



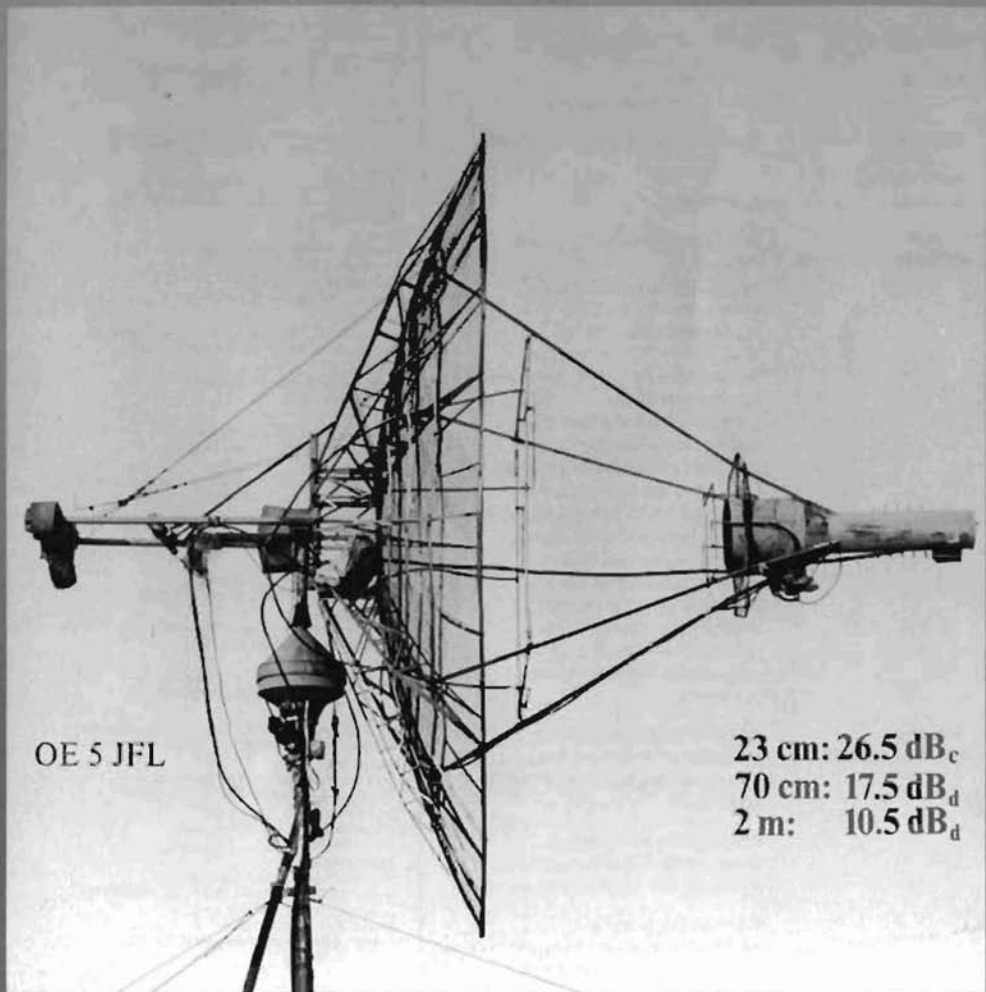
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*A Publication  
for the Radio-Amateur  
Especially Covering VHF,  
UHF and Microwaves*

VHF

# *communications*

Volume No. 20 · Winter · 4/1988 · DM 7.50



OE 5 JFL

23 cm: 26.5 dB<sub>c</sub>  
70 cm: 17.5 dB<sub>d</sub>  
2 m: 10.5 dB<sub>d</sub>



# VHF communications

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

Volume No. 20 · Winter · Edition 4/1988

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**Publishers:** TERRY BITTAN OHG

**Editors:** Corrie Bittan  
Colin J. Brock (Assistant)

**Translator:** Colin J. Brock, G 3 ISB / DJ O OK

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Verlag UKW-BERICHTE, Terry D. Bittan  
POB 80, D-8523 BAIERSDORF / W. Germany  
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Phone 5285962.

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Terry Bittan, Schweiz, Kreditanstalt ZÜRICH,  
Kto. 469.253-41, PSchKio ZÜRICH 80-54.849

Leo Kälin, HB 9 CKL, Funktechnik  
Ahe Landstr. 175, CH 8708 Männedorf  
Tel. 01-9203535

### United Kingdom

Mike Wooding, G 6 QIM, 5 Ware Orchard  
Barby, nr. Rugby, Warks CV23 8UF, Tel. 0788 890385

### USA

Timekit, P.O. Box 22277,  
Cleveland, Ohio 44122, Phone: (216) 464-3820

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## DSP Techniques for Radio Amateurs by Matjaž Vidmar, YT 3 MV

Unfortunately, it was not possible till now to publish the continuation of this interesting project, but it will be included in the next editions. Two other articles are planned about principles and interesting circuit details for a suitable computer, and constructional details with component plans.

At this opportunity, the staff of VHF COMMUNICATIONS would like to wish all readers a **Merry Christmas** and a **Happy and Prosperous New Year!**





John Wood, G3 YQC

## FM Television for the Amateur

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### 1. INTRODUCTION

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In recent years there has been a considerable move towards FM for transmitting television pictures in the amateur UHF and microwave bands. This has not been possible on 70 cm because of the restricted band space available.

FM offers several distinct advantages over the more traditional AM system: –

- a) Modulation takes place at a low level, usually by applying the signal to a varactor diode in the transmitter carrier oscillator.
- b) RF amplifiers for FM do not need to be linear, therefore they can work at higher efficiencies.
- c) There can be a substantial improvement in signal-to-noise ratio at the receiver over that obtained from a similar strength AM signal
- d) Interference to broadcast radio and TV transmissions is markedly reduced since frequency modulation is not demodulated by the usual AM rectification method. Consequently

interference to AM receivers, tape recorders, record players and deaf aids is very much reduced and, in many cases, eliminated.

Of course there are disadvantages too –

- a) The channel bandwidth can be much greater than that required for AM.
- b) The receiver demodulator is more complex to build and align.
- c) The use of readily available broadcast television receivers is limited due to the lack of an FM demodulator.

It is interesting to note that the FM mode is exclusively used by television broadcasters when operating through satellites.

FM television for amateurs has become popular in recent years – particularly in the 1.3 GHz band – due in no small part to the introduction of several successful TV repeater stations throughout the U.K.

Because FM is a relatively new mode for television amateurs, there have been many mistakes, misconceptions, theoretical inaccuracies and a general lack of understanding of the mode. These have been propagated by word-of-mouth and through the amateur press. The BATC's Techni-



cal Liaison Committee has, however, recently addressed itself to the subject and, after much discussion, deliberation and practical experimentation, has been able to propose a set of standards, recommendations and guidelines for all important parameters. These are explained in this article and much useful advice is given to those setting up an amateur FM-TV station for the first time, and for those who constantly strive to improve their stations by the application of new information and techniques as they become available.

### 1.1. Theoretical Definitions

Many misconceptions and theoretical inaccuracies amongst amateurs have arisen because of the difficulty in relating one's thinking to FM and trying to link it in some way to AM. For example: a 6 MHz wide video signal, when amplitude modulated onto an RF carrier, will occupy a channel bandwidth of 12 MHz. The same signal, when frequency modulated onto the same RF carrier, can produce a channel bandwidth less than 12 MHz or, as is more likely, considerably more. The point being that FM bandwidth is not strictly related to the bandwidth of the modulating signal as might be supposed.

It is useful to spend time in examining the more important fundamental parameters of an FM system in order that the mind can grasp the concepts and examples put forward in this article. Additional theory will be introduced as required later.

Frequency modulation is produced when the carrier oscillator's frequency is changed by a modulating signal. The amount of frequency change is directly proportional to the amplitude of the modulating signal and is called DEVIATION. We can therefore say that:

**PEAK CARRIER DEVIATION ( $\Delta f_p$ )** is the largest deviation of the carrier from its average frequency and is caused by the maximum amplitude excursion of the modulating signal in either direction.

Any AC signal is said to have a peak-to-peak amplitude. A black and white video signal's peaks are at peak white and peak sync levels. If this

signal is used to frequency modulate an RF carrier, its frequency will be a maximum during peak white, minimum at peak syncs (sync tips) and at an average somewhere in between. When the RF carrier frequency moves from its average to either of its peak values, it is said to have reached **PEAK DEVIATION**. The total frequency change from the positive to negative peak deviation is called **PEAK-TO-PEAK DEVIATION** and is usually around twice the peak deviation.

**HIGHEST MODULATING FREQUENCY ( $f_m$ )** is the maximum frequency of any component within the modulating frequency band. This will usually be the upper 3 dB limit of the video bandwidth. In practice, most television pictures will produce highest instantaneous frequencies considerably less than maximum. Highest frequencies are generated only during very sharp transitions between black and white, such as the highest set of multiburst definition bars.

**MODULATION INDEX (M)** is the ratio between the peak carrier deviation and the highest modulating frequency

$$\text{Modulation index } M = \frac{\Delta f_p}{f_m}$$

This is a very important parameter and can be likened in some ways to the modulation depth of an AM signal. Note that peak carrier deviation is used and not the peak-to-peak value. The highest modulating frequency is usually taken to be the highest frequency which will be encountered during transmissions. In actual fact, the modulation index is different at any instant during transmission because the video frequency and amplitude (and hence peak deviation) is continually changing. Modulation index therefore is usually quoted as the highest or 'worst-case' figure which will ever be achieved.

**CHANNEL BANDWIDTH (B)** is the amount of frequency spectrum in MegaHertz required to accommodate the FM television signal. We are usually concerned with the maximum bandwidth which will ever be required to accommodate the signal, in practice, for 'normal' television scenes used by amateurs, the bandwidth required is usually considerably less.

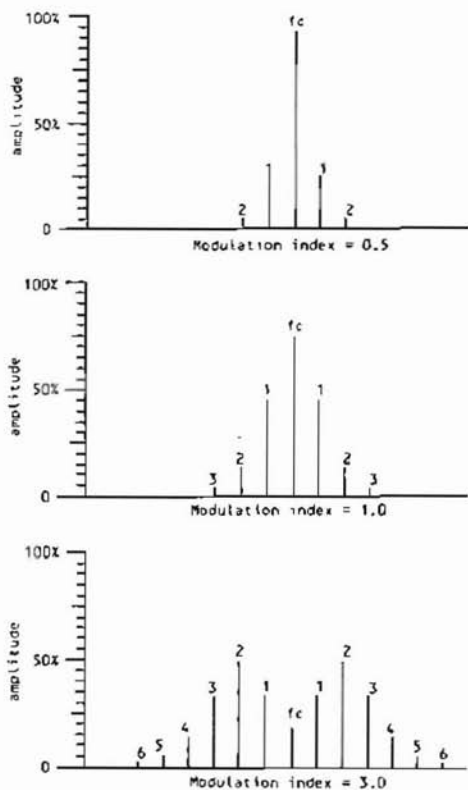


Fig. 1 Relative carrier and side frequency levels for three different modulation indices.

Fig. 1: Relative carrier and side frequency levels for three different modulation indices

Fig. 1 shows the spectrum of FM transmissions with different modulation indices. For simplicity it is assumed that the modulating signal is a single 1 MHz sine wave. It can be seen that the spectrum includes the carrier and a number of side frequencies spaced 1 MHz apart. The amplitudes of the side frequencies change with the modulation index and at higher indices a larger number of side frequencies become significant.

Amateurs use a modulation index of around 0.5 and fig. 1 shows that in this case only one set of side frequencies is significant, the remainder being 50 dB down or more, relative to the amplitude of the unmodulated carrier

If the 1 MHz sine wave is replaced by a television waveform, the spectrum becomes very complex indeed. Calculation of channel bandwidth therefore becomes a very advanced exercise in higher mathematics and not something which could be attempted by most amateurs (or indeed the author!). Fortunately, a fellow, called Carson, has simplified the calculations for us and has produced 'Carson's Rule'.

Carson's rule provides an approximation of the channel bandwidth required for a carrier which is frequency modulated by a complex television signal:

$$\text{Channel bandwidth } B = 2(\Delta f_o + f_m) \text{ MHz}$$

Note that  $f_m$  is the instantaneous modulating frequency, it is usually taken to be the highest possible modulating frequency since this will give a maximum or worst-case figure. In practice, the channel bandwidth is likely to be considerably reduced due to scenes having low lighting levels or without a significant number of sharp black/white transitions.

Carson's bandwidth is calculated such that all out-of-channel signals are considered to be insignificant. This method of determining channel bandwidth is most quoted and used by commercial broadcasting organisations and by operators of wideband satellite communication links.

## 2. CONSIDERATIONS OF SIGNAL PARAMETERS

### 2.1. Video Bandwidth

Broadcast quality signals generally require a video bandwidth of the order of 5.5 MHz. The upper limit is determined principally by the requirement to reproduce sharp transitions between black and white. In a video signal, these transitions produce a sharp stepped waveform and restriction of HF bandwidth will tend to lessen the steepness of the step and reduce the definition when viewed on a television screen (fig. 2)

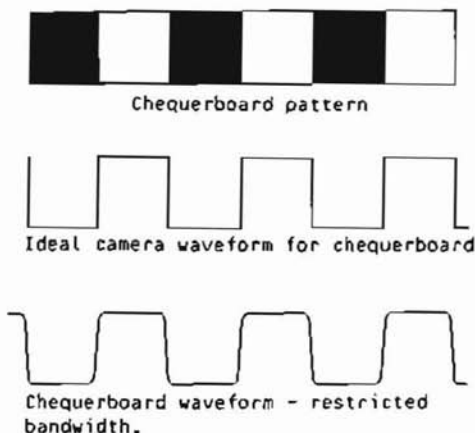


Fig. 2: Effects of bandwidth on a chequerboard

A further consideration is the need to accommodate a colour subcarrier on a frequency of 4 433618 MHz.

Modern amateur video equipment often approaches broadcast quality and can include commercially available video generating and monitoring instruments. The use of computers and digital generating devices is widespread. For these reasons it is reasonable to expect the fidelity of amateur video signals to approach that of the broadcasters. Video bandwidth must be wide enough to comfortably accommodate the colour subcarrier but should fall off early enough to avoid interference with the sound subcarrier at 6 MHz (U.K.). A video bandwidth of 5 MHz is considered adequate for amateur television.

## 2.2. Modulation Index

The available signal-to-noise ratio at the input to the receiver demodulator is a function of modulation index. The larger the modulation index the higher the signal-to-noise ratio. However, the larger the modulation index the greater is the required channel bandwidth to accommodate the signal. Modulation index therefore needs to be large enough to provide an acceptable signal-to-noise ratio but small enough to keep channel

bandwidth to a minimum.

I have conducted experiments to determine what is an 'acceptable signal-to-noise ratio'. A modern solid-state transmitter was constructed which generated the FM signal at around 100 MHz and mixed it up to the 1.3 GHz band. A balanced mixer was followed by a number of RF amplifiers.

Linearity of the frequency modulator was checked by applying a variable DC bias to the varactor diode, and measuring the resulting carrier frequency using a frequency counter. The transmitter was adjusted to use the most linear portion of the frequency modulator's characteristic.

Bandwidth and flatness of the video amplifiers which drive the modulator were checked by applying a constant-amplitude sinusoidal signal from an RF signal generator to the video input socket. The amplified signal was monitored at the varactor diode using an oscilloscope.

Initial tests attempted to adjust the transmitter in such a way as to produce an acceptable signal-to-noise ratio at the receivers of several amateur stations at different locations.

Peak deviation (hence modulation index) was adjusted to produce an acceptable signal-to-noise ratio at the receiver. The output power was varied to produce strong, medium and weak (but locked) video signals at the receiver. A modulation index was thus determined which was considered overall to be the lowest necessary to achieve a reasonable signal-to-noise ratio under actual operating conditions.

Without changing any adjustments on the transmitter, the modulation index was then measured as follows:

A 5 MHz sinusoidal signal was applied to the video input socket at a peak-to-peak amplitude equal to the largest video signal which would normally be applied. The RF output from the transmitter was monitored using a spectrum analyser and the modulation index was determined by measurement of the FM side frequencies. The results were compared with a table of Bessel functions and the modulation index was found to be very close to 0.5

Now that we have established the highest possible modulating frequency and the modulation

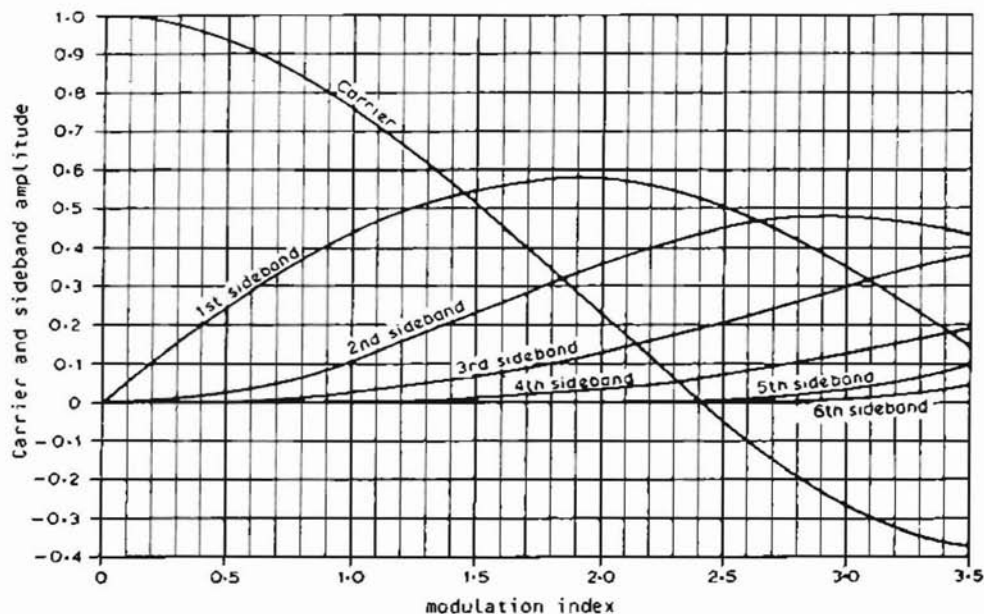


Fig. 3: Bessel curves showing variation in carrier and sideband amplitude with modulation index

index, the peak carrier deviation can be calculated:

$$\Delta f_p = M \times f_m \text{ (MHz)}$$

$\Delta f_p$  thus becomes  $0.5 \times 5 \text{ MHz} = 2.5 \text{ MHz}$

### 2.3. Bessel Functions

Bessel functions are a set of curves on a graph depicting the relative amplitudes of carrier and side frequencies with different modulation indices (fig. 3). The amplitudes are given relative to the amplitude of the unmodulated carrier.

Bessel functions can be shown in other ways such as the spectrum diagrams in fig. 1 or simply as a set of tables like those in table 1.

Using any of these methods it is easy to look-up the modulation index for any given signal. It is interesting to note that when the carrier amplitude

drops by half (3 dB), a modulation index of 1.5 is produced whilst when the carrier actually disappears (reduces to zero), the modulation index is close to 2.4. These facts are often used to calibrate measuring instruments but are of little use to amateurs since not only are we interested in modulation indices of around 0.5, most of us don't have a spectrum analyser with which to observe the carrier.

### 2.4. Channel Bandwidth

Bandwidth of an FM signal is usually determined by looking at the amplitude of the side frequencies and judging when they have become 'insignificant'. In theory of course there are an infinite number of side frequencies.

This is very well for simple sinewave modulating frequencies because the sideband amplitudes can be looked-up in a table of Bessel functions





index	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	1.0
carrier	- 0.02	- 0.09	- 0.2	- 0.35	- 0.55	- 0.8	- 1.1	- 1.45	- 1.86	- 2.32
f1	- 26.0	- 20.0	- 16.6	- 14.16	- 12.3	- 10.85	- 9.66	- 8.66	- 7.83	- 7.49
f2	-	-	- 39.0	- 34.1	- 30.29	- 27.19	- 24.6	- 22.41	- 20.48	- 18.8
f3	-	-	-	-	-	-	-	- 39.83	- 36.83	- 34.15

All figures are in dB relative to the level of the unmodulated carrier. Only figures above - 40 dB are given.

Table 1: Bessel functions

or measured with a spectrum analyser. Because of the complex nature of a video modulating signal, however, it is difficult to accurately determine the channel bandwidth required for a frequency modulated television transmission. A survey of related works in the professional technical press shows that the use of 'Carson's Rule Bandwidth' is universally accepted as giving a close approximation to the actual channel bandwidth required. This is particularly true when modulation indices of less than unity are used – as in our case.

Carson's Rule Bandwidth (B) is given by:

$$B = 2 (\Delta f_p + f_m) \text{ MHz}$$

In determining channel bandwidth, the 'worst-case' figures of peak carrier deviation and highest modulating frequency are used. It is recognised that in actual amateur service, the calculated bandwidth will seldom be achieved, since most scenes will have relatively low lighting levels and contain little or no energy at the high frequency end of the video band. With a television station, however, operating near the band edge, it is important to consider the effect of fast switching transients which may be present – particularly in digitally derived pictures. These will normally be reduced by a video filter and it's also advisable to incorporate a peak clipper to eliminate any remaining spikes. Filters and clippers are discussed later in this article.

Using Carson's rule, the channel bandwidth necessary to accommodate an amateur FM-TV transmission can be calculated:

$$B = 2 (2.5 \text{ MHz} + 5 \text{ MHz}) = 15 \text{ MHz}$$

## 2.5. Sound

Sound in an FM-TV system is produced in much the same way as on AM. A 6 MHz subcarrier is frequency modulated by the audio, but this time the subcarrier is used to frequency modulate the main carrier to produce a modulation index of approximately 0.2.

The amplitude of a sound subcarrier signal should be great enough to ensure an adequate signal-to-noise performance at the receiver, but not so great that it adds unacceptable intermodulation distortion products to the video signal.

Broadcast FM-TV transmissions typically have the sound subcarrier amplitude between 20 and 30 dB below peak video amplitude. It must be remembered, however, that broadcasters expect strong pictures at the receiver. Because the signal strength is so great, it is not necessary for the sound subcarrier to be any larger since an adequate signal-to-noise ratio will always be produced.

Amateurs, however, deal mostly with signal strengths considerably less than those of the broadcasters, often so weak that the picture only just locks. It is important therefore for the sound subcarrier to be as large as possible (without causing intermodulation distortion with video signals) so that an acceptable signal-to-noise ratio is produced at the receiver – even under relatively weak signal conditions.

