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Extending the range of a Racal Dana 1991 Frequency Counter to 12GHz

Ralph Berres, DF6WU

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When a magazine starts to reprint articles from years ago you know it is running out of material. I have had to resort to that tactic in this issue. I am desperate for new articles to keep the pages of this magazine full of interesting articles. If you have something that you think would be suitable please contact me.

The remainder of the articles are very interesting with three articles directly from readers of VHF Communications Magazine.

73s - Andy



K M Publications, 503 Northdown Road, Margate, Kent, CT9 3HD, UK

Telephone / Fax +44 (0)1843 22080, email : andy@vhfcomm.co.uk

web : <http://www.vhfcomm.co.uk>



Gunthard Kraus, DG8GB

Quadrature Amplitude Modulation (QAM)

Quadrature Amplitude Modulation is an indispensable modern transmission tool but the working principle is somewhat mysterious. This article hopes to create clarity using LTspice on a PC to give the interested reader proof.

1.

Basic information

Very often in communications technology, two completely independent pieces of information must be transmitted at the same time using the same carrier frequency on the same channel.

Using a mathematical example, a coordinate system can be arranged with two axes perpendicular to one another. If the two signals use the two axes then it suddenly becomes completely simple: If these two are summed then only one total signal is transmitted.

Technically this means that the two different information signals modulate two separated carrier signals (using AM) with a 90° phase shift but with same frequency. Each individual carrier changes both its amplitude and its sign in sympathy with its axis information. These two carrier signals (called the I = in phase signal and Q = quadrature phase signal)

simply add and the QAM signal is the result.

This QAM signal can now be sent without problems as an individual carrier over the existing channel, although it actually consists of two completely different pieces of information. Its amplitude and phase change nonstop and the receiver must evaluate these two characteristics continuously in order to recover the original signals.

This trick originates in vector analysis that has been well known for many centuries. Everything is represented by a point in a coordinate system, either by its X coordinate and Y coordinate or by a phase and magnitude of the point from the zero point of the diagram.

In the communications this was used for the first time with the transition from the black-and-white to the colour television. The colour information was packed into the same television channel; a Chroma signal was added to the black-and-white picture (Y signal). This was at the upper frequency of the Y signal of 5MHz, in Europe this was at 4.43MHz so that the two signals affect each other as little as possible.

This Chroma signal consists of information about the blue (more correctly B-Y) on the horizontal axis and the red (more correctly R-Y) on the vertical axis. But a colour image is built from three portions;

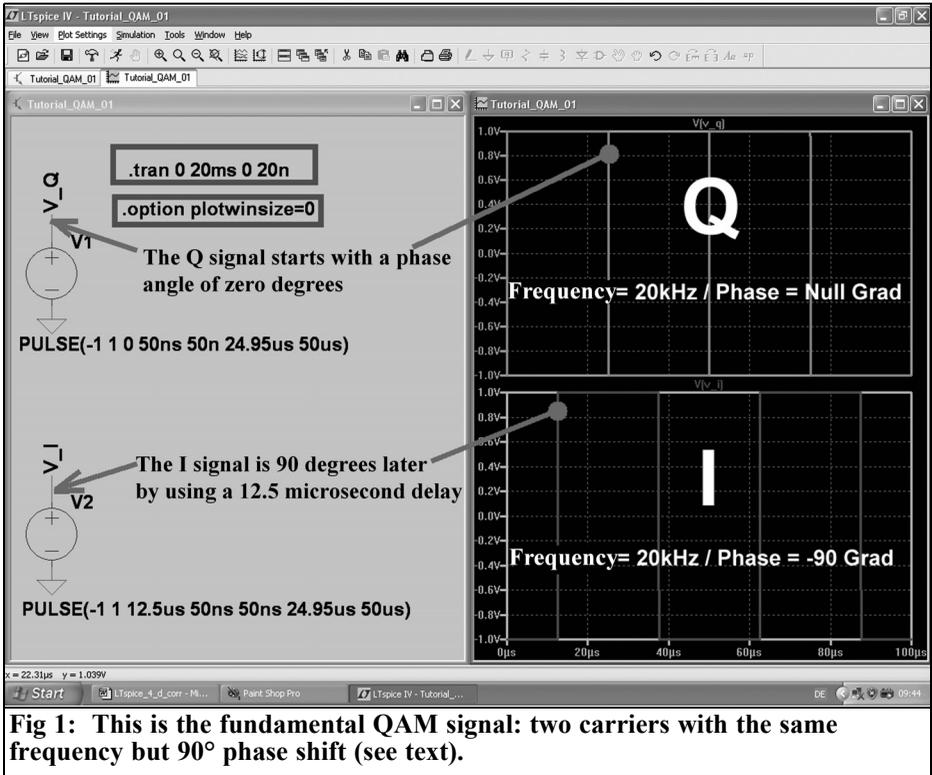


Fig 1: This is the fundamental QAM signal: two carriers with the same frequency but 90° phase shift (see text).

red, green and blue and it would seem that at the receiver the green information is missing. It is not like that because the radiated black-and-white picture consists of the three portions red, green and blue in an accurately known relationship (in accordance with the relative luminosity curve). If the red and blue are recovered using quadrature demodulation of the Chroma signal, the remainder is very easy: If the accurately known portions of red and blue are subtracted from the black-and-white picture only green remains.

The actual QAM problems lie on the receiver side. It must be supplied nonstop with sufficient information about the reference phase, which sounds more complicated than it is. The receiver needs exactly the same coordinate system as the transmitter to decode the momentary value of the QAM signal as phase and

magnitude to read off the values of the X coordinate and Y coordinate from both axes.

For colour television this is achieved by transmitting 10 to 12 cycles with the zero phase position in the horizontal synchronising signal of the transmission. Because the screen is blanked during the synchronising signal this reference signal (burst) does not cause any picture disturbance. The burst synchronises a crystal oscillator and with an additional circuit it produces two reference carriers for the QAM demodulator with a 90° phase shift to demodulate the I and Q signals.

For digital transmissions the problem of the reference phase is clearly more difficult. Many channel transmissions use one reference channel to recover the reference in the receiver from a simple and visible but accurately specified bit sequence. Naturally there are other possi-

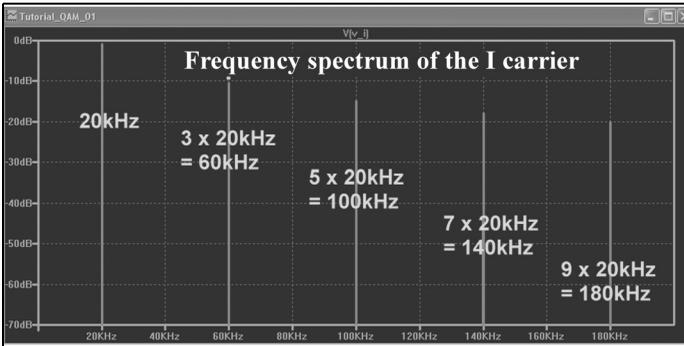


Fig 2: The attenuation of the harmonics can be seen best with the symmetrical square wave signal.

bilities that do not use an additional reference channel, however these are a more complex topics outside the scope of this article.

2.

Producing a QAM signal

The tasks: Two information signals are to be transmitted by QAM on a 20kHz carrier. The following data applies:

- Signal A (I channel) is a sine wave $f = 1000\text{Hz}$ and a peak value of 0.5V .

- Signal B (Q channel) is a square wave $f = 500\text{Hz}$ with a minimum value of 0V and a maximum value of 1V .
- The two carrier components (I and Q) are symmetrical square wave signals. Their frequency is $f = 20\text{kHz}$, the minimum value is -1V the maximum value is $+1\text{V}$. The I signal is 90° behind the Q signal.

Using LTSpice the voltage supplies for the carrier are programmed first.

Two symmetrical square wave signals with the same frequency ($f = 20\text{kHz}$) and an amplitude of 1V (no DC portion) are used with a phase difference of 90° . Rise and fall times are 50ns in each case.

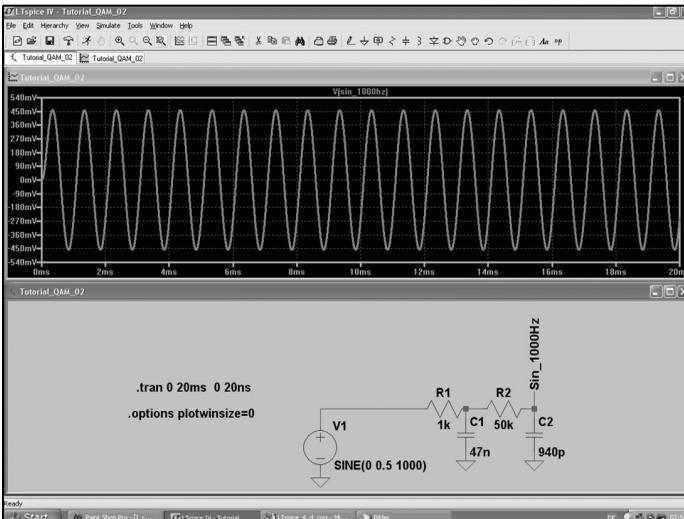


Fig 3: A 1000Hz signal source with a low pass filter for the I channel.

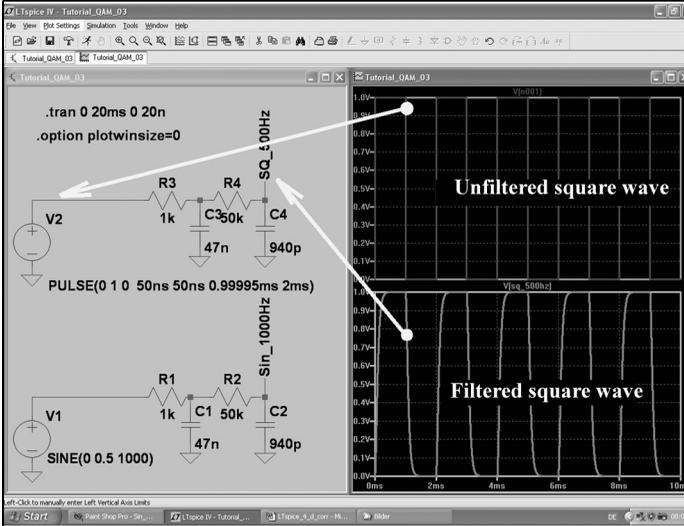


Fig 4: The same technique is used for the Q channel to generate a 500Hz square wave

Note: An advance of the phase in degrees cannot be programmed in SPICE; therefore the I signal is delayed by 12.5µs for f = 20kHz. This gives the desired phase difference of 90°.

This results in the following source programming of the Q signal:

PULSE(-1 1 0 50ns 50ns 24.95us 50us) and the I signal:

PULSE(-1 1 12.5us 50ns 50ns 24.95us 50us)

Simulation is for t = 20ms with a maximum time step of 20ns. Clicking the button with the hammer opens the SPICE

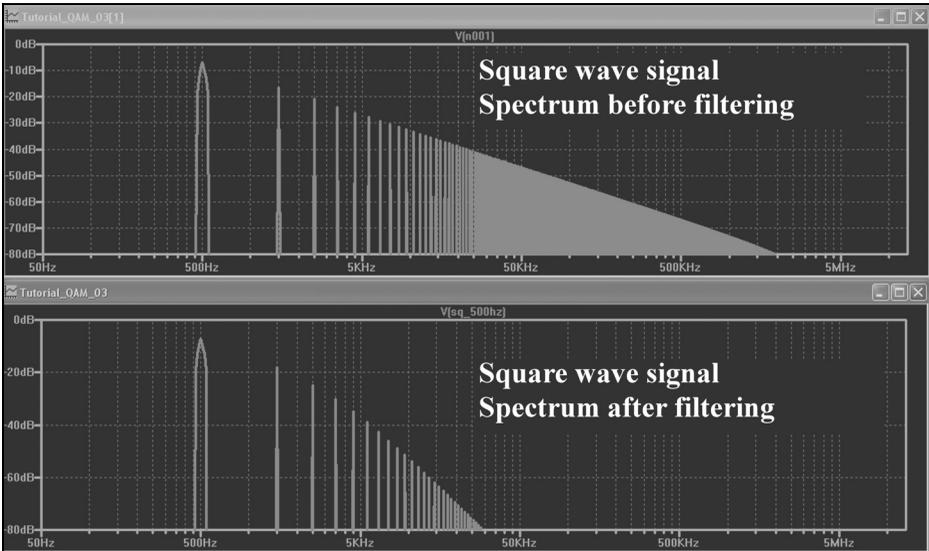


Fig 5: A simple RC lowpass filter miraculously reduces the harmonics at 12dB per octave.

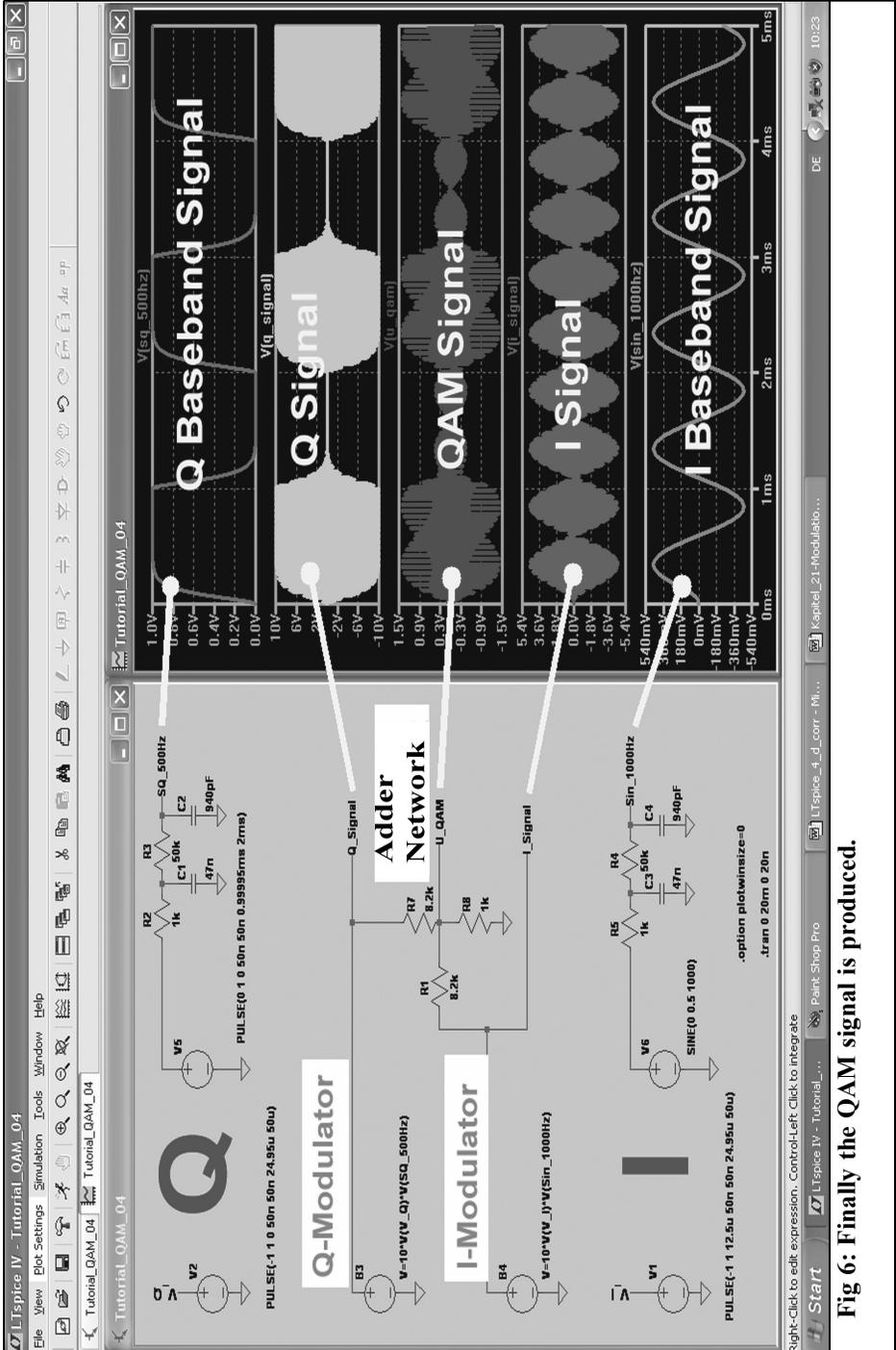


Fig 6: Finally the QAM signal is produced.

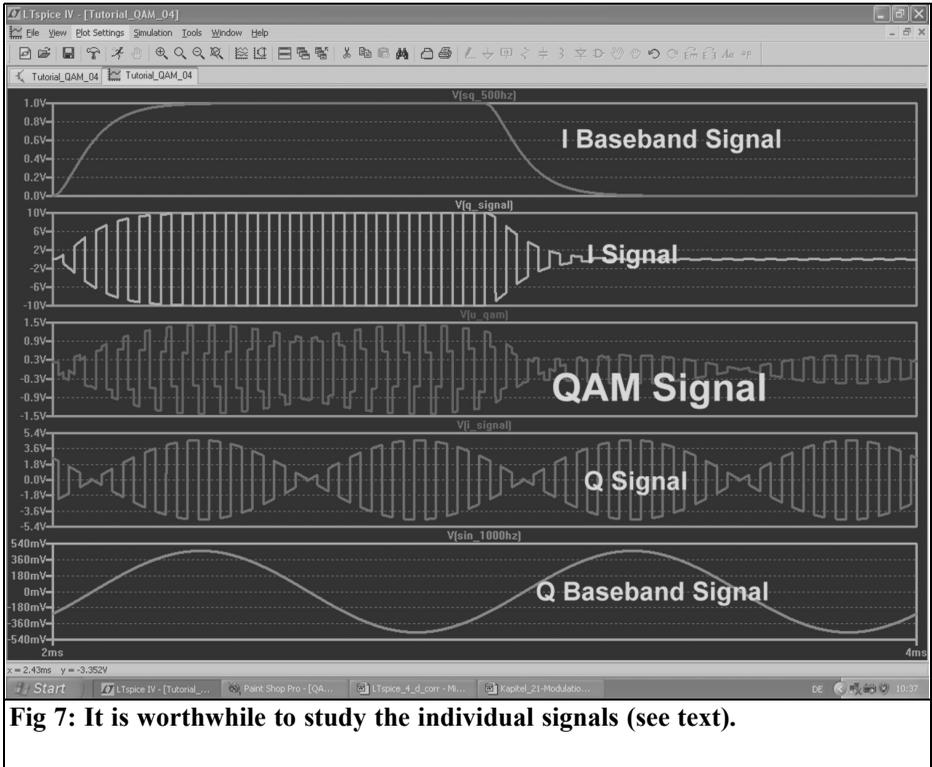


Fig 7: It is worthwhile to study the individual signals (see text).

control panel; select “Gear” under “Default Integration Method” (this increases the stability of the simulation).

With $t = 20\text{ms}$ the FFT spectrum shows frequencies from $1/20\text{ms} = 50\text{Hz}$. The maximum time step of 20ns guarantees that the simulator routine cannot ignore the rise and fall time of 50ns of the pulse signals.

The SPICE data compression is switched off with the directive “option plotwinsize = 0”. This guarantees a genuine number of samples for the FFT of:

$$20\text{ms} / 20\text{ns} = 1,000,000 \text{ samples}$$

Who is surprised by the $24.95\mu\text{s}$ instead of $25\mu\text{s}$ for the pulse length? Consider:

The pulse length is actually only the time when the signal exhibits the positive maximum value. Therefore the rise time must be taken off if the signal is to

remain symmetrical. Symmetry control is quick and easy if an FFT with $524,288$ samples is used for one of the two signals (that is allowed because $20\text{ms} / 20\text{ns} = 1,000,000$ samples and more samples for the FFT result in a greater frequency range and dynamic range).

The result of the simulation in the time domain is shown in Fig 1.

