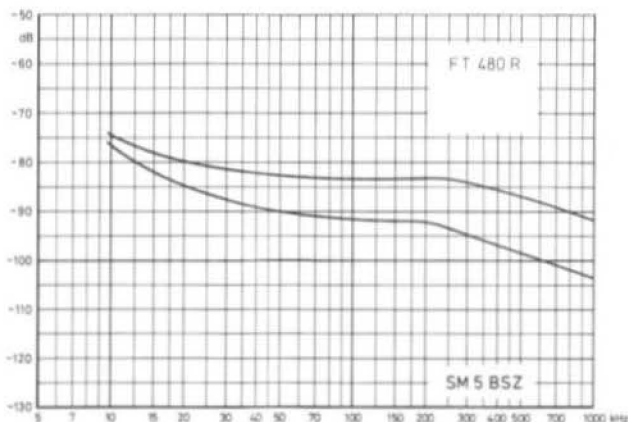


Fig. 6:
As in Fig. 1 –
two different transceivers
type FT 480 R in their
original state

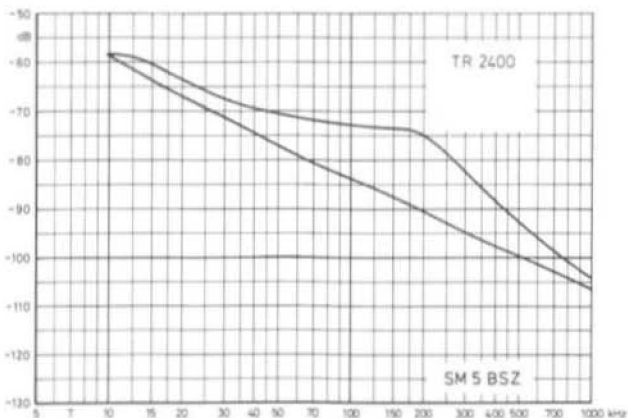


Fig. 7:
As in Fig. 1 –
two different transceivers
type TR 2400 in their
original state

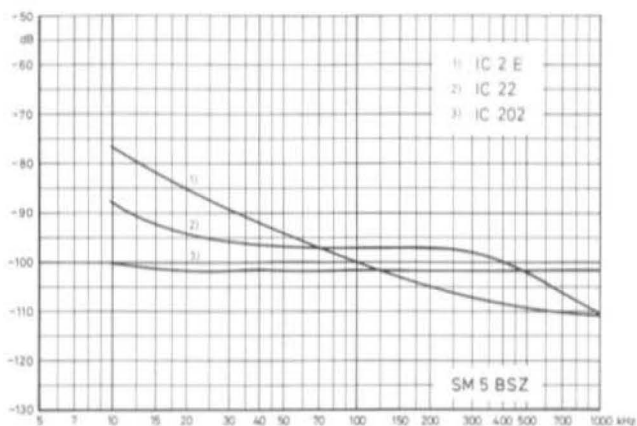


Fig. 8:
The noise sidebands of many individual transceivers of types 1 and 3 exhibit always the same characteristics