During the experimental transmissions already described, the amplitude of the 6 MHz sound subcarrier signal was varied to determine what

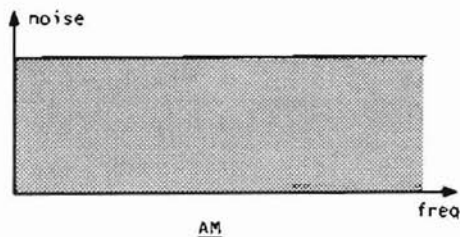


Fig. 4a: AM noise spectrum after demodulation

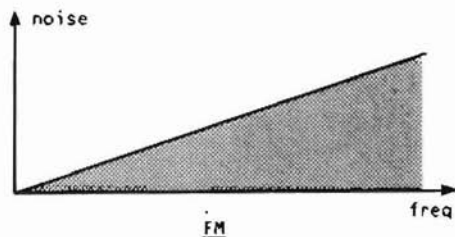


Fig. 4b: FM triangular noise spectrum after demodulation

was an 'acceptable' sound signal-to-noise ratio at the receiver. Subsequent measurement of the subcarrier amplitude revealed it to be very close to  $-14$  dB with reference to peak video amplitude.

The presence of a 6 MHz sound subcarrier will add slightly to the required channel bandwidth. It has been found by measurement that the levels of RF sidebands due to the sound carrier, fall by about 20 dB per octave. Under these conditions the total channel bandwidth will increase by about 2.5 MHz. This must be taken into consideration when operating near the band edges.

## 2.6. Pre-emphasis

During demodulation an effect, known as 'FM triangular noise', is present and is caused by a linear increase in noise as the modulation frequency increases (fig. 4). The effect is seen as a reduction in overall signal-to-noise ratio and thus a noticeably noisy picture.

To correct for this problem it is usual, at the transmitter, to emphasise the higher modulation frequencies linearly and thus equalising the signal-to-noise ratio across the video band. This technique is known as 'pre-emphasis' and a

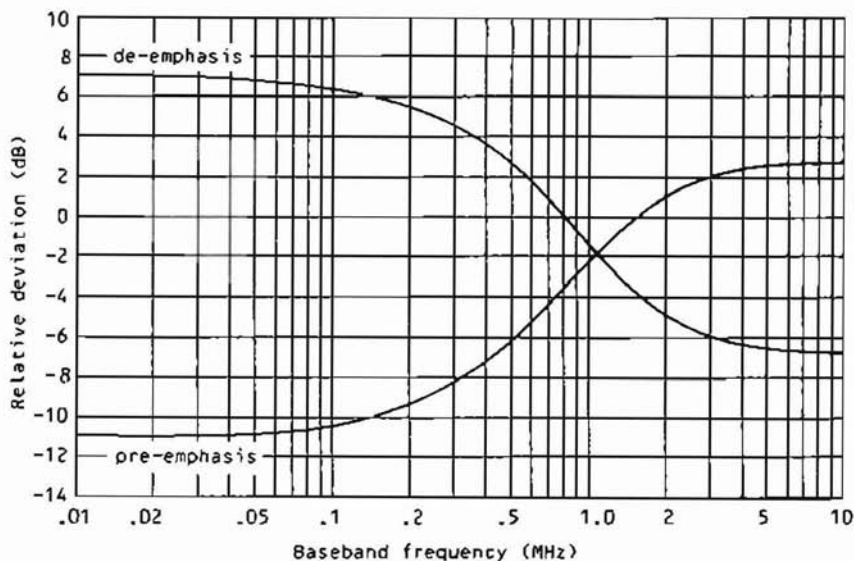


Fig. 5: Characteristic curves for emphasis on 625-line television systems



corresponding 'de-emphasis' network is necessary in the receiver.

Amateurs are permitted to use the CCIR recommendation 405-1 characteristic for pre-emphasis. This characteristic lifts a 5 MHz signal by nearly 3 dB whilst attenuating LF signals by up to 11 dB (fig. 5)

If the amplitude of a worst-case HF video signal at 5 MHz is raised by 3 dB,  $\Delta f_p$  rises to  $1.4 \times 2.5 \text{ MHz} = 3.5 \text{ MHz}$  and channel bandwidth thus becomes 17 MHz. Since a potential improvement in overall signal-to-noise ratio of up to 14 dB is available, and since amateurs are usually dealing with signal strengths considerably lower than that required for broadcast quality reproduction, the improvement in signal-to-noise ratio due to pre-emphasis is significant and well worth the extra 2 MHz of channel bandwidth (see photographs a and b)

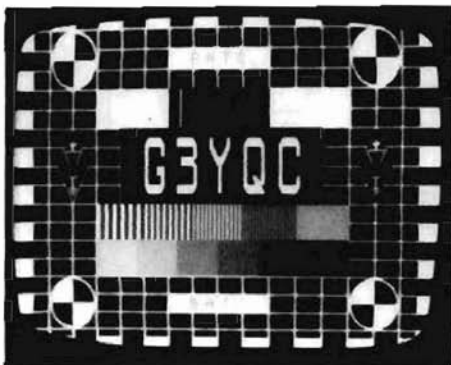
## 2.7. Standards

FM television is a relatively new art for the amateur and it is therefore all the more important that a suitable set of standards is adopted at this early stage. These standards should reflect the special requirements of FM-TV communication under weak-signal conditions; they should not simply be a 'watered-down' version of those used by the broadcasters.

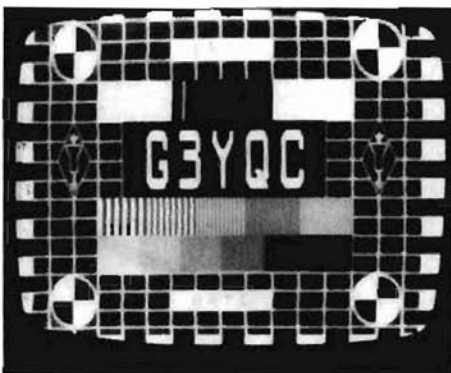
The BATC's Technical Liaison Committee has done a great deal of work towards establishing suitable standards and has submitted its findings and recommendations to the RSGB and DTI. It is hoped that they will become the accepted specifications for the amateur service.

The proposed standards for amateur FM-TV are as follows:

Mode of emission	F5/F3
Video bandwidth (3 dB)	5 MHz
Colour subcarrier frequency	4.433618 MHz
Maximum instantaneous modulation index (video)	0.5
Peak deviation	
(without pre-emphasis)	2.5 MHz
(with CCIR pre-emphasis)	3.5 MHz
Channel bandwidth	18 MHz
Sound subcarrier frequency	6 MHz



Photograph a: With pre-emphasis



Photograph b: Without pre-emphasis

Sound subcarrier amplitude (with respect to peak video)	- 14 dB
Sound subcarrier modulation index	0.2

It is usual to arrange modulation in such a way that peak syncs will shift the carrier lower in frequency.

When considering FM for amateur TV repeaters, there are several other points worthy of consideration:

- A video filter having a bandwidth of 5 MHz should be included in the modulating amplifier. This will stop any HF signals beyond the speci-

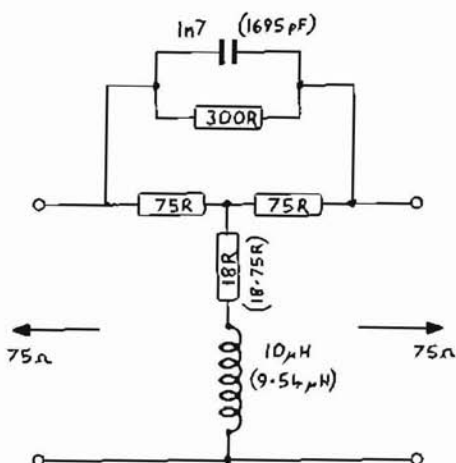


Fig. 6a: Pre-emphasis network 625-line CCIR

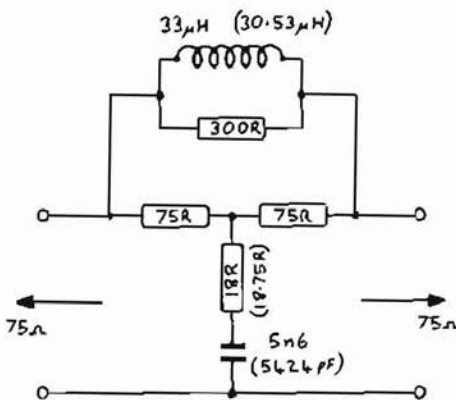


Fig. 6b: De-emphasis network 625-line CCIR

fication, these may typically be produced by computer generated video

- A video peak clipper should be included after the video filter. This will cut any odd high-amplitude spikes which may be present
- DC clamping of the video signal should be included to prevent the nominal carrier fre-

quency from changing with different television scenes. The carrier frequency is fairly stable anyway but this would tighten it up when operating near band edges.

- An RF output filter should be included to prevent out-of-band energy from whatever source from reaching the aerial system.

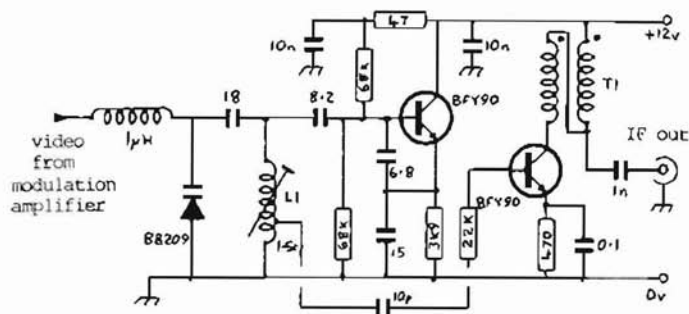
### 3. CIRCUIT TECHNIQUES OF FM-ATV TRANSMITTERS

#### 3.1. Pre-emphasis

As we saw earlier, the demodulation of an FM-TV signal reveals an inherent undesirable effect; the linear increase in the noise signal as the modulation frequency increases. This is known as the FM triangular noise.

As the amplitudes of the higher-frequency video signals from amateur equipment tend to be lower, the result of this FM triangular noise is a noisy picture. The solution to this phenomenon is to emphasise the higher modulation frequencies linearly with frequency, thus compensating for the noise and tending to equalise the signal-to-noise ratio across the band. This procedure is known as **pre-emphasis** and, to be effective, must be matched at the receiver by a network which attenuates the high modulating frequencies, thus restoring the relative amplitudes to those of the original from the vision source. The network at the receiver is known as the **de-emphasis** network and is shown, together with the pre-emphasis network in figs. 6a and 6b. The component values shown are the nearest preferred ones commonly available to amateurs, whilst the values in brackets are the actual ones required for full specification performance as recommended by the CCIR.

The term "pre-emphasis" is something of a misnomer because, considering that both circuits use passive components and are therefore incapable of amplification, the only way to tailor the relative amplitudes of an FM signal in a particular way is to attenuate the appropriate



L1 - Toko S18 blue, with can and core.  
 T1 - Bifilar wound. As many turns of thin twisted wire as will fit through a Fair-Rite two-hole ferrite core No.26-43006302 or similar core. Windings connected in series as shown.

Fig. 7:  
 Typical frequency  
 modulated oscillator  
 (100 MHz)

lower frequencies, leaving the higher ones unaltered.

Both circuits may be built on a piece of strip-board (Vero) and inserted in the 75-Ω video line. Since both networks are attenuators though, this must be taken into account when adjusting video gain. Therefore both receiver video gain and transmitter video deviation will need to be increased to compensate for losses in the emphasis networks.

### 3.2. Modulation

Because of the relatively wide deviations used in a television system, the modulation method employed must be of the frequency type, rather than the more usual indirect or phase method favoured for NBFM telephony transmissions.

The most appropriate method of achieving frequency modulation is to vary the carrier-oscillator frequency by the modulating signal. It is important that the amount of frequency change is directly proportional to the amplitude of the modulating signal, otherwise the resulting demodulated picture at the receiver will not be a true representation of the original. To do this the carrier oscillator must change its frequency linearly according to modulation and the modulating diode must operate over the linear part of its characteristic.

Fig. 7 shows a typical Colpitts oscillator intended for use as a frequency modulated IF generator in a heterodyne or mixer-type transmitter. The circuit is designed for a nominal frequency of 100 MHz, although selection of frequency sensitive components will allow it to be used anywhere between (say) about 50 and 400 MHz.

Video/sound modulation is applied to the varactor diode via a radio frequency choke. This prevents the carrier signal from feeding back into the video circuits which could cause interference. A standard Toko S18 coil (blue) is used for L1 and is fitted with a screening can to ensure best stability. The output tap is taken close to the ground end and should be soldered with care.

The oscillator output is lightly coupled to a buffer amplifier which uses a wideband transmission-line transformer as its collector load. The output amplitude is flat over the required frequency range (+/- at least 10 MHz) and the level obtained is sufficient to drive a conventional transmit mixer such as the USM-3. This 1.3 GHz mixer is particularly suitable for wideband ATV applications since it employs wideband input circuitry and its RF amplifiers are not unduly narrow in their frequency response. It is a fundamental requirement of FM television that sufficient RF bandwidth is maintained in both transmission and reception equipment, otherwise the full potential of the FM system will not be realised.

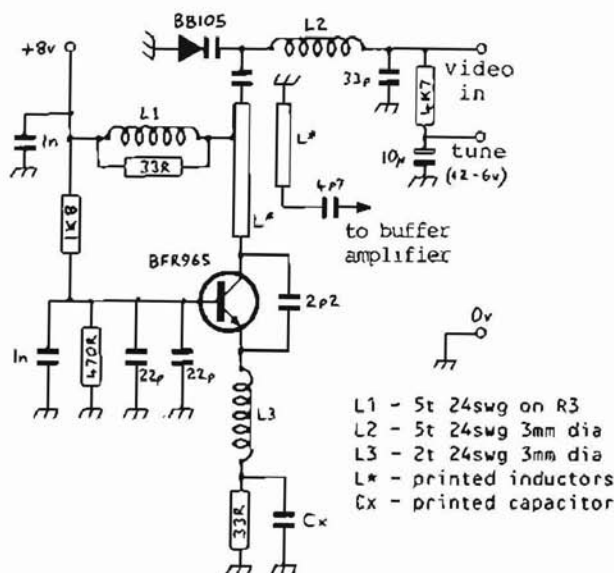


Fig. 8:  
1.3 GHz oscillator used in Solent  
Scientific 24 cm ATV transmitter

Apart from the heterodyne transmit system referred to, there are at least two other methods in common use: The first uses a free-running oscillator on the same frequency as the transmitter output (typically in the 1.3 GHz band). The advantages are that no frequency changing circuitry is needed and a relatively low video amplitude is needed to provide the required deviation of the oscillator. The major disadvantage is one of relatively poor stability. It is very difficult to maintain a 1300 MHz free-running oscillator to a reasonable degree of stability, although, considering the channel bandwidth of the transmission (around 18 MHz), oscillator stability need not be to a particularly high specification.

Fig. 8 shows the circuit of the 1300 MHz oscillator used in the Solent FM-TV transmitter. This oscillator – when constructed according to the instructions! – is particularly stable and is therefore quite suitable for its intended application.

The second method is to use a fairly high frequency oscillator (say 420 MHz), frequency-modulate it and then frequency-triple the signal to the required frequency in the 1.3 GHz band

The advantage is that stability is easier to maintain on lower frequency oscillators – although the rate of drift will be magnified by the rate of frequency multiplication, similarly the modulator deviation will be increased by the same factor.

### 3.3. Video Clamping

Because the picture content will vary as the video picture moves or changes scene, the 'average' level of the signal will vary accordingly. For example, the video signal shown in fig. 9 will have a different average corresponding to the peak amplitude of the video. If this constantly shifting average level were used in an FM-TV modulator, the result would be that the nominal carrier frequency would move with scene changes, consequently the overall video deviation would be wider than it need be.

To prevent this, clamping of the video signal to a reference is required. The reference used is logically the part of the signal which remains constant, regardless of scene content such as the sync tip or black level which is the point where the blanking pulse and sync pulse meet.

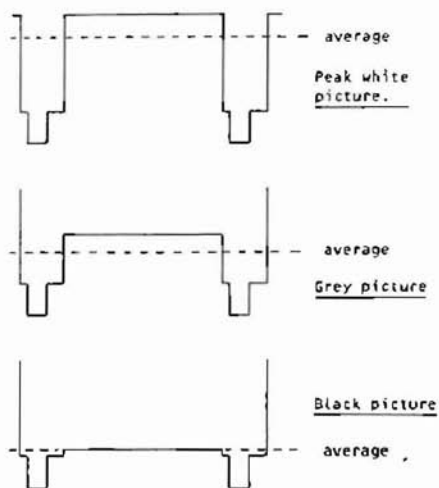


Fig. 9: Average video amplitude levels shown at different scene brightnesses, related to a constant p-p reference amplitude

The simplest form of clamping is DC restoration using a diode clamp. This is a passive circuit which relies on the ratio of the diode resistance when conducting against the value of parallel resistance when the diode is not conducting. DC clamping is used in the transmit video circuit shown in fig. 10. This method is particularly

suitable for TV repeaters where the incoming video signal may be quite weak and it would therefore be difficult to lock a clamp circuit to its sync pulses.

A rather better system for TV signals is the 'sync tip' or 'active' clamp and this method is shown in fig. 11. Since the point of interest is at sync tip, it is desirable to have a clamp which operates during the blanking/sync interval.

The composite video signal is presented to a sync separator (Q7) where the video is removed. The resulting sync pulses are fed to a clamping transistor (Q3) and the  $2.2 \mu\text{F}$  tantalum capacitor is charged during blanking by forcing the transistor into heavy conduction during this interval. In this circuit the negative side of the capacitor is brought to ground, and when the clamping pulse is removed, the transistor is non-conducting and therefore at a very high impedance.

### 3.4. Video Processing

Fig. 10 shows the circuit of a typical television video strip. Composite video is applied to a simple low-pass filter and then on to a wideband video amplifier. The signal is DC clamped by the diode circuit, the level of which is adjustable. From an emitter follower buffer the video signal passes through a 6 MHz sound trap which prevents any video from interfering with the inter-carrier sound signal. 6 MHz sound is introduced after the sound trap and the resulting composite

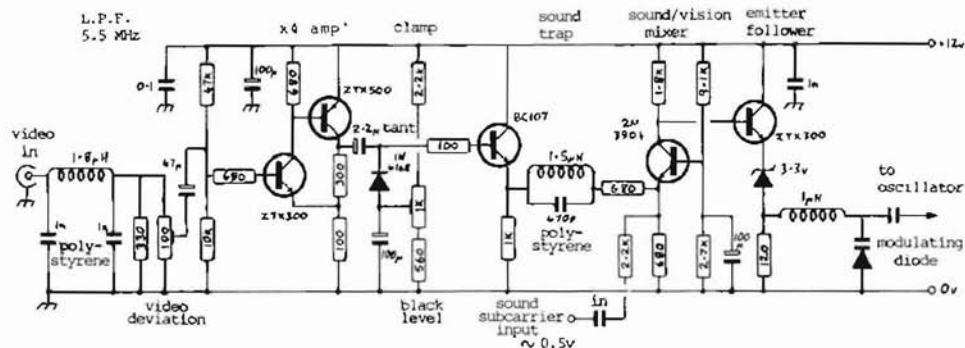


Fig. 10: FM-TV transmitter video circuit (diode clamping)

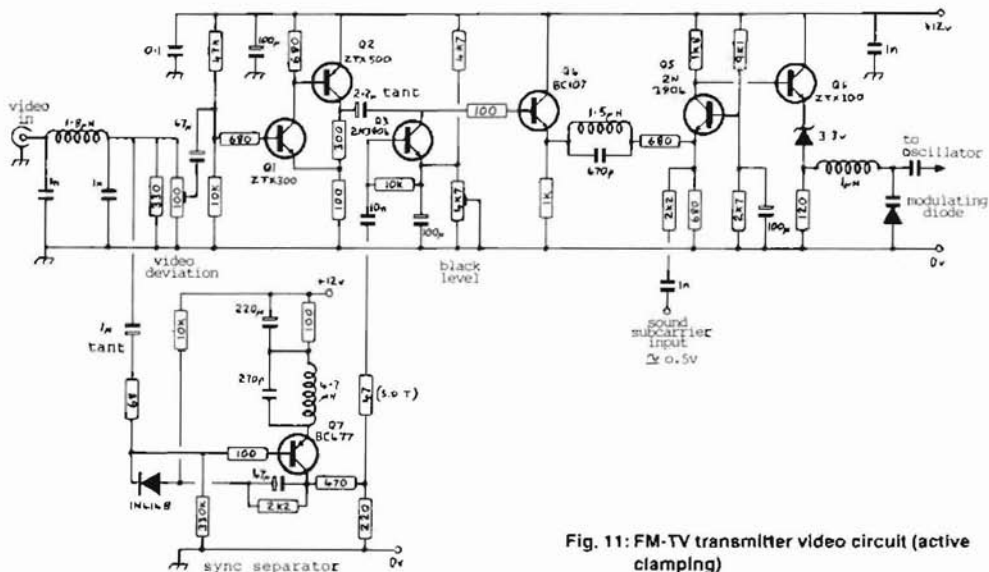


Fig. 11: FM-TV transmitter video circuit (active clamping)

(sound and video) signal is then buffered by an emitter follower, and DC shifted by a 3.3 volt zener diode. This method of DC shifting the signal is essential to maintain the video DC restoration provided by the clamp circuit. The output gives a swing of up to 6 volts p-p, adjustable by a combination of the deviation and black level controls

A similar circuit is shown in fig. 11, except that a sync separator and active sync clamping is used instead of the simple DC restorer. This method is preferred where closed-circuit quality video is to be used.

### 3.5. Peak Clipping

Although a video signal normally has a certain maximum overall amplitude (1 V p-p for normal distribution), it is possible under certain conditions for the maximum video amplitude to be exceeded. This is sometimes caused by digitally derived pictures or signals with excessive HF peaking. If such signals were to be presented to a video modulator, then the instantaneous

maximum deviation would exceed that which the transmitter had previously been adjusted to. It is particularly important in TV repeater stations that this is not allowed to happen, therefore some form of peak clipping is needed.

Fig. 12 shows a peak clipping arrangement which may be incorporated into the circuits in figs. 10 and 11. The circuit prevents the base of the final transistor from exceeding the maximum voltage set by the preset control.