Fig 2 shows the spectrum of the I signal after the FFT (Help: with the right mouse button on the potential gradient click “View” and “FFT”). Only odd number harmonics can be seen, unwanted even harmonics are attenuated by more than 70dB . That is a good indication of almost perfect symmetry of the square wave signal.

Now the baseband signals are programmed. Two more voltage supplies are

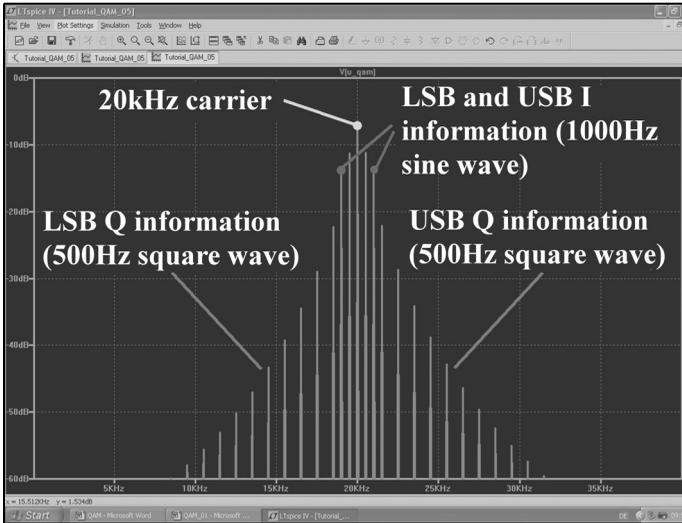


Fig 8: The frequency spectrum around the 20kHz carrier.

used with simple lowpass filters having a cutoff frequency of approximately 3.4kHz to limit the bandwidth (the upper cutoff frequency of a telephone channel). This avoids problems of frequency band overlaps (Aliasing) at the following carrier modulation stage.

Start with a sine wave signal ($f = 1000\text{Hz}$ / peak value = 0.5V) for the I channel. Fig 3 contains all the necessary information, the entry for this voltage supply is:

SINE (0 0.5 1000).

The voltage supply for the square wave signal for the Q channel can be added to the same simulation diagram (minimum value = 0V, maximum value = +1V, frequency = 500Hz). This is:

PULSE (0 1 0 50n 50n 0.99995ms 2ms)

the rise time must be taken off of the pulse length again for perfect symmetry

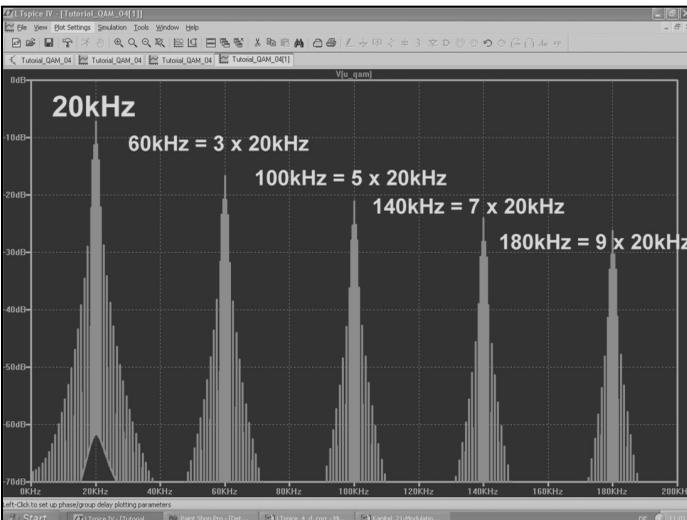


Fig 9: The complete QAM spectrum is exactly as the theory but adjacent channels would suffer from interference.

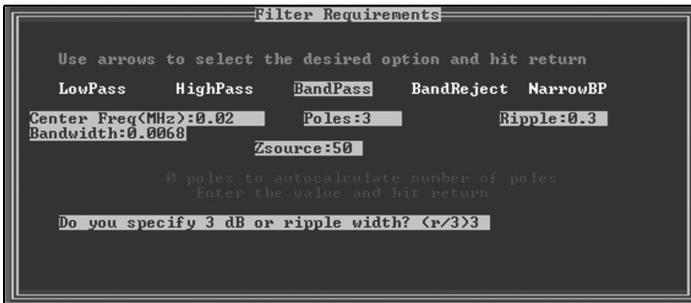


Fig 10: The input window for the filter program "fds".

(Fig 4). A view of the effect of the filter is interesting (lowpass filter at the output) showing the harmonics of this square wave Fig 5. The signals above 5kHz are attenuated by more than 40dB.

Fig 6 is interesting because it shows the complete QAM modulation process using two "bv voltage sources" and so some explanations are necessary:

The bv voltage supply is an LTspice component which represents an ideal analogue multiplier to multiply each carrier component with its baseband signal. So every baseband signal is modulated to its belonging diagram axis resp. the belonging carrier signal with $f=20\text{kHz}$. As required this multiplier inverts the phase of the carrier if the baseband is negative.

The addition of the I and Q signals is carried out using a simple resistance network following each modulator that attenuates the signal amplitude by around a factor of 10. A times ten amplification is included in each multiplier to compensate for this attenuation.

The necessary formula inputs for the two multiplying stages are therefore:

for the I path:

$$V = 10 * V(V_I) * V(\text{Sin_}1000\text{Hz})$$

and for the Q path:

$$V = 10 * V(V_Q) * V(\text{SQ_}500\text{Hz}).$$

The details shown in Fig 7 are interesting. The phase change for both information signals (I signal and Q signal) is in the rhythm with the carrier frequency of 20kHz. After adding the two carrier components their phase difference of 90° is clearly visible in the QAM signal.

But how does the spectrum of such a QAM signal look? Using the FFT with 524,288 samples the result is shown in Fig 8. There are not many surprises because as expected there is a carrier frequency of 20kHz and the sum and the difference frequency for the baseband sine wave of 1kHz in the I channel. The square wave on the Q channel with its familiar (however already somewhat filtered) collection of odd harmonics produces the easily recognisable lower sideband (LSB) and upper sideband (USB).

Since the carrier is a square wave it must repeat itself with each odd harmonic of the carrier frequency as shown in Fig 9. Some more must surely be done to minimise the interference spectrum.



Fig 11: The calculated values for the filter components (see text).

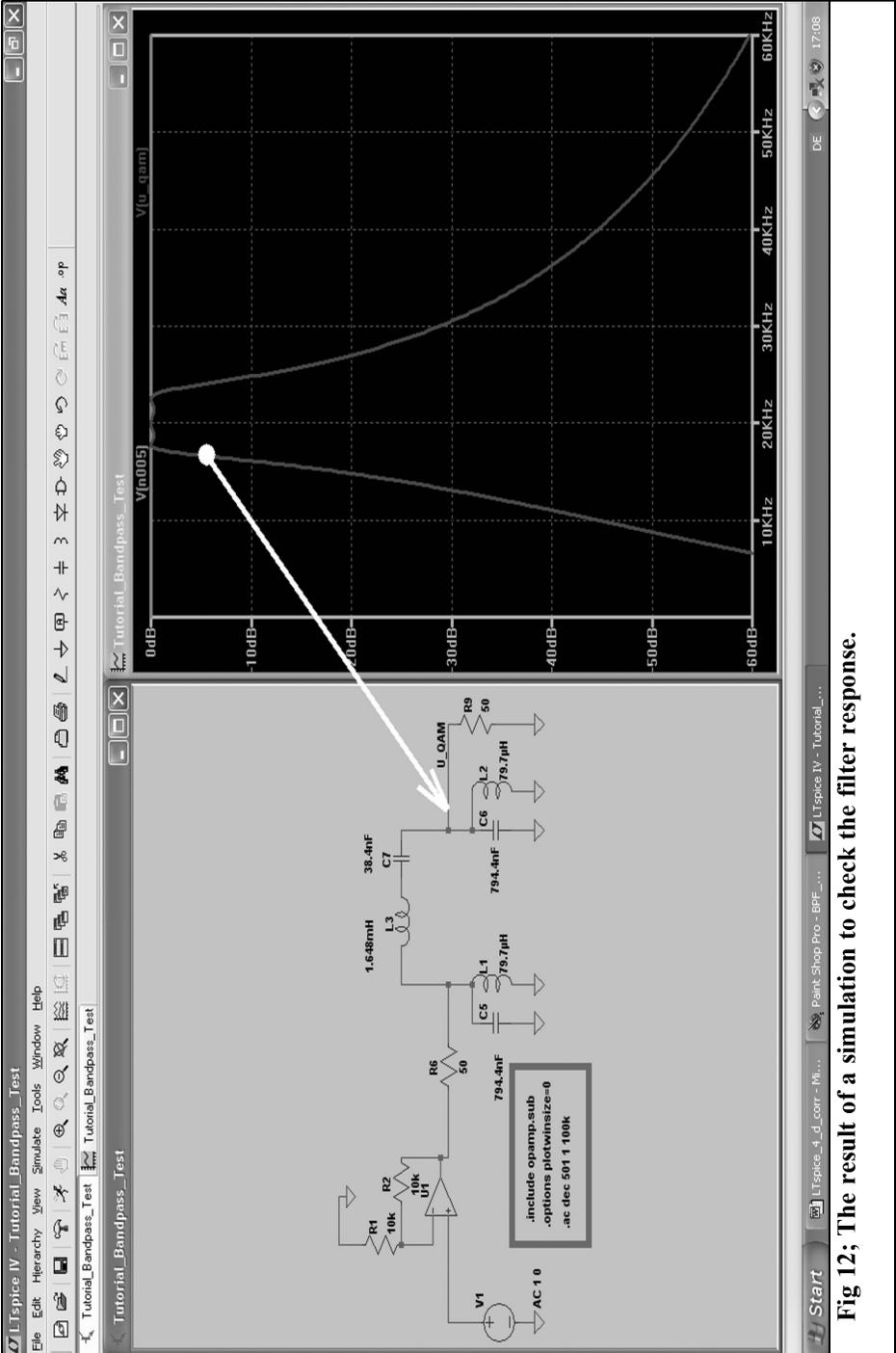


Fig 12; The result of a simulation to check the filter response.

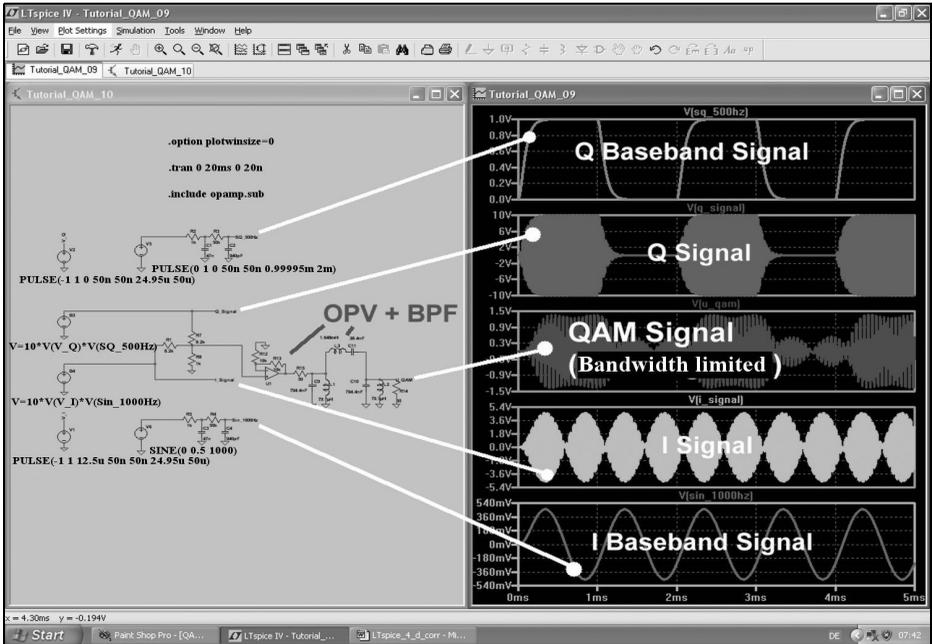


Fig 13: The QAM signal after the filter has been inserted.

3. Filtering the QAM signal to adjust the channel spacing

As shown in Fig 9 the spectrum of the QAM signal in this form is not suitable for transmission because adjacent channels are often needed (for FDM = frequency division multiplex). The channels should not disturb each other therefore each channel requires a suitable sharp bandpass filter to isolate it from its neighbours. A non-inverting operation amplifier with a gain of two follows the adding network for QAM production. It has an additional internal resistor of 50Ω and can feed a bandpass filter with a range of 6.8kHz and $R_i = R_A = 50\Omega$.

The bandpass filter design is child's play with today's programs (from the Internet). The choice of program is large but I use (probably from habit, because the

program has been installed for many years and because it is quicker to use) an old DOS program named "fds.exe". And these are the necessary inputs:

- Bandpass filter, Tschebyschef type
- Filter degree = number of the poles = $n = 3$
- $Z_{in} = Z_{out} = 50\Omega$
- Ripple = 0.3dB
- Centre frequency = 20kHz
- 3dB bandwidth = 6.8kHz

The input screen for the program is shown in Fig 10. Fig 11 gives the calculation results and "PLC" means "Parallel resonant LC Circuit" resp. "SLC" means "Series Resonant LC Circuit".

A test circuit is developed in LTSpice and the filter curve including attenuation is

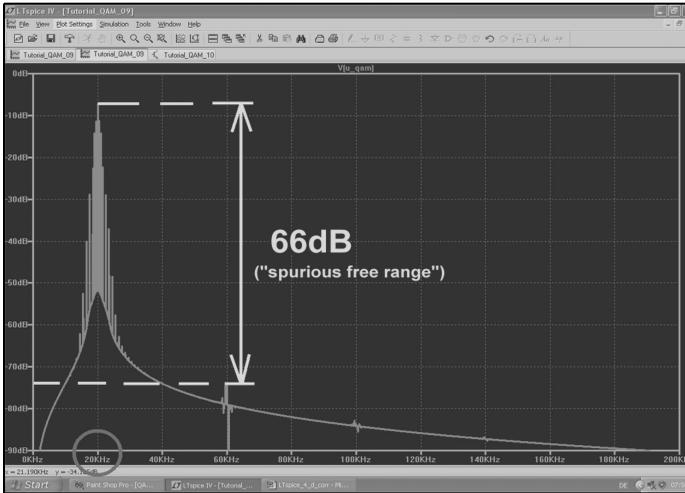


Fig 14: The spectrum shows that the filter technique has been successful.

examined as shown in Fig 12. The following things require an explanation:

- options plotwinsize = 0 means that the data compression is switched off.
- .ac DEC 501 1 100k results in a decade AC sweep from 1Hz to 100kHz and 501 points per decade
- For this AC sweep the amplitude is 1 and the phase must be set to 0 for the voltage supply V1 (AC 1 0) using the Properties of the voltage supply
- The operation amplifier comes from the library folder "Opamps" as "opamp"
- The Opamp must be activated in the SPICE model using the ".include opamp.sub" directive.
- The output voltage at the load resistance of 50Ω is attenuated (because of the attenuation in the filter) by 6dB = factor 2. Therefore the non-inverting operation amplifier circuit with a gain $G = 2$ is required so that at the filter output is the same level as at the input of the Opamps.

Clicking the button with the hammer opens the SPICE control panel; select "Gear" under "Default Integration Method" (this increases the stability of the simulation). Fig 12 shows the result giving the "all-clear" for this simulation.

Some work is required because the Opamp including the bandpass filter must be inserted in the large diagram for QAM production and a new simulation be started (Fig 13). A view of the spectrum is interesting, the band-limited QAM signal is shown in Fig 14. Now the adjacent channels cannot disturb each other, as they could with Fig 9 because the minimum attenuation of the unwanted signals amounts to 66dB.

4. Demodulation of the QAM signal

4.1. Principle of synchronous demodulation

On the receive side the two signals must be separated again and in addition the coordinate system used on the transmitter side is required. As already mentioned

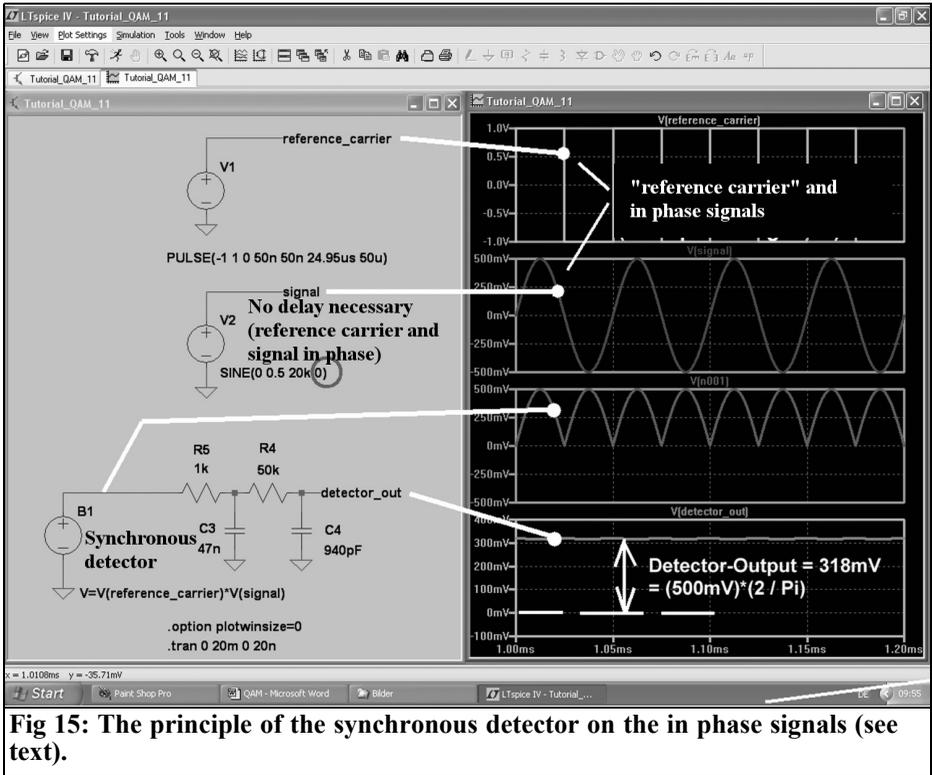


Fig 15: The principle of the synchronous detector on the in phase signals (see text).

this is the first and most important task in the receiver but that is a complex area.

Therefore it is simply assumed that this problem is already solved and the two carrier signals required with a 90° phase shift are available.

For the recovery of the two signals, I and Q, two product detectors are required. The received QAM signal is supplied to them together, but each detector is also fed with its associated carrier signal. This supplies the two different phase positions between QAM signal and carrier signal as demodulated in Fig 15 and 16. In Fig 15 an in phase signal is demodulated and after the multiplication of the two signals it can be seen that the circuit behaves like a bridge rectifier. The negative half waves of the QAM signal are commutated by the negative carrier square wave signal (third diagram from the top) with a QAM peak value of 500mVs the arithmetic

average value is produced at the output of the detector after the filtering.

$$V_{DC} = \frac{V_{peak} \cdot 2}{\pi} = \frac{500mV \cdot 2}{\pi} = 318mV$$

It is just the same as the text book says...

Fig 16 shows the case for those QAM signal components out of phase in relation to the carrier by 90°. There the voltage at the output of the detector reaches zero because a pure alternating voltage is produced after multiplication with equal positive and negative transitions without any DC voltage component (third diagram from the top).

4.2. The simulation of the complete QAM Demodulator

The main items required are two “bv” voltage sources for the usage as analogue multipliers.