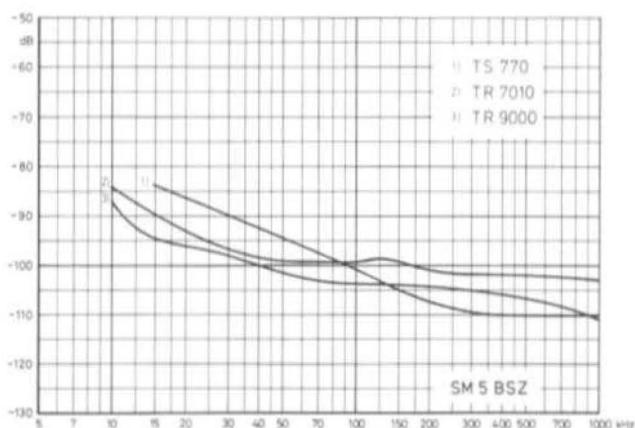


Fig. 9:
The noise sidebands of 3 further types

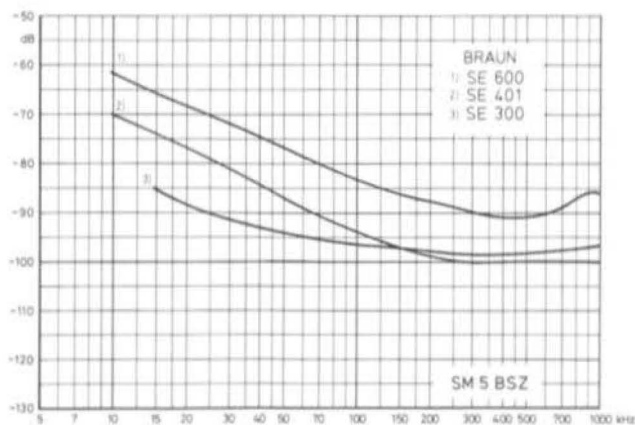


Fig. 10:
Noise sidebands of 3 well-known BRAUN-transceivers



Fig. 11:
The noise sidebands
of type 1 are
remarkably high

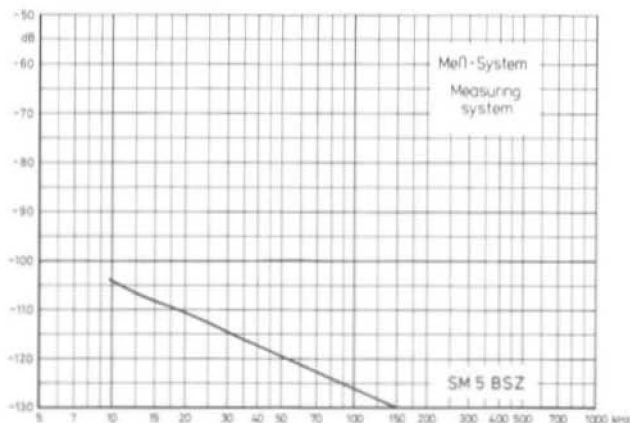


Fig. 12:
An IC 202 with a
144 MHz crystal
filter connected
to the output was
used as the beat-
frequency oscillator

In general, the two-tone dynamic range of the receivers is similar to the measured values for the purity of the transmit signal. However, measurements were only made on certain types.

Additional interference may be produced when the transmitters are modulated. Several models produce very strong key clicks, and others produce a strong wideband splatter in the SSB-mode. Generally, such problems can be cured easily and suitable modifications will be described for certain transceivers.

Finally, the importance of a clean transmit signal, and good receiver should be underlined by studying column 4 of Table 1. The number of radio amateurs will most certainly

increase – although the situation in the densely populated areas is bad enough today, which is understandable when one compares the values given in Table 1 with the measured values of modern transceivers, as given in the diagrams of this article.

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VHF COMMUNICATIONS
Edition 2/1981, pages 66-82



Leif Åsbrink, SM 5 BSZ

Dynamic Range of 2 m Transceivers

Part 2: Modifications to the TS 700

The noise sidebands that accompany the transmit signal of the TS 700 are relatively strong at a spacing of ± 100 kHz or less. They originate in the DC-converter of the power supply circuit. The converter produces a 600 Hz square-wave signal. The steep slope results in spikes that cause an amplitude modulation of the transmitter, which in turn increases the signal bandwidth to ± 100 kHz.

The modifications described in this article ensure that the 600 Hz signal cannot reach the transmitter, thus eliminating the amplitude modulation and the interference sidebands.

Figure 1 shows the noise sidebands of some TS 700 transceivers before modification (continuous lines), and after modification (dashed lines).

The curves designated with »A« are from an older unit, which was defective during the first measurement (A 1). The cause for this were loose nuts used for mounting the board in the power supply. After tightening the four nuts, the transceiver exhibited normal, typical values (A 2). After publishing this article in »Radio och Television«, the author received a number of letters that indicated that many such transceivers have this problem.

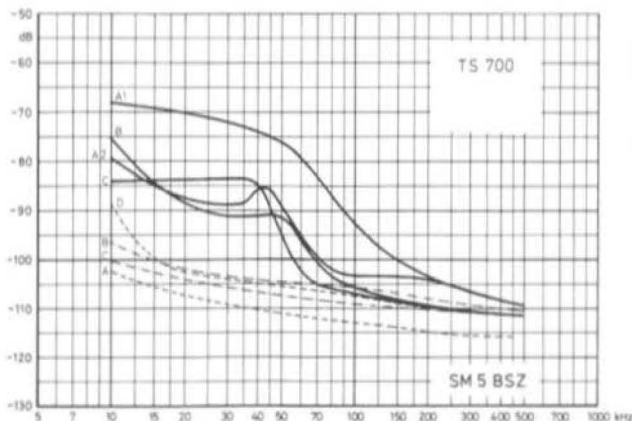


Fig. 1:
In the TS 700 the noise sidebands are caused by 600 Hz pulses from the DC converter. Massive lines = 4 different transceivers in their original states; broken lines = after modification



1. The first step in improving the interference sidebands of the TS 700 is thus simple: Tighten these nuts so that the power supply has a good ground connection.