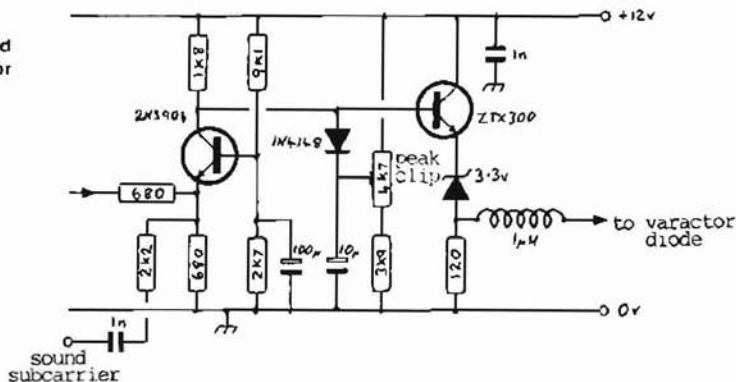
A different type of clipper is shown in fig. 13. This circuit is incorporated into the Solent Scientific 1 W 24 cm ATV transmitter, now being marketed by the Worthing ATV Repeater Group. This circuit not only prevents excessive video excursions but also prevents sync pulse overshoot, usually found in un-filtered digital sync circuits.

Some small advantage may be obtained from the use of a clipping circuit in that it allows the video deviation control to be turned up a little and thus increase the average video level. Of course if this were too high, then a noticeable impairment in the white parts of the picture will become apparent.





Fig. 12:  
Simple peak clipper fitted  
to an FM video modulator



#### 4. CIRCUIT TECHNIQUES OF FM-ATV RECEIVERS

To receive FM-TV correctly, a suitable demodulator must be provided. It is true that an FM signal can be received using an ordinary AM-TV demodulator, this is called 'slope-detection'. However, because the received signal needs to be positioned part way down the IF passband or 'on its slope', the resulting picture will be considerably weaker than if it were received using an FM demodulator.

The first requirement of an FM demodulator (discriminator or detector) is that the applied IF signal should be adequately limited. The limiter ensures that any FM signals containing amplitude variations are clipped, thus maintaining the FM carrier level at constant amplitude. Therefore, the limiter provides a high degree of AM rejection before the FM detector.

##### 4.1. Demodulators

There are not very many FM demodulators having a wide enough bandwidth to make them suitable for receiving television pictures. Fig. 14 shows a typical diode discriminator found in certain satellite TV receivers. The circuit is a form of ratio detector and relies on the IF signal being split into two. One of them is phase-shifted

through a delay line and each signal then applied to a diode bridge which provides a baseband video output. This sort of discriminator (in common with other types such as the Foster Seeley) requires a very high level of signal to drive them – usually several volts. Obviously this level, at relatively high IF frequencies (typically 70 MHz or more) is not easy to achieve and the gain circuitry must be laid out with great care to avoid instability. Also it is difficult to achieve the required bandwidth using straight-forward tuned circuits. For this reason, wideband transformers are used in this design. The transformers comprise a very few turns (8 or so) wound

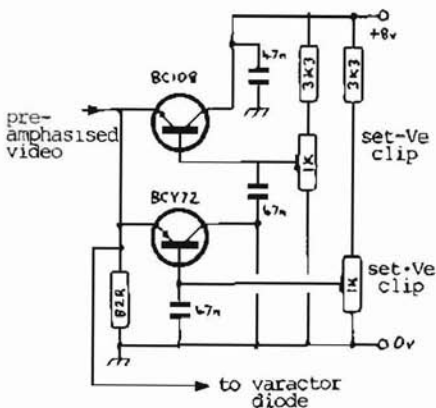


Fig. 13: Circuit of video peak clipper used in the Solent Scientific transmitter

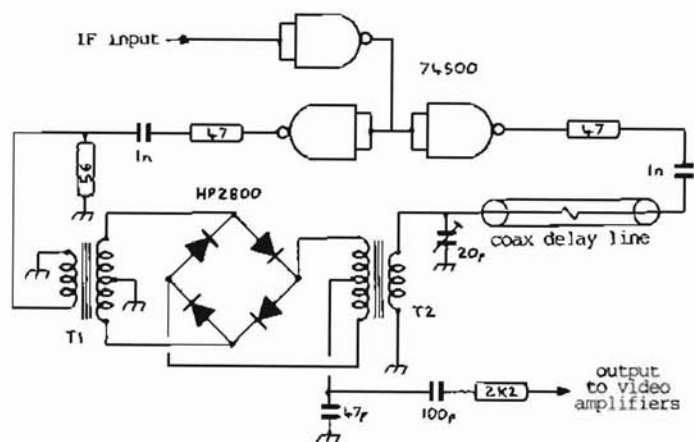
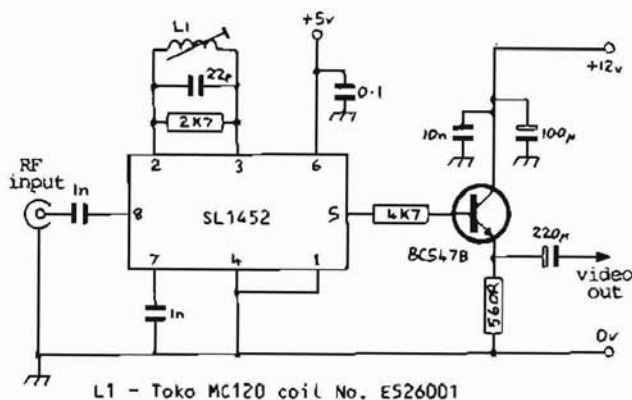


Fig. 14:  
A typical wideband diode discriminator

on a small ferrite former, or even just a bead, the low impedance winding (just a couple of turns) matches the external circuitry to the demodulator. The circuit uses Schottky TTL gates to ensure hard signal limiting and to present a high enough voltage to the demodulator. It also makes it easy to split the signal into two. Plessey have developed a quadrature detector chip specially for use in satellite TV receivers and this is almost ideal for amateur applications. It is easy to use, has a minimum of external components and its working bandwidth may be tailored to suit the incoming signal deviation.

Fig. 15 shows a practical circuit diagram, but don't forget that this too needs a fairly high level (around 5 mV) limited RF input signal to drive it correctly. The RF input frequency can be anywhere between 400 and 1000 MHz, although this design is for the standard IF frequency of 479.5 MHz.

The IC includes a divide-by-four circuit at the input, therefore the demodulator tuned circuit actually resonates at one quarter of the input frequency. A simple emitter follower buffers the video output signal and brings it to near 75 Ω

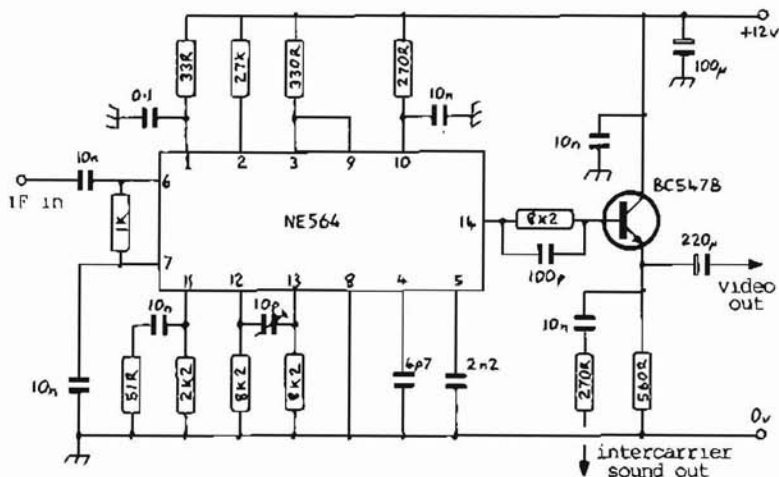


L1 - Toko MC120 coil No. E526001

Fig. 15:  
Quadrature FM demodulator using Plessey SL 1452 custom chip



Fig. 16:  
PLL TV  
demodulator  
circuit



output impedance. However, since the amplitude of output video is directly proportional to the deviation of the received signal, it is probable that in amateur service some further video amplification will be required in order to bring the level to around 1 V peak-to-peak.

#### 4.2. The Phase-Locked Loop

Amateurs have, for some time, relied heavily on the Signetics high frequency, wideband phase-locked loop IC – the NE 564 – for demodulation. The benefits of this IC are that it exhibits around 5 dB improvement in signal-to-noise ratio and only requires a relatively low-level drive signal. The chip has its own IF limiting amplifier although the IF signal should still be hard limited before being presented to the PLL.

A typical application circuit is shown in fig. 16. There are, however, one or two disadvantages with using the PLL. The maximum specified working frequency of the NE 564 is 50 MHz, although some can be got going at 70 MHz but these are liable to be unstable. The bandwidth is only around 22 MHz which causes broadcast signals to display ragged edged pictures, although for normal amateur purposes 22 MHz is probably about right. The internal vision signal-to-noise ratio is only around 40 dB, therefore the

received picture is likely to be slightly noisy, even under the best reception conditions. Commercial TV reproduction demands a signal-to-noise ratio of around 60 dB. The NE 564, despite its internal limiter, can still be easily overloaded, so care should be taken to drive it correctly, hence the need for a good limiter in the preceding circuitry.

There is a custom-designed PLL chip intended for demodulation of satellite TV signals; this is the Plessey SL 1451 and a practical application circuit is shown in fig. 17. The IF input range for this chip is 400 to 700 MHz and the circuit shown is intended to operate at around 612 MHz. All circuit values are shown on the diagram including details of the two inductors. Care should be taken in the layout and construction of this circuit. All component leads should be as short as possible and a double-sided PC board should be used – the component side being a ground plane.

Capacitors C1 and C2 connected between the differential inputs and outputs of the video amplifier determine the loop filter response and, in this design, are optimised for the DBS standard 13.5 MHz peak-to-peak deviation at a centre frequency of 612 MHz. An RF input level of typically – 10 dBm (minimum – 25 dBm) is required for correct locking and, as with the SL 1452 circuit, the video output may need amplifying to provide a 1 V p-p level.

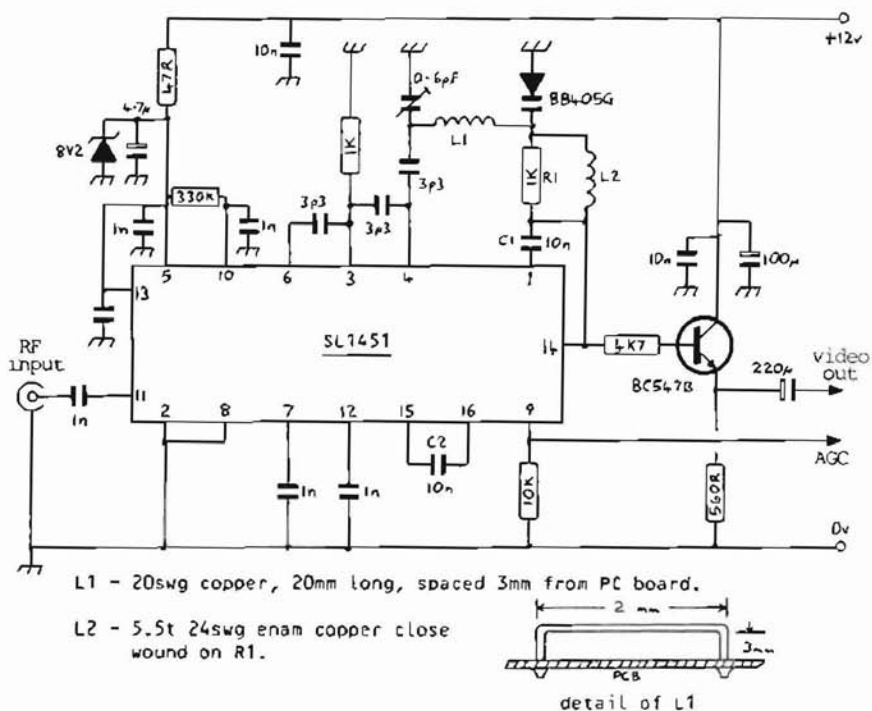


Fig. 17: Wideband FM demodulator using Plessey SL 1451 custom chip

### 4.3. TVRO Modules

There are a number of satellite TV modules which are quite suitable for amateur applications. Of these, perhaps the most popular are those by Astec and Mitsumi. It should be remembered that broadcast satellite TV equipment is designed for much wider deviations than amateurs typically use, therefore, in order to get the best sensitivity from a system, it may be necessary to reduce the IF bandwidth response and increase the level of video amplitude.

The Astec AT 1020 tuner head tunes the band between 950 and 1450 MHz and provides a fixed IF output at 612 MHz. The Astec AT 3010 IF/demodulator module incorporates IF filtering, limiting and demodulation (using the Plessey SL 1452). This module has a narrow-bandwidth facility which makes it particularly suitable for amateur applications. I understand though that these two modules are soon to be discontinued

and replaced with others. A complete receiver project was described in R & EW Magazine for August 1986.

The Mitsumi system comprises a set of three modules: a tunable down-converter covering between 950 and 1750 MHz with a fixed IF of 479.5 MHz, a fixed down-converter to convert the first IF to 70 MHz and a video demodulator. Interconnection details and circuitry were described in CO-TV Magazine 140.

## 5. REPEATERS

There are a number of amateur television repeaters in the UK and these are shown in table 2.

## U.K. ATV Repeater Stations

Call sign	Location	Channel	Contact
GB3AF*	Durham	RMT-2	G1FBY
GB3CT	Crawley	RMT-2	G4ZPP
GB3ET	Emley Moor	RMT-2	G3PTU
GB3GT*	Glasgow	RMT-2	GMØGIB
GB3GV	Leicester	RMT-2	GØCND
GB3HV	High Wycombe	RMT-3	G4CRJ
GB3NV	Nottingham	RMT-2	G6YKC
GB3RT	Rugby	RMT-2	G6IQM
GB3TV	Dunstable	RMT-2	G4ENB
GB3UD	Stoke-on-Trent	RMT-2	G8KUZ
GB3UT	Bath	RMT-1	G4JQP
GB3VI+	Hastings	RMT-1	G3ZFE
GB3VR	Brighton	RMT-2	G4WTV
GB3ZZ	Bristol	RMT-2	G8VPG

\* Temporary beacon mode only.

+ Not yet operational at the time of going to press

Table 2

Input and output frequencies for the three channels at present available in the U.K. are as follows:

Channel	Input/MHz	Output/MHz
RMT-1	1276.50	1311.50
RMT-2	1249.00	1318.50
RMT-3	1248.00	1308.00

Table 3

Repeater stations on RMT-1 will accept either AM or FM signals, whilst inputs on RMT-2 and RMT-3 are in FM only. Repeater output on RMT-1 is AM

only, and the outputs on RMT-2 and RMT-3 are FM.

Repeater video identification is standard 625-line Fast-Scan TV (system I), with negative going video and positive going syncs. A morse code ident is also present on the 6 MHz audio carrier detailing callsign and location. FM transmissions are limited to a maximum peak deviation of  $\pm 6.5$  MHz and they can employ CCIR pre-emphasis as an option. The intercarrier sound subcarrier is at 6.0 MHz and repeater access is by the presence of a standard 625-line video signal at the input.

Although the actual coverage of each repeater will depend on local geography, and in some cases aerial directivity, the average area covered is in the order of a 30 km radius from the repeater. Apart from one or two exceptions, aerials are omni-directional. All ATV aerials operate horizontally, regardless of band.

## 6. REFERENCES

- (1) 'Frequency Modulated Amateur Television (ATV)' by Josef Grimm, DJ 6 PI VHF COMMUNICATIONS Vol 18 (1986) Ed 3, pages 165 - 176
- (2) 'Pre-emphasis characteristics for frequency modulation radio-relay systems for television' - CCIR Rec. 405-1
- (3) 'Television Handbook for the Amateur' by Biagio Presti.
- (4) CQ-TV MAGAZINE No. 140, 141, 144

Interested in AMATEUR TV ?

....then you need CQ-TV Magazine.

For further information see page 256

Hannes Fasching, OE 5 JFL

## A Compact Hybrid Antenna for 2 m, 70 cm and 23 cm

Many UHF amateurs would like to work VHF, UHF and SHF in various modes. This means, of course, a certain outlay in the provision of antennas. In my case, additionally, there is the problem concerning the lack of available space – and I am certainly not the only radio amateur to suffer this! It is for this reason that the antenna form selected (fig. 1) could also be of interest for others.

Out of the following requirements placed upon the antenna system, a suitable design crystallized:

- Able to work 2 m, 70 cm and 23 cm
- Provision to work even higher frequencies
- Switchable polarizations (horizontal/vertical)
- Compact form
- Enough gain for satellite working
- Within a limited framework, should be possible to work 23 cm EME

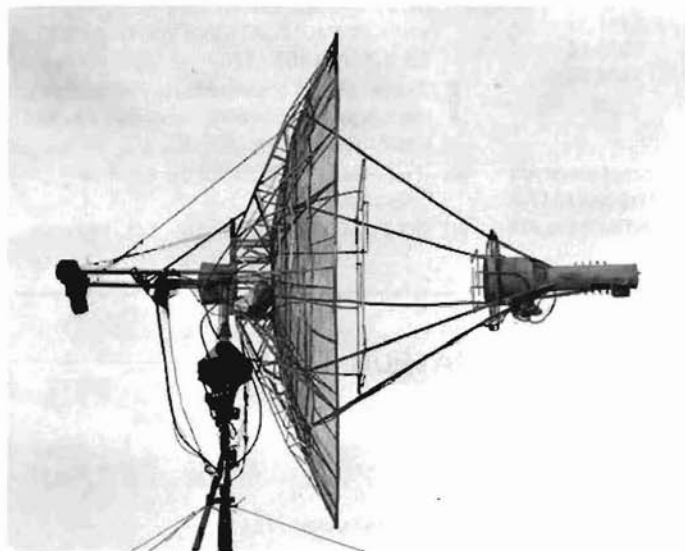


Fig. 1:  
OE 5 JFL hybrid antenna  
for 2 m, 70 cm and 23 cm

## 1. CONCEPTUAL CONSIDERATIONS

The last of the above demands dictated a parabolic dish antenna for the 23 cm band (1296 MHz). The smallest diameter of such an antenna, capable of working 23 cm EME, is 2 metres. Yagi antennas for 2 m (144 MHz) and 70 cm (432 MHz) could then be constructed about the dish antenna. This was successfully tried for the 70 cm band antenna but only for one polarization and there was still a problem of where to site the 2 m band antennas.

With this, a further round of considerations was undertaken with the idea in mind to increase the main dish diameter until it was also viable for the 70 cm band

The calculation was quite simple: If a dish, greater than 1 metre diameter, gives good 23 cm results, a 3 metre-diameter dish would produce about the same results on 70 cm. The increase of 1 m from 2 metre to 3 metre would also increase the gain on 23 cm by some 3.5 dB.

Using a universal (hybrid) radiator comprising a horn for 23 cm and a switchable polarization



Fig. 2: Feed arrangements for 23 cm (circular) and 70 cm (H or V switchable)

NBS antenna (fig. 2) for 70 cm, the demands for these two bands were amply fulfilled inasmuch that a 3 metre dish with an  $f/D \approx 0.5$  may be very well illuminated.

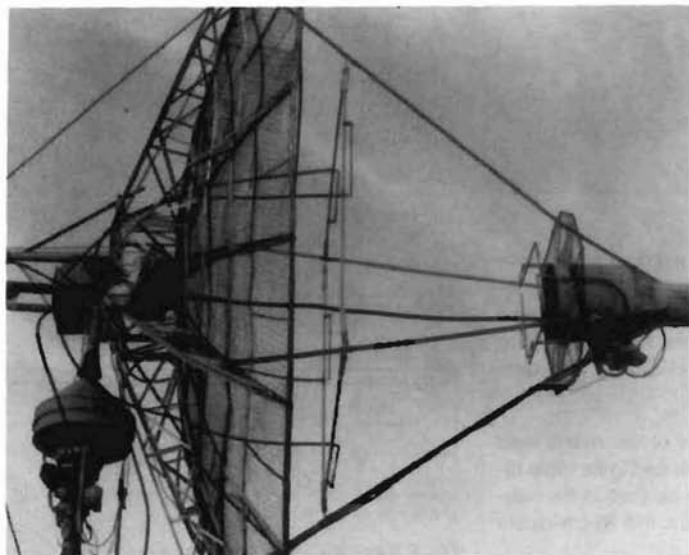


Fig. 3: Right against the surface of the dish are located the four 2 m dipoles for H and V polarization



In order that the installation be equipped for the 2 metre band, Yagis can be mounted around the dish. The parabolic dish diameter of 3 metres entailed having a stacking distance of some 4 metres. Sixteen-element long Yagis would have been very suitable for this application. As the elevation of the antenna had also to be variable with its attendant mechanical problems – quite apart from considerably higher weight and space considerations – the following solution was arrived at:

The parabolic dish can also serve very well as a reflecting surface for the 144 MHz dipole. The 2 metre antenna then becomes very similar in conception to that of the 70 cm feed system. The NBS antenna has two stacked dipoles ( $\lambda/2$  long,  $\lambda/2$  apart) and  $\lambda/4$  from the reflector surface mounted about it. The gain of these antennas approximates 8 dB<sub>d</sub>. Increasing the stacking distance to  $0.6\lambda$  will increase the gain – and that is an option. In addition, the curvature of the reflector also proves itself to be beneficial. A disadvantage is, however, a certain deleterious influence of the four 144 MHz dipoles upon the gain at 70 cm and 23 cm. The dipole mounting is effected quite simply by means of plastic support insulators affixed to the four horn support struts. The latter, in turn, are supported at a suitable distance from each other by the fixations at the reflector surface (fig. 3).

## 2. IMPLEMENTATION

The theme of this article is not how a parabolic reflector with an  $f/D$  of 0.5 is designed and constructed but it is intended largely as an inspiration for related ideas. For example, for the feed horn is a design taken from the RSBG's "VHF-UHF MANUAL"

Fig. 4 shows a frontal view of the hybrid-feed system for 23 cm and 70 cm giving its main dimensions. Further detail can be seen in the side-view diagram of fig. 5. Finally, the 70 cm-dipole matching is shown in fig. 6.