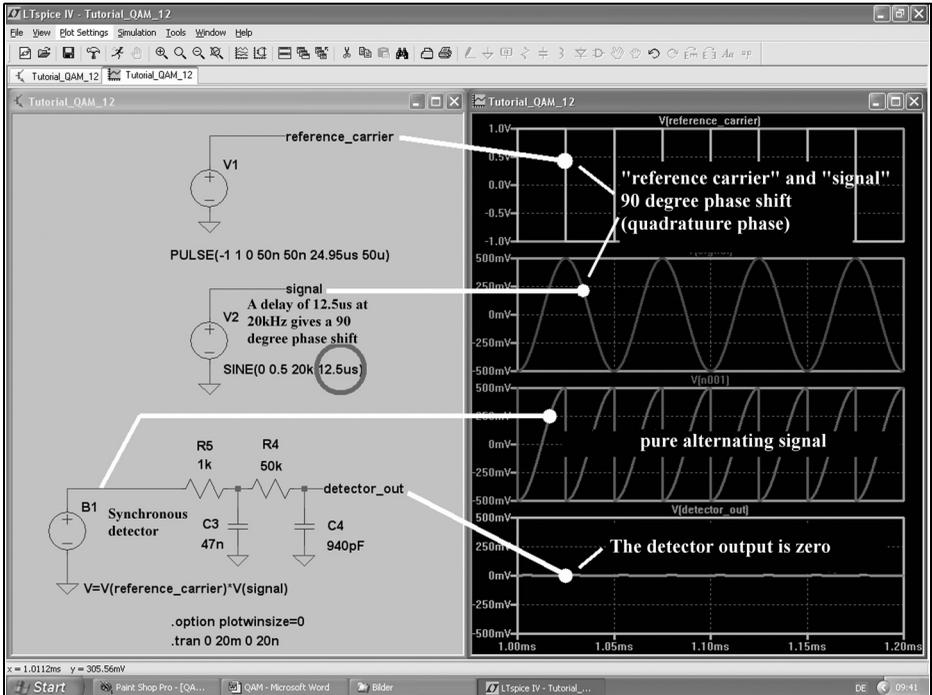


Fig 16: Demodulation of the quadrature phase signal is not more difficult to understand.

As discussed, each circuit receives the QAM signal as well as the appropriate carrier component with the suitable phase position. By multiplication the signals relating to the correct phase position are retrieved (the phase position agreeing).

The necessary formula inputs for the two bv sources are therefore:

for the Q path:

$$V = V(V_QAM) * V(V_Q)$$

and for the I path:

$$V = V(V_QAM) * V(V_I)$$

Thus the output of each multiplier will be the appropriate baseband signal.

A lowpass filter circuit at the output of each demodulator will remove the carrier frequency harmonics as shown in Fig 17. It can be seen that after demodulation the two information signals transferred are

identical to the baseband signals supplied. Only the rounded off corners and the tired edges of the square wave indicate that its harmonics were reduced. Also there is a time delay of the two output signals that indicates the time delays caused by the limited bandwidths of the filter circuits in each channel. But this is exactly the well known "group delay".

5.

Literature

[1] LTSpice Tutorial (volume 1 and volume 2) of the author. Can be downloaded from www.elektronikschule.de/~krausg from the Internet

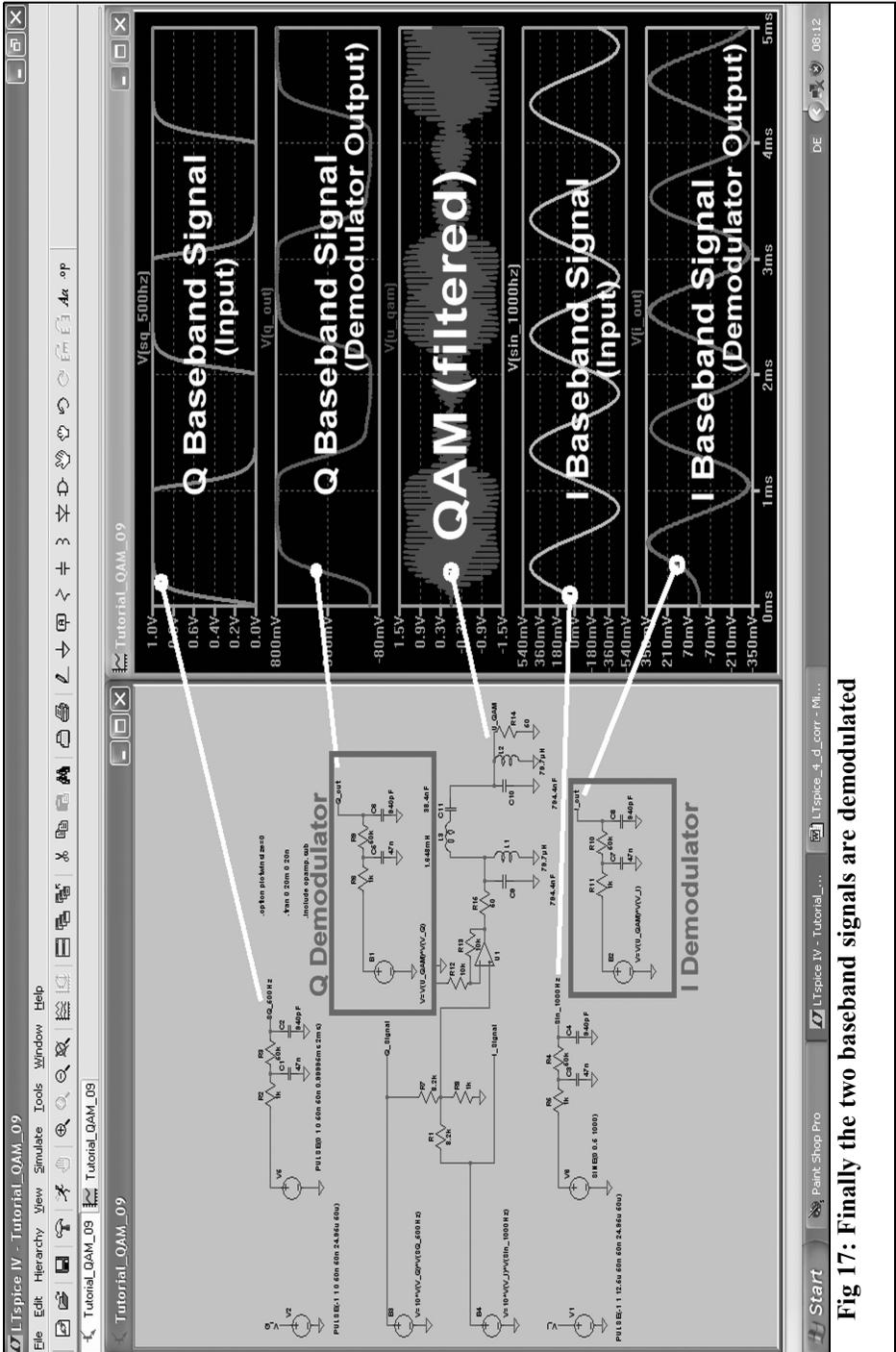


Fig 17: Finally the two baseband signals are demodulated



Ralf Rudersdorfer, OE3RAA

Sources of external noise and its effects on radio reception

The spectrum is heavily loaded with atmospheric and cosmic radiation that is dependant on the receive frequency, the receive mode and the antenna. Additionally ever increasing sources arise from the so-called industrial noise or technical noise. These can impair or prevent reception of the smallest signals. The specification for a receiver sensitivity determined by the self-noise are partly to blame. Therefore it is important to know the external sources of noise that contribute to the quality of radio reception.

The following article considers this for the frequency range between 1Hz and 100GHz. The definition and meaning of receiver sensitivity is described in [1] as well as the different types.

1.

Introduction

For most receive modes the actual usable sensitivity of a radio receiver depends greatly on the receive frequency. This is caused by the sum of all the noise like signals received by the antenna. Because the noise from the antenna is often higher than the thermal noise of the radiation resistance of the antenna; it is called

external noise. The primary sources of radiation causing the noise can be classified as follows:

- Atmospheric noise; for low receive frequencies this is due to world-wide lightning discharges, Fig 1 (a high number of lightning strikes happen each second particularly in the tropics). On higher receive frequencies the cause is atmospheric gases and hydraulic meteors (a generic term for all forms of condensed water - "hydraulic" - that are observed in the atmosphere).
- Galactic and cosmic noise; on high receive frequencies where the ionosphere is permeable the noise is generated by radioactive heavenly bodies, the centre of the Milky Way and the Sun. This is known as the cosmic background radiation.
- Earth or ground noise due to the earth and the earth's surface (Fig 2). This is because each body that absorbs radiation also produces radiation, dependent on its temperature, the roughness of the surface, the humidity and the frequency. A concrete surface or a stony mountain face pours out radiation compared to a lawn [2]. A hypothetical ideal black body [3] would have the greatest capacity to absorb ra-



Fig 1: Energy sources such as high voltage lines and suns (right) contribute to the external noise level.

radiation and would also produce the greatest thermal radiating power at the given temperature. Since such a body radiates equally in each direction the radiation produced by it is the same in all directions. It does not depend on the material condition, only on the temperature. The thermal radiation produced by a black body has a universal character and no material body can radiate stronger at any frequency than a black body. Therefore the radiating power of a material body is referred to the maximum value given by the black body.

The relationship of the radiation intensity delivered by a body to the radiation intensity of a black body at the same temperature is the emissivity. This can take values between 0 and 1 [3]. The earth shows different colouring explaining the different noise intensities; it is not a black body

- Thermal noise or Johnson noise, Nyquist noise or resistor noise; caused by ohmic loss resistance due to the varying speed of the electrons and from irregular lattice vibrations that transfer charge carriers by impact. (The resistor noise is not, by definition, among the sources of external noise, it is however produced by the material of an antenna. Its actual contribu-

tion determines the antenna efficiency).

- Technical or man made noise; caused by all kinds of electrical switching transitions, sparks, impulses etc. due to insufficient or defective electromagnetic shielding or noise suppression. (Causes include, igniting sparks, switching in the lighting systems, phase controls, neon tubes, fluorescent lamp ballasts, fast switching edges in switch mode power supplies, power transformers, plasma displays etc.)

Signals received over the air from a distance have a noise characteristic. The sum level can be so strong from low frequencies up to VHF that the weakest receivable information signal is not determined by the self-noise of the RF amplifiers used. In modification of the definition for noise figure/factor ([1] and Fig 3) it can be determined by the external noise figure and is universally represented as the relationship of the entire received noise power to the thermal noise power. Because the external noise figure (Fig 4) in dB, according to the definition from [6] is more or less the noise power of the antenna at the receiving location represented as the ohmic resistance of the real part of the feed impedance of the antenna at 290K. (Also known as the noise temperature as shown in Fig 3 and Fig 4).

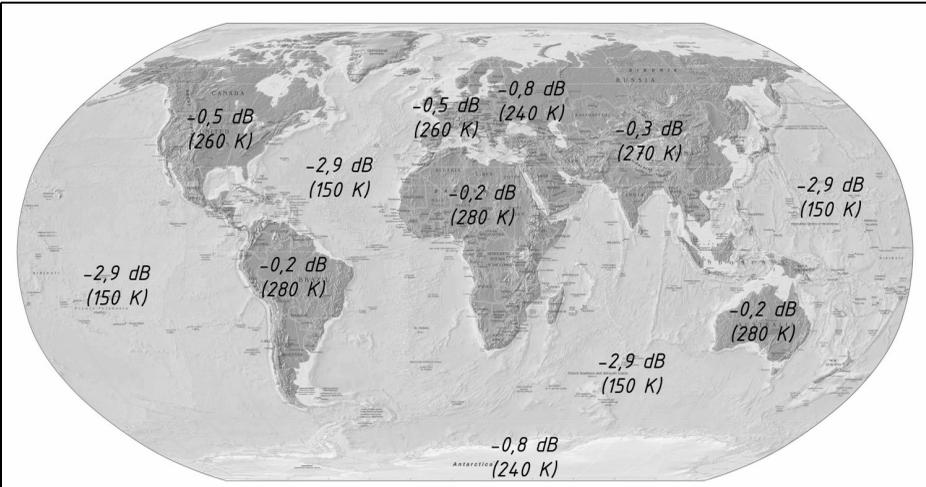


Fig 2: Noise levels of the earth's surface, the figures shown are for 14GHz in accordance with the ESA/EUTELSAT model [4]. Frequently if the noise temperature of the earth is 290K. [5] states that "The earth can be approximated as a black body radiator with an average temperature of 290K. If an antenna had its entire radiation pattern directed towards the ground, the temperature would be approximately 290K. The precise temperature will vary somewhat with the local terrain and physical temperature of the earth surface".

2.

System calculation

Consider a receiving system and receiving antenna, Table 1 shows the receiver coupled to the antennas so that the smallest perceptible signals are determined by the external noise if they are greater than the receiver noise level. The following is desirable:

$$F_{dB} < F_{AdB} \tag{Gl. 1}$$

where

F_{dB} = receiver noise figure in dB

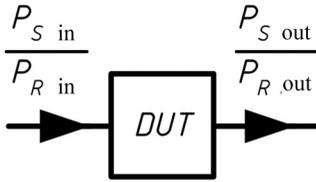
F_{Adb} = external noise figure, in dB

The intensity of the external noise for the receive mode depends upon:

- the receive frequency
- the receive location (for example quiet rurally, urban or industrial environment)
- the time of operation (day/season)

Table 1: Changes in receiver performance with noise level.

Receiver noise level	Sensitivity degradation
$F_{dB_i} = F_{AdB_i}$	Receiver performance degraded ≈ 3 dB
$F_{dB} = F_{AdB} - 6$ dB	Receiver performance degraded < 1 dB
$F_{dB} < F_{AdB} - 10$ dB	small degradation in receiver performance



Two-port

$$F = \frac{P_{R \text{ DUT}} + k \cdot 290 \text{ K} \cdot B_R}{k \cdot 290 \text{ K} \cdot B_R}$$

$$F_{dB} = 10 \cdot \lg(F)$$

$$T_R = 290 \text{ K} \cdot (F - 1)$$

Fig 3: The definition of noise factor/figure/temperature for a two-port. The interpretation of external noise factor/figure is shown in Fig 4. The available thermal noise power at the reference temperature is also considered. Table 1 shows simple examples of noise received from the environment to show this distinction. ($k = 1.38 \times 10^{-23} \text{ W s/K}$; $P_{R \text{ DUT}} = P_{R \text{ out}} - V \times P_{R \text{ in}} \text{ W}$; $B_R = \text{equivalent noise bandwidth in Hz}$)

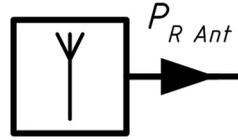
and the propagation conditions (ionospheric)

- the receiving antenna used, possibly directional and the elevation of the lobe,
- the antenna adjustment.

3.

Estimating the intensity of the external noise

The data and evaluation from [6] can be used to make an educated quantification



Receive antenna

$$F_A = \frac{P_{R \text{ Ant}}}{k \cdot 290 \text{ K} \cdot B_R}$$

$$F_{A \text{ dB}} = 10 \cdot \lg(F_A)$$

$$T_{R A} = 290 \text{ K} \cdot F_A$$

Fig 4: The definition of external noise factor/figure/temperature (compared with Fig 3) at a receive location above the available noise power $P_{R \text{ Ant}}$ of a lossless receiving antenna. High efficiency receiving antennas that are cooled can give lower received noise than the thermal noise of the reference temperature (in the denominator) so that negative external noise figures (for instance the cosmic background radiation) are explained. ($k = 1.38 \times 10^{-23} \text{ W s/K}$; $B_R = \text{equivalent noise bandwidth in Hz}$)

of the external noise. Data taken from an extensive worldwide measuring campaign (Figs 5 to 7) replace the CCIR report 322 data that has been used for many years.

This data serves as a reference point for receive system concepts. The data in the three diagrams are principally for antennas with Omni directional characteristics. A deviation of approximately $\pm 5 \text{ dB}$ can be expected for highly directional HF antennas for atmospheric noise due to lightning; this is dependent on the antenna adjustment and operating location.

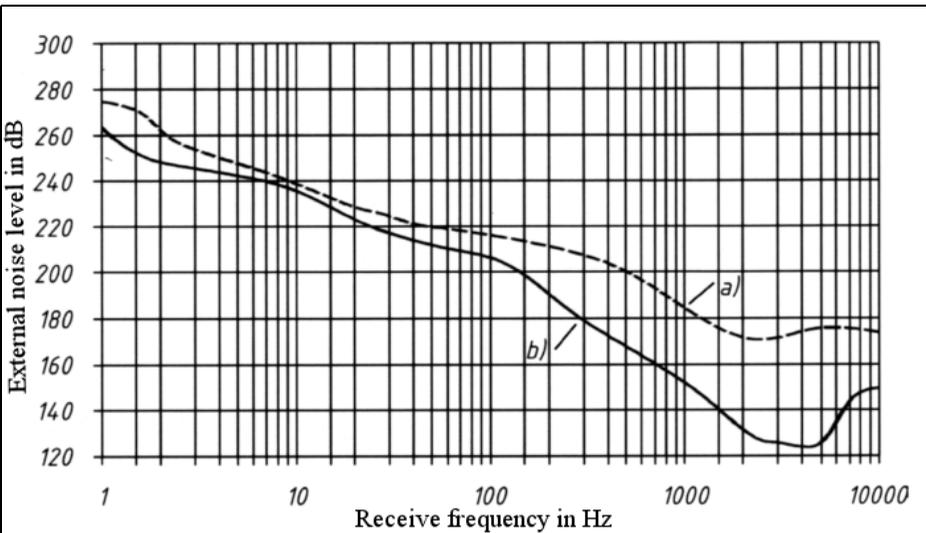


Fig 5: External noise level in the frequency range from 1Hz to 10kHz in accordance with [6]. It shows atmospheric noise: a) The maximum intensity expected that is rarely exceeded; b) The minimum intensity expected that it rarely falls below.

In each case the solid curves show expected minima. These models are discussed in [7]. The interpretation margin is large, since the basic conditions are variable and unstable which makes more exact characterisation difficult!

For example for the frequency range around 227MHz measurements shown in [8] compared with those in [6] indicate the values for the antenna noise figure in a business environment to be 4.2dB and 8.5dB higher than in a residential environment. This data was taken from about 3.3 million individual mobile measurement values along 480km of roads in the Swiss canton Zurich. Current measurements from the British Broadcasting Company (BBC) and the Munich Institute for Radio Technology (IRT) would confirm these values. The introduction of electronic devices and increase in building in recent years has led to a great increase in man-made noise. For the computations in [8] another definition of the antenna noise figure is used then in [6] and in Fig 4 is fixed. Because of the

large radiation angle for omni antennas above 1.5m the measurements in [8] show the earth noise received is mainly dependant on the ambient temperature. Additionally the levels determined for the man-made noise lie far above thermal noise.

Due to these two facts the difference is due to the different definitions used for the antenna noise figure in [8] and therefore no further correction is required.

4.