Curves B and C are from transceivers of more modern construction. Transceiver D was only measured after modification; it is not known whether it is an older or newer model.

Very high currents of the 600 Hz spikes flow within the power supply. The selection of the ground points is therefore very critical. The ground points of the 9 V stabilizer have not been carefully selected, and for this reason, the second and third steps of the modifications are in improving the ground connections of the

9 V stabilizer circuit. These are carried out as follows:

2. The ground pin of the integrated stabilizer (Q 11, pin 1) is connected to one of the previously mentioned nuts by connecting a copper wire of at least 2.5 mm diameter from the upper side of the IC to the nut, and soldering it into place. That nut should be used that is nearest to the 144 MHz power amplifier.

3. The ground end of R 28 is now connected to pin 1 of Q 11. The upper end of R 28 is now cut and a new 1 k Ω resistor connected from the loose end of R 27 to pin 1 of Q 11 (from above).

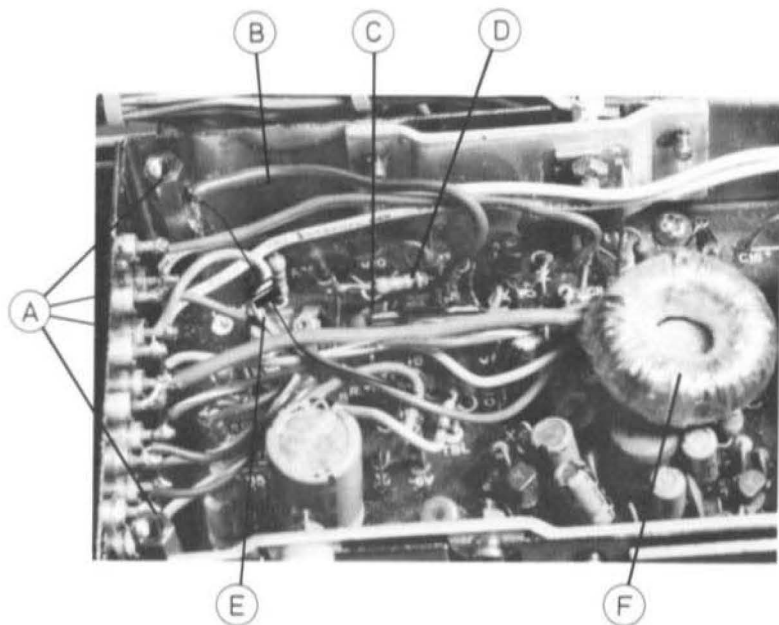


Fig. 2: These are the modifications:

- A: Tighten all 4 bolts
- B: Solder a heavy wire from the bolt in the upper left-hand corner to pin 1 of Q 11
- C: Cut the old resistor
- D: Put in the new 1 k Ω resistor
- E: This is where the 220 μ F electrolytic capacitor finds its place
- F: The toroid-core choke



Ferdi Schmehr, DC 8 EC

Stabilizing the Operating Point of Preamplifiers and Linear Amplifiers for an Output Power up to approx. 1 W

The described circuit maintains the collector current of an RF-transistor virtually constant in a wide temperature range from -50°C to $+100^{\circ}\text{C}$.

The advantages of this circuit are:

- No thermal contact is required to the RF-transistor
- Collector current and voltage are variable separately in a wide range.

The RF-transistor is shown in the circuit diagram without resonant circuits and transformation links. For safety, the connections of base and collector voltage are bypassed with the aid of capacitors C 1 and C 2 whose value is dependent on the frequency range of the

stage to be stabilized. This ensures that no RF-noise voltages can be injected from the control stage into the RF-stage.