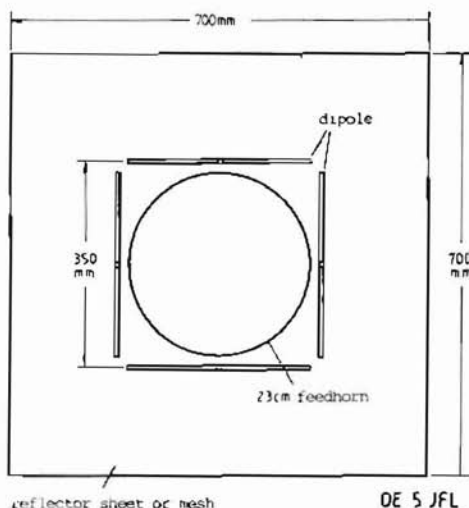


Fig. 4: View looking into the hybrid feed for 23 cm and 70 cm for an  $f/D = 0.4...0.6$

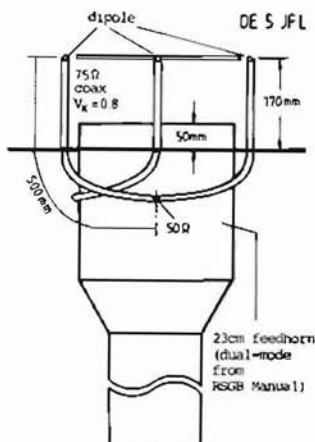
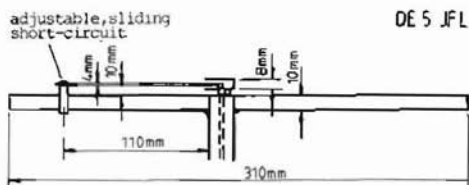


Fig. 5: The side elevation of Fig. 4 showing how each pair of 70 cm dipoles is connected



insert depth C: 5 to 10 mm  
insulation: 1 mm Teflon

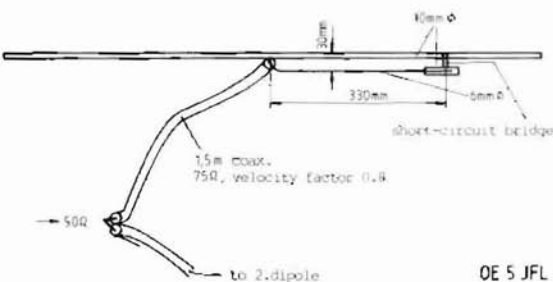
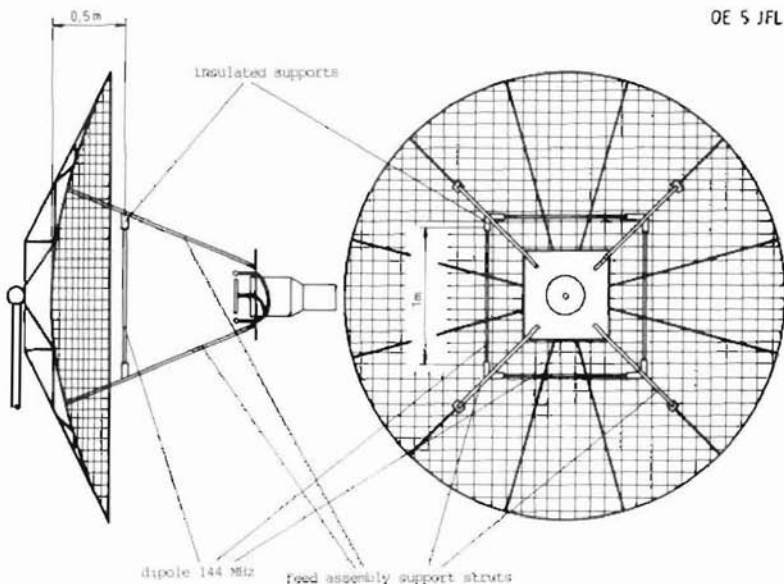
Fig. 6: Matching the 70 cm dipoles





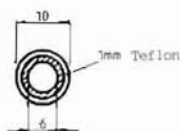
OE 5 JFL

Fig. 7:  
The complete  
arrangement



OE 5 JFL

Fig. 8: Matching the 144 MHz dipoles



'inner conductor' intrudes 5cm approx.

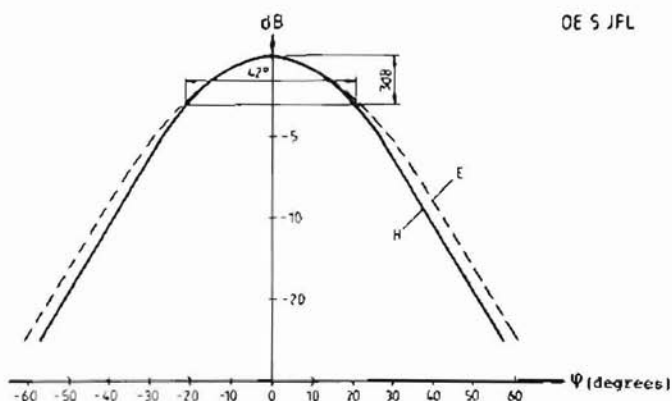
Fig. 9: The gamma-match capacitor

Figure 7 shows the arrangement of all three feed systems mounted looking in to the parabolic dish. The four 144 MHz dipoles are matched by a "gamma-match" and via a one-wavelength 75 Ω cable between dipole pairs, see fig. 8. The necessary capacitors are constructed in accordance with fig. 9

### 3. WORKING EXPERIENCE

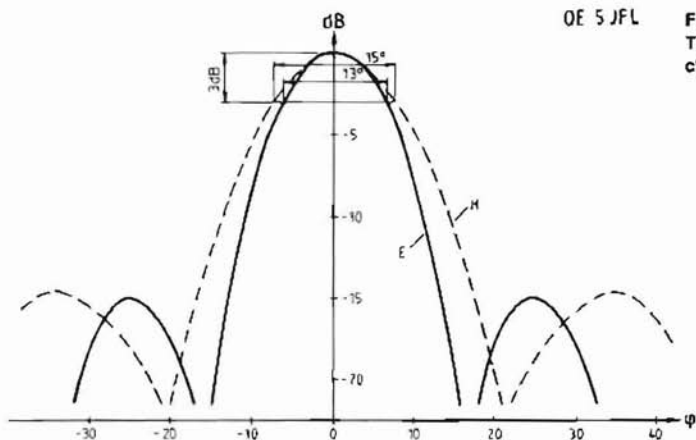
My parabolic dish has a diameter of 3.3 m. Working via the OSCAR 10 satellite is easily possible but the receiver pre-amplifier should be mounted immediately in the vicinity of the antenna. The switching of the polarization has proved a worth-while provision and is particularly useful for experimentation in ES and MS.

Moon-bounce (EME) contacts are possible on 1296 MHz but only if the distant station is using a parabolic dish of at least four metres in diam-



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Fig. 10:  
The 2 m antenna polar  
characteristic taken at 144 MHz



OE 5 JFL

Fig. 11:  
The 70 cm antenna polar  
characteristic taken at 432 MHz

eter. Using 200 Watts of RF power, it is sometimes possible to hear one's own moon-reflected signal

In the 70 cm band, EME signals can be heard, from stations with large installations, in a 200 Hz bandwidth and with a 10 dB (S + N)/N.

If the shadowing effect of the feed and the influence of the 144 MHz dipole is not taken into account, the parabolic system using a 3.3 metre

dish at an efficiency of 55 %, is calculated to give the following gains:

23 cm : 28 dB<sub>c</sub>

70 cm : 18.5 dB<sub>c</sub>

2 m : 8 dB<sub>c</sub> (approx. using NBS standard)

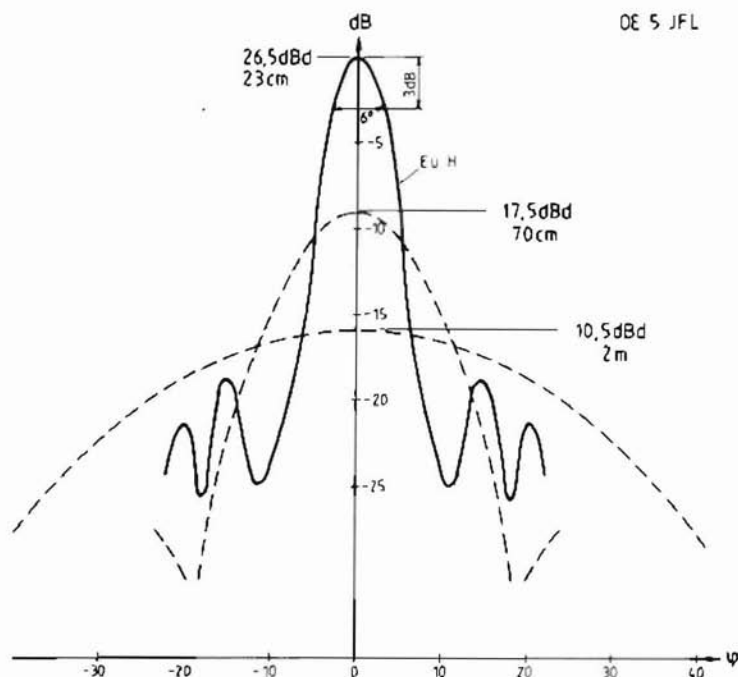
### 3.1. Polar Diagrams

The polar diagrams were measured with the results shown in figs. 10, 11 and 12. The fol-



OE 5 JFL

Fig. 12:  
The 1296 MHz polar characteristic. The main lobes of the 2 metre and 70 cm antennas are shown superimposed for comparison purposes.



lowing gains were determined from them by integration, taking the side lobes into account:

23 cm:	26.5 dB <sub>c</sub>
70 cm:	17.5 dB <sub>c</sub>
2 m:	10.5 dB <sub>c</sub>

At 23 cm, the gain deterioration is around 1.5 dB, comprising 0.5 dB due to shadowing effects and another 0.5 dB due to non-optimal reflector illumination (i.e.  $l/D = 0.45$ ). The remaining 0.5 dB can be put to the account of the 2 m dipole.

At 70 cm, the 1 dB loss is divided equally between shadowing and the influence of the dipoles.

These losses seem to me to be bearable as the saving of additional antennas for 144 MHz has been made.

If it should be desired to obtain the same antenna

performance in the normal manner, i.e. using a dedicated mirror for 23 cm and Yagis for 70 cm and 2 m, the following antenna specifications would be required:

- 23 cm: Dish diameter of 2.8 metre (approx)
- 70 cm: Two stacked 4 m long Yagis for each polarization
- 2 m: An 8 element, 3 m long Yagi for each polarization

Altogether, six long Yagis and a parabolic dish antenna – certainly not conducive to a compact construction! In addition, there would be mutual coupling interference between antennas leading to a high likelihood of gain reduction.

P.S. The damage to the parabolic reflector which is evident on the photographs, was caused by ice rain last winter.



*Willi Rass, DF 4 NW*

## An Introduction to Moonbounce (EME)

Earth-Moon-Earth (EME) is a propagation means which enables world-wide amateur contacts to be established preferably using amateur band frequencies from 144 MHz upwards. These communication distances are extremely large for these frequencies and the technique is fascinating a growing number of radio amateurs owing to the challenge presented by its highly technical nature.

The principle is really quite simple. The transmitter antenna is directed at the moon and the signals, reflected from the moon's surface, are picked up at the distant receiver. At a suitable position for the moon (common window), two stations separated by up to 20,000 km can achieve communication i.e. over a distance of half the world's circumference.

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### 1. HISTORY

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The first attested case of an EME contact was reported in July 1960 between two US stations, W6HB and W1BU using 1296 MHz. The first inter-continental EME contact came in April 1964 be-

tween Finland (OH 1 NL) and the USA (W 6 DNG).

The following year saw EME experiments using high-powered radio astronomical installations with large parabolic antennas, by various institutes. Foremost among these were the universities of Stanford (WA 6 LET) and Arecibo in Puerto Rico (KP 4 BPZ). In particular, the series of EME experiments conducted by American radio amateurs in June 1964 from the Arecibo radio telescope (KP 4 BPZ) with a 300 meter diameter dish, were very widely known. This is because the unusually high antenna gain of this gigantic antenna installation enabled amateurs to receive SSB and CW signals when using quite simple Yagi antennas.

The initial years of this new propagation method, represented a very large financial and technical outlay for the radio amateur which could only be alleviated by group undertakings. The necessary parametric amplifiers for the receive pre-amplifier, highly stable oscillators, large antennas and AZ/EL antenna control, had to be home-constructed.

The first moonbounce contacts, using purely amateur-constructed equipment and with one-man efforts at each station, took place in the 70's. The burgeoning development in HF and



semi-conductor techniques enabled reliable and cost-effective VHF/UHF, low-noise pre-amplifiers as well as transmit amplifiers capable of high power to be developed. The somewhat difficult, parametric amplifier technology was rendered meaningless by these developments and became obsolete. In short, the best part of an EME station could be purchased from a dealer

This article will now go on to examine the minimum requirements for the development of an EME station, by way of technical specifications and operating practices, which are particular for this form of communication.

## 2. BASIC TECHNIQUES

The serious enquirer must first of all be informed that EME contacts, because of the large path distances involved, are undertaken for the most part, with extremely low-level, received signals. The construction of an EME station has, therefore, to be conducted with the utmost care. This applies particularly to the send and receive antenna losses which must be reduced to the absolute minimum possible. In order to be able to optimally equip an EME station, the following factors must first of all be given an airing:

- Frequency range
- Path loss
- Signal/noise ratio
- Transmit power
- Antenna gains

$$\text{path\_loss\_dB} = 10 \cdot \text{LOG}(\text{received\_pwr\_watt}/\text{erp\_watt})/\text{LOG}(10)$$

Formula 1: Total path loss calculation

$$\text{free\_space\_db} = 37 + 20 \cdot \text{LOG}(\text{freq})/\text{LOG}(10) + 20 \cdot \text{LOG}(\text{dist})/\text{LOG}(10)$$

Formula 2: Free space attenuation calculation

where: freq = frequency (MHz)

dist = distance in miles (km x 0.62)

### 2.1. Frequency Range and Path Loss

The first consideration for a newcomer to moon-bounce is the important choice of a suitable frequency band. When this has been resolved, in principle, the whole station equipment, EME transmitter and receiving installation being dependent upon frequency, can then be considered. The propagation loss, earth-moon-earth, also has a bearing upon the choice of band etc.

The basis for calculation is tabulated:—

Frequency MHz	Perigee dB	Apogee dB
144	- 251.5	- 253.5
432	- 261.0	- 263.0
1296	- 270.5	- 272.5
2304	- 276.0	- 278.0

Table 1: EME path losses for various amateur bands

It may be seen from the table 1 that enormous propagation losses are caused by both the path distances involved and the reflection loss from the moon's surface. See formulae 1 and 2.

Owing to the nature of the moon's surface, rocky, sandy and dry, it has a very low reflection characteristic. Experiments in USA (2) have revealed that the moon's surface reflects only 7 % of the energy falling upon it, the remaining 93 % being



$$\text{erp} = \text{tx\_pwr\_watt} * \text{EXP}((\text{ant\_gain\_db} - \text{cable\_loss\_db}) / 10 * \text{LOG}(10))$$

Formula 3: The ERP derived from sender-power, antenna-gain and cable-attenuation calculation. (ERP = effective radiated power)

totally absorbed. This proportion varies with frequency, and is more favourable at the lower bands.

Additional reflection losses are caused by the ragged nature of the moon's surface on the incident and the reflected angle (should be equal) of the incoming energy. A massive part of the transmitted energy is also dispersed into space by the electromagnetic field surrounding earth. The total reflection loss can be deduced from the difference between the total loss and the calculated theoretical free-space loss.

The 144 MHz band exhibits the smallest loss on the amateur bands under question, and therefore presents the best pre-requisites for a satis-

factory S/N. This band therefore possesses the greatest world-wide EME activity at present. The article will now concentrate on this band as it is clearly the one to be preferred for an introduction to EME.

## 2.2. Signal/Noise Ratio

In order to cover the 790.000 km (approx.) distance and to overcome reflection losses, a certain radiated power is required. This is known as "effective radiated power" (ERP) and is the product of the transmitter power and the antenna gain (cable losses being taken into account). See formula 3.

$k = 1.38E-23$ (Boltzmann-Konstante $1.38 \times 10^{-23}$ Joule/Kelvin)
Eingabe-Routine - Systemparameter
path_loss_db = Gesamte Streckendämpfung - dB rx_width_hz = 2F/NF-Empfängerbandbreite - Hz rx_noise_fig_db = Empfänger-Rauschzahl - dB rx_ant_gain_dbi = Empfangseitiger Antennengewinn - dBi rx_cable_loss_db = Kabeldämpfung - Ant./Vorverstärk. - dB ant_noise_temp = Antennen-Rauschtemperatur - K tx_pwr_watt = Sender-Ausgangsleistung - Watt tx_ant_gain_dbi = Sendeseitiger Antennengewinn - dBi tx_cable_loss_db = Kabeldämpfung zwischen Ant. u. PA - dB
Berechnungs-Routine
tx_pwr_dbw = $10 * \text{LOG}(\text{tx\_pwr\_watt}) / \text{LOG}(10)$ total_rx_noise_db = $\text{rx\_noise\_fig\_db} + \text{rx\_cable\_loss\_db}$ rx_noise_temp = $290 * (\text{EXP}(\text{total\_rx\_noise\_db} * \text{LOG}(10) / 10) - 1)$ sys_noise_temp = $\text{rx\_noise\_temp} + \text{ant\_noise\_temp}$ rx_noise_pwr = $10 * \text{LOG}(k * \text{rx\_width\_hz} * \text{sys\_noise\_temp}) / \text{LOG}(10)$ s_n = $\text{tx\_pwr\_dbw} - \text{tx\_cable\_loss\_db} - \text{tx\_ant\_gain\_dbi}$ s_n_ratio = $\text{s\_n} - \text{path\_loss\_db} + \text{rx\_ant\_gain\_dbi} - \text{rx\_noise\_pwr}$
Print-Routine
PRINT "Sender-Ausgangsleistung "; tx_pwr_dbw; "dBW" PRINT "System-Rauschtemperatur "; sys_noise_temp; "K" PRINT "Empfänger-Rauschleistung "; rx_noise_pwr; "dBW" PRINT "Signal/Rausch-Verhältnis "; s_n_ratio; "dB"

Fig. 1:  
Program proposal for  
calculating signal/noise



System data input	
Path loss	251.5 dB
Receiver IF bandwidth	200 Hz
Pre-amplifier noise figure	1.5 dB
Receive antenna gain	22.1 dBi
Receive cable loss	0.2 dB
Antenna noise temperature	170 K
Transmit output power	750 W
Transmit antenna gain	22.1 dBi
Transmit cable loss	0.5 dB
Calculated values	
Transmit output power	28.75 dBW
System noise temperature	308.94 K
Receiver noise power	-180.69 dBW
Signal/noise ratio	+1.64 dB

Fig. 2:  
Calculation example with program  
from fig. 1

The signals at the receiver input, in order that they are intelligible, must be above the system noise power. This may be considered as being the total received signal-to-noise (S/N)

The signal-to-noise ratio is derived from values for transmit power, path loss, receiver noise figure, receiver bandwidth, antenna gain, antenna noise temperature and coaxial cable losses. The formulae in fig. 1 have been prepared for a direct input into a calculator in order to obtain the S/N from these factors. Varying the parameters in the program input of fig. 1 will enable various transmission situations to be simulated. Fig. 2 gives an example. It will be seen that received EME signals, for the most part, lie about the limit of intelligibility. This must be borne in mind during the construction of an EME station. The following will take this into account in the practical considerations for a moonbounce station.

### 2.3. The Transmitting Equipment

The send power is limited by the authorities for B-licence holders, to 750 Watt at present. This power represents, in connection with the feasible antenna gains, a usable compromise in the achievement of an effective radiated power output. Final power amplifiers in this power class may be obtained directly from a dealer. Before

obtaining one, however, it should be ascertained that the rated power output is not attained at the very limits of the linearity characteristics of the final tube(s). Driving such an amplifier to its limits will result in the production of unwanted (and illegal) side frequency products which could interfere with other radio services (TV, radio broadcasting etc.). Furthermore, the longevity of the final tube(s) will also be drastically curtailed. CW allows full ratings to be achieved with reasonable dimensioning of PA cooling and power supplies and is almost the only mode which can be considered.

Home-construction is very often a better technical and economical solution, the necessary components all being available from specialized firms. Some parts may be already in the station store or can be obtained from other amateurs. Nevertheless, in connection with some mechanical parts, particularly the construction of the coaxial tuned circuits, there are a few demands which have to be met.

Suitable construction kits are obtainable from any of the referenced sources (1, 4, 6). The PA driver can be any modern 2 m or a short-wave transceiver, the latter being employed with a down-converter. These units may be regarded as being pre-requisites for a normal amateur station and do not therefore have to be specially home-constructed for EME use.



$\text{noise\_temp} = 290 * (\text{EXP}(\text{noise\_fig} * \text{LOG}(10)/10) - 1)$
$\text{noise\_fig} = 10 * \text{LOG}(1 + (\text{noise\_temp}/290)) / \text{LOG}(10)$
$\text{noise\_factor} = 1 + (\text{noise\_temp}/290)$

Fig. 3:  
Calculator input formula for the conversion to noise temperature, noise figure and noise factor. The value 290 is the reference temperature 290° K i.e. 17° C.

#### 2.4. Receiving Equipment

The following points are to be given special attention in the setting-up of the receive side of an EME station:-

- 1 Receiving system noise figure (NF)
- 2 Selectivity of the receive converter
- 3 Frequency accuracy and stability

For successful EME reception, the optimal noise matching of the receiver to the antenna gain is of crucial importance and has a decisive impact upon the overall signal-to-noise ratio (S/N).

Depending upon both the quality and length of the coaxial feed cable between antenna and receiver, there will be some attenuation of the signal before it reaches the receiver pre-amplifier. This attenuation adds directly to the receiver

noise figure (NF). In order to reduce this loss to a minimum, it would be a good idea to place the receiver directly at the antenna output terminals, but this is not practical! As the receiver pre-amplifier is the main item which determines the system noise figure, it is a logical step to install it at the output terminal of the antenna, well away from the receiver.

This philosophy is behind the current state of technology where a so-called mast amplifier combined with transmit-receive switching, is widely employed. This concept enables an optimal signal/noise ratio and offers at the same time the convenience of having the receiving unit in the shack.