Possible field strengths disturbance

The possible field strength disturbances due to external noise expected can be shown from the approach in [7]

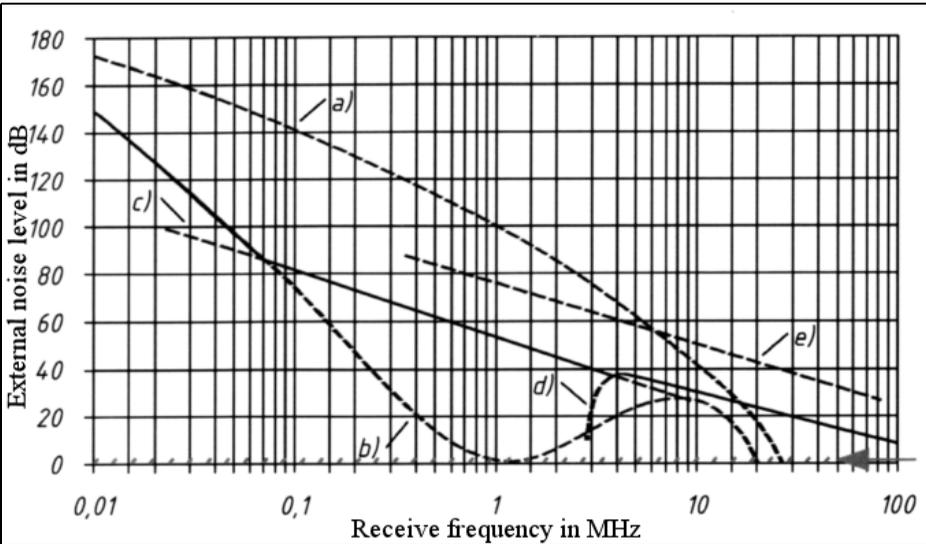


Fig 6: External noise level in the frequency range from 10kHz to 100MHz in accordance with [6]. They are shown in by intensity:
a) Atmospheric noise (Maximum, overrun for more than ≈ 0.5% of the time)
b) Atmospheric noise (Maximum, overrun for more than ≈ 99.5% of the time)
c) Technical noise (Quiet rural location)
d) Galactic noise
e) Technical noise (Noisy industrial location)
For example the external noise that can be expected below 6MHz for 99% of the time will be between a) and b).

$$E_A(B_{-6dB}) \approx F_{AdB} + 20 \cdot \lg\left(\frac{f_A}{10^6 \text{ Hz}}\right) + B_{RdB} - 100.3dB\left(\frac{1}{m}\right) + G_{Ant}$$

(Gl. 2)

where

- E_A = from the receive bandwidth (B_{-6dB}) dependig external noise field strength in dB ($\mu\text{V/m}$)
- F_{AdB} = external noise figure in dB
- f_A = working frequency Hz
- B_{RdB} = equivalent noise bandwidth of the receive bandwidth in dBHz
- G_{Ant} = antenna gain in dBi

For reception in the 41m band using a short vertical receiving antenna with a gain of approximately 1.8dBi and 9kHz

receive bandwidth an external noise figure of at least 34dB is estimated in accordance with Fig 6.

The field strength disturbance level that can be expected is:

$$E_A(B_{-6dB} = 9kHz) \approx 34dB + 20 \cdot \lg\left(\frac{7.325MHz}{1MHz}\right) + 39.5dBHz - 100.3dB\left(\frac{1}{m}\right) + 1.8dBi \approx -7.7dB (\mu\text{V/m})$$

Expressed as a voltage this is 0.4 $\mu\text{V/m}$. At an urban location, which is strongly influenced by technically generated radiation emittances it must be increased by up to 22dB field strength disturbance as shown by curve e) in Fig 6. For

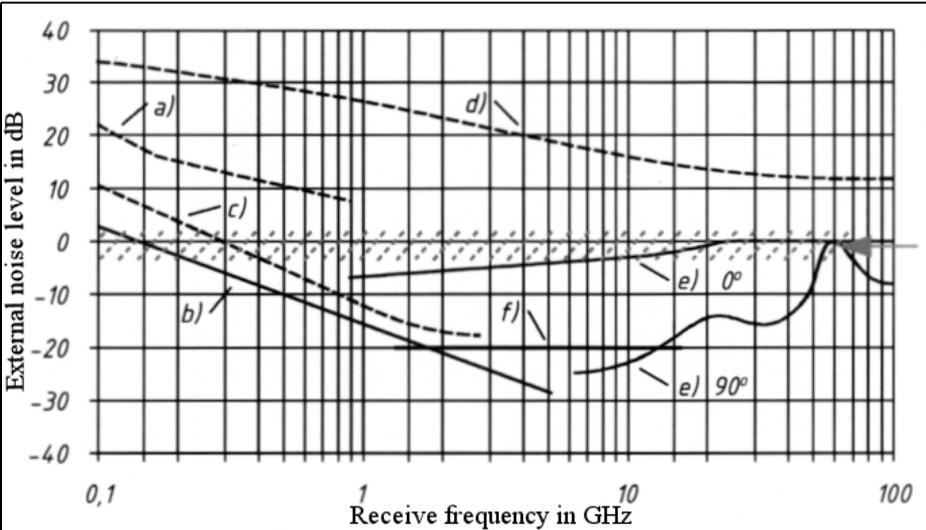


Fig 7: External noise level in the frequency range from 100MHz to 100GHz in accordance with [6]. They are shown in by intensity:

- a) Technical noise (noisy industrial location)
- b) Galactic noise
- c) The galactic centre (with narrowest antenna lobe)
- d) Quiet suns (antenna with 0.5° beamwidth directed at the centre)
- e) Oxygen and water vapour (with narrowest antenna lobe at 0° or 90° elevation)
- f) The cosmic background radiation as an absolute minimum with (a noise temperature of 2.7K)

reception at the centre of the VHF broadcast band with an appropriate receive bandwidth of 120kHz the conditions change:

$$E_A(B_{-6dB} = 120kHz) \approx 10dB + 20 \cdot \lg\left(\frac{97.75MHz}{1MHz}\right) + 50.8dBHz - 100.3dB(1/m) + 1.8dBi \approx 2.1dB(\mu V / m)$$

Expressed as a voltage this is 1.3μV/m. With very strong industrial noise on the receiving channel and with 120kHz bandwidth gives a field strength disturbance level of approximately 17dB (μV/m). For the receive bandwidth the equivalent noise bandwidth for calculations is the -6dB bandwidth of the receiver with a small (good) shape factor were used in each case. The following conversion applies:

$$B_{RdB} = 10 \cdot \lg\left(\frac{B_R}{1Hz}\right) \quad G1.3$$

where

- B_{RdB} = equivalent noise bandwidth for the receive bandwidth used in dBHz
- B_R = equivalent noise bandwidth for the receive bandwidth used in Hz

The background for this is described in more detail in [1].



5.

Conclusion

It can be determined: For reception under 35MHz the weakest receivable signals today by using a state of the art designed receiver are only rarely determined by the receiver sensitivity. The equation (Gl. 1) is practically almost always fulfilled.

The requirements regarding good sensitivity and high intermodulation performance [1] are diametrical opposed to each other and inappropriate small receiver noise level leads to losses for the large signal compatibility [1]. For higher receive frequencies the weakest receivable signals are actually determined by receiver sensitivity! In these cases the advantages of a noise-optimised receiving system are apparent, for example in chapter V.1 in [1] this is introduced and explained.

I must thank Dipl. Ing. Harald Wickenhäuser and Dipl. Ing. Markus Ortner for the technical discussion in the course of producing this article.

6.

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Ralph Berres, DF6WU

Extending the range of a Racal Dana 1991 Frequency Counter to 12GHz

1.

Introduction

There are low priced frequency counters from Racal Dana on the surplus market. The type 1991 frequency counter has 2 inputs, the main input is specified to 160MHz and the second input is specified to 100MHz. The type 1992 has a third input specified to 1.2GHz. Both frequency counters have a 9-digit display. There are various options, for example an IEEE488 interface. There are more counters, 1 or 2 channel, in the series 19.91 – 1998; all use the same motherboard so they have the same connection pins for the channels. Similarly the board locations for the auxiliary PCBs is the same so they can be modi-

fied for additional channels.

The specified measuring range of the type 1991 is only up to 160MHz but it actually works up to nearly 220MHz. The function of the third channel was examined using the service manual [1]. It is essentially a preamplifier with a passive level control and a divide by 64 counter. The divide by 64 counter is only useful to 1.3GHz. There must be still more possible.

To investigate how the frequency counter can actually perform the appropriate pins on the motherboard are activated as if the third channel was present. An RF signal generator with a level of +13dB was connected to the pin that the divide by 64 signal uses and the frequency was increased until the counter began to miss-count. The frequency limit was determined to be 240MHz.

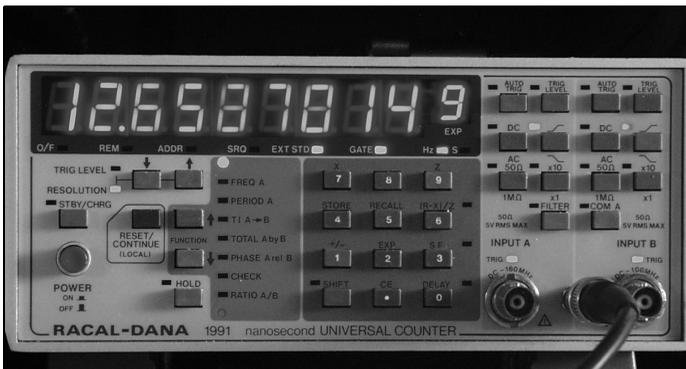


Fig 1: A model 1991 with the modifications described showing 12.65GHz.

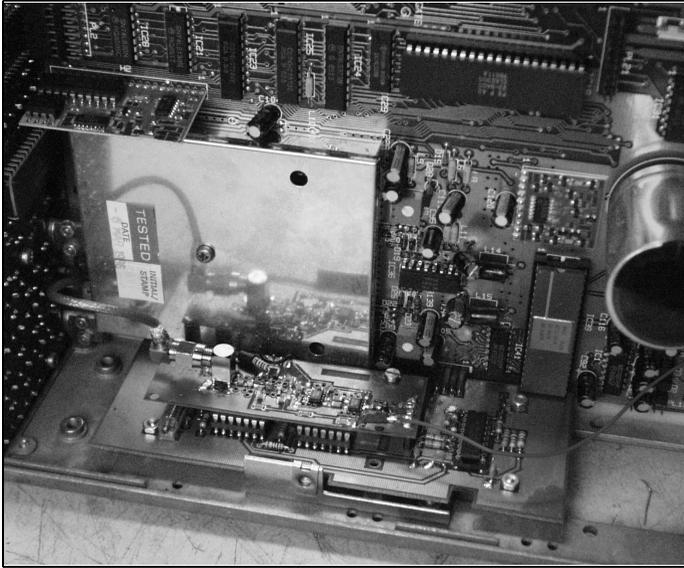


Fig 2: A side view showing the two PCBs installed. The upper PCB is the RF PCB. To the left is the front panel PCB with the input channels A and B in a shielded box.

So the maximum possible input frequency is: $240\text{MHz} \cdot 64 = 15.36\text{GHz}$. Why not build a third input port that still counts at least to the 3cm band? After some development of the Racal Dana 1991 a third channel with a frequency limit increased by a factor of 10 was produced.

Modification of the counter is described in the following article.

2.

The modification

After studying the original circuit it was decided to use the control circuit section and put the new RF module on a separate PCB piggybacked on the control PCB. But the RF PCB had to be designed.

The control PCB uses an MC10115 tri-state ECL receiver. It contains 3 receivers with 2ns switching time and schmitt trigger characteristics. Commercial IC cannot simply replace this IC. The supply resistors for the MC10116 were changed

to cope with the higher frequency range.

Like the original circuit, the new circuit contains a signal detector that only enables the counter if the signal level is high enough for counting. On the control PCB the level measurement was immediately before the divider. The first prototype signal detector did not work reliably between 7 and 10GHz. Therefore the signal detector has an adjustment trimmer R27 to adjust for reliable operation.

A 2.2pF capacitor was later added to the control PCB. It is in parallel with the 180Ω R21 to compensate for a drop in level above 150MHz. That corresponds to 12.5GHz at the RF input. The two capacitors C51 and C52 (15pF) are removed. A 180Ω SMD resistor was soldered between pin 4 and pin 5 of IC U2A to prevent overloading IC U2A. This became apparent because a double frequency was displayed occasionally at low frequencies (under 1GHz). Resistors R39, R68, R69 were removed. The hysteresis of the signal detector was too large so it did not trigger occasionally. R62 was changed to 100Ω and R29 to $18k\Omega$.

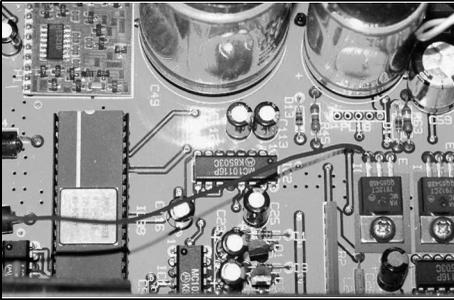


Fig 3: The wire running across the PCB is a direct connection to the 12V power supply.

The photograph in Fig 12 shows the old PCB layout with the components soldered in place. The PCB layout has been changed for a later version (R29 is on the circuit diagram is 100k Ω instead of 18k Ω).

The RF PCB is made from of 0.8mm thick FR4 material. The conductor width is not exactly appropriate for 50 Ω impedance. The impedance is approximately 65 Ω but that is still acceptable.

All coupling and decoupling capacitors for 12GHz consist of parallel connected capacitors of: 100pF + 10pF + 1pF. The 1pF capacitor is particularly important at 12GHz. The larger capacitors behave inductively at this frequency.

The input amplifier uses two-cascaded NBB300 amplifiers from RFMD. They have a gain of 12dB up to 12GHz. Two-cascaded HMC363S8G divide by 8 counters from Hittite follow; these are specified up to 12GHz.

An ERA3 broadband amplifier that raises the signal from the divider to the level required by the control PCB follows the divider. The supply voltage of 12V is taken directly from the power supply of the counter (Fig 3).

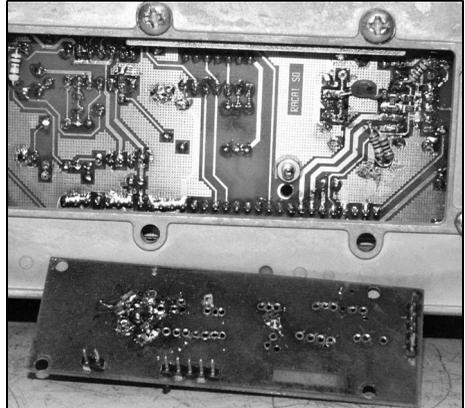


Fig 4: A view of the right hand side of the counter. The connections for the RF PCB can be seen on the main PCB.

3.

Construction

The control PCB was designed with two SMD components and wired components.

There are IC bases with precision contacts with plastic bodies that overhang the PCB by 1.5mm. It is advisable to solder these on both sides first with a fine point soldering iron. The contact strip that connects to the motherboard requires a bent socket strip. Some components were added later in the photograph (Fig 4), e.g. the 150 Ω SMD resistor at the input of IC U2A.

The RF PCB consists of double-sided 0.8mm FR4 PCB material and only uses 0602 SMD capacitors and 0804 SMD resistors. All of the RF ground connections to the ground surface are first drilled with a 5mm hole and hollow rivets used as plated-through holes. It is important that the lower ground surface has the same potential at 12GHz as the ground connections on the topside otherwise the PCB will not work correctly.



R39, R68, R69 only fitted if necessary

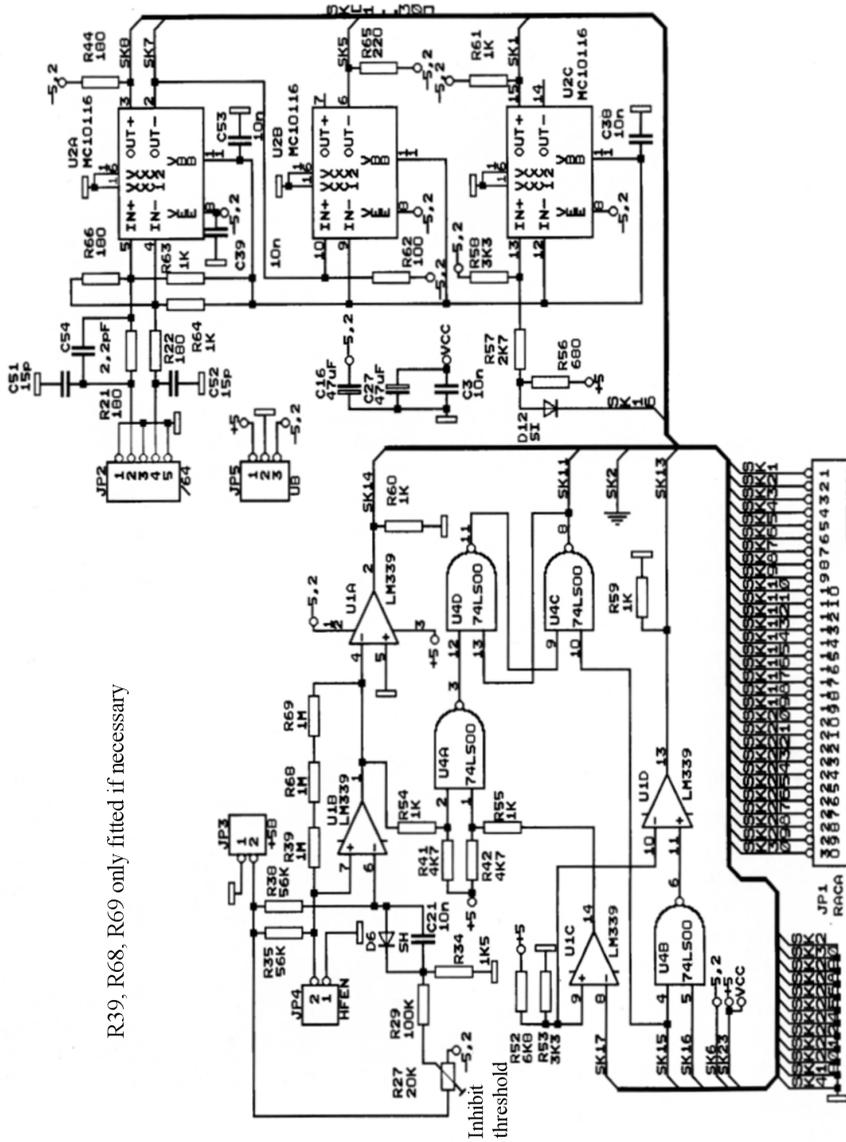


Fig 5: Circuit diagram of the control logic.

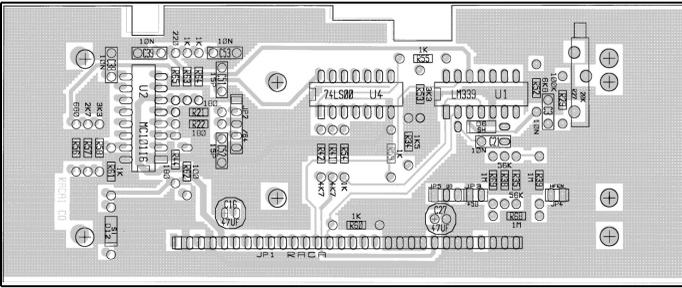


Fig 6: Component side PCB layout of the control board.

The SMD components are then fitted.

The 12GHz coupling and decoupling capacitors consist of a 100pF capacitor and 10pF capacitor soldered next to one another then the 1pf capacitor is soldered on top. The two counter ICs are tacked to the PCB by two corner legs after the ground surface under them has been tinned and covered with flux. The surface under the ICs is then heated from below until the solder under their ground pad melts ensuring that not too much solder has been used because that may short out other connections. The remaining legs can then be soldered in place. The RF tracks should be tinned sparingly.

The SMA socket is soldered to the PCB with a clearance of approximately 2mm from the PCB. The 4 ground connections are soldered from both the top and bottom. Then the centre conductor pin is heated from the bottom side and solder fed in to fill the gap. Only a branded socket specified for 12GHz should be used, it should be gold plated with Teflon insulation.

The RF PCB is attached to the control PCB using two 10mm long hexagonal

spacers. The holes for these must be drilled in the PCB and insulating washers used on at the connection to the control PCB. The control PCB is screwed to the side plate with spacers. The control PCB has to be cut out along the outside edge so that it fits into the counter.

4. Notes

Removing the rear framework opens the case of the counter; it is fastened with 2 screws. The cover can then be removed towards the rear of the counter. Next the IEC bus plate is removed. It is fastened at the front with 2 screws holding a bracket and with 2 screws on the IEC bus socket. The plate can then be removed carefully. It is still connected with a ribbon cable to an IC socket this can be removed carefully. Next the front panel is removed. The two nuts attaching the BNC sockets and 4 countersunk screws that attach the front panel to the sides are removed. The printed circuit board at the front is fastened with 4 screws these are removed.