A PNP-audio transistor type is used as stabilizer transistor T 1. The recommended type 2 N 2905 A (TO-5-case) allows the operating point stabilization of RF-amplifiers up to 1 W output. Attention must be paid, however, to ensure that the rating of resistor R 3 is sufficient. The power dissipation across R 3 is calculated as follows:

$$P_{R3} = \frac{(U_B - U_{CE2})^2}{R_{3a} + R_{3b}}$$

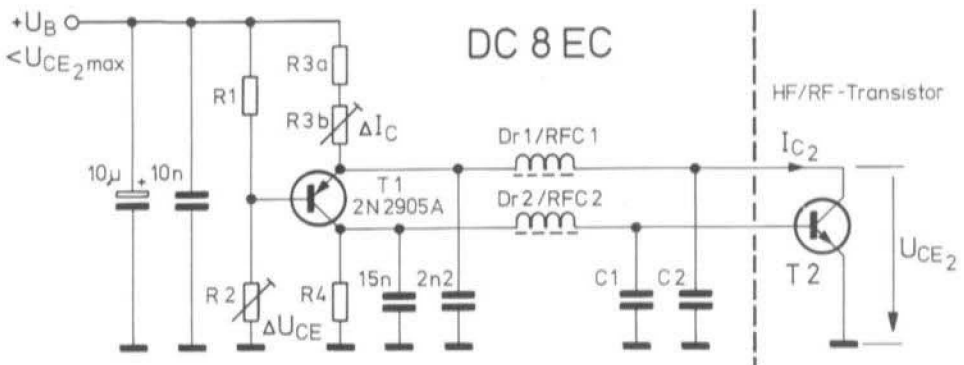


Fig. 1: Operating point stabilization of an RF-stage between -50 and $+100^{\circ}\text{C}$



The collector emitter voltage of the RF-transistor can be adjusted with the aid of trimmer resistor R 2, and the collector current with the aid of R 3.

To calculate all four resistance values approximately, the following values are required:

- Operating voltage U_B *)
- Required collector voltage of the RF-transistor: U_{CE_2}
- Required collector current of the RF-transistor: I_{C_2}
- Current gain of control and RF-transistor: B_1 and B_2
- Base-emitter voltage of both transistors for the collector current: U_{BE_1} and U_{BE_2}

*) With some limitation, the following is valid for the operating voltage: It should not be higher than the maximum permissible collector-emitter voltage of the RF-transistor ($U_{CE\ max}$ of T 2).

It is then possible to use the four following equations:

$$R\ 1 = \frac{[U_B - (U_{CE_2} - U_{BE_1})] \times B_1 \times B_2}{110 \times I_{C_2}}$$

$$R\ 2 = \frac{(U_{CE_2} - U_{BE_2}) \times B_1 \times B_2}{121 \times I_{C_2}}$$

$$R\ 3 = \frac{U_B - U_{CE_2}}{I_{C_2} \times [1 + 11/B_2 (1/B_1 + 1)]}$$

$$R\ 4 = \frac{U_{BE_2} \times B_2}{10 \times I_{C_2}}$$

The values for a transistor HXTR-6101 (HP) are to be given as an example. This transistor is to be used as a low-noise preamplifier in a receiver for the 13 cm band.

The calculation is made from the following values obtained from the data sheet or estimated:

$$U_B = 15\ V, \quad U_{CE_2} = 10\ V, \quad I_{C_2} = 4\ mA,$$

$$B_1 = 100, \quad B_2 = 150, \quad U_{BE_{1,2}} = 0.7\ V.$$

This results in the following resistor values:

$$R\ 1 = 194\ k\Omega \quad R\ 2 = 287\ k\Omega$$

$$R\ 3 = 1.2\ k\Omega \quad R\ 4 = 2.6\ k\Omega$$

and $P_{R3} = 22\ mW$

Finally, attention must be paid, since the RF-transistor T 2 can be destroyed when disconnecting its base line !

It is, of course, clear that each RF-transistor requires its own stabilizer circuit.

REFERENCES

- (1) Hewlett-Packard Application Note 967
- (2) D.E.Schmitzer, DJ 4 BG:
Stabilizing the Operating Point of Transistors with Directly Grounded Emitter
VHF COMMUNICATIONS 9,
Edition 2/1977, pages 100-103