A successful EME station should have an NF of better than 1.5 dB. This high specification might

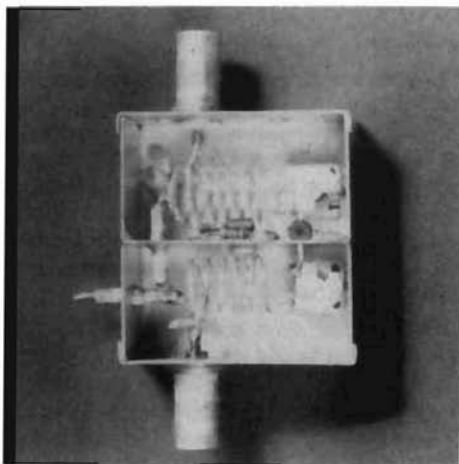


Fig. 4: 144 MHz pre-amplifier using a BF 981

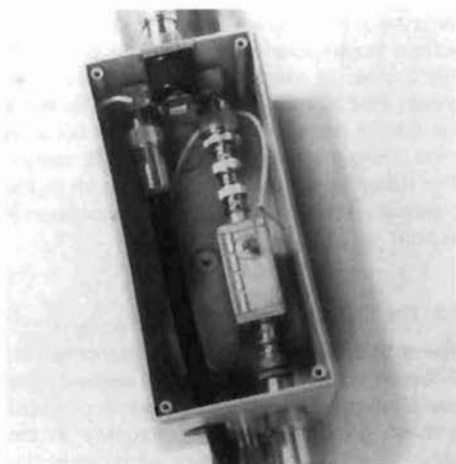


Fig. 5: The pre-amplifier enclosed together with a coaxial relay





seem to be unattainable but the most modern semiconductors achieve an NF of under 1 dB. It must be borne in mind, however, that at 144 MHz, the minimal cosmic noise temperature is already 150 K, i.e. equivalent to a noise-figure of 1.8 dB (fig. 3) so it seems pointless to aim for a much lower figure than this. The receiver noise above 1 GHz is higher than the cosmic noise and therefore the lowest possible receiver noise figure should be aimed for (8, 9).

Suitable pre-amplifiers for 144 MHz may be constructed quite economically using the MOSFET BF 981 (fig. 4). A constructional article can be found, for example, in (7). Proprietary pre-amplifiers possess, in the main, a really outstanding noise figure. They are fitted with a send/receive relay which all too often does not afford a sufficient degree of isolation between send and receive. This defect has caused the author's pre-amplifier to fail on a few occasions.

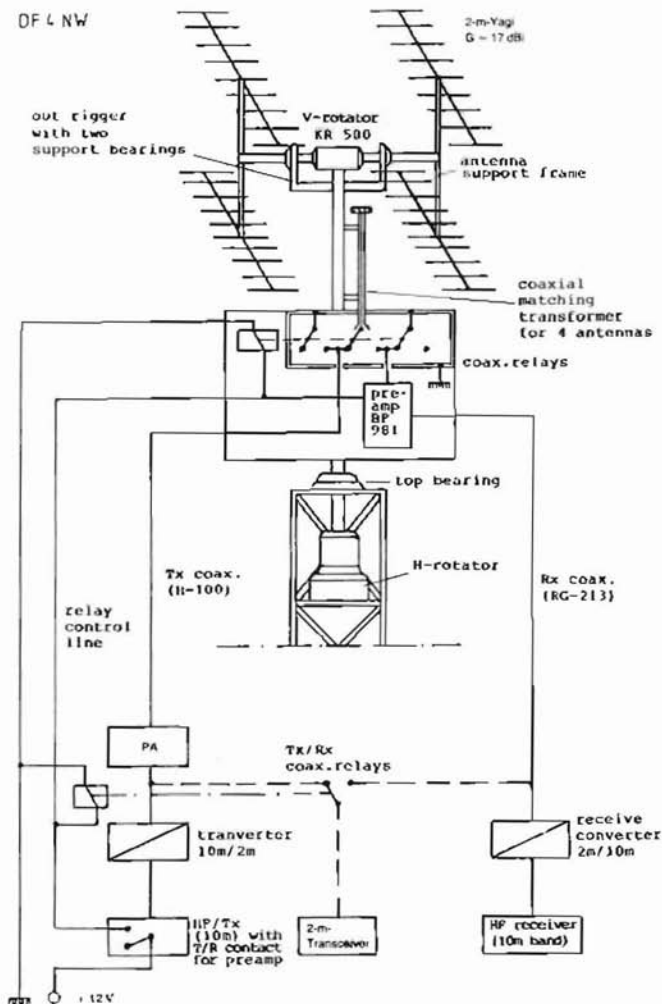


Fig. 6:  
Example of a 144 MHz, EME  
Installation with either an HF  
transceiver and 10 m - 2 m  
transverter or with a 2 m  
transceiver

The only way round the problem is to change the antenna relay for a high-quality type (fig. 5).

By way of protection, the wiring and cabling of this relay's transmit path is taken via an idle contact. This ensures that the output of the PA is switched to the antenna and the receiver input to earth when the relay drops out. A further effective measure to isolate HF from the pre-amplifier, is to use a separate transmit cable from the power amplifier to the antenna relay. A second coaxial relay is necessary at the equipment end to switch the transceiver over the two paths. The arrangement is shown in fig. 6.

On grounds of cost, and the low outlay in terms of labour, it is recommended that the pre-amplifier should be home-constructed.

Any modern 2-metre transceiver, having two VFOs or calibrated receiver fine-tuning, could be suitable. An arrangement such as that shown in fig 6, with a converter translating 144 MHz down

to a short-wave receiver at 28 MHz, is also just as good. The main requirement is good selectivity and frequency stability. As the aerial gain is responsible, in the first instance, for a good signal-to-noise ratio, additional measures are required in order to boost the signal over the noise level. This is achieved by reducing the received bandwidth to that necessary for the reception of CW signals i.e. 300 Hz. This may be done either by filters in the IF or in the audio stages of the receiver.

## 2.5. Antenna System

As the available RF power is curtailed by law, and the noise figure of the receiver by natural limitations, the most important element of the EME station is the antenna. The well-known axiom that the antenna is the best pre-amplifier, is particularly apt in this application. The construction of



Fig. 7:  
The author's 4 x 17-el. Yagi,  
EME array.



a high-performance antenna should utilize all of the available space. The best solution would be to mount it on a dedicated, self-supporting mast. It is possible, however, with not much extra difficulty, to mount an EME antenna on a house or garage roof (fig. 7). The height above ground is not the most deciding factor. What is important, however, is the screening of the antenna by nearby objects, such as buildings and trees etc., should be avoided right down to the horizontal plane ( $0^\circ$ ) if possible. The high-performance EME antenna would then be also suitable for direct terrestrial communications.

Experience has shown that for the 144 MHz band 20 dB<sub>0</sub> antenna gain is the minimum that should be aimed for. On the higher bands, a greater antenna gain is necessary in order to compensate for the larger path losses, see table 1.

This gain cannot be realized with one single Yagi antenna. The ideal would be a dish antenna with a suitable paraboloid aperture. This solution is normally precluded on grounds of cost and space considerations.

An economical compromise solution is presented by the antenna array. This can comprise several proprietary Yagi antennas together with the necessary combining and matching networks. In order to fulfill the 20 dB<sub>0</sub> gain requirement, four 15 dB<sub>0</sub> element Yagis are required for the array. The boom must be at least 5 wavelength long i.e. about 10 metres. A 4-element array, theoretically, has a 6 dB gain advantage over any of the elemental Yagis which compromise it. In practice, this is reduced to 5 dB<sub>0</sub> which still amounts to the required total of 20 dB<sub>0</sub>.

The 4-antenna array is constructed quadratically in the so-called H-form. This arrangement enables a well-formed, symmetrical radiation lobe thus ensuring the proper illumination of the reflecting zone on the moon's surface. A rule to follow for the location of the elemental Yagis is that they should be disposed symmetrically about the central support and 5/7ths of the boom length apart in both the horizontal and the vertical planes (10).

This construction allows very uncomplicated mounting arrangements for the elevation rotator on the cross support boom. High quality H-100



Fig. 8:  
The pre-amplifier/  
relay unit from fig. 5  
affixed to a coaxial  
matching trans-  
former

coaxial cable and N-type plugs and sockets should be used for the feeder and for the transformer connections, see fig. 8.

When planning an antenna installation, it may be just as well to consider personnel expectations of the likelihood of EME contacts, with the reality. This is presented in table 2 in the form of what antenna installation the stations at the two ends of a link are likely to have for a 100 % solid contact.

Basically, it may be deduced that the more attention given to the antenna, thereby increasing

Station A	· EME ·	Station B
Number of antennas	Traffic between	Number of antennas
4-ant. array	↔	16-ant. array
8-ant. array	↔	8-ant. array
16-ant. array	↔	4-ant. array

Table 2: EME contact possibilities for various link array combinations



the ERP and the received signal-to-noise ratio, the greater will be the possibility of working large numbers of stations. Naturally, it cannot be ruled out that certain favourable positions of the moon allow a solid contact between two parties both having only four stacked antennas.

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### 3. ANTENNA CONTROL

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For EME contacts it is imperative that the antenna is directed accurately at the moon and that it follows the moon in the course of its movements. The construction of the antenna control is based upon the following two principles:

- Polar mounting
- AZ/EL mounting

The polar mounting has the vertical axis (horizontal rotator) pointing towards the earth's- or polar axis, and the vertical rotator towards the axis of the declination. This system is in widespread use for the mounting of astronomical instruments. The system allows, above all, for the antenna to be directed by inputs of only declination and the Greenwich Hour Angle (GHA) for that particular station latitude and longitude. This information may be obtained from astronomical or nautical tables (11). This system has, however, the important disadvantage that the antenna array is limited for terrestrial applications.

The almost universally accepted alternative to the polar mounting is the AZ/EL arrangement. The horizontal axis of the antenna is orientated vertically about the centre point of the earth in this method of mounting.

This is the normal way in which terrestrial antennas are mounted on masts and poles etc. Proprietary motor drives, but always in a more robust version, serve for the AZ and the EL orientation of the antenna array.

Movement in elevation (EL) is particularly well carried out by the rotator KR-500 (A, B). It is recommended, however, that the shaft is extended and extra support bearings fitted to an

out-rigger in order that the load is taken off the rotator's die-cast housing and the rotator gearing. Larger arrays require more rotator systems working on a single axis in order to provide the necessary torque.

A large choice is available in dealer outlets for the horizontal rotator. Again here, it is advisable to use an out-rigger and extend the driven shaft to allow the fitting of extra support bearings in order to relieve the load on the rotator mechanism.

To allow a greater starting torque for large arrays, and also to obtain a more precise control over the setting in both AZ and EL, it is desirable to increase the gear ratios of the driver mechanisms. This can be done mechanically by changing the gear train or, the same effect can be achieved by using an electronically regulated supply to the motor. The electronic method is, of course, the one to be preferred, as the mechanical rearrangements to the gear ratios can cause all sorts of problems. The electronic control of the motor supply is preferably carried out by the pulse technique. The principle is based upon a Triac placed in series with the motor supply current which is controlled at a varying rate by a pulse generator. By changing the impulse duration, a smooth control will be effected.

The highly concentrated beams from these arrays, demand a high setting and monitoring accuracy for the AZ and EL headings. The normally scaled instruments of 5° per division are not adequate for this application. In order to increase the read-out resolution, it would be better to lay out a modest investment in a digital display based upon a digital voltmeter. This can bring up the resolution to 1°. By suitably switching in preset potentiometers, it is quite a simple matter to get the AZ indicator to read 0 - 360° and the EL to read 0 - 90°.

#### 3.1. Tracking the Sun and Moon

Most radio amateurs in the world are almost always obscured by clouds from sight of the sun and moon, when they are required. For this reason, an optical antenna alignment can be ruled out and a mathematical method developed. This is based upon the longitude and latitude



of the EME station and calculated from the Greenwich Hour Angle (GHA) and the declination

Owing to the continuous movement of the moon in its orbit, this calculation has to be continuously updated. This can be done with a pocket calculator but can be enormously time-consuming. However, personal computers are now becoming popular owing to the burgeoning progress in data-processing and find a home in large numbers of radio amateurs' shacks. These computers are ideal instruments to undertake the task of calcu-

lating this sort of thing. Programs, such as "EME" for PCs, by DF 4 NW, (figs. 9 and 10), can process the data with great precision. The DF 4 NW programs can calculate in real-time, once the date and time have been given in GMT (UTC). For two freely chosen locations, the following data is presented

- AZ and EL angles for sun and moon
- Angle between moon and sun from observer
- GHA and declination for sun and moon
- Semi-diameter and distance to the moon

TUE, 16. FEB. 1988		TIME-UTC 07:31:30	
1ST. LOC. LAT. 49.71		LONG. -10.82 DEG	
EL-MOON	8.33	AZ-MOON	140.99 DEG
EL-SUN	8.69	AZ-SUN	121.38 DEG
SPACING BETWEEN MOON & SUN		19.61 DEG	
2ND. LOC. LAT. 50.00		LONG. 100.00 DEG	
EL-MOON	-55.25	AZ-MOON	47.14 DEG
EL-SUN	-51.79	AZ-SUN	14.84 DEG
SPACING BETWEEN MOON & SUN		32.49 DEG	
GHA-MOON	307.17	DEC-MOON	-21.84 DEG
GHA-SUN	289.34	DEC-SUN	-12.57 DEG
SD-MOON	16.675	DIS-MOON	358332 KM

Fig. 9: Screen print-out from the "EME for PCs" computer program. The 12 lines show the following data:

- 1) date and time
- 2) first location, latitude and longitude
- 3) first location, elevation and azimuth
- 4) first location, sun's elevation and azimuth
- 5) first location, angle between sun and moon
- 6)...9) data as above for the second location
- 10) GHA (hour angle) and declination of the moon
- 11) GHA and declination of the sun
- 12) Semi-diameter (SD) and distance (DIS) of the moon (elevation of the moon over the horizon)



DATE : TUE, 16.FEB.1988		E M E - COPYRIGHT BY DF4NW - JAN.1988								
		LAT 49.70 LNG -10.82			LAT 50 LNG 100			DEG		
TIME UTC	EL-MOON	AZ-MOON	SPACING MOON/SUN	EL-MOON	AZ-MOON	SPACING MOON/SUN	DIST. KM	S.D. MOON	DECL. MOON	
07:30	8.174	140.8	19.61	-55.4	46.64	32.58	358333	16.675	-21.84	
07:45	9.651	143.7	19.59	-53.6	51.44	31.52	358315	16.676	-21.79	
08:00	11.02	146.8	19.59	-51.7	55.91	30.34	358297	16.676	-21.74	
08:15	12.29	150.0	19.60	-49.7	60.09	29.12	358279	16.677	-21.69	
08:30	13.45	153.2	19.61	-47.7	64.01	27.90	358261	16.678	-21.64	
08:45	14.49	156.5	19.62	-45.6	67.70	26.71	358244	16.679	-21.59	
09:00	15.41	159.8	19.62	-43.4	71.19	25.57	358227	16.680	-21.54	
09:15	16.20	163.2	19.61	-41.2	74.51	24.50	358210	16.680	-21.48	
09:30	16.86	166.7	19.59	-38.9	77.67	23.50	358193	16.681	-21.43	
09:45	17.38	170.2	19.55	-36.6	80.71	22.58	358176	16.682	-21.38	
10:00	17.76	173.7	19.48	-34.3	83.65	21.74	358159	16.683	-21.33	
10:15	17.99	177.2	19.40	-32.0	86.49	20.97	358143	16.684	-21.28	
10:30	18.08	180.8	19.28	-29.7	89.25	20.27	358127	16.684	-21.22	
10:45	18.03	184.3	19.15	-27.3	91.96	19.64	358111	16.685	-21.17	
11:00	17.83	187.9	18.99	-25.0	94.61	19.07	358095	16.686	-21.12	
11:15	17.49	191.4	18.81	-22.7	97.23	18.55	358080	16.686	-21.06	
11:30	17.01	194.9	18.61	-20.4	99.82	18.08	358064	16.687	-21.01	
11:45	16.39	198.4	18.40	-18.1	102.4	17.67	358049	16.688	-20.96	
12:00	15.64	201.8	18.18	-15.8	104.9	17.29	358034	16.688	-20.90	
12:15	14.75	205.2	17.96	-13.5	107.5	16.96	358019	16.689	-20.85	
12:30	13.75	208.5	17.74	-11.3	110.1	16.68	358004	16.690	-20.80	
12:45	12.62	211.8	17.53	-9.17	112.6	16.40	357990	16.691	-20.74	
13:00	11.38	215.0	17.32	-7.02	115.2	16.17	357975	16.691	-20.69	
13:15	10.04	218.2	17.12	-4.91	117.8	15.97	357961	16.692	-20.63	
13:30	8.598	221.3	16.93	-2.85	120.6	15.79	357947	16.693	-20.58	

Fig. 10: Print-out from "EME for PCs" program

In this program mode the same data is given as was displayed in the screen mode. Following the input of a freely selectable time-frame, a calculation of the movements of sun and moon may be obtained for the planning of QSO schedules.

#### 4. PREPARATION FOR INITIAL EME TESTING

At this point, it is as well to find out whether or not the whole system is capable of EME operations. The following tests will show, without the use of test equipment, the equipment's capability in this respect.

- 1 Measurement of the sun's noise
- 2 Echo trials

The simplest, but also the most effective, method of analyzing the receive system sensitivity is to measure the sun's noise (13). For this measurement, the antenna is directed towards the sun. A

high elevation angle (mid-day) will ensure the least atmospheric and man-made interference with the test results. The received noise should clearly increase by at least 5 dB using a previously calibrated S-meter. This figure depends upon the sunspot activity in the surface of the sun as indicated exhaustively in ref. (13).

The so-called echo tests are a little more problematical. The antenna is aimed at the moon and a short burst (2 secs) of CW pulses are transmitted at full power. After a period of 2 1/2 seconds, the echo pulses should be detectable in the receiver output if the conditions are optimal. That the return pulses are extremely weak, or indeed, not to be heard at all, should not be taken as poor indication of the system's performance - at least, not yet! There are various factors which could be responsible for a non- or weak return



when using only a 4-element array. An EME contact could still be established under such conditions, but only if the distant station is using a 16-element array and/or with high power. Before undertaking either an echo test, or attempting to contact other EME stations, it is advisable to be acquainted with these external influences. The following factors are to be particularly taken into account:

#### 4.1. The Sun's Noise

When the sun is at its closest approach to the moon (new moon), the sun's emitted noise will increase to such an extent, that it will influence the EME reception until it becomes impossible to establish a radio link. The sun/moon separation will cause noise to an extent depending upon the antenna beamwidth. Practical experience has shown that angle distances below  $10^\circ$  are to be avoided. In addition, winter and night-time operations are, of course, favourable.

#### 4.2. Perigee and Apogee

As may be seen from table 1, the path-loss at perigee is 2 dB smaller than at the moon's apogee. This 2 dB, in the context of EME working, can mean the difference between a contact or no-contact! Typical values for a perigee is 360,000 km (SD = 16.5). See fig 9.

#### 4.3. Angle of Reflection

The optimal antenna orientations at the two stations participating in a contact, is when they have a common elevation angle. This follows from the angle of incidence at the moon's surface being almost the same as the angle of reflection for radio waves.

#### 4.4. Low Angles of Elevation

Low angles of elevation should be avoided in order to minimize reception of earth noise and other terrestrial disturbances. Under favourable conditions, however, a kind of tropo-propagation mode may be encountered, leading to extremely long-distance contact possibilities.

#### 4.5. Declination of the Moon

Owing to the smaller distances involved between observer and moon, northern declinations of the moon, for stations in the northern hemisphere, bring better path conditions. The same applies for southern moon declinations for stations in the southern hemisphere.

#### 4.6. Galactic Noise

Up to 170 K noise temperature is to be expected when the moon, twice monthly for a period of 4 days, crosses the galactic plane. When the crossing occurs in the declination of Orion and Gemini for the northern hemisphere, even higher galactic noise temperatures are to be expected. Likewise, when the moon occults Scorpio and Sagittarius in the southern hemisphere, a higher noise will be received (3, 9, 11, 12).

#### 4.7. Delay Distortion

Owing to the reflection taking place on the rough and uneven moon's surface, the signal experiences path-time variations which are within the reflected wave. This causes random phase distortions such that the return CW signals from the moon sound very rough. SSB voice signals are, nevertheless, readable but only under good signal conditions.

#### 4.8. Faraday Polarization Reversal

A polarity reversal occurs in the earth's atmosphere. This can cause the signal to disappear entirely under certain conditions. This effect can be countered by employing circularly polarized antennas (parabolic, or crossed-Yagis). The extra complication involved at the antenna, particularly using Yagi based arrays, is fairly high but unavoidable.

#### 4.9. Doppler Effect

The moon follows an elliptical orbit around the earth. Within the orbit, however, the moon wobbles, causing a periodical distance variation at a speed of up to 1575 km/h (see par. 4.2.). This causes a doppler phase shift of  $\pm 210$  Hz max. in the 144 MHz band.



## 5. OPERATING TECHNIQUES

In the development of EME contact procedures, a few departures from the normal form of amateur contacts take place using the CW mode. Owing to the use of narrow-band selective receiver filters and the prevailing signal distortion, it is necessary that only low keying speeds are employed. The highest transmitted keying speeds are from 30 to 50 BpM. The use of very low keying speeds also has the effect of a reduction in intelligibility, as a combination of both signal drop-outs and delay distortion could result in a dash sounding like several dots.

EME traffic is conducted mainly at frequencies at the beginning of the band. At 2 metres, this extends from 144 000 MHz to 144 050 MHz. Most activity appears to be concentrated, at or around 144.010 MHz. A similar band plan exists for EME working at 70 cm.

As EME stations were not too frequent a few years ago, it was necessary to pre-arrange a contact. The system went as follows:-

The most eastern station begins to call on the full hour (when other arrangements have not been made). This takes place in an unvarying cycle of exactly two minutes calling, followed by an exact period of two minutes listening. Should a contact ensue, the call-sign is repeated for 1 1/2 minutes and a further 30 seconds is taken up by the report (fig. 11). Following an exchange of reports, both stations send a final confirmation in turn, by sending a series of 'R's' followed immediately by the report, for a period of 1 1/2

minutes. The next 30 seconds is followed again by the call sign and the final 'K' or 'SK'.

Contacts in the 432 MHz band are conducted in 2 1/2 minute periods i.e. two minutes followed by an information change for the remaining 30 seconds. This procedure is repeated as long as necessary in order to achieve an intelligible contact. The contact is regarded as being satisfactorily completed when, both stations have confirmed the distant call sign and their report by sending an 'R' to the partner station. A comprehensive exchange of information is only possible, normally, under good signal-strength conditions. It happens very frequently that the whole QSO process, as outlined above, could take over an hour under poor path conditions.

Because of the very weak signals and, as a result, the difficult nature of communication, a simple reporting system has been devised for EME working. Instead of the RST system, the TMO system (fig. 11) is employed.

A communication net (EME-net) has been established on 14.345 MHz for those wishing to find partners for an EME contact attempt. This net finds its greatest activity at the weekends or holiday periods. If a schedule (sked) cannot be adhered to, it is, of course, morally obligatory to try and let the other party know in good time. This can be achieved by telephone, telegram or via other OMs on the band. Nothing is more annoying, than to listen for hours to receiver noise only to discover later that the other party was not QRV. Be reliable, is the catch-phrase here!