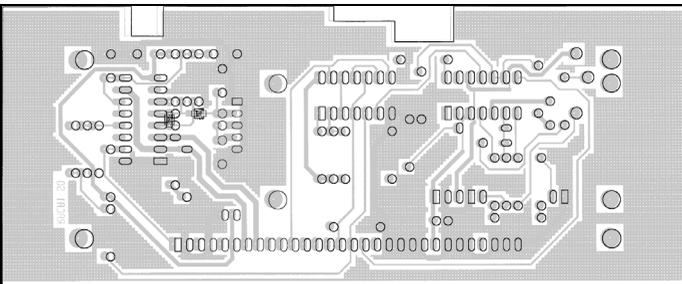


Fig 7: Reverse side PCB layout of the control board.

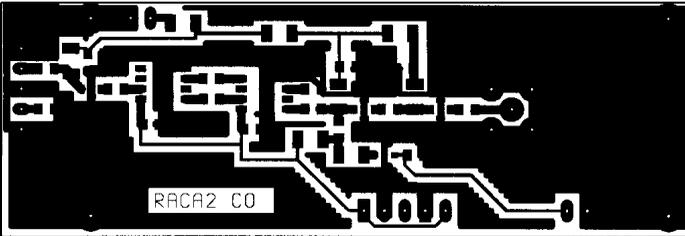


Fig 9: Track side layout of the new RF PCB.

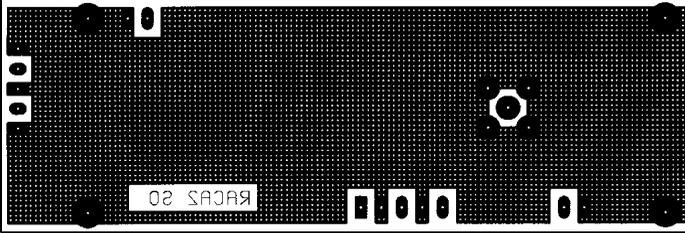


Fig 10: Reverse side layout of the new RF PCB.

After the printed circuit board is removed it reveals holes next to the LEDs that display the input functions. The surface layer can be removed with a glass fibre pin using a turning motion in this hole until the foil becomes visible. The relevant LEDs are already fitted. The Racal 1991 only differs from the 1992 by the missing components for third input and a different front panel foil.

Next a suitable hole for an SMA socket is drilled in the foil using the existing pilot hole of the front panel. Make sure the centre mark is used because the cutout in the front panel has an anti-twist hole for the BNC socket. It is easy to make the hole in the wrong place. An SMA socket should be used having a cable attached with a plug to connect to the RF PCB.

When the socket including the cable is fitted the assembly of the front panel is carried out in reverse order. Be careful

with the plug connecting the front panel to the motherboard. Try the counter after this is complete to ensure that everything works correctly.

Now fit the control board connecting it to the correct row of pins. The counter should now announce itself as a 1992. Two new functions should be available indicated by the front panel LEDs. The third input should be selected when moving up and down and the frequency relationship between channel 2 and 3 should be correct.

Select the third input port. Attach a signal generator to JP2 pin 2 via a 1nF coupling capacitor with the ground on pin 5. The level should be about 0.5V. 64 times the frequency of the signal generator should appear on the display. Adjust R27 so that the counter is just disabled. So far everything is correct, switch the counter off and fit the RF PCB on to the control

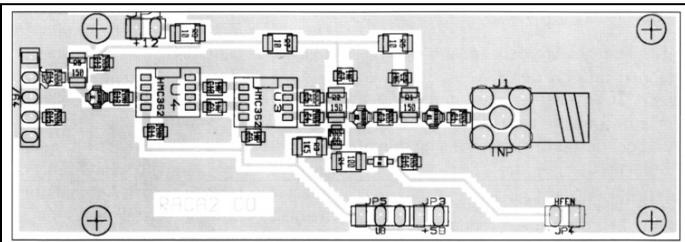


Fig 11: Component layout of the new RF PCB.

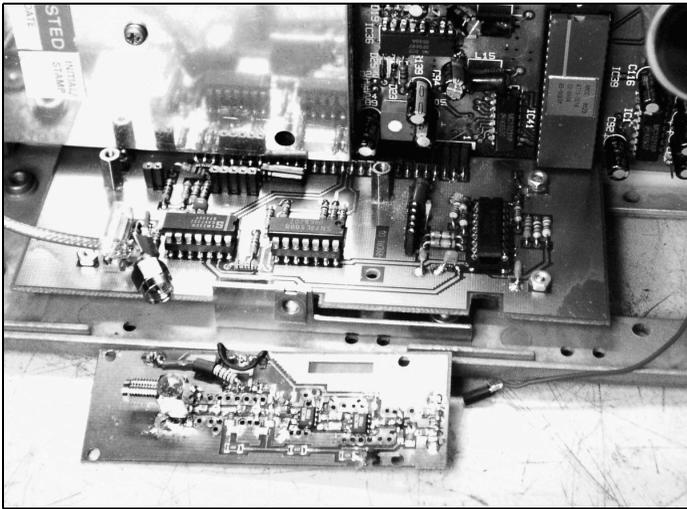


Fig 12: View showing the control PCB fitted without the RF PCB mounted.

PCB. Also attach the cable from the front panel SMA connector to the RF PCB.

Switch the counter on and select channel C. The display should swing wildly, something between 6 and 8GHz. Attaching a microwave signal generator should stop that if the PCB is working correctly. It should indicate the frequency supplied by the signal generator. Sensitivity may vary between approximately -30dbm at 100MHz and -15dbm at 12.5GHz. See

Fig 13 showing the response curves. Potentiometer R27 adjusts the threshold where no swinging is indicated. The sensitivity is not so high everywhere, in particular between 7 and 10GHz the level is already -9dbm before the counter is enabled. The cause is the level detector on the RF PCB it is much less sensitive between 7 and 10GHz.

If everything is satisfactory, assemble the counter in reverse order. A comprehen-

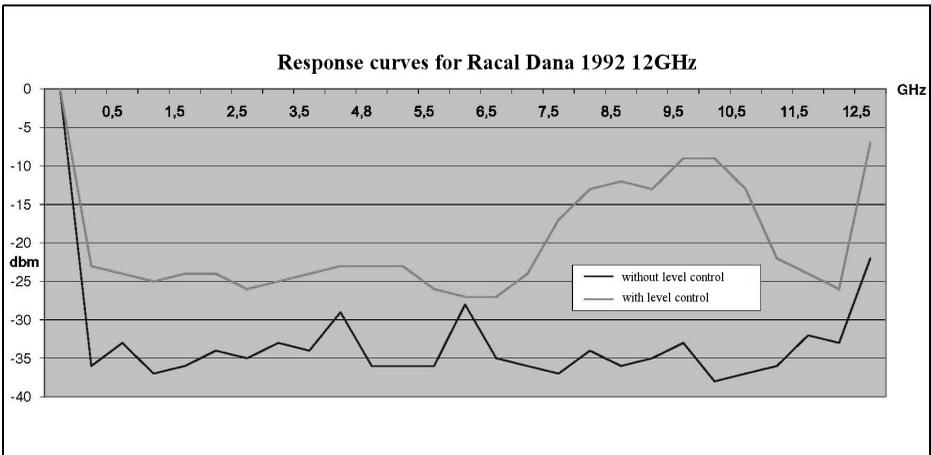


Fig 13: The response curves, one without level control and the second with the level control set so that "0" was displayed with no signal input. At each frequency the signal level was increase until a stable reading was obtained.

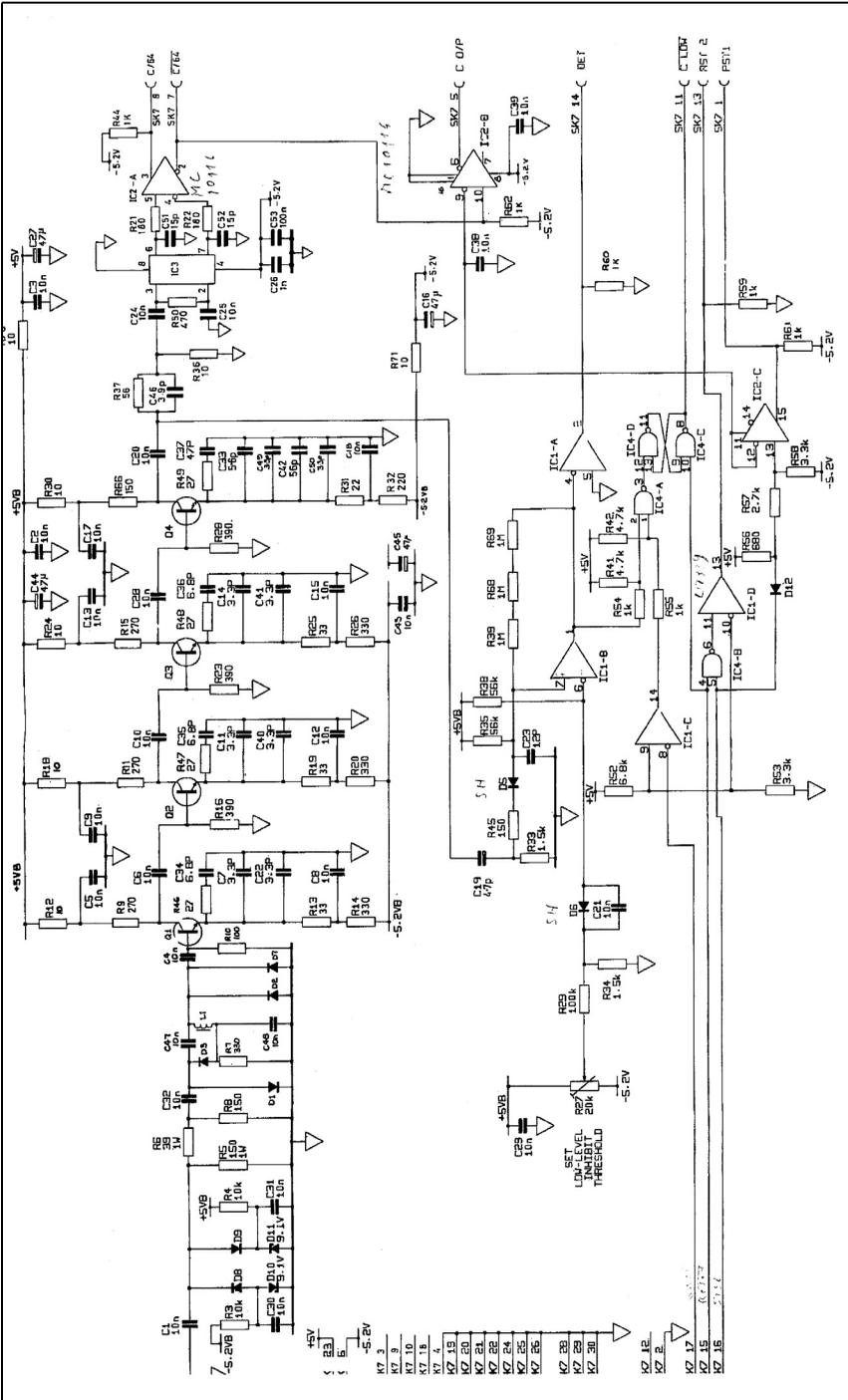


Fig 14: The original circuit for channel 3 of a Racal Dana model 1992 counter.



sive function test of the completed unit should be carried out.

HMC363S8G; <http://www.hittite.com/>
 Hollow rivets; <http://www.reichelt.de> -
 Part no.: RIVETS 0,6MM

5.

Sources of supply of the special parts

Possible suppliers are:
 MC10116; <http://www.abra-electronics.com>
 NBB300; <http://farnell.com/>

6.

References

[1] Service Manual for the Racal Dana 1991:
<http://bama.edebris.com/manuals/racal/1991/>



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Paolo Antoniazzi, IW2ACD and Marco Arecco, IK2WAQ

High Performance 10GHz Dish Tests Using a Simple 20dB 1W Amplifier

How to measure more than 40dB F/B and side lobes of quality antennas with a simple, but powerful amplifier

Most manufacturers of high performance microwave antennas offer paraboloids with a typical Front-to-Back (F/B) ratio of 40 to 65dB. A high performance microwave antenna may also use a shroud absorber and a radome in addition to the components normally used in quality dishes.

Quality antennas with such very high performance are not usual on the 10GHz amateur band, but this is a new interesting field for simulation and testing. We are working with simulation and preparing for measurements that today are very difficult. Our first step was the use of more output power for the test signal. Our solution is realised with the Toshiba module TMD1013-1, a simple 1W 10GHz amplifier with gain >20dB.

Assuming an electrically transparent structure supporting the feed and no ohmic losses in the system, the ideal antenna would be 100% efficient, i.e. the aperture will produce the full theoretically available gain. In the real world that is not possible and it may not even be desirable to achieve these characteristics. For example, the system engineer may want very low sidelobes, in which case he may have to pay dearly in efficiency. Losses are introduced by:

- Phase error across the aperture
- Amplitude distribution taper
- Spillover (energy lost by radiation from the field not incident on the reflector)
- Aperture blockage
- Cross-polarisation (unwanted polarisation)
- Radome and ohmic losses

1.

The Paraboloid

Simple and inexpensive, the paraboloid is the most widely used microwave antenna reflector. The illumination pattern should yield uniform phase and amplitude distribution over the surface of the dish, without spillover (no energy missing the reflector). The reflector of this antenna would not deviate from a true paraboloid.

The departure from uniform phase across the aperture of a paraboloid can be ascribed to the scattering from, and blocking due to, the primary feed and the systematic deviation from a true parabola of the reflector. Surface tolerances normally are responsible for between 0.1 and 0.5dB of the loss in microwave antennas. Perfect accuracy is unattainable, and construction of a precise sur-

**Table 1: Typical Paraboloid Antenna Losses.**

Losses (dB)	Horn Feed	Scalar Feed	Shaped reflector
Phase error of a good reflector	0.20	0.20	0.20
Phase error of feed	0.50	0.30	0.20
Amplitude taper	0.70	0.80	0.20
Spillover	0.50	0.20	0.20
Aperture blockage	0.15	0.20	0.20
Cross polarisation	0.15	0.15	0.20
Ohmic loss	0.10	0.10	0.10
Total gain loss (dB)	2.30	1.95	1.30
Efficiency	59%	64%	74%

face is very expensive. The generally accepted rule-of-thumb figure is that the permissible deviation is 1/32 to 1/16 wavelength.

As an example, a paraboloid reflector of circular aperture, having a diameter of 25 wavelength and typically tapered illumination will have an additional sidelobe level of -20dB when the rms deviation of the reflector is 1/32 wavelength (about 1mm at 10GHz) if the correlation interval is one wavelength.

The ideal paraboloid antenna would include a feed with a well-defined phase centre. In practice, the familiar feed horn does not have a phase centre that is well-defined for all angles subtended between the centre and edge of the reflector. This shortcoming introduces a phase error that is usually greatest at the edge of the dish. The positions of the feed must be a compromise.

A uniformly illuminated circular aperture yields a radiation pattern with a first sidelobe that is about 18dB below the main beam. To improve the first few sidelobes, to reduce spillover and to improve the front-to-back ratio of the dish, the designer employs a primary feed that provides a tapered distribution across the face of the dish. Also it is very desirable from the stand-point of efficiency to illuminate the parabola uni-

formly, it is very undesirable and inefficient to allow the feed to radiate in such a way that energy spills over the edge of the reflector. Losses due to taper plus spillover are typically 1.0 - 1.2dB.

Energy reflected from the parabola, and then incident upon the feed supports, generally is reflected in arbitrary directions and is essentially lost. The magnitude of this effect depends on the ratio of the blocked aperture to the total aperture. The blockage loss can be approximated by:

$$\text{Gain reduction (dB)} = 10 \log_{10}(1 - B)$$

where B is the ratio of aperture blockage to the total aperture area.

An offset paraboloidal reflector can avoid all aperture blockage, but the complexity of the feed make it useful only under special conditions. Other gain losses are likely to exceed the saving associated with avoiding aperture blockage.

Due to the feed pattern, and the geometries of both the feed and the parabolic reflector, some energy is radiated with incorrect polarisation. The cross-polarised energy is lost as far as a gain measurement is concerned (0.1 - 0.2dB) and therefore reduces the total antenna efficiency.

Antenna materials are less than perfect



Fig 1: Chaparral feeder for TV Sat (10.7 - 12.75GHz).

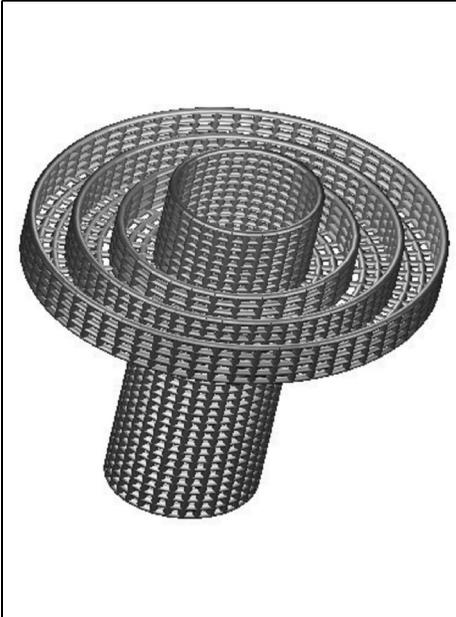


Fig 2: 3D view of the feeder described by NEC-Win Pro simulation software.

and a small part of the RF signal is converted to heat in waveguide walls and on the reflector surface. This loss is normally small and essentially independent from the feed design.

To conclude: for many years, a horn feed at the focal point has been considered the standard for paraboloidal antennas, but a help can be obtained today from modern simulation and manufacturing technologies. An estimate of the loss budget for modern 10 - 12GHz antennas (2 - 3m diameter) is given in Table 1.

For an important analysis of the losses in microwave parabolas see also [1].

Non paraboloidal dish surface and precise shaped sub-reflector are needed to obtain over 70% efficiency, but it is not a difficult task to realise a Chaparral style feed to obtain 60% efficiency 10GHz dish.

2.

Simulation of the Feeder

Starting from the need of an efficient paraboloid dish (not lower than 60%) for 10GHz experimentation and thanks to a modern simulation tool like NEC-Win Pro 1.6, the decision was taken to accurately analyse a circular feed optimised for a direct focused dish with F/D of 0.38. All the simulation patterns are relative to a circular waveguide horn with internal diameter of 20mm driven from a vertical and nominally $\lambda/4$ probe.

The dimensions of the traps that are movable on the waveguide are similar to the values of a commercial version of Chaparral style feed used in the past for the 10.7 - 12.75GHz TV Sat band (Fig 1 & 2).

Starting from a 108cm paraboloid dish we selected a 7dB edge attenuation for

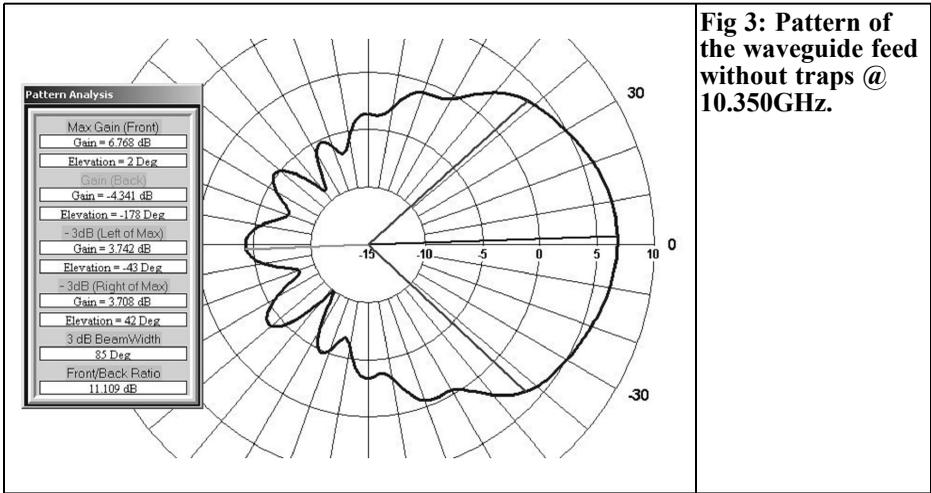


Fig 3: Pattern of the waveguide feed without traps @ 10.350GHz.

the Radiation Angles because of the about 3dB of added space attenuation (for -10dB amplitude taper).