MATERIAL PRICE LIST OF EQUIPMENT

described in Edition 1/1982 of VHF COMMUNICATIONS

Art.No.	DJ 7 VY 005	Wideband Amplifier with 44 dB Gain between 1 and 30 MHz	Ed. 1/1982
6702	PC-board	DJ7VY 005 Double-coated and etched, undrilled	DM 29.—
6719	Semiconductors	DJ7VY 005 1 x P 8002, 1 x BFW 16 or 2 N 5109, 1 x VN 88, 1 Diode	DM 37,50
6720	Parts for	DJ7VY 005 3 x toroid core R10-N30 with wire, 11 different chokes, 17 different capacitors, 3 feedthrough caps, 1 trimmer pot., 1 relais, 2 BNC sockets, 1 casing with heat sink, 1 cooling fin for T 2	DM 79,50
6721	Kit	DJ7VY 005 complete with above parts	DM 142.—
	DJ 1 JZ 004	METEOSAT-GOES-Converter – Local Oscillator Module	Ed. 1/1982
6710	PC-board	DJ1JZ 004 Double coated and etched, silver coated	DM 29.—
6711	Parts	DJ1JZ 004 8 transistors, 2 diodes, 3 coil formers with slug, 5 different wires, 1 ferrite bead, 7 foil caps, 1 ceramic trimmer, 2 feedthrough caps, 1 chip cap., 1 tantalum- and 28 ceramic caps, 22 resistors, 1 relais, 30 cm PTFE-coax cable, 1 pc. of insulating tubing, 1 casing	DM 93.—
6712	Crystal	97.3125 MHz KVG: XS 2101	DM 34.—
6713	Crystal	97.09375 MHz KVG: XS 2101	DM 34.—
6714	Kit	DJ1JZ 004 complete with the above parts	DM 185.—
6715	Kit	METEOSAT converter consisting of DJ1JZ 003 and 004 – packet-price:	DM 360.—
6565	Low-noise preamplifier DJ6PI 010, as described in Ed. 1/1980, 13 cm band version, aligned for 1700 MHz – special price:		DM 225.—
<hr/>			
Semi-rigid cable SR-3	50 Ω, VF = 0.695; diam. 3.58 mm	per cm	DM –,50
N-connector, socket	50 Ω		DM 6.80
N-connector, plug	50 Ω for cable type RG-213		DM 6.90
N-connector, plug	50 Ω for cable type RG-58		DM 6.90
Parabolic antenna 1.1 m diam.	12 segments prepared for riveting or screwing as described by DC 3 NT in Edition 3/1979		
	Set of 12 segments, already drilled		DM 180.—
	Riveting machine and rivets		DM 93.—
1.7 GHz Tubular radiator for above antenna:			
	Tube, lid, N-connector, radiator, screws		DM 90.—
	Ready-to-operate		DM 150.—
Set of drawings for the DC3NT FAX-machine, described in Edition 1/1981 of VHF COMMUNICATIONS			DM 8.—

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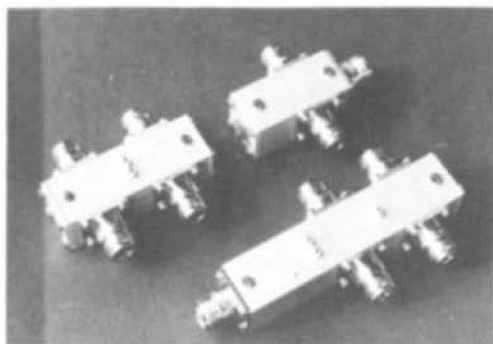


Fig. 1

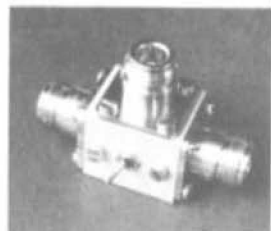


Fig. 2

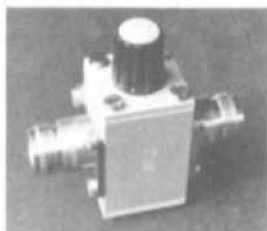


Fig. 3

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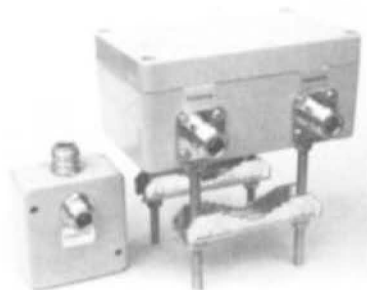
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Accuracy: 0 to + 0.1 x 10⁻⁶ + 20 to 35°C
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Frequency drift: \leq 0.1 x 10⁻⁶/month
Readout: 9 red digits
Power supply: 9-14 V / 120 mA to 400 mA
Dimensions: (mm): 215 x 85 x 230
Weight: 800 g
Price: DM 825.—

Accessories:

Telescope ant. T 1200 DM 33.—



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Frequency ranges

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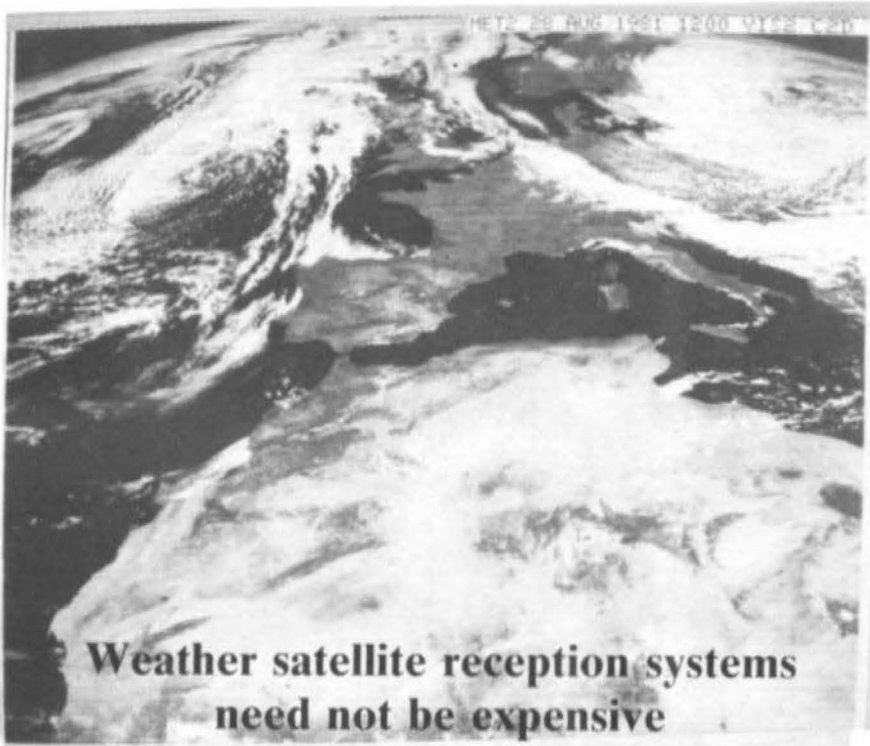
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1 M Ω /30 pF 50 Ω /25 pF

Time base: 3.57954 MHz TCXO
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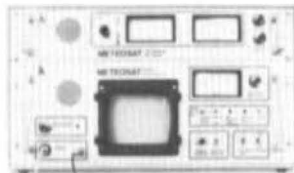




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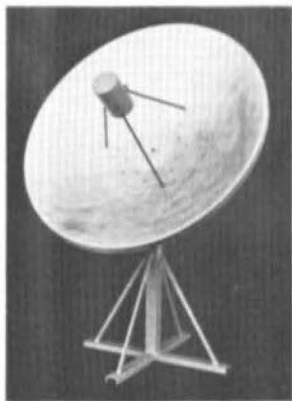
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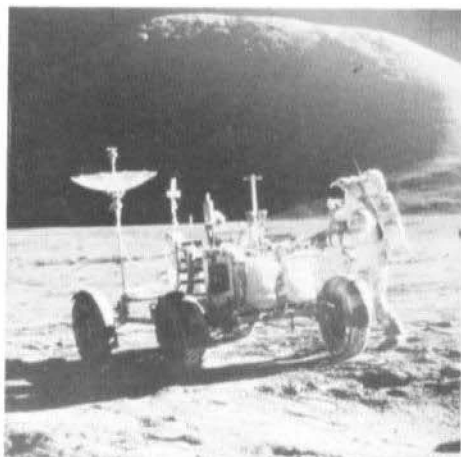
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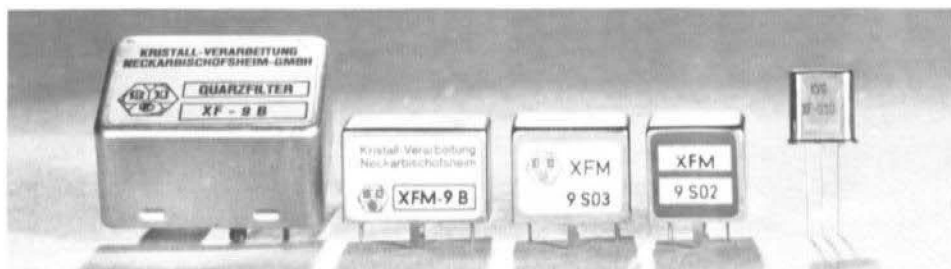
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