Skeds on the lower-frequency bands (144 and 432 MHz) are only very seldom possible in the

TMO reporting system
T = Signal heard but not readable M = Signal partly readable O = Everything readable
R = Call-sign and report received SK = End of contact

Fig. 11: The EME reporting system "TMO"





weekdays At 144 MHz, many of these QSOs occur just on the off-chance when the moon is at its perigee. Recently, an annual autumn EME contest has been organized. These contests should be very interesting for any prospective EME operator to hear the extremely rough sounding signals and to become acquainted with the fascinating EME *modus vendi*.

## 6. DESCRIPTION OF AN EME STATION

During the course of tests conducted with many stations by the author, the following station equipment was used –

### Receive equipment

1. Home-constructed mast-preamplifier using a BF 981
2. Receive down-converter 144/28 MHz – Microwave Modules 144 MHz converter.
3. HF receiver – IF bandwidth 4 kHz to 300 Hz – Drake R7
4. AF processor – 3 kHz to 100 Hz; Signaltrans Nachrichtentechnik – 7901 Bollingen
5. Headphones – Sennheiser

### Transmit equipment

1. Transmit driver – Transceiver: Braun SE 402
2. Transmitter final – Dressler d200s
3. Morse-key (electronic memory) – Accu Keyer
4. Morse-key (bug) – Bencher

### Antenna Installation

1. Test 4 x 13 element, 4.5 m boom-length – Tonna
2. Test 4 x 17 element, 6.6 m boom-length – Tonna
3. Matching network for 4 Yagi antenna – Andes
4. Horizontal rotator – CDE T<sup>2</sup>X
5. Vertical rotator – Kenpro type KR 500
6. Vertical rotator support bearings – Kenpro
7. Horizontal rotator support bearings – Kenpro

Owing to the local authorities' reluctance to grant a building permit for the construction of a lattice mast, the antenna had to be mounted on the roof of the author's house. The mast itself, is a proprietary steel telescopic with a diameter of 48/60 mm and 2 x 3 metres long. The array main-frame

is constructed from high-strength aluminium tubing (AlMgSi 0.5-tube F22 – 40 x 2 mm dia.) using the necessary cross clamps. A guy-rope system of six wires ensures that the installation stays in place on the roof.

All the HF cabling uses RG-213/U co-axial. In order to avoid losses in the 30 m co-ax between the cellar shack and the mast, the power amplifier was mounted in the attic and remotely controlled. On the receive side also, the cable losses were reduced to the absolute minimum by mounting the pre-amplifier as close to the array as possible. It is to be expected that another few tenths of a dB could have been saved by using a low-loss cable, such as H 100, in the antenna change-over switching and in the transmit leg.

As may be seen from the list of equipment used, tests were carried out using two different arrays. A very interesting trans-continental contact was carried out using the 4 x 13 array but the number of stations contacted increased noticeably when the array was modified to a 4 x 17 antenna.

Finally, it may be said that EME working is entirely feasible with a minimum outlay and offers the experimentally-minded amateur a very wide field. This includes antenna gain, transmit power, receiver noise-figure and selectivity, all of which demand the utmost performance and therefore the maximum endeavour.

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## The DB 1 NV Spectrum Analyser

(VHF Communications 3 and 4/1987)

### Specifications:

Frequency range:	0 - 500 MHz (1000 - 1500 MHz image)
Sensitivity:	– 100 dBm
IF bandwidth:	1 kHz / 3 kHz / 10 kHz / 50 kHz / 300 kHz
Video bandwidth:	100 Hz / 1 kHz / 10 kHz / 100 kHz
IM-free dynamic range:	≥ 60 dB
Display bandwidth:	10 / 20 / 50 / 100 / 200 kHz / per cm 0.5 / 1 / 2 / 5 / 10 / 20 MHz / per cm also 0 - 500 MHz screen display 0...70 dB in 10 dB steps (DC-2 GHz)
Input attenuator (built in):	

### Special Features:

The video filter can be automatically switched to accommodate the selected IF bandwidth. When displaying ≤ 200 kHz/cm, PLL stabilization of the 1st local oscillator may be switched in.

<b>Ready-made unit</b>	<b>DM 6.350. –</b>	(export price DM 5.571. –)
<b>Modules (complete set)</b>	<b>DM 2.995. –</b>	(export price DM 2.628. –)



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Tel. West Germany 9133 47-0. For Representatives see cover page 2



*Dragoslav Dobričić, YU 1 AW*

## 432 MHz Linear PA using 3 x 2C39BA

The linear 70 cm power amplifier described here can be built in little more than a weekend. It delivers high-frequency power as required in satellite working, for small antennas and short cables or larger antennas and longer cables. Three disc-seal triodes, which are universally obtainable and are members of the 2C39 series, work in parallel delivering 300 W output for an input drive of 15 W.

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### 1. CIRCUIT

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As may be seen in fig. 1, the three triodes operate in a grounded-grid circuit which entails the cathodes being driven in parallel. The amplifier requires only two supply voltages for reliable operation: the anode and the filament voltages. The anode voltage may be between 1.3 and 1.5 kV and the filament between 5.8 and 6.0 VAC (at 3 A).

With 1.3 kV on the anode, the anode current can be driven up to 400 mA giving an HF output power of some 300 W for 15 W drive power. It is quite possible that, if good tubes are employed, the output power will be even more but they should not be overdriven. Also, a good axial air-blower should be used for the anode cooling.

As mentioned in earlier articles, the author computes his HF circuits with the aid of a special program on the personal computer. Special attention was given to optimize the half-wave anode line so that with the given impedance the lowest possible loaded Q was obtained. Through a careful selection of the parameters, a loaded Q of 39 was achieved which, for this application, is the lowest possible value. This ensures that the anode tank circuit and the amplifier as a whole, work with the maximum efficiency.

The variable capacitors, shown in fig. 1, should have the following calculated values for optimum operation: -

C1: 4.3 pF

C2: 5.3 pF

C3: 1.4 pF

C4: 5.3 pF

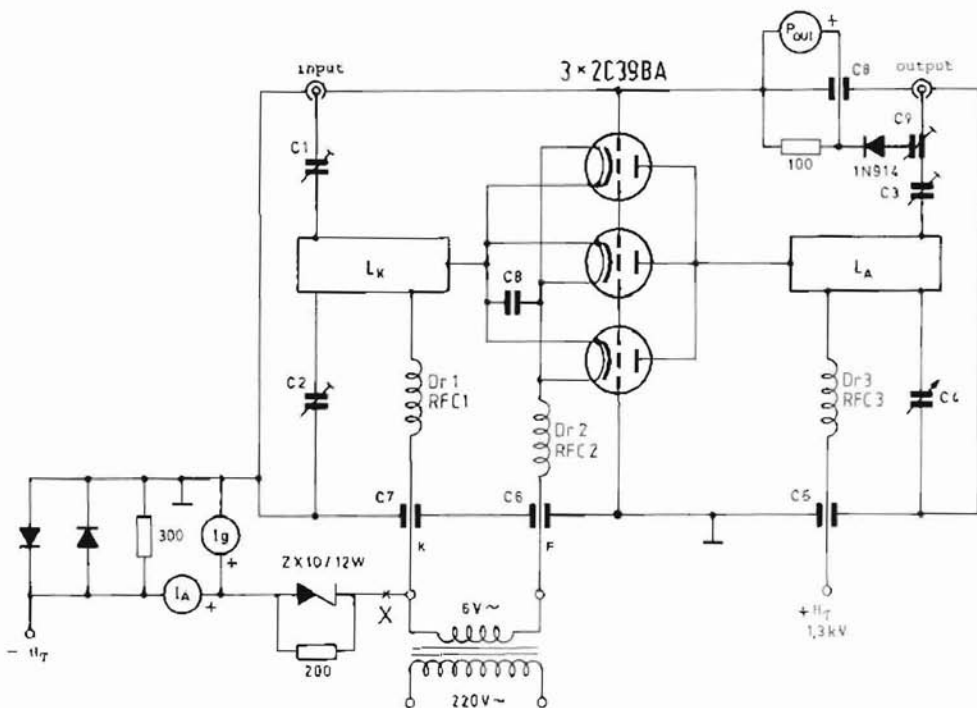


Fig. 1: This PA can deliver some 300 W at 13 dB gain in the 70 cm band

## 2. CONSTRUCTION

The dimensions for both the cathode ( $L_K$ ) and the anode ( $L_A$ ) line resonators as well as the coupling and tuning plates (C1 to C4) can be found in fig. 2.

The construction is very simple as may be seen from figs. 3 to 5. A few circuit specialities should, however, be explained.

In order that the anode resonator ( $L_A$ ) can be properly connected to the tubes, the latter should be modified in the following manner. — The cooling-fins are taken off and tapped, 4 mm are then removed from the tapped flange. The strip line can then be held tightly between the cooling-fin body and the tube's anode. The other end of the strip line is supported by a 25 mm ceramic pillar

The grid-ring contact must be made by strip contact fingers. The cathode contact, on the other hand, can be fashioned from 10 mm outer diameter copper tube of 0.5 mm wall thickness. This tube is 12 mm long and slit longitudinally down to the middle. The slotted half is then press-fitted over the cathode contact and the other end soldered to the cathode strip line  $L_K$ . The remote end of the strip line is secured to a PTFE or ceramic pillar.

The strip lines for the anode and cathode resonators are cut from 1 to 1.5 mm stock and sivered, if at all possible

The amplifier is built into a housing home-made from 1 mm thick brass plate, see figs. 3, 4 and 5. The walls are soldered together

The tuning capacitors C2 and C4 are made from 0.5 mm thick brass plate (fig. 2) and are hinged and the rotors actuated by nylon fishing lines. A

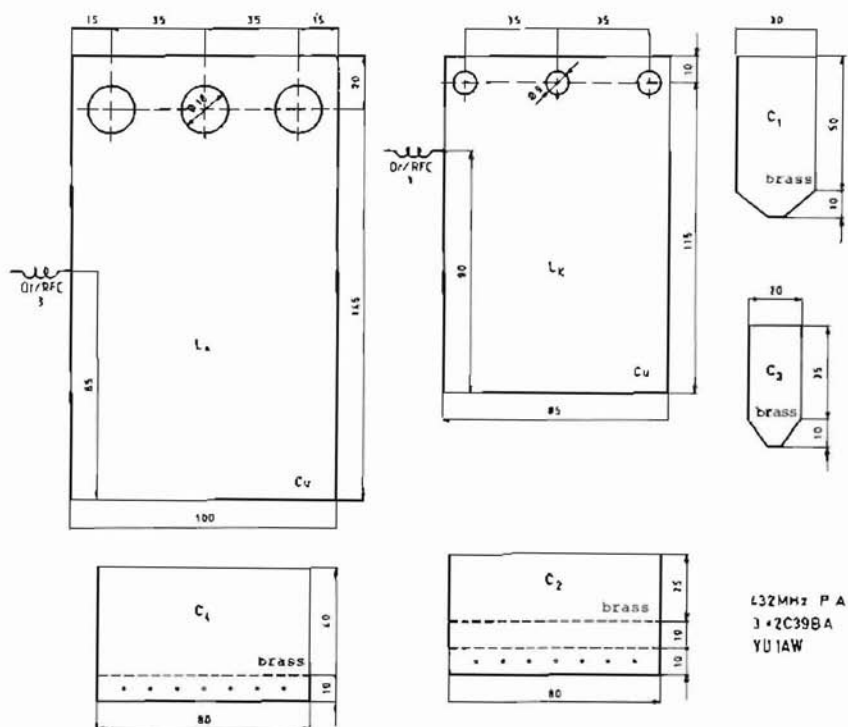


Fig. 2: The dimensions of the simplest possible housing construction

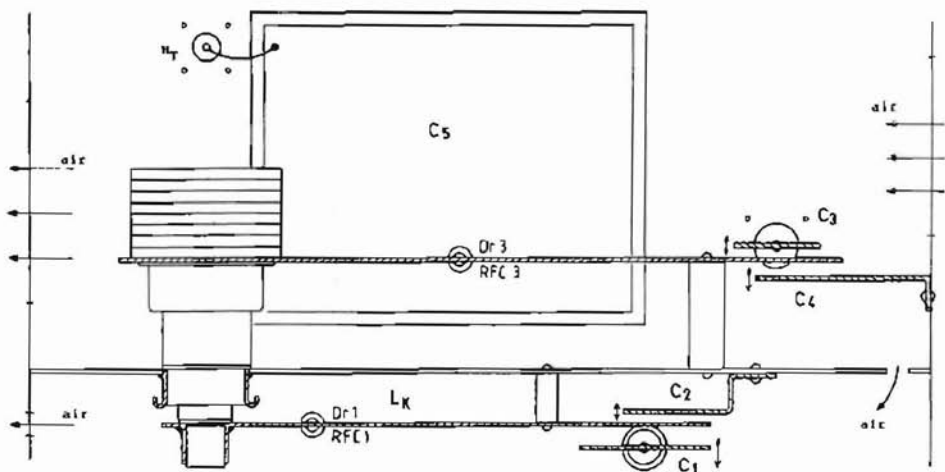


Fig. 3: PA side view

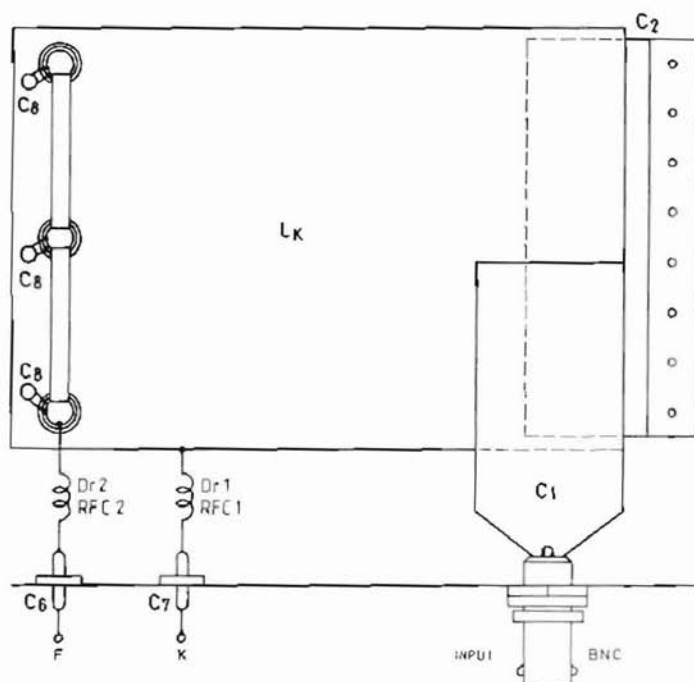


Fig. 4: View from below

piece of insulating material – PTFE or polystyrene – is suitably placed as a stop to prevent direct contact with the opposite electrode. A couple of thick knots tied in the fishing line would serve the same purpose.

These tubes require a lot of cooling air if they are to work reliably over a long period. The air blast must also be powerful in order to achieve sufficient cooling over all the cooling-fin's surfaces. The forced air comes in from above via C4 and cools both the anode resonator and the anode itself and is then vented out of the anode area. It is recommended that a couple of not too small holes be provided in the screening wall between anode and cathode enclosures (fig 3) in order to allow a weak flow of air from the mainstream to flow over the cathode resonator and cathode.

The HT supply as with the drive power is introduced into the amplifier via BNC panel sockets

An N socket is, however, used for the RF output line.

The tube heaters are connected in parallel. Between the inner heater contact and the cathode lead of every tube, a 1 nF disc ceramic (C8) is fitted using the shortest possible connections.

The RF chokes (RFCs) are wound using a 6 to 8 mm shaft with 0.8 to 1 mm dia. copper wire. They are 6 to 7 turn coils, supported from their soldered ends.

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### 3. TUNING

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The tuning of this amplifier is very straight forward – simply tune for maximum output power. This

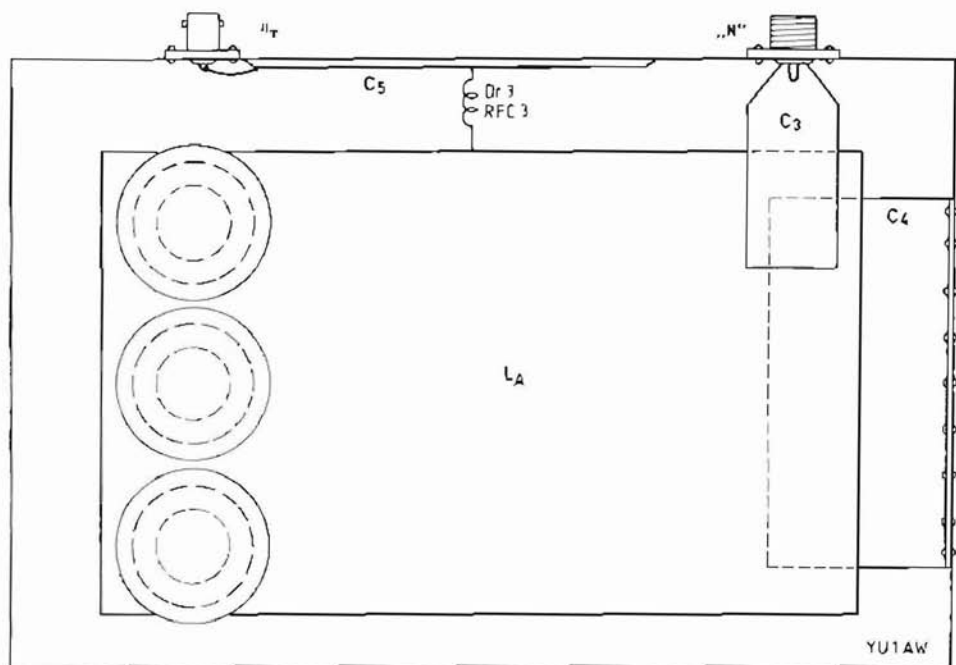


Fig. 5: View in the anode enclosure

may be accomplished with the aid of a UHF SWR-meter or by the use of the detector circuit shown

in fig. 1. The coupling (C9) to the detector is adjusted by varying the distance of the silicon diode to the N socket centre pin.

$H_T$	= 1300 V
$V_{g1}$	= - 10 to - 12 V
$V_i$	= 5.8 to 5.9 V
$I_i$	= 3 A
$I_{Ao}$	= 120 mA (per valve 40)
$I_{Amax}$	= 400 mA (per valve 130)
$I_{g1max}$	= 100 mA (per valve 32)
$P_{out}$	= 280 to 300 W
$P_{Diss}$	= 210 W (per valve 70)
$P_{Inp}$	= 520 W
$\eta$	≈ 60 %
G	≈ 13 dB

Table 1

The first tuning attempt should take place with very low input drive power and then gradually increase it to maximum when the amplifier is in tune. Table 1 shows the sort of data which can be expected

In the author's experience, the anode voltage can remain on during transmit breaks and receive periods. If a noise interference be heard in the receiver, a 10 k $\Omega$  resistor can be included in the circuit at the point marked X. This resistor must, of course, be short-circuited during transmit. Any type of available relay (post office) will do this job.



Dr. (Eng.) Ralph Oppelt, DB 2 NP

# Stabilizing the VCO Frequency by Means of Monostables

## Part 1

Expressions such as „monostable multivibrator“ or „monoflop“ have, since digital chips became cheap and ubiquitous, encroached into the mainly analogue world of HF circuit technology, e.g. (1) and (2). It is therefore appropriate to conduct a short overview of the principles of working and particular characteristics of such modules in their function as frequency stabilizers as well as – and without any claim to completeness – a series of application possibilities. A second article will concern itself with a constructional project of an SSB VCO for the 2-metre band.

### 1. PRINCIPLE

Fig. 1 shows the block diagram of a VCO stabilized with the aid of a monostable. The frequency range  $f_L \dots f_H$  of the VCO is translated by means of a stable reference frequency  $f_{x0}$ , derived from a

crystal oscillator (XO), to a relatively low frequency. The output of the first low-pass filter LP<sub>1</sub> suppresses the sum frequencies  $f_L + f_{x0} \dots f_H + f_{x0}$  and passes the difference frequencies  $f_L - f_{x0} \dots f_H - f_{x0}$ .

Instead of the „upright“ mixing occasioned by  $f_{x0}$  being smaller than  $f_L$ , a small modification to the feedback loop allows an „inverted“ mixing process whereby  $f_{x0}$  is greater than  $f_H$ . [The terms „upright“ and „inverted“ refer to the products of the modulation process, whether they occupy the same relative dispositions as the untranslated range of frequencies, low to high, i.e. upright, or run high to low; inverted. – G3ISB]

The following treatment will presume an upright mixing process. This results in the output of the limiter/amplifier containing a symmetrical rectangular pulse having a frequency spectrum:

$$f_{min} = f_L - f_{x0} \leq f \leq f_{max} = f_H - f_{x0} \quad (1)$$

This triggers the following monoflop with either the positive or the negative pulse flank.

As shown in figure 2, the monoflop's output is a train of rectangular pulses  $U_a(t)$  with a constant



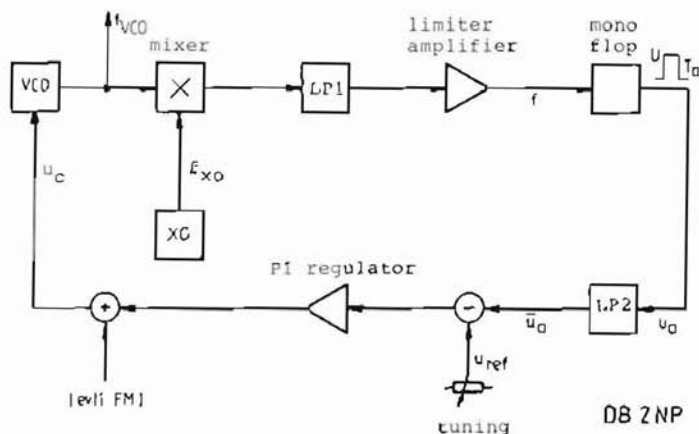


Fig. 1: Principle of frequency stabilization with a monostable

pulse duration of  $T_a$  and which have a varying pulse repetition frequency  $f$  in the range:

$$f_{\min} \leq f = 1/T \leq f_{\max}$$

The low-pass filter (LP 2) has a limit frequency  $f_l \ll f_{\min}$  therefore its output is  $\bar{u}_a$ , i.e. the integrated version of  $u_a(t)$  taken over time period  $T$ . With exactly rectangular signals, as depicted in figure 2, the average of  $u_a(t)$  is given simply by the ratio  $T_a/T$

$$\bar{u}_a = U T_a/T = U T_a f \quad (2)$$

In order that the monoflop is always reset to zero within the period  $T$ , the condition  $T_a < T$  must be fulfilled for the total frequency range  $f = 1/T$ , or in other words:

$$T_a = k T_{\min} = k/f_{\max} \quad (3)$$

with  $0 < k < 1$



Fig. 2: Typical output signal of a monostable being triggered at a frequency  $f = 1/T$

The following Performance Index regulator (PI) controls the VCO frequency  $f_{VCO}$  as long as  $\bar{u}_a$  coincides with the (as stable as possible) reference voltage  $u_{ref}$ . From the equation (2) it follows for  $f_{VCO}$ :

$$f_{VCO} = f_{x0} + f = f_{x0} + \frac{\bar{u}_a}{UT_a} = f_{x0} + \frac{u_{ref}}{UT_a} \quad (4)$$

As  $f$  is directly proportional to  $u_{ref}$ , the VCO frequency can be varied linearly by a linear potentiometer controlling  $u_{ref}$ .