The circular waveguide projection of the feeder can be optimised to have about the same radiation angle in both perpendicular directions: azimuth and elevation. The system is relatively broadband as shown from the diagram at 10.350GHz (Fig 3 & 4). The SWR is low and the imaginary part of the impedance practically flat versus frequency (Fig 5).

3.

Gain Field Tests

To obtain an accurate far field (or Fraunhofer) pattern of a dish antenna [2] – [6] a first requirement is that the measurements must be made at a distance at which the field at the antenna under test approximates to a uniform plane wave.

This occurs when the distance "d" between the source and the receiving aerial

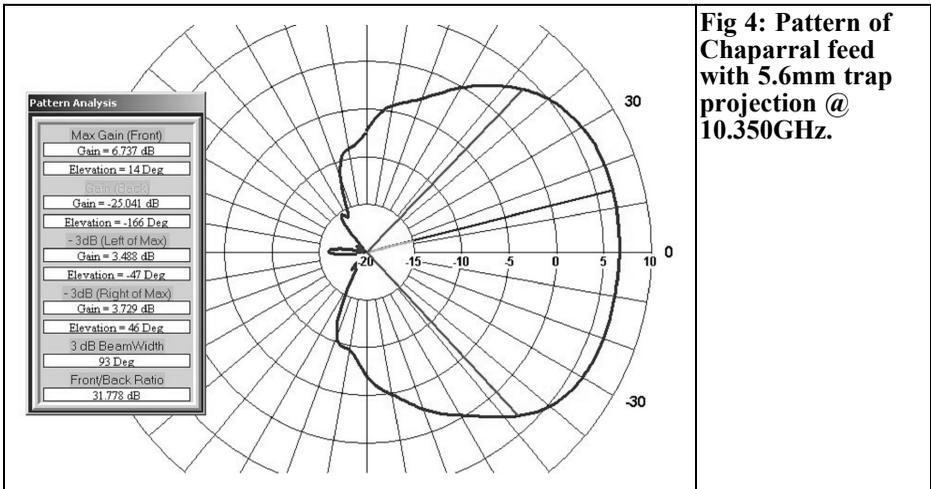


Fig 4: Pattern of Chaparral feed with 5.6mm trap projection @ 10.350GHz.

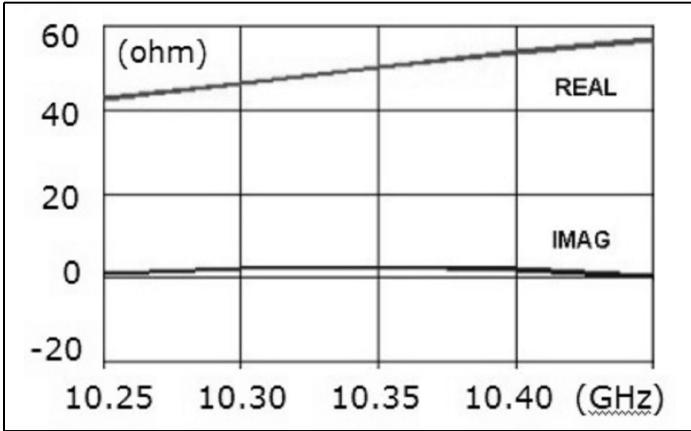


Fig 5: Chaparral feed input impedance.

satisfy the relation:

$$d \geq (2 D^2) / \lambda$$

Here “D” is the diameter of the dish and λ the operating wavelength (in metres).

In this case it is accepted that a maximum phase difference between the path at the centre and at the edge of the antenna will be $\lambda/16$ or, in other words, 22.5° electrical degrees.

Applying these considerations to our case the minimum distance becomes:

$$d \geq 80.7\text{m (for } D=1080\text{mm and } f=10.368\text{GHz)}$$

This distance is very large and not compatible with the area available for our experiments, my house garden. So we decided to make the tests with a more reasonable distance ($24\text{m} \approx \lambda/5$) taking into account the error introduced by this decision: -3dB as can be seen in the Fig 6 [6].

In the Fig 6 there are vertical lines that define the threshold between the near field or Fresnel zone on the left and the far field zone on the right. The area between the two vertical lines is the transition zone between the near field and the far one.

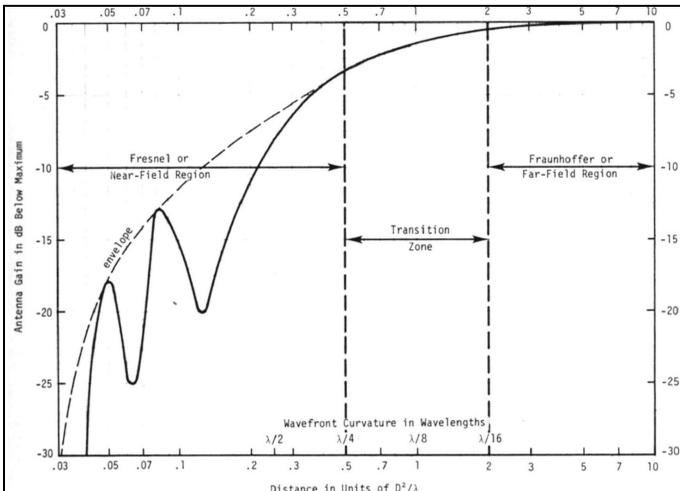


Fig 6: Antenna gain loss vs source to dish normalised distance.

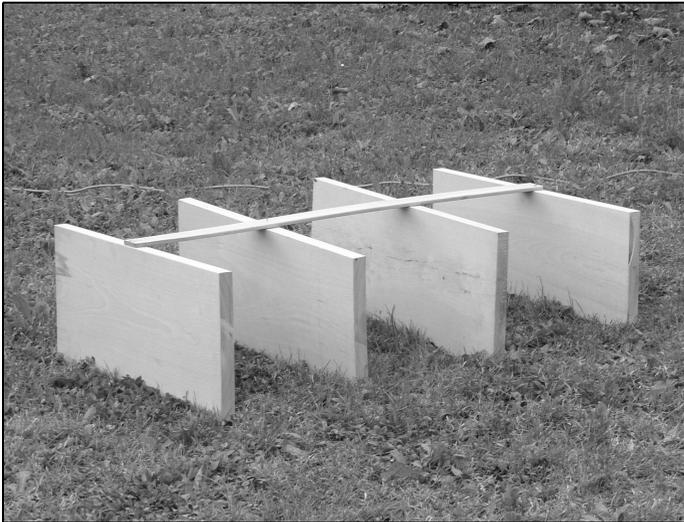


Fig 7: Beech panel microwave absorbers.

To reduce the possible effect of ground reflected waves, the transmitting antenna (horn with 17dB gain) has $h = 4\text{m}$. To have the maximum flexibility in the source positioning and to evaluate possible ground reflections, a photographic tripod with continuous height adjustment has been used.

The dish under test has been put at $h = 1\text{m}$ to help the pointing operations.

The free-space attenuation, in decibels, between transmitting and receiving antenna can be computed with the following:

$$\begin{aligned} \text{Loss} &= 10\log(4\pi d/\lambda)^2 - G_T - G_R \\ &= 32.45 + 20\log(f) + 20\log(d) \\ &\quad - G_T - G_R \end{aligned}$$

Here, “ f ” is frequency in GHz, “ d ” is the distance in m, “ G_T ” and “ G_R ” in dB.

Considering that Narda Reference Horn gain $G_T = 17\text{dB}$ (measured) and calculated Dish $G_R = 38\text{dB}$ the total free-space attenuation Loss became = 25.4dB.

Pointing the Horn has been performed by finding the maximum received signal. Pointing the dish has caused two problems which made the operation difficult and not so accurate: the parabola dimension and the lack (in the testing phase) of

a fine adjustment system for both azimuth and elevation angles.

The 10.4GHz power level available at the output of the 1W amplifier (the input is from a Marconi 6058B Generator), to feed the horn source antenna is completely satisfactory for all the measurements normally difficult because of the necessary filters, cables, isolators and attenuators.

We changed the source height (+/- 18cm) to find the possible maximum and minimum of the received signal in order to detect the presence of reflections by the ground. Good results have been obtained with and without the beech absorbers put in the wave reflection area (along the radio path and at about 5m from the parabola site). Remember that according to J.F.Aubin [2] the measurement error is of +/-1.5dB max if the test area offers at least 15dB of Direct to Reflected Path Ratio. The field test results are very good considering that the parabola positioning is very difficult because its radiation angle is lower than 2° , at -3dB power gain. To verify this result we confirmed the power gain evaluation through sun noise measurements.

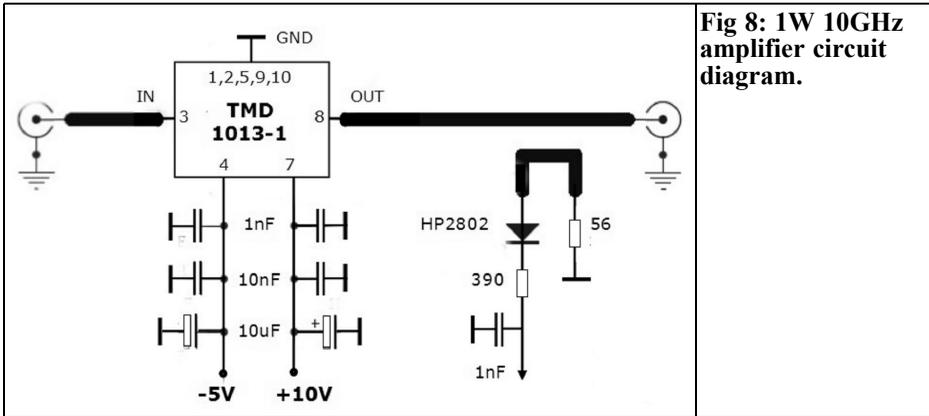


Fig 8: 1W 10GHz amplifier circuit diagram.

4. Wave Reflection from the Ground and Microwave Absorbers

We made many checks on the reflected wave question. The complete results are in a processing phase.

In the field tests of dish gain microwave absorbers are very useful. A good microwave absorber material must have several qualities. The reflectivity should be low to avoid reflection on the surface of

the absorber and the penetration depth should be small in order to have a reasonably thin absorber. Matzler, in 1970 (Reference not available) measured the permittivity of various materials at 10.5GHz. His results lead to the conclusion that beech wood has optimal properties to construct a microwave absorber.

The reflection of the electromagnetic waves on a plane boundary depends not only on the material, but also on the incidence angle. For normal incidence the polarisations do not differ. For all other angles, the horizontally polarised wave is reflected more strongly. If possible, the Brewster angle of 57.4° is used (complex reflection index: $n = 2.4 +$

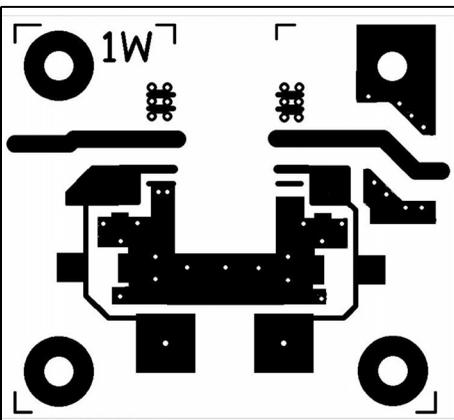


Fig 9: PCB layout for 1W 10GHz amplifier.

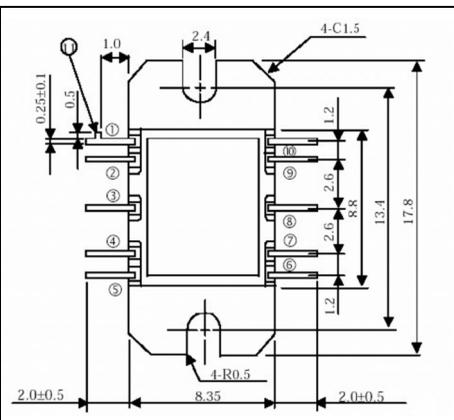
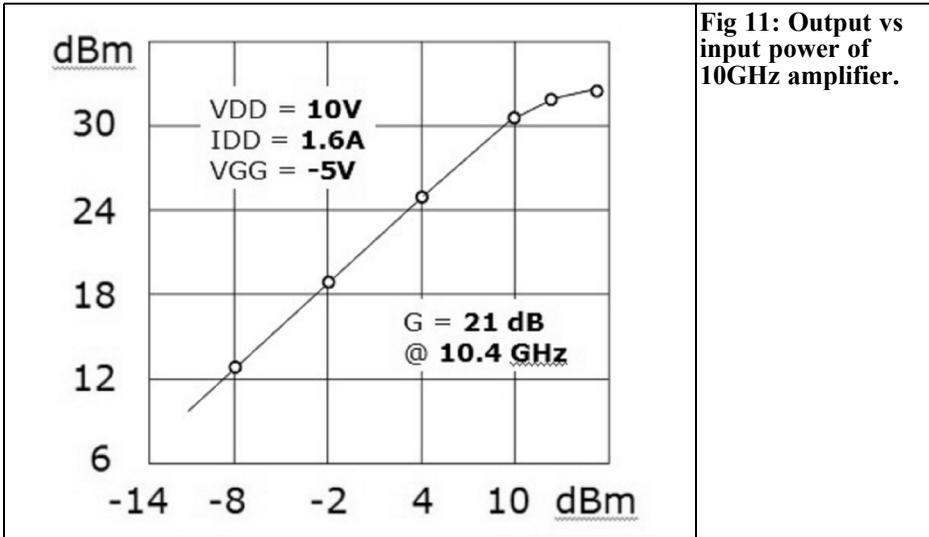


Fig 10: TMD1013-1 pin connections.



j0.38), where the vertically polarised reflection becomes zero.

Practically from 0° to about 45° the reflectivity is lower than 10% both for horizontal and vertical polarisation. We made a lot of beech panels of about 70 x 30 x 3 cm (Fig 7). One single panel in the direct path attenuates the 10GHz signal by about 10dB. The panels were very useful to limit the reflected wave in outdoor gain tests.

5.

A Powerful Amplifier

Following many unsatisfactory measurement with the 10mW output from the Marconi 6058B oscillator, many losses related to cables, isolators, etc, and the availability from a telecom surplus supplier of a pair of 1 - 2W Toshiba modules, we started to manufacture the amplifier. In Fig 8 shows the circuit diagram (the output SWR circuit is exter-

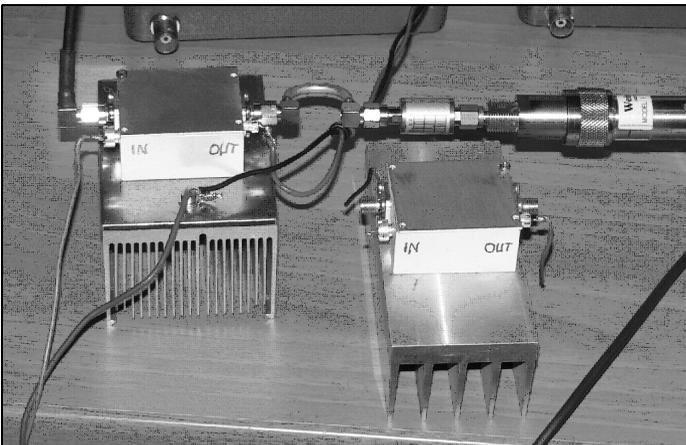


Fig 12: Measuring the amplifier.



Fig 13: Very high quality dish to be tested.

nal) and Fig 9 shows the simple PCB layout made using 0.8mm FR4. We include Fig 10 showing the pin connection of the TMD1013-1. According to the datasheet it gives a typical 33dBm output. Also see the measured values in Fig 11.

6.

Conclusion

Measurements of F/B and Side Lobes of Paraboloid Antennas are not easy, but testing of the new High Performance Dishes is quasi impossible for Hams. An important aid is a power amplifier. With 20dB higher levels in the Reference Horn, the use of bandpass filters and ground reflection absorbers we can in the near future start to test new very high performance paraboloid dishes (Fig 13) that today are only simulated at both 10GHz and 2.4GHz band.

7.

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André Jamet, F9HX

No, I do not want to use lead free solder, especially for microwave projects!

1.

RoHS

The European RoHS directive is to limit the use of six hazardous substances. RoHS means “Restriction of the use of a Hazardous Substances in electrical and electronic equipment”. Most of the states have adopted it or similar versions.

Lead, cadmium and mercury relate directly to the electronics industry and in particularly tin soldering that should no longer contain lead.

2.

Lead

The harmfulness of lead is known and ingestion of low doses leads to neurological disorders and even birth defects. A high dose of lead causes poisoning of the body (lead poisoning). While it is true that the alloys of lead can cause risks, it is limited (low temperature use) for assembly of electronic components.

Note that the lead used in the electronics industry represents about 0.5% of the total production. Here is the distribution of the main uses of lead:

- Batteries 79%
 - Ammunition 5%
 - Paints + chemistry 4.7%
-

3.

Tin/lead solder

Before this directive, the electronics industry used alloys containing mainly tin and lead. The eutectic alloy, 37% Sn / 63% Pb, has a remarkable property because it goes directly from the solid state to the liquid state without going through the pasty state (Fig 1). This greatly facilitates soldering items that are not fully locked together and gives it a shiny appearance.

While tin melts at 232°C and lead at 327°C, as most alloys, it melts at a temperature lower than those of its constituents. Its melting temperature of 183°C is moderate. It is possible to use a thermostatically controlled iron at 320°C.

Other metals, such as 1 or 2% of copper and silver, can be added to improve the mechanical strength or electrical conductivity. They do not substantially change the above essential eutectic property. They give the possibility of soldering on surfaces covered with these metals with-

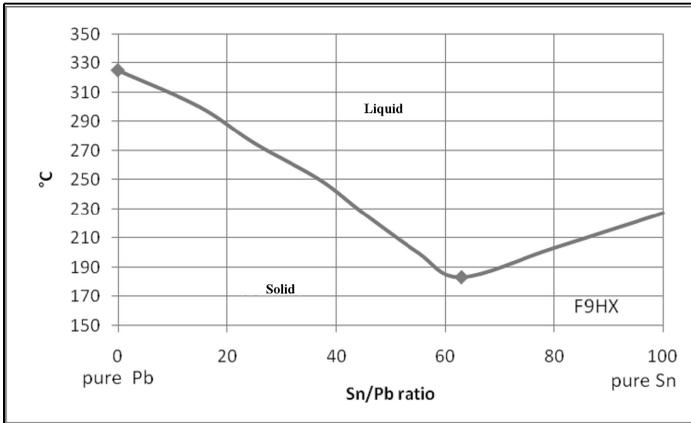


Fig 1: The phase diagram of tin-lead solder.

out dissolving them and creating brittle inter-metallic areas. For the same purpose gold is also added for joints on golden surfaces.

On the other hand, plumbers use lead rich alloys because the pasty state phase allows a degree of plasticity to “work” the solder.

4.

Lead free solders

The objective is to find these soldering characteristics with little or no lead. Unfortunately all of the alloys without lead offer a melting point higher than classic tin/lead alloys. One of the most used alloys is the SAC 305 alloy:

- 95.5% Tin, Silver 3.5%, 0.3% copper
- melting temperature 217°C
- recommended iron temperature 340°C to 420°C

This means that the soldering iron temperature must be higher than usual and its life will be reduced. The component and the printed circuit must withstand these higher temperatures.

The performance of a solder without lead

is radically different from that of the classic tin/lead alloy. The general contact and sprawl on the circuit offers more stiff contact angles on the periphery of the joint.