The practical advantage of this characteristic is self evident. The potentiometer can be directly calibrated in terms of frequency. This can be arranged easily in a transceiver, for example, as the indicator has only to deal with DC instead of RF. Preset channels (e.g. transponder frequencies) can also be arranged easily by means of fixed potentiometers as well as the construction of sweep-oscillators. The latter manual tuning application requires that the tuning voltage be applied in the form of a saw-tooth or a triangular wave.

As fig. 1 shows, frequency modulation is easily carried out by the application of AF at this point. Finally, it should be mentioned, that a control loop containing monoflops, as those with PLL chips, are also suitable for the demodulation of FM.



## 2. THE STABILIZING EFFECT

Given the pre-condition that the monoflop is able to work with the high VCO frequency, the control concept of the downwards mixing process presented in fig. 1 is not, in principle, required. Because a relative VCO frequency error  $\Delta f/f_{VCO}$  is converted at the input to the monoflop, into the considerably larger frequency error of  $\Delta f/f$ , the frequency stability is noticeably increased ( $f_{xo}$  can be regarded as being absolutely stable). The error sensitivity increases by a factor:

$$\frac{\Delta f/f}{\Delta f/f_{VCO}} = \frac{f_{VCO}}{f} = 1 + \frac{f_{xo}}{f} = 1 + \frac{f_{xo}}{f_{VCO} - f_{xo}} \quad (5)$$

The frequency shift  $f_H - f_L$  of the VCO is normally very much smaller than the lowest VCO output frequency  $f_L$ . For example, the VCO shifting over the 2 m band: 2 MHz  $\ll$  144 MHz. According to equation (5),  $f_{xo}$  must be disposed closely below  $f_{VCO}$  in order to obtain greater error sensitivity. The main problem now arises, that there is insufficient decoupling between the two mixer inputs resulting in frequency  $f_{xo}$  appearing in the VCO output signal, which should, in principle, be free of any other spurious frequencies. Also, because of the proximity of the  $f_{xo}$  to the output frequency  $f_{VCO}$ , the following buffer stage is not able to assist in the suppression of  $f_{xo}$ . This disadvantage can be clearly seen in (1).

Some assistance may be afforded here by the use of an isolating amplifier in the form of a directional coupler between the VCO and mixer, or – easier to implement at UHF – a double-downward translation to bring the frequency of the first XO to a greater spectral distance from  $f_{VCO}$ .

In some circuit examples, the downward translation is replaced by a divider chain. This is only worth-while when the VCO itself already covers a large frequency range ( $f_H/f_L \geq 2$ ) as shown, for example in (2). As the error frequency  $\Delta f$  of the

VCO is also divided down by the same division factor as the signal  $f_{VCO}$ , the sensitivity gain given in equation (5) is no longer valid.

### 2.1. The Effect of Interference

By analyzing the influence of spurious signals, a few general observations may be made.

Assuming that a disturbing quantity is a noise voltage which is superimposed upon the VCO control voltage  $u_c$ . The effect of the noise upon  $u_c$  and  $\bar{u}_a$  will be smaller as the deviation is increased. A large deviation of  $u_c$  means that the control sensitivity of the VCO should not be made too great (usually designated as  $[K_i]$ , which is the same as Hz/V).

The VCO control voltage  $u_c$  is usually applied to a varicap diode. With a portable equipment using a battery voltage of 12 V, a 3 to 8 Volt variation should be necessary to cover the bandwidth  $f_L$  to  $f_H$ . A larger deviation for  $\bar{u}_a$  demands, first of all, the largest possible value for  $T_a$ , i.e. the highest value for  $k$  in eq. (3).  $T_a/T_{min}$  is then only just short of unity.

Secondly,  $T_a/T_{max}$  must approach zero as nearly as possible. This can be accomplished by the correct choice of  $f_{xo}$ . The voltage deviation range  $\bar{u}_{a,max} - \bar{u}_{a,min}$  can be obtained from eq. (2) and (3).

$$\begin{aligned} \bar{u}_{a,max} - \bar{u}_{a,min} &= \frac{Uk}{f_{max}} \cdot (f_{max} - f_{min}) \\ &= Uk \left( 1 - \frac{f_{min}}{f_{max}} \right) \quad (6) \end{aligned}$$

Given the VCO boundary frequencies  $f_L$  and  $f_H$  the required monoflop pulsewidth  $T_a$  and the oscillator frequency  $f_{xo}$  can be determined.

$$T_a = \frac{\bar{u}_{a,max} - \bar{u}_{a,min}}{U (f_H - f_L)} \quad (7)$$

$$f_{xo} = \frac{f_L \cdot \bar{u}_{a,max} - f_H \cdot \bar{u}_{a,min}}{\bar{u}_{a,max} - \bar{u}_{a,min}} \quad (8)$$

The maximum obtainable deviation  $0 \leq \bar{u}_a \leq U$  should only be utilized to 80 %. For example,



$u_{a, \min} = 0.1 U$  and  $u_{a, \max} = 0.9 U$  could be chosen. For these practical values inserted in eqs. (7) and (8):

$$T_a = \frac{0.8}{f_H - f_L} \quad (9)$$

$$f_{co} = (9 f_L - f_H)/8 \quad (10)$$

Of course, the calculated value for  $f_{co}$  need not be strictly adhered to, but it does serve as a working basis. A suitably cheap crystal which comes closest to that calculated above is used.

## 2.2. The Effect of Drift

Following these more basic dimensioning considerations, the effects of disturbing influences – above all drift – will be analyzed. The most important factor for stability is the reproducibility of:

- the monoflop pulsewidth  $T_a$
- the Voltage  $U$
- the reference voltage  $u_{ref}$

Short-term instabilities, in the form of needle impulses, can be attributed to VCO phase-noise. This type of variation is manifest as phase-jitter in (a) and superimposed noise voltage in (b) and (c). Low phase-noise is ensured by low-noise voltages  $U$  and  $u_{ref}$ , as well as the provision of a low-noise operational amplifier for the PI regulator.

Contrary to the practice in sweep oscillators, the loop control speed of the manually tuned VCO plays a relatively unimportant role. This allows a low-pass filter to be placed behind the PI regulator which allows only the low-frequency noise components through and thereby concentrating the VCO noise pedestal directly about the VCO output frequency.

Disturbing elements also take the form of the above discussed quantities but they are long-term and are caused by thermal drifts. As the regulator is forced continually to make  $\bar{u}_a = u_{ref}$ , a variation of the pulsewidth  $T_a$ , to  $T_a' = T_a + \Delta T_a$  eq. 2 for the frequency error  $\Delta f$  is:

$$\bar{u}_a/U = u_{ref}/U = T_a f = (T_a + \Delta T_a) (f + \Delta f) \quad (11)$$

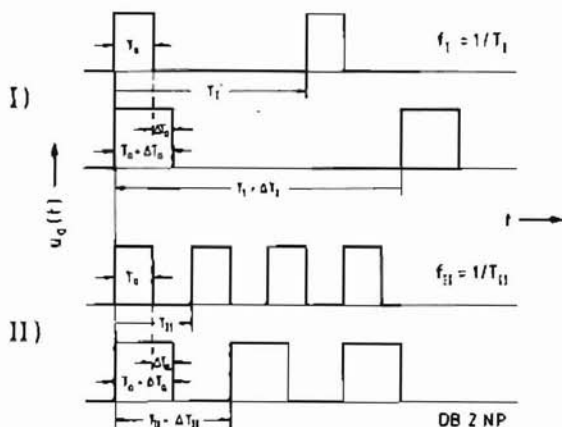


Fig. 3: Frequency drift as the consequence of a – greatly exaggerated – drift in the monostable pulsewidth  $T_a$  to  $T_a + \Delta T_a$ , by a factor 2.5 for the two monostable frequencies  $f_1$  and  $f_{II}$ . As the PI regulator always retains the same average value  $\bar{u}_a$  (here  $U/5$  or  $U/2$ ), the duty factor remains constant:  $T_a/T = (T_a + \Delta T_a)/(T + \Delta T)$ . With  $f = 1/T$  and  $f + \Delta f = 1/(T + \Delta T)$  results directly in eq. 12. This means that with any given drift  $\Delta T_a$ , that  $\Delta f/f$  remains constant, i.e. a constant relative frequency error. With  $f_1 = 2.5 f_1$ , it follows that  $\Delta f_{II} = 2.5 \Delta f_1$ .

$$\Delta f = -f \frac{\Delta T_a}{T_a + \Delta T_a} \quad (12)$$

The VCO frequency is lowered when  $T_a$  increases and, interestingly enough, more sharply the higher the monollop's input frequency  $f$  was before the  $T_a$  drift took place. It follows, that from the previously mentioned reasons for  $f$  to have a large deviation i.e.  $f_{max}/f_{min} \gg 1$ , that the VCO's stability is noticeably better at the lower than at the upper band-limit

Fig. 3 demonstrates these effects in diagrammatic form  $\Delta T_a$  is therefore made disproportionately large in order to show the effects clearly. Working through this example shows the proportions between  $\Delta f$  and  $f$  as found in eq. 12.

Next, the effect of the drift in the monollops supply voltage  $U$  is considered as it changes to  $U' = U + \Delta U$ . Since  $U$  in eq. 2 may be treated in exactly the same way as  $T_a$ , one receives directly by substitution of  $U$  and  $T_a$  in eq. 11

$$\dot{u}_a/T_a = u_{ref}/T_a = Uf = (U + \Delta U)(f + \Delta f) \quad (13)$$

The solution for  $\Delta f$  yields again eq. 12:

$$\Delta f = -f \frac{\Delta U}{U + \Delta U} \quad (14)$$

This may be seen in the numerical example of fig. 4

The effects of both, a drift in  $U$  and in  $T_a$ , lead to a higher frequency drift at the VCO's higher frequency  $f_H$  than at its lower frequency  $f_L$ . If the frequency stability at the upper-band end is not satisfactory, the range  $f_L \dots f_H$  can be divided into  $n$  equal band segments in order to make it better by a factor of  $n$ . The  $n$  switchable crystal frequencies would represent an additional complication.

Now, the drift in reference voltage  $u_{ref}$  is considered.  $u'_{ref} = u_{ref} + \Delta u_{ref}$ . This analysis of the attendant frequency error  $\Delta f$  is quite simple as  $f$  is directly proportional to  $u_{ref}$  (as it should be), so eq. 4 follows directly:

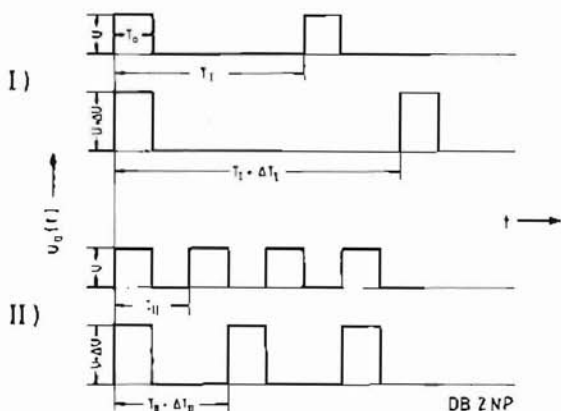


Fig. 4: Frequency drift as the consequence of a - greatly exaggerated - drift of the monostable output amplitude from  $U$  to  $U + \Delta U$ , by a factor of 2.5, of the two monostable frequencies  $f_1$  and  $f_H$ . Here, also, the PI regulator forces the average value  $\dot{u}_a$  to be maintained ( $U/5$  at  $f_1$  and  $U/2$  at  $f_H$ ). As, this time,  $T_a$  is an assumed constant, it follows that the same amplitude/period ratio prevails prior to, and following the drift:  $U/T = (U + \Delta U)/(T + \Delta T)$ . With  $f = 1/T$  and  $f + \Delta f = 1/(T + \Delta T)$  results directly in eq. 14.



$$\Delta f = \frac{\Delta u_{ref}}{UT_a} \quad (15)$$

A drift of  $u_{ref}$  has the same effect in all parts of the VCO's tunable range  $f_L$  -  $f_H$ .

It is now conceivable that the frequency drift component, caused by a change in temperature, can be countered by superimposing a temperature-compensation voltage on to  $u_{ref}$ . This could be done with an NTC-resistor network, for example. This compensatory voltage would correct the VCO frequency by the same amount, higher or lower, and independent of frequency  $f$  according to eq 15.

Frequency drifts caused by  $\Delta T_a$  or  $\Delta U$  have to be corrected additionally, with  $f$  taken into account. See eq. 12 and 14. As, nowadays, supply voltages  $U$  and reference voltage  $u_{ref}$ , may be made very stable and reproducible owing to the availability of high-grade voltage-regulator chips, the main cause of frequency drifting is due to the drift in the monoflop's pulse-width  $T_a$ . This is countered normally by making an external RC network, a small part of which contains an NTC or PTC resistive element. This offers a simple means of temperature compensation of the  $T_a$  drift.

### 3. THE USE OF MONOFLOPS IN FM-FIXED FREQUENCY OSCILLATORS

Fixed-frequency-channel, AM equipments using crystals, employ transmitter concepts involving the RF output signal being derived from the receiver L.O. (e.g. 135 MHz) being mixed with the IF frequency (10.7 MHz). This device enables the receiver crystal (about 45 MHz, working in 3rd overtone) to be used for the transmitter, thus saving both space and money.

Operation with FM, however, requires that the crystal be a relatively low frequency and therefore must be multiplied many times to achieve 135 MHz entailing the use of inter-multiplier filtering. This multiplication process ensures that a sufficiently large system deviation is ob-

tained (which normally requires a multiplication factor of 9 or 12). As before, the frequency channelling requires a large number of FM crystal oscillators to be switched and all multiplied to the same frequency band - not an easy matter.

Commercial hand-held transceivers using fixed (crystal controlled) channels usually employ unmodulated crystal oscillators. A phase modulator stage follows the crystal oscillator and the AF input is taken through a I/F network. The resultant modulated RF output then simulates an FM waveform which has been generated in a more conventional manner. This method also relies heavily upon frequency multiplication (at least  $\times 9$ ) in order to achieve the required deviation. The Oberon project (ref. 3) solved this problem by using a complete PLL with a fast pre-scaler. This is actually more by way of being a frequency synthesizer as opposed to the discrete transmit and receive crystals considered here.

There is a concept which carries out the modulation process at IF. This uses a monoflop stabilized, fixed-frequency oscillator and is capable of easily generating the required deviation without recourse to multiplication according to the principle of fig. 1. For relay work, however, the IF must be able to be switched to 10.1 MHz from the nominal 10.7 MHz IF. If the crystal oscillator XO of fig. 1 is required to have only one crystal frequency (e.g.  $f_{xo} = 10$  MHz), the direct channel ( $f = 0.8$  MHz) will experience noticeably greater drift problems than that of the radio relay channel ( $f = 0.1$  MHz) (see eq. 12 and 14). This may be optimized by choosing a middle value for  $f_{xo}$  from both IFs, namely 10.4 MHz.

The use of 10.1 MHz involves a downward translation process resulting in an "inverted" output as  $f_{xo} > f_{VCO}$ . In order that the control loop works in a negative feedback manner, this case must be provided with a pre-inversion.

This can best be done by switching from the monoflop's output  $u_a$  to its complementary output  $u'_a$ . A further switch-over process is not necessary as the monoflop frequency is  $f = 300$  kHz in both cases when  $f_{xo} = 10.4$  MHz. This means, of course, that  $u_{ref}$  does not have to be changed.



A crystal with an awkward frequency of 10.4 MHz can prove difficult to obtain and might be expensive. Considerably easier to obtain would be 10.245 MHz crystals and 10.345 MHz PLL crystals which also lie more or less between 10.1 and 10.7 MHz. Their use, also, does not entail the necessity of switching  $u_{rel}$ , if  $T_a$  and  $u_{rel}$  are suitably chosen:

$$T_a = \frac{1}{f_H - f_L} \tag{19}$$

$$u_{rel} = U \frac{f_H - f_{xo}}{f_H - f_L} = \frac{U}{T_a} (f_H - f_{xo}) \tag{20}$$

Assuming a chosen crystal frequency  $f_n$  between  $f_L = 10.1$  MHz and  $f_H = 10.7$  MHz, the monoflop frequencies are:

$$f_1 = f_H - f_{xo} \text{ (upright-downward mixing)} \tag{16a}$$

$$f_2 = f_{xo} - f_L \text{ (inverted-downward mixing)} \tag{16b}$$

From this, the average value of the monoflop output voltage results: -

$$\bar{u}_{a1} = UT_a (f_H - f_{xo}) \tag{17a}$$

$$\bar{u}'_{a2} = U [1 - T_a (f_{xo} - f_L)] \tag{17b}$$

It should be noted that in eq. 17b, the inverted output of the monoflop is selected.

If the change-over of  $u_{rel}$  is to be avoided, both voltages must be equal in amplitude i.e.:

$$u_{rel} = \bar{u}_{a1} = \bar{u}'_{a2} \tag{18}$$

From eq. 17a and 17b it follows that:

The monostable pulsewidth  $T_a$  is, interestingly, not only independent of the crystal frequency  $f_{xo}$  in use, but also fixed upon the reverse value of the relay i.e.  $1/(600 \text{ kHz})$ .

The method by which a frequency modulated 10.1/10.7 MHz oscillator can be realized is demonstrated by the example of fig. 5. This circuit sketch is a little less than a constructional project but it does serve to show that by using cheap consumer ICs, the component count and therefore the constructional size, can be kept small. With a little PCB-design experience, the whole arrangement of fig. 5 can be placed on a board of only  $4 \times 7 \text{ cm}^2$ .

The AM receiver chip Siemens TCA 440 functions here as a downwards mixer, crystal oscillator, limiter amplifier and buffer between VCO and mixer - this buffering is to suppress the  $f_{xo}$  spurious frequencies, see chapter 2.

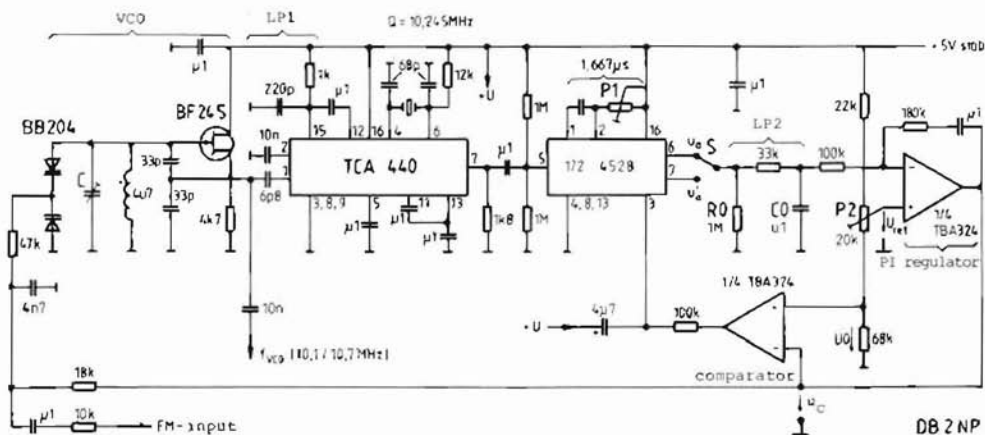


Fig. 5: Circuit diagram of a sample, frequency-modulated 10.1/10.7 MHz oscillator.



The IF, square-wave signal at pin 7 has about  $2 V_{DD}$  amplitude and is sufficient to trigger the CMOS monostable 4528, the input of which has been biased at  $U/2$ .

The pulsewidth  $T_a$  at pin 6 of the monostable is adjusted to  $1.667 \mu s$  with the aid of P1 (see eq. 19). Potentiometer P2 adjusts the reference voltage  $u_{ref}$  of the PI regulator in accordance with the crystal frequency  $f_{co} = 10.245$  MHz (see eq. 20). P1 and P2 are then fine trimmed with  $f_{VCO}$  being monitored on a frequency counter. When switch S is in the  $u_a$  position, the  $f_{VCO} = 10.7$  MHz exactly. Likewise with S in position  $u'_a$ ,  $f_{VCO} = 10.1$  MHz.

In order that the VCO does not jump out of lock during the switching of S, a comparator has been added to the circuit. Immediately following every switching action, the previous VCO frequency is always on the wrong sideband so that the loop has a positive feedback. This means that the initial drift of the control voltage  $U_c$  — irrespective of direction — is amplified. At the moment of switching, switch S has neither contact with  $u_a$  nor with  $u'_a$ . The voltage at  $C_o$  sinks momentarily due to  $R_o$  thus causes the control voltage  $U_c$  to drift upwards.

When switching the lower to the upper frequency, the favourable direction is: The VCO frequency gets to the correct sideband so that the VCO locks firmly on to 10.7 MHz. Switching back to 10.1 MHz, the VCO would run to its upper frequency limit as, this time, the control-voltage drift during the switching action is going in the wrong direction. Because of the comparator, a further triggering of the monostable is inhibited when the control voltage exceeds  $U_o$ . The monoflops output  $u'_a$  assumes the voltage  $U$ . As a consequence, the control voltage  $U_c$  sinks quickly, and the VCO locks to the frequency of 10.1 MHz following the resetting of the com-

parator. The trimmer C in the VCO tuned circuit must be adjusted such that the control voltage  $U_c$  at  $f_{VCO} = 10.7$  MHz and is about 20 % below the value of  $U_o$ .

In conclusion, for the individualist, it is mentioned that in the construction of a complete FM transmitter, the receiver forms part of the transmit-frequency locked loop in order to improve transmit-frequency stability. The function of the monostable can be replaced by a normal FM demodulator or a coincidence detector if a fixed IF is employed. A good temperature stability is afforded by the FM-receiver chip TCA 420 with a push-pull output which facilitates any temperature drift to be compensated by means of a subtractor stage. The more popular S 041 and TBA 120 are not recommended for this purpose as they only possess a multiplier output.

(To be continued)

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Wilfried Hercher, DL 8 MX, Jochen Jirmann, DB 1 NV

## A Stabilized Power Supply for Valved PAs

Traditionally, valved PA final stages have been supplied by unstabilized power supplies comprising a HV transformer, diode rectifier network and a smoothing capacitor. Under full load, the nominal HT voltage falls to about 75 % of its quiescent value. For amateur use, employing SSB syllabic modulation, this represents a continually varying load on the power supply. The anode voltage varies in sympathy with the modulation giving rise to the following disadvantages: - The final stage's components must be dimensioned for the maximal voltage, i.e. the quiescent voltage under a mains over-voltage condition, in order to avoid a voltage break-down. The PA tube (s) is continually in a mis-matched condition during this operation because for tune-up purposes, they have been

delivering maximum power at a low HT voltage, and the match is only valid for this condition. This gives rise to non-linearity.

In the following article, a simple thyristor (SCR) will be described which may be retro-fitted into an existing power supply without too much difficulty. It will hold the anode voltage constant to within about 3 %, suppressing the effects of both load and input mains voltage variations. The only condition for its use is that the tube's HT and heater supply powers are delivered via separate mains transformers. This is so that the heaters maintain their nominal voltage when the anode voltage is being regulated.

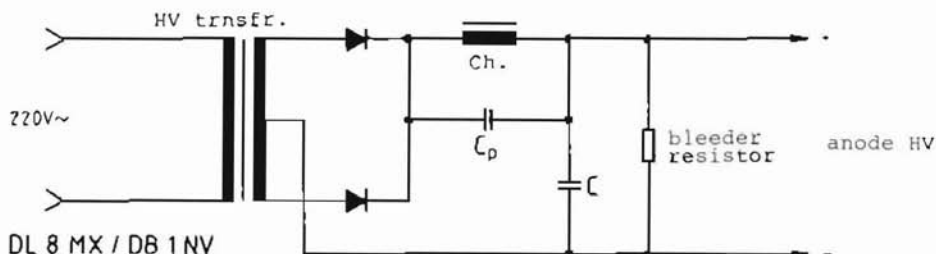


Fig. 1: A conventional choke-input, HV power supply



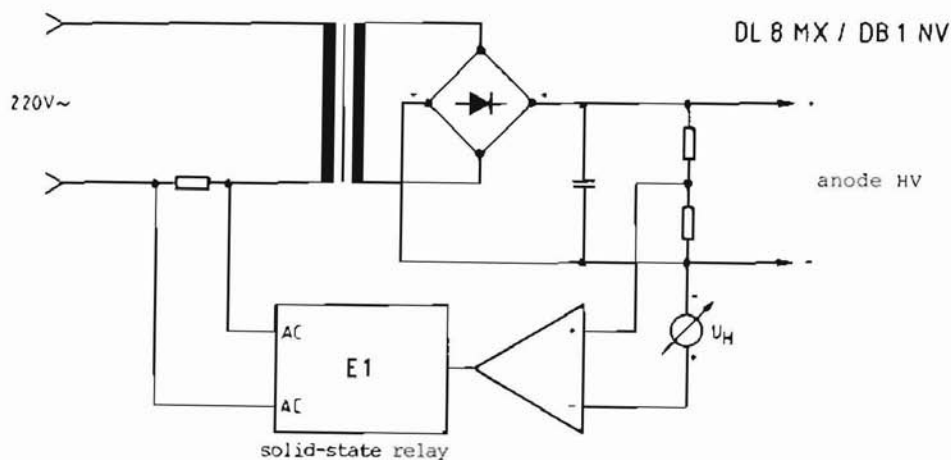


Fig. 2: Mains input, TRIAC-regulated power supply (AA 5 C)

## 1. EARLIER SOLUTIONS

The manufacturers of high-grade commercial HF PAs, such as Henry Radio or Ameritron, have recognized this problem years ago and have tried "passive" solutions to keep the HT voltage stable under all load conditions. This was accomplished using the "swinging choke" inductor input filter shown in fig. 1. The swinging choke's characteristic of variable inductance at all rated load currents is achieved by the omission of an air gap in the core. For greater suppression of the ripple frequency at 100 or 120 Hz, this choke can form part of a parallel resonant circuit with  $C_p$ . By this technique, the anode voltage remains sensibly constant over a varying load range of some 10 : 1. In order to reduce the high voltage developed across C (and the output) at low loads, bleeder resistances are placed across the output to absorb some 10 % of the load current. The great disadvantage of the swinging-choke input technique is the heavy and bulky nature of the component needed, especially if it is worked in a resonant condition. The capacitor  $C_p$  must also

be capable of withstanding the heavy circulating currents and voltages in the kilo-volt region entailed in resonant working.

A large power supply built regardless of volume weight and with good regulation, would be the best solution. The author realized this during the construction of a 1.2 kW power supply with an output voltage of 4 kV, the effort involved being quite reasonable. The danger of these high-power, HV power supplies lies, however, in the break-down of components caused by a faulty selection or installation. This factor could put the constructor's life at risk and therefore the construction of such heavy duty power supplies is not to be recommended.

## 2. HV POWER SUPPLY WITH MAINS-INPUT REGULATION

As the regulation does not to be of the same order as that provided by a laboratory power supply, it was decided to regulate the mains

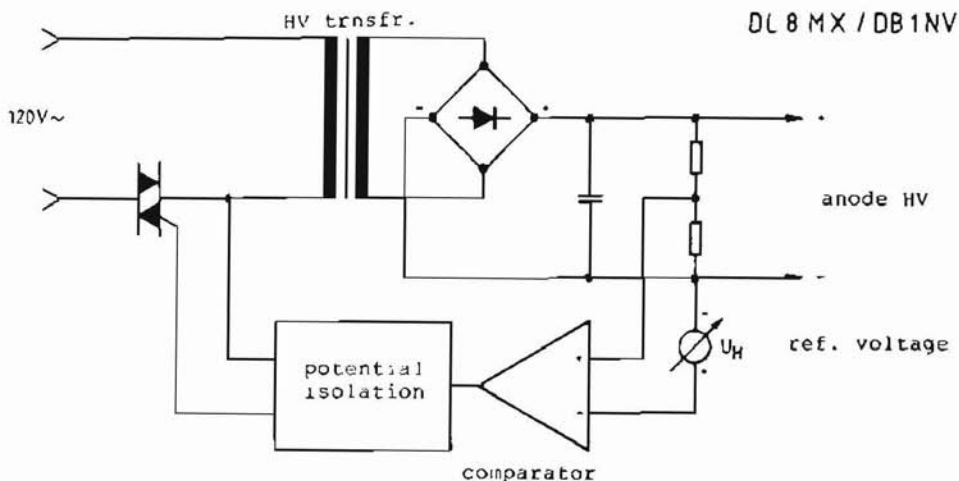


Fig. 3: A HV power supply with a solid-state relay as the control element

input to the HV transformer by means of a thyristor or silicon controlled rectifier (SCR). The control circuitry for the thyristor, despite the availability of integrated phase-chopper circuits, is relatively complex and Greg McIntire (AA 5 C) in ref. (1) went as far as not providing a continuously regulated energy supply to the mains transformer.

It can be seen from fig. 2 that an output voltage sample is compared with a reference voltage. If the output voltage sample exceeds the reference, a TRIAC is caused to isolate the mains from the mains transformer. The output voltage then falls slightly thus resetting the comparator and firing the TRIAC. The output voltage is then restored.

This very simple concept caused concern to many constructors who complained of loud noises emanating from the transformer. The cause of this is that the firing time of the TRIAC does not correlate with the mains frequency. It can occur, for example, that the TRIAC conducts on a positive mains half cycle, then it blocks the negative half-wave and then conducts again for the following positive half-wave. This produces an average value of more than zero for the input current i.e. there is a strong DC component flowing through the transformer primary which

tends to saturate the core. This leads to sharp impulses of current flowing through the primary winding. This becomes manifest in the form of strong magnetostriction in the core and this is responsible for the production of core noise. In addition, the high current peaks stress the transformer and the TRIAC unnecessarily.

### 3. AN IMPROVED CIRCUIT

One could perhaps consider a circuit which blanks individual half-waves and thereby preventing a DC component from being developed. There is, however, an easy way of doing this:

One of them uses a proprietary solid-state relay with a zero voltage switch (concept rejected by AA 5 C owing to energy feedback into the mains) instead of the more usual opto-coupler controlled TRIAC. Another solution is not to fully isolate the supply input but to partially isolate it by means of a series resistance. This is shown in fig. 3

The resistor is so chosen that at maximum mains input voltage and at minimum load on the supply



unit, an output voltage is available which is just under the desired nominal value. The exact resistance of the resistor is not critical but tests have indicated that it must lie between 200  $\Omega$  and 500  $\Omega$  and have a dissipation of 100 W (approx.). This approach ensures that the mains transformer is continually pre-magnetized which causes no current impulses to be generated when the solid-state relay switches on.

Figure 4 shows the developed circuit of the high-voltage regulator. A potential divider provides an HV sample voltage of 5 V. This potential divider must certainly be constructed of HV type resistors. The author used 5 x 4.7 M $\Omega$ , 1 W types. These are best supported, in air, on insulated pillars such that they are not within 1 cm, at least, from any other surrounding object.

The LM 393 comparator compares the high-voltage sample with a reference voltage provided by a Z-diode ZPD 10. The latter is preset to the exact value by means of the potentiometer P. A small amount of positive feedback is applied to the comparator in order that a hysteresis of 50 V (w.r.t. the high-voltage) is obtained. The out-

put emitter-follower of the comparator drives the electronic load relay and an LED indicates the state of the switch. The supply voltage of about 15 V can be taken either from a 1 VA transformer, bridge rectifier and smoothing capacitor or from one of the supplies used on the PA itself. The electrolytic capacitor across the Z diode plays a special role. It enables the HV power supply to be given a "soft start" from switch-on, i.e. it reaches its full potential gradually. This feature prevents the house fuses or breakers from operating every time the supply unit is switched on.

The solid-state relay can be of any type which can work with a DC control potential of from 12 to 20 volts and a rated current of about twice that of the maximum current of the PA at full output. This relay has the approximate dimensions of a cigarette packet, possesses lugs for affixing purposes and must be screwed with its underside against a heat-sink which has been isolated from the control voltage and the mains. The heat to be dissipated by the heat-sink lies between 10 and 20 watt. The author used the Siemens V 23100-S 0302-A 225 rated at 25 Amps.

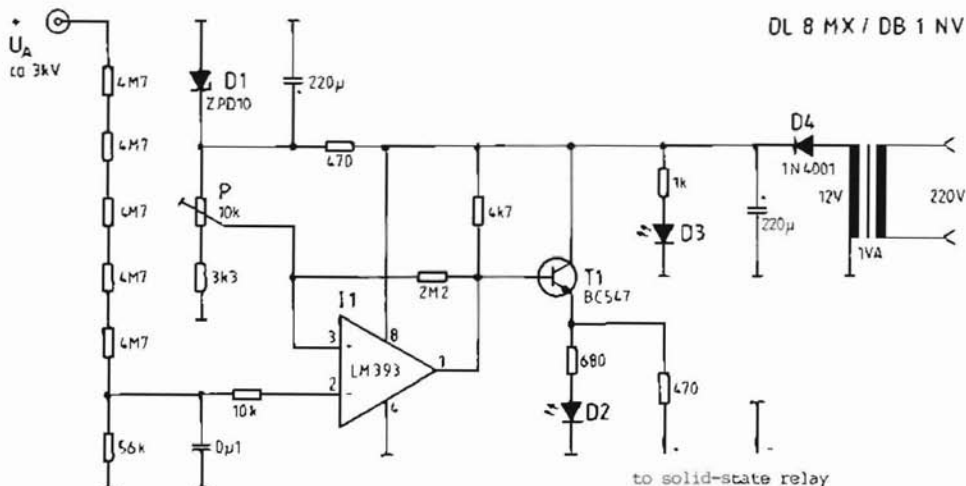


Fig. 4: The control circuit of the solid-state relay



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## 4. CONSTRUCTION AND COMMISSIONING

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The few components required for this project can easily be placed on a VERO-BOARD or similar, but the high-voltage divider should be located in the vicinity of the PA stage.

### 4.1. Component List

Semi-conductors II:	LM 393, Motorola, National.
D1:	ZPD 10
D2, D3:	Red and green LEDs
D4:	1N 4001
T1:	BC 547
E1:	Solid-state relay, e.g. Siemens V 23100-S 0302-A 225 rated 25 A and DC control
Resistors:	1/8 W except the HV 4.7 M $\Omega$ 1 W (exact value uncritical)
	Pre-set pot. 10 k $\Omega$
Capacitors:	Electrolytic, 35 V Foil, 63 V
Miscellaneous:	Mains trnsfr. 12 V 1 VA (approx.) Vero-board 100 x 100 mm (approx.) Insulated pillars (HV divider)

### 4.2. Warning: High Voltage!

Before putting the power supply into operation, the anode voltage of the PA stage, under load, is determined. The built-in voltmeter, normally fitted to the PA, is sufficiently accurate for this purpose. Then, a variable high-wattage resistor of about 100 W is placed in series with the mains input line and adjusted until the unloaded voltage of the power supply is just under that required for full load working.

The variable resistor should be connected to the HV primary circuit only and not to the filament supply transformer as well! Otherwise, the

filament primary supply voltage would also be inadvertently reduced.

Following this preparatory work, the solid-state relay is connected across the primary series resistor. The control circuit voltage is then connected to the solid-state switch. The HV control is then turned to obtain minimum output voltage.

The HV supply and that to the control circuit can now be switched on. If everything has been correctly connected, a voltage should appear at the (unloaded) output which is just under the full load value and the LED is off, indicating that the solid-state relay is inoperative. The high-voltage control is now turned up slowly until the LED begins to blink and the HV (still unloaded) starts to rise. The voltage begins to oscillate about 50 V in sympathy with the LED indicator. The point at which this occurs is the correct setting of the regulation.

The PA stage is now driven to full output and the LED indicates that the solid-state relay is operative. The anode HV potential must remain sensibly constant. The setting-up is now finished and it only remains for the power supply to be made safe with proper protective covering for both mains and HV points.

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## 5. OPERATIONAL EXPERIENCE

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The circuit, as described, was installed into the Drake L7 linear PA of DL 8 MX at the beginning of 1988. Since then, there has been neither operational failure nor other undesirable effects such as HF interference or signs of component stress to report. Indeed, the mains input transformer runs noticeably cooler since the regulator was installed. This is certainly something to do with lower iron losses in the core, as the transformer is now subjected to a lower input voltage for a larger portion of the operating time. The additional loss in the mains input resistor (50 W approx.) cannot, however, be left out of the efficiency calculations.



The above PA final had, originally, an unloaded HV of 3600 V and a fully loaded (1.3 kW CW carrier) HV of 2900 V. Following the installation of the HV regulator, the quiescent voltage is now 3000 V and the full load voltage is again 2900 V. It should be noted that, owing to the mains internal resistance at DL 8 MX, the supply voltage sank by some 5 % when the PA was transmitting full load.

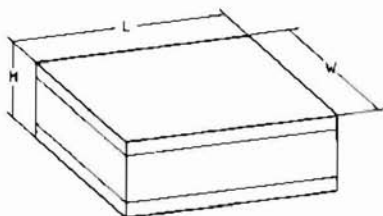
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111	x	37	x	50	10704	4.50
148	x	37	x	30	9820	4.50
148	x	37	x	50	10705	5.15
74	x	55.5	x	30	10706	3.80
74	x	55.5	x	50	10707	4.50
111	x	55.5	x	30	10708	5.15
111	x	55.5	x	50	10709	5.70
148	x	55.5	x	30	9489	6.30
148	x	55.5	x	50	10710	6.85
74	x	74	x	30	9498	5.15
74	x	74	x	50	10711	5.70
111	x	74	x	30	9499	6.30
111	x	74	x	50	9761	6.85
148	x	74	x	30	9501	7.40
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WA 5 VJB with KF 5 N in Texas and WA 7 CJO with KY 7 B in Phoenix, Arizona succeeded in establishing the first EME contact using the 3 cm band (10.368 MHz) on the 27th August 1988. Owing to the continually varying path propagation times, the reflected signal had a bandwidth strewn over 1 kHz and sounded like a beginner's lesson on the flute - more hiss than tone! In addition, this was no large institute achievement, only surplus and amateur equipment was used for this record. The antenna dimensions were between 4 and 5 metres and the output powers between 50 and 100 W! Congrats de DL 3 WR es OE 6 AP.

STOP-PRESS information for constructors of the DL 0 HV frequency counter (VHF Comms. 3/88)

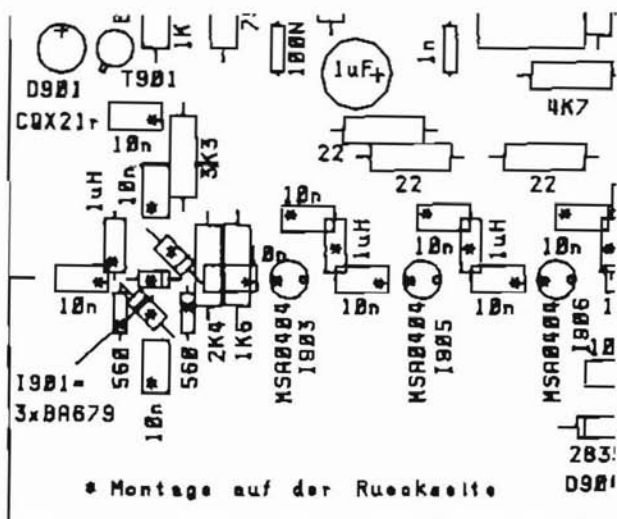
Unfortunately, the production of the integrated PIN-diode attenuator TDA 1053 A has been discontinued by Telefunken and is no longer obtainable. Experiments have shown that this attenuator pad can be replaced by using 3 discrete PIN diodes. The extract from the layout plan (Fig. 1) shows how the 3 diodes are arranged. They are mounted on the underside of the board and soldered directly to the board tracks. When using the PIN diodes BA 479 S (glass-type), small alterations in the dimensioning of the control network are required

These modifications have no notable effect upon the frequency bandpass of the input circuits. Tests on the prototype indicated that the area of highest sensitivity was shifted some 50 MHz higher in frequency

Also, in the component plan of the input stage a mistake has been discovered. The diode D 905 (ZTE2) is shown 180° reversed. It should be positioned so that the ring marker is nearest the multiway plug. This component is not a genuine zener diode but a series of normal silicon diodes.

D. Schwarzenau

Fig. 1





## Heavy-Duty Support Housing for KR 5400/KR 5600 Satellite Rotators

This heavy-duty, stable profile replaces the L-bracket which forms part of the KR 5400 or KR 5600 rotator systems. The use of this component ensures a much higher degree of antenna-installation stability.

### Material:

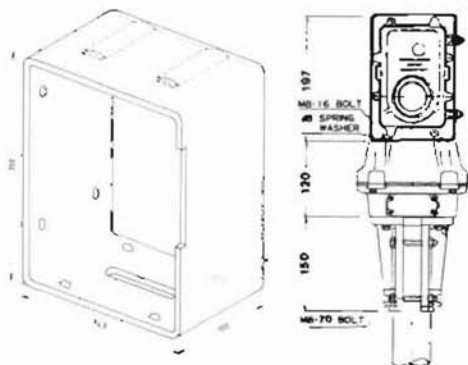
Square -section tubing  
200 x 149 x 100  
6 mm thick steel plastic coated

Weight: 3.1 kg

Art.Nr.: 1068

Price: DM 148.00

Post and package charges upon request.



## Satellite Rotators

	Art.No	Export price(!)
Including the enhanced dates of elevation rotator KR-500 A: (100 Nm/200 Nm) and KR-500 B: (140 Nm/40 Nm)		
KR-5400 A	1013 DM	785.-
KR-5600 A	1014 DM	1026.-
KR-5400 B	1015 DM	956.-
KR-5600 B	1016 DM	1132.-

Available from:



Telecommunications, VHF-Communications  
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*Note: By using the TNC unit into the RS232C connector, PACKET RADIO (PR) can be worked and a standard RTTY converter allows teletype operation.*

Art.No.: 3403/9.x

DM 190.00

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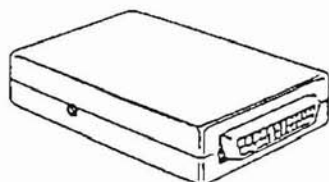
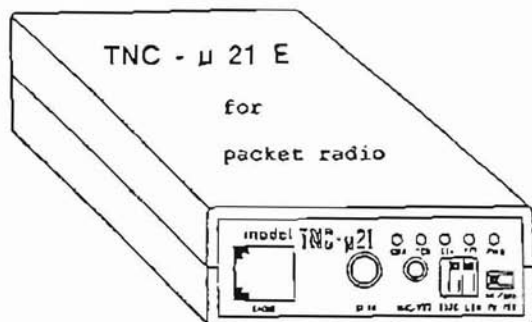
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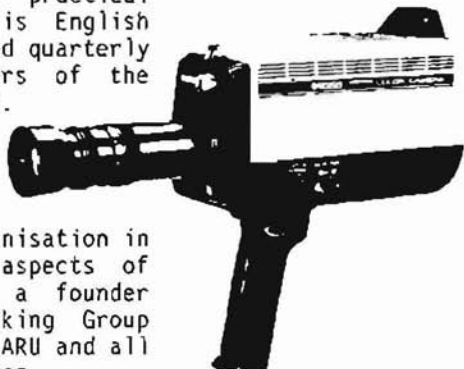
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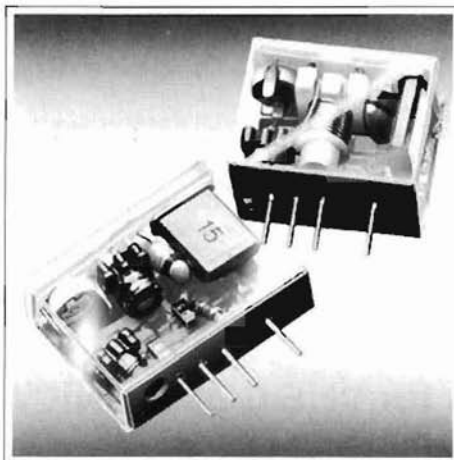
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