5.

Joints using lead or lead free

Professional guidelines are very restrictive. Contamination in a solder without lead significantly reduce the reliability.

However [3] describes how a product with lead solder can be repaired manually with lead free solder and vice versa. The method is to remove the former alloy in a harsh manner with de-soldering braid and re-soldering with the new alloy. However this is discouraged for BGA components.

6.

Other difficulties

Because of the temperature increases, moisture accumulated in components can generate a greater number of popcorn type defects (component explosion) or de-lamination of PCB substrates. Soldering of components with thin pins can no



longer be done continuously.

The whiskers: under the effect of mechanical stress outgrowth of tin develops in the form of very fine wire. Growth occurs some time after soldering and it can cause short circuits in connectors or component pins. Although their appearance remains largely unexplained, ways to resolve the problem have been researched. Several ways have been experienced, but the results remain limited. It is one of the reasons for the restriction on the use of tin/lead solder in severe applications.

7.

Soldering RoHS components with ordinary solder

According to the indications given in [3], already cited in paragraph 5 above, this is possible. The most recent components are virtually all RoHS. For my part, I use compliant SMD components and I have not found any particular difficulty in soldering them with my ordinary lead solder. I use the Sn + Cu alloy or better, the variant Sn + CU + AG.

As for old components that have stayed for ten or twenty years in a drawer, it may be useful to scrape them with a scalpel to refresh the RoHS surface when their appearance seems little poor!

8.

Conclusion

A few years after that legislation has decided on the main deadlines to replace lead in solder alloys, demand remains vague and no standard substitute for tin/lead has emerged.

The problem of the whiskers is not yet

solved and although some techniques can delay or eliminate the appearance of this phenomenon, it may be fatal in some applications such as military, medical and aerospace electronics. Consumer electronics, whose life and reliability are not the major criteria, can accept these criteria.

Under these conditions, as an amateur scientist, I have neither obligation nor interest to use lead free solder for my projects that are not marketed. In addition, if I compare the kilograms of lead from the battery in my car, changed every three years, with the weight of my thousands of solder joints on SMD components, there is comparison. The major distributors of components and materials for electronics, professional and the general public, should restore tin/lead solders in their catalogues.

9.

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André Jamet, F9HX

A Versatile VHF Bandpass filter

1.

The component

A component that is very useful for VHF bandpass filters. It is a two coils arrangement housed in two metal enclosures welded together. The coupling between windings is by a hole in the joining wall. The mode of coupling is mixed, inductive and capacitive.

It is the NEOSID 25252/5118 36 (Fig 1). I have used this component for years, for example in a 2m-band transceiver [1] and synthesisers [2,3]. The characteristics are the following.

- $L = 53$ to 72nH adjustable by ferrite core

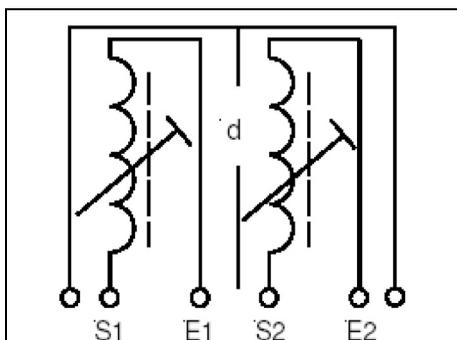


Fig 1: Neosid filter diagram.

- $Q \geq 150$ at 120MHz
- Case: $7 \times 14 \times 12 \text{ mm}$

This filter is usable at VHF from 100MHz up to 160MHz .

2.

Tuning

Each coil can be tuned by one or two capacitors to get low impedance (about 50 to 150Ω) or high impedance (a few $\text{k}\Omega$), for input and/or output.

To tune at 145MHz we need $18 + 100\text{pF}$ for LZ (Low impedance) and 15pF for HZ (High impedance) [including all external capacities] see Fig 2.

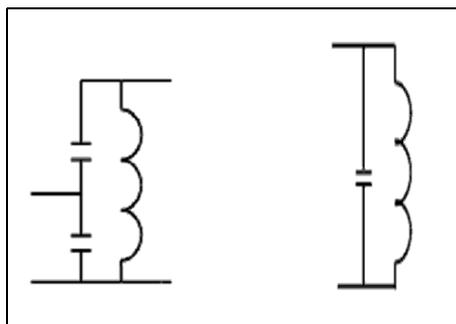


Fig 2: Arrangement to get LZ or HZ

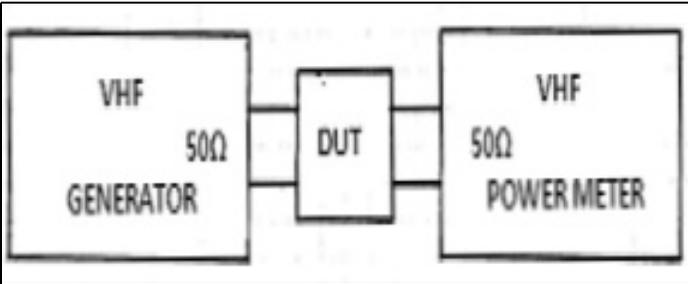


Fig 3: Test bench.

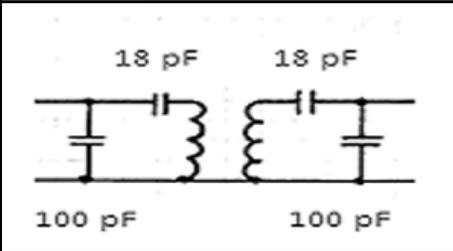


Fig 4: LZ / HZ arrangement.

3.

Test bench

Fig 3 shows the test bench used to test in situ various filter arrangements.

4.

LZ/LZ Application

Fig 4 shows an LZ/LZ arrangement. Fig 5 shows the attenuation versus frequency curve. It is normalised at 0dB for the central frequency. A light over coupling can be seen producing a flattened curve at 145MHz. This filter is low loss at the nominal frequency. Both LZ input and output is convenient to link two MMICs and an antenna to a MMIC.

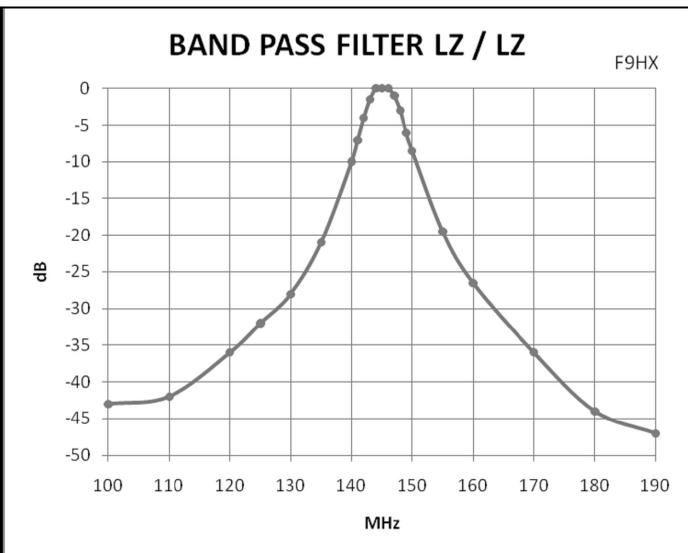


Fig 5: Attenuation versus frequency curve for LZ input and output.

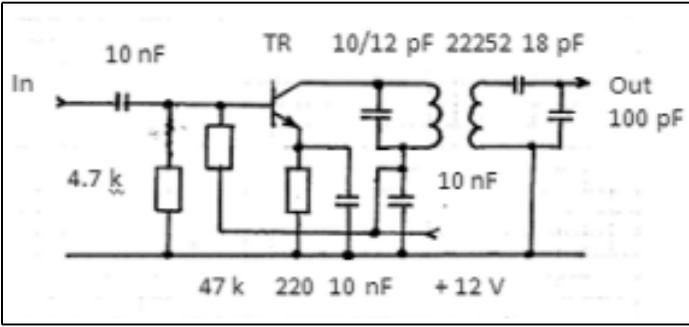


Fig 6: Circuit diagram of an HZ / LZ arrangement.

5. HZ/LZ applications

Fig 6 shows an HZ/LZ arrangement. It is valuable to link a junction transistor and an MMIC. Fig 7 and 8 show the attenuation versus frequency curves. A light over coupling can be seen producing a flattened curve at 145MHz. Fig 7 is obtain with a 2N2369 5MHz transistor, Fig 8 with a BFR 92 5GHz transistor. The collector capacitance of the first transistor is 4pF instead of 0.6pF for the later. Therefore tuning needs 10 and 12pF respectively. Moreover, a damping effect appears with the first that upsets the attenuation for low frequencies. In

both cases the power gain is about 10dB. Both curves are normalised for 0dB at 145MHz.

6. LZ/HZ applications

A typical application of LZ/HZ is to match a 50Ω antenna and a dual gate MOS-FET like BF 998.

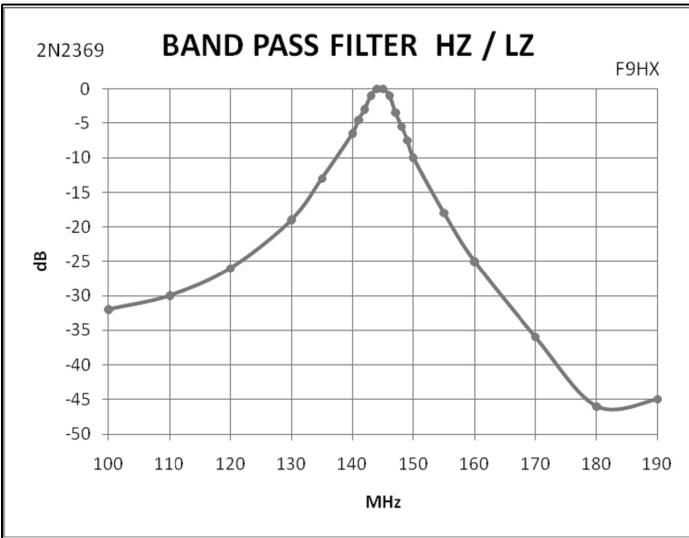


Fig 7: Bandpass filter HZ / LZ with 2N2369 transistor.

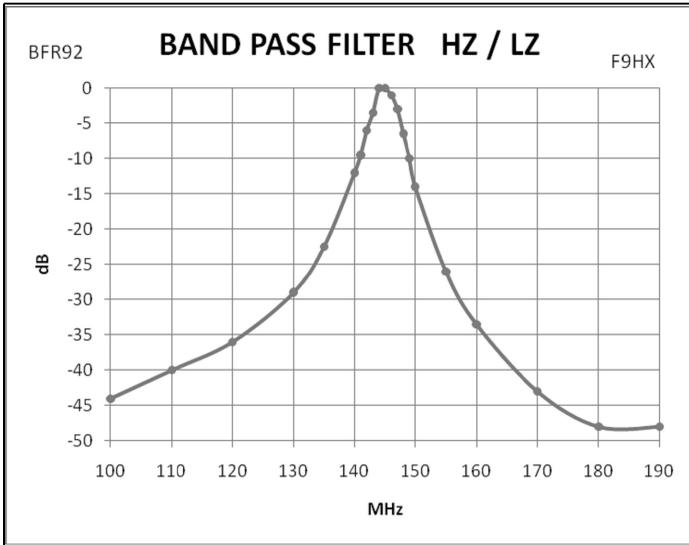


Fig 8: Bandpass filter HZ / LZ with BFR90 transistor.

7.

Conclusion

This VHF filter is useful for a lot of applications.

8.

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R. Lentz, DL3WR

Remembering 30 Years Ago: Circulators and Isolators

An article from VHF Communications Magazine issues 1/1972 and 2/1972

Circulators and isolators are used extensively in the professional microwave technology and are also becoming more and more popular at UHF and VHF. These components are expensive and are difficult for the radio amateur to obtain or manufacture, and are therefore only used where unavoidable, e.g. in parametric amplifiers.

The number of radio amateurs experimenting at UHF and SHF is increasing and they will be very interested in circulators that can be manufactured inexpensively according to a new process.

1.

Terms

Circulators are electromagnetic components having three or more connections (ports) in which RF-energy circulates in one direction from one port to another

whereas a relatively high attenuation occurs in the opposite direction. In addition to this, all ports are matched. The clear symbols for three-port and four-port circulators are given in Fig 1.

Four-Port Circulators operate according to the principle of a non-reciprocal phase-shifter, and also according to Faraday's law of induction.

They are used in the radar technology due to their high power capability.

Three-Port Circulators have become popular for microwave applications, especially simple Y-type **Junction circulators**. Such junction circulators use a symmetrical Y (or T) junction of three lines (waveguide or stripline). A ferrite body is arranged at the centre of the junction.

Lumped-Constant Circulators (with lumped inductances and capacitors) are used for frequencies below approximately 1GHz since they are smaller than junction types for such low frequencies. Both lumped-constant and junction circu-

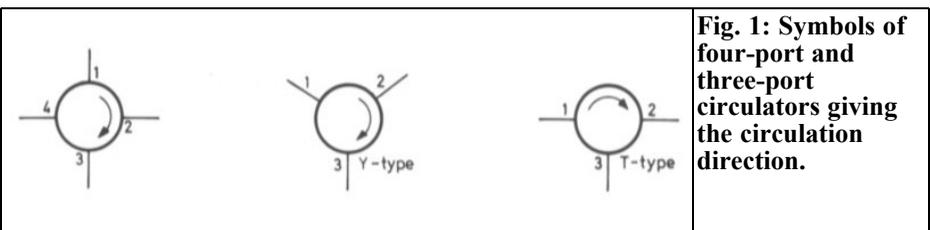


Fig. 1: Symbols of four-port and three-port circulators giving the circulation direction.

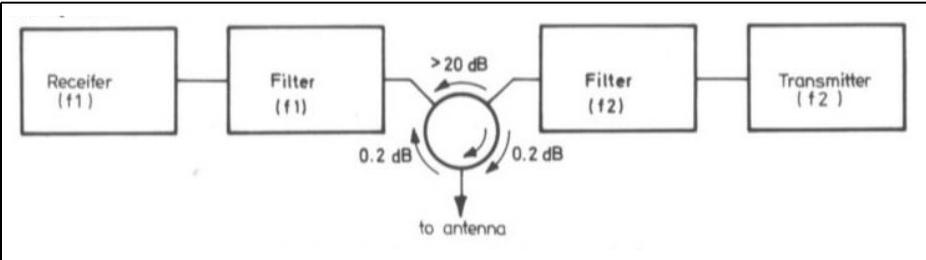


Fig. 2: Application of a three-port circulator as a duplexer.

lators are to be described in more detail in the following sections.

The (low) attenuation in the circulation direction is called **Insertion Loss** and is in the order of tenths of a dB. The attenuation in the opposite direction is called **isolation** and amounts to 20dB or more. Further important magnitudes are the **Standing Wave Ratio (VSWR)** at the ports and the **Bandwidth** of the circulator. Due to the method of operation, resonance effects occur which limit the useful frequency range. However, even the types with the lowest bandwidth have a bandwidth sufficient for amateur applications.

Isolators have only two ports: The insertion loss is very low in the forward direction whereas a high isolation (attenuation) is present in the opposite direction. **a magnetically biased ferrite**

core which absorbs the electromagnetic waves in one direction but does not influence them in the other direction, is the most important part of both isolators and circulators.

2. Application

The most popular application of the circulator is probably as duplexer where a transmitter and receiver are connected to a common antenna. In this manner, the circulator decouples the receiver from the transmitter and provides a low-loss path from the transmitter to the antenna and from the antenna to receiver (Fig 2). Normally, transmitter and receiver oper-

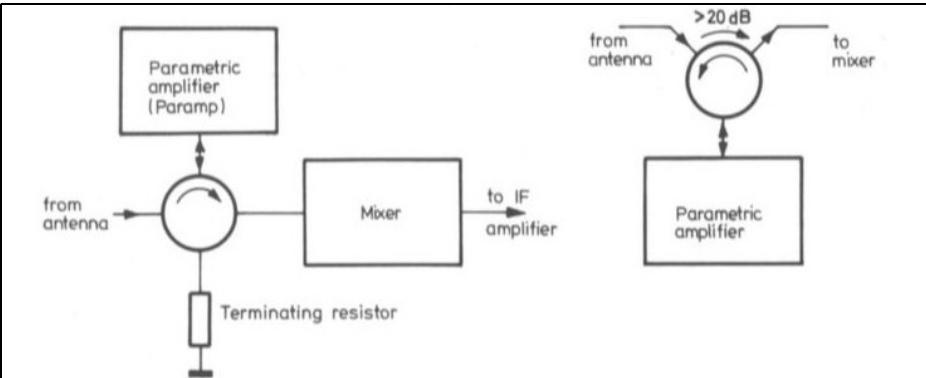


Fig 3: Parametric amplifiers are provided with a defined input and output impedance by addition of a circulator.

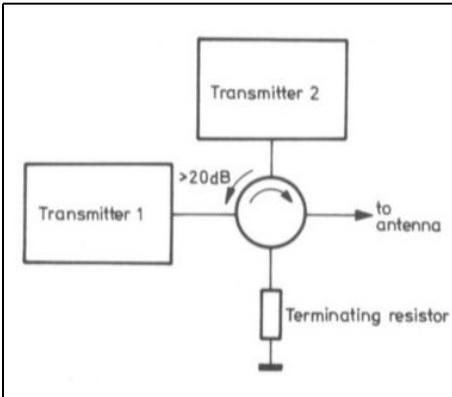


Fig 4: Two transmitters working into a common antenna via a circulator.

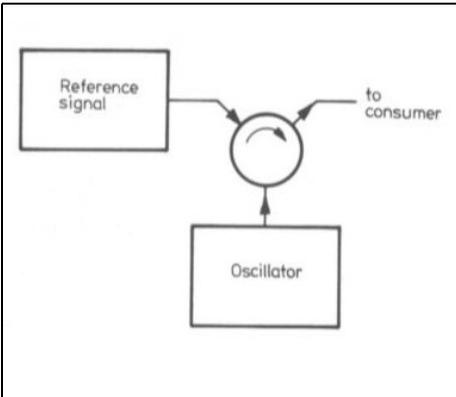


Fig. 5: Phase-locking of an oscillator.

ate at different frequencies within the bandwidth of the circulator so that additional filters can be used to increase the limited isolation in order to allow simultaneous transmission and reception.

Circulators are also used as coupling links between a signal source and the input of circuits possessing a negative impedance (e.g. tunnel diodes or parametric amplifiers) as well as for passing the amplified signals to the output (Fig 3). Since the circulator separates the reflected signal from the forward signal and feeds the former to a terminating resistor where it is absorbed, the result in an amplifier with only two ports which operates very constantly even when the source and load impedances vary. In principle this arrangement would also be possible with a three-port circulator, however, without terminating resistor.

A further application for terminated circulators is to decouple two or more transmitters connected to the same antenna in order to reduce intermodulation distortion (Fig 4). If no such isolation was present, the energy from one transmitter could pass to the other and cause third-order intermodulation products. The terminating resistor need only be dimensioned for the reflected power and can be deleted completely in some cases. The transmitters can either operate at differ-

ent frequencies within the bandwidth of the circulator or on the same frequency (for doubling the power).

Circulators can also be used for phase-locking a diode oscillator to an external signal (Fig 5). In this case, the oscillator is connected to the first port, the reference signal (at lower power) to the second; the synchronized output signal is then taken from the third port.

The application of an isolator need not be explained in any great detail: it is used where the reaction of a fluctuating load is to be isolated from an oscillator or amplifier. A typical example of this is the pump oscillator (e.g. klystron) of a parametric amplifier which must be kept from the fluctuating impedance of the varactor diode (Fig 6). However, an isolator will not be required in this position if the power level of the pump oscillator allows a basic attenuation setting of 6dB or more for the attenuator.

3.

Principle and Construction

3.1 Lumped-Constant circulators

The construction and basic relationships

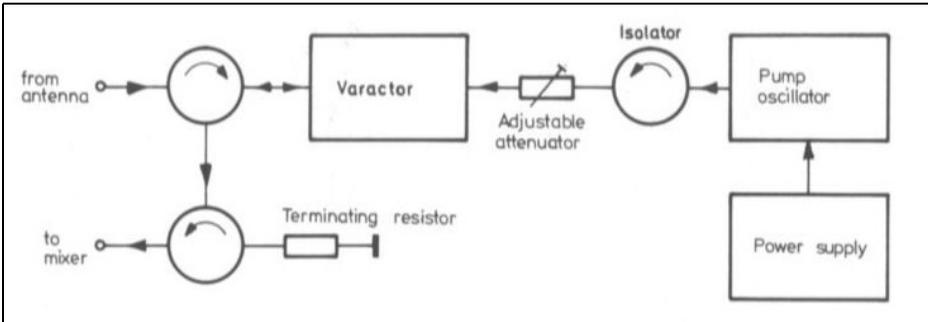


Fig 6: Block diagram of a parametric amplifier using two three-port circulators and a isolator.

can be shown in conjunction with lumped-constant circulators usually used for frequencies below approximately 1GHz [1], [2], [9]. This type represents a balanced, non-reciprocal configuration of three inductances that are arranged on a ferrite disc so that the magnetic fields of the three inductances are shifted by 120° to another (Fig 7). Additional discs for the return path of the RF-field are arranged above and below the main disc. In addition to this, a magnetic DC field exists perpendicular to the surface of the disc for providing magnetic bias up to the vicinity of the ferromagnetic resonance

ω_0 . This frequency has the following dependence on the magnetic bias H_A :

$$\omega_0 = \gamma \times H_A - 4\pi \times M_S$$

Where:

γ = gyromagnetic relationship; a constant that is usually assumed to be 2.8MHz/Oe (Oersted).

$4\pi M_S$ = magnetic saturation of the associated ferrite material in G (Gauss).

If the ferrite disc were not present, the voltage fed to port 1 would induce

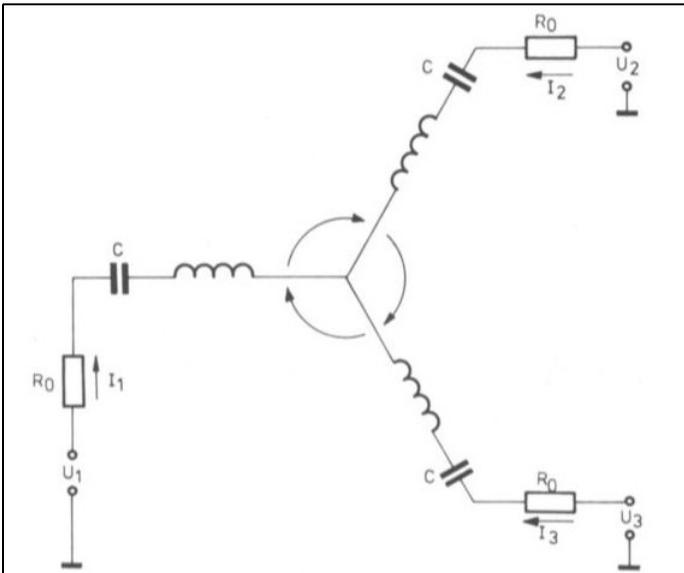


Fig 7: Equivalent circuit of a three-port circulator with lumped elements. The inductances are brought to resonance with the aid of series capacitors.

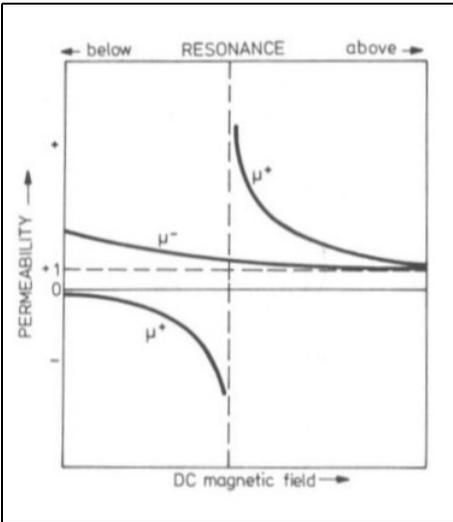


Fig 8: The magnitude of the magnetic bias determines isolation, insertion loss and bandwidth.

μ^+, μ^- : effective scalar permeabilities of the ferrite material for positive or negative circular polarized fields.

A typical run [1] of the characteristic curves as a function of the magnetic bias is shown in Fig 8

These equations still do not provide the complete basis for practical construction. For example, the difference between μ^+ and μ^- at a given frequency depends on the degree of magnetic bias in the ferrite material. The value required for L_0 is dependent on this. The operating point of the magnetic bias can by no means be selected at will since the bandwidth and also the insertion loss will increase when the magnetic bias approaches resonance. On the other hand, the insertion loss also increases when the operating point is far from resonance.

equally great voltages into inductances 2 and 3 due to the balanced arrangement. Due to the magnetic disc, the magnetic coupling between the inductances becomes non-reciprocal; this means that the coupling from inductance 1 to inductance 2 is not the same as from inductance 2 to inductance 1. This non-reciprocity causes the attenuation or isolation in the reverse direction.

The equations therefore only guarantee a good isolation at one certain frequency. At other frequencies, the isolation is reduced and the insertion loss increased. The bandwidth over which the isolation is greater than 20dB - the minimum for practical purposes - is given by the following equation:

An analysis of the lumped constant circulator shows that the following conditions must be fulfilled for ideal circulation at any frequency ω :

$$\frac{\Delta\omega 20dB}{\omega} = 0.2 \times \sqrt{3} \left(\frac{\mu^+ - \mu^-}{\mu^+ + \mu^-} \right)$$

The lowest operating frequency in MHz for practical purposes can be expressed as:

$$\frac{\omega \times L_0}{2} (\mu^+ - \mu^-) = \frac{R_0}{\sqrt{3}}$$

$$\frac{\omega \times L_0}{2} (\mu^+ + \mu^-) = \frac{1}{\omega C}$$

$$f_u \approx 2.8 \times 4\pi M_s$$

Where:

- R_0 = characteristic impedance
- C = capacitance in series with each inductance
- L_0 = proportional to the inductivity of one of the coils without ferrite

The magnetic saturation value $4\pi M_s$ for normal materials [3], [4] is between 100 and approximately 5000 Gauss.

It will be seen that a material with a low magnetic saturation is required for VHF applications. A value of less than 100 Gauss is required for operation below 280MHz. Unfortunately these materials tend to have very poor temperature char-

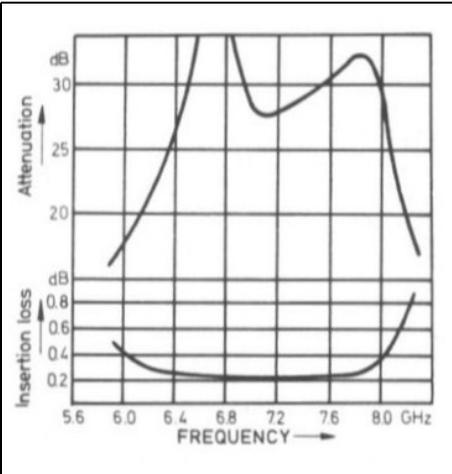


Fig 9: Extension of the bandwidth of an X-band stripline circulator using double-tuned circuits.

acteristics.

Circulators with magnetic operating points above the ferromagnetic resonance exhibit lower bandwidths and require considerably greater magnetic bias for higher frequencies.

Good operating points are to be found just below the resonance frequency. The isolation tends to drop here but the insertion loss as well. The final selection of the operating point naturally depends on the special demands of the application in question. Lumped-constant circulators have been constructed and operated at frequencies from UHF down to 35MHz. The insertion loss is 0.3dB at 400MHz, over 0.8dB at 100MHz and up to 2dB at 35MHz. Typical values are given in Fig 9 for a double-tuned microstripline circulator

3.1.1 Isoductor or Gyrator

One special type of circulator with lumped elements is a component which has been designated **Isoductor** (Melabs) or **Gyrator** (Microwave Ass.). This component does not have one important characteristic of a circulator: the matching at the connections. One can assume it

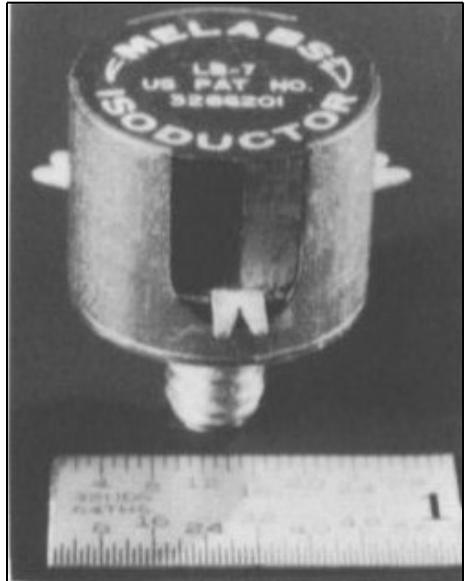


Fig 10: An isoductor.

to be the core of a circulator with lumped elements but without transformation links, connections and matching elements (Fig 10). The isoductor or gyrator is therefore smaller and less expensive than a circulator. However, it will be able to be used for the same applications as a circulator if external matching links are provided. For this reason, this component is extremely suitable for use in stripline circuits. Another advantage is that the resonant frequency of a circulator built up in this manner is determined by the matching links and can thus be shifted in a relatively wide frequency range. Of course, it is also possible for such a circulator to be matched with the aid of the matching links to other impedance values than 50Ω .

The impedance at the three connections of a gyrator or isoductor consists of the ohmic resistance r_o and a parallel inductivity l_o . For this reason, it is necessary for matching links to be provided at connection 1 and 2 whereas connection 3 is terminated with a resistor (R3) and a

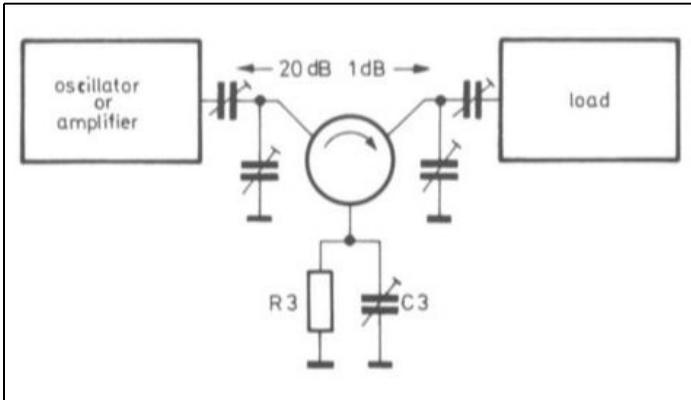


Fig 11: Matching of an isolator or gyrator.

parallel capacitance (C3) as shown in Fig 11. The values of R3 and C3 can be determined with the following formulas:

$$R3 = r_0$$

$$C3 = \frac{1}{\omega^2 \times l_0}$$

Whereby $\omega=2\pi f$; with f = required operating frequency or centre frequency of the required frequency range. The isoductor or gyrator will operate as a circulator within the frequency range in which the inductivity l_0 is compensated for by C3. At resonance, the stopband attenuation is theoretically infinity; at other frequencies, the stopband attenuation will fall. The bandwidth is determined by the loaded Q_L :

$$Q_L = \frac{r_0}{\omega \times l_0}$$

The lower the value of Q_L , the wider will be the bandwidth.

Typical calculated values for an isoductor covering the 70cm band are given in Fig 12. This allows the values for R3 and C3 and for the transformation networks to be calculated. Fig 13 gives the stop-band and passband attenuation as a function of frequency. As an example, the frequency characteristics of an isoductor type LB-3 when tuned to 435MHz have been drawn below this. It can be seen

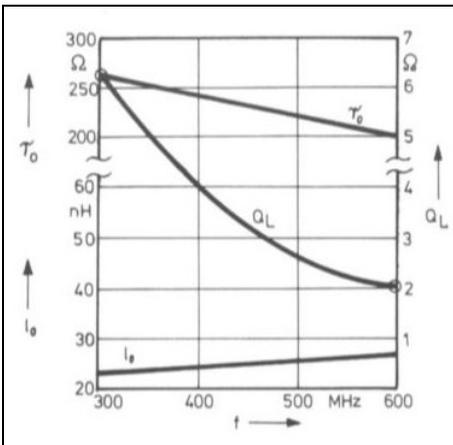


Fig 12: Typical characteristics of an isoductor LB-3.

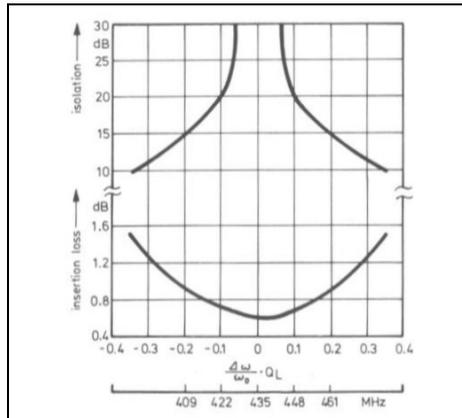


Fig 13: Standardized attenuation curves of the isoductor series LB-1, LB-3, LB-7. Below: Frequency characteristics calculated for a type LB-3 tuned to 535MHz.

**Table 1: Gives the nominal specification of the available isoductor types.**

Frequency Range	100 - 200MHz	200 - 400MHz	300 - 600MHz
Insertion loss	1dB	<1dB	<1dB
Stopband attenuation	>20dB	>20dB	>20dB
r_0	225 Ω	250 Ω	230 Ω
l_0	32nH	26.5nH	24nH
Q_L	7.5	5	2.7
Designation	LB-1	LB-1	LB-3
Price (Dec 1969)	300DM	300DM	300DM

Further specifications and applications can be taken from [6], [7] and [8]

from this that the bandwidth of this component is more than sufficient for radio amateurs.

Table 1 gives the nominal specifications of the available isoductor types.

3.2. Junction circulators

Junction circulators are closely related to circulators with lumped elements. This relationship can be compared to that between cavity resonators and resonant circuits constructed from inductances and capacitors. Fig 14 shows a junction circulator

manufactured in stripline technology. The designation of this type of circulator originates from the symmetrical Y-shaped line junction.

The ferrite disc in the centre of the junction represents a dielectric resonator. The smaller the diameter, the higher the frequency. The relatively high dielectric constants of ferrite materials, which are 12 to 15 times as great as that of vacuum, allow the construction of circulators for microwaves that are very compact. In the VHF and UHF range, on the other hand, it is possible to obtain more compact dimensions when using concentrated elements since concentrated elements are very much smaller at lower frequencies than waveguide resonators or $\lambda/4$ transformation links.

Junction circulators [1], [5], [9] are constructed in stripline, waveguide and micro-stripline configurations. At microwave frequencies, bandwidths of up to an octave are covered; at VHF and UHF, the bandwidth is limited to approximately 40% of the operating frequency. However, if only a relatively low bandwidth is required, more favourable specifications, lower cost and more compact dimensions are possible. Of special interest in this context is a special circulator manufactured by Melabs for parametric amplifiers on the 70cm, 24cm and 12cm bands. Table 2 gives the specifications and price (1965) of these three circulators.

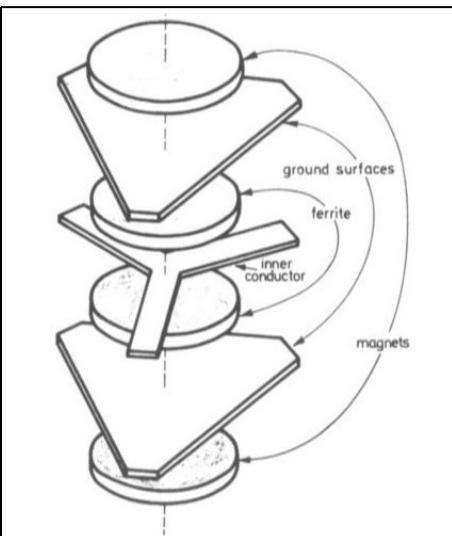


Fig 14: Construction of a stripline junction circulator.



Table 2: Specification and price (1965) for Junction Circulators.

Frequency	432MHz	1296MHz	2300MHz
Bandwidth (min)	2MHz	4MHz	4MHz
Insertion loss (max)	0.5dB	0.3dB	0.3dB
Stopband attenuation (min)	20dB	20dB	20dB
VSWR (max)	1.5	1.2	1.2
Impedance	50Ω	50Ω	50Ω
Connectors	TNC	N	N
Designation	J-6969	J-6970	J-8017
Price in US \$	97.50	95.50	88.00

Waveguide junction-circulators exhibit lower losses and a better standing wave ratio for the same stopband attenuation and bandwidth than coaxial types (e.g. an SWR of 1.05 and an insertion loss of 0.1dB over a bandwidth of 15%). However, waveguide circulators are larger than the appropriate stripline types at VHF anti UHF.

3.3. Switching circulators

Since the direction of circulation is dependent on the polarity of the magnetic bias, it is possible for the circulator to be used as a simple switch (without moving parts) by controlling this magnetic field. The field is then generated with the aid of an electromagnet or by a permanent magnet whose polarity is reversed by a short, high-current surge through a magnetizing coil. The symbol of a switching circulator is given in Fig 15.

The radio frequency characteristics of

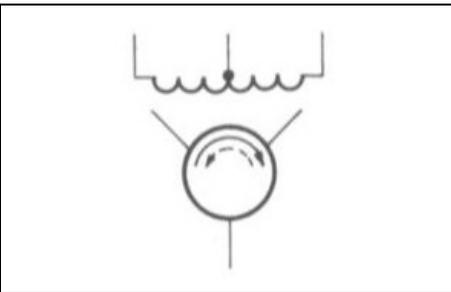


Fig 15: Symbol of a switching circulator.

switching circulators correspond to those of equivalent conventional circulators. Further details are given, for instance, in [1], [5] and [9].

4.

Isolators

Isolators consist of a RF line (stripline, coaxial line or waveguide) into which a magnetically biased ferrite body is placed. The RF-energy is absorbed in one direction by the ferrite material and is not affected in the other direction. For optimum attenuation values the operating frequency must coincide with the gyro-magnetic resonant frequency of the ferrite material. In this context, the operation of an isolator coincides to that of a circulator. In principle, a three-port circulator will work as an isolator as long as one port is terminated with the correct impedance.

5.

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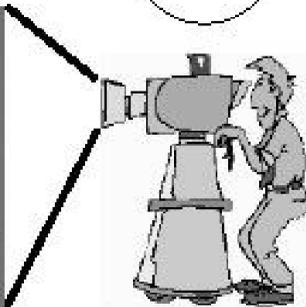
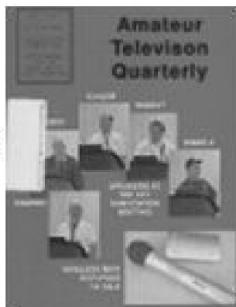
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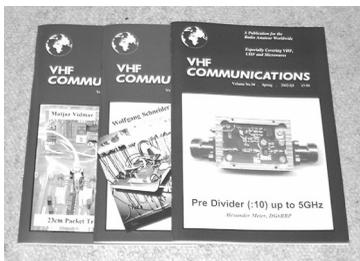
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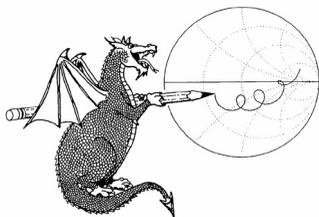
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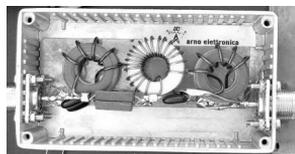


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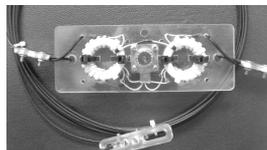


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