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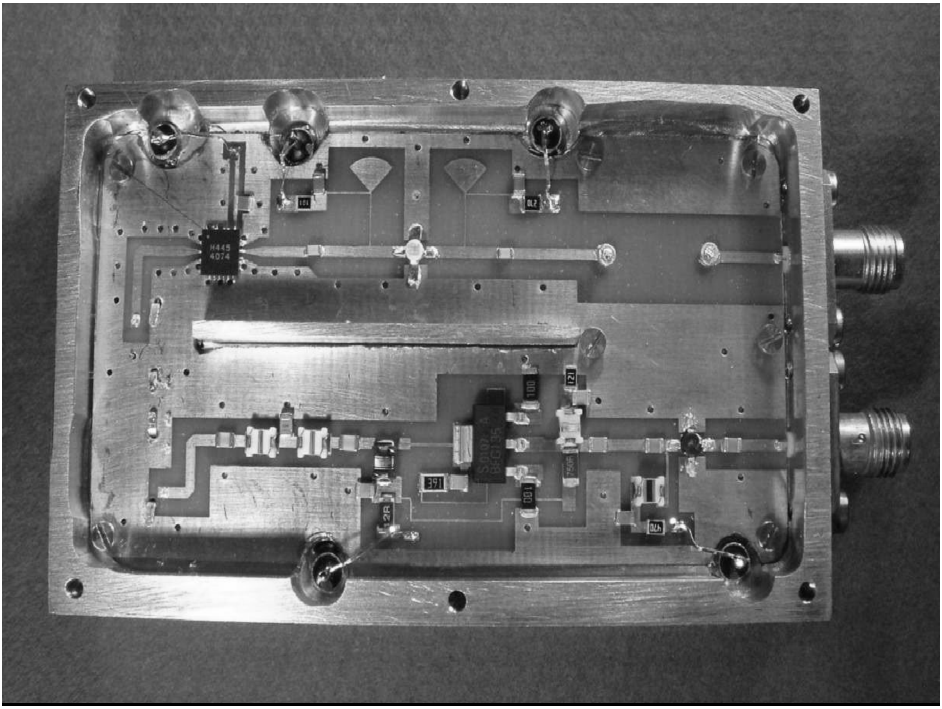


A modern technique for microwave oscillator generation: Part 1

Sigurd Werner, DL9MFV

Contents

There is no Franco's Finest article in this issue but Franco is working on an ultra low noise pre-amplifier with a 0.9 - 1 db NF from 30MHz to 2.5GHz. It will require no tuning and have a high dynamic range but low price. This article will be in issue 3/2005.

Once again there are some new authors in this issue with new ideas, there is additional information on the web site for the Transverter adapter and Lightning Scatter articles.

73s - Andy

K M Publications, 63 Ringwood Road Luton, Beds, LU2 7BG, UK

Telephone / Fax +44 (0)1582 581051, email : andy@vhfcomm.co.uk

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

Sigurd Werner, DL9MFV

A modern technique for microwave oscillator generation Part 1

This article describes a simple and up-to-date technique for oscillator generation. It is used to generate LO signals in a range between 9.5 and 12GHz.

The advantage is obvious: you'll just need two multiplier stages and only one printed circuit board layout for various models. A high degree of flexibility is attained, the number of components is reduced and easy reproduction of the project is guaranteed. Two variants of the oscillator module are described (A1 and A2). Higher frequency LO signals (e.g. 24 and 38GHz) can be obtained by additional frequency multiplication. The corresponding modules will be described in a later article (Part 2).

1. Introduction

For about fifteen years now, the various oscillator techniques used by M. Kuhne, DB6NT [1] have received a warm welcome from those interested in microwaves. These modules use the following technique: the oscillator signal (typically between 100 and 125MHz) is initially trebled and then continuously doubled until a fixed multiplication factor of 96 is attained. A filter system is set up between the individual multiplier or amplifier stages. The output frequencies obtained vary between 9.5 and 12GHz. Other designs such as those from P. Vogel, DL1RQ [2] and M. Vidmar, S53MV [3], work in a similar way.

The purpose of this two part article is to develop an alternative technique relating to frequency multiplication in the microwave range covering the entire spectrum that interests radio amateurs.

The advantage, as against other solutions, is obvious from two arguments alone :

- You are working with only one board layout.
- There are just two multiplier stages in use in the circuit.

The design presented here has the following advantages, as against other developments:

- Only two semi-conductors are used for frequency multiplication (the power supply is designed on correspondingly simple lines).
- The modules can be assembled in a very compact manner and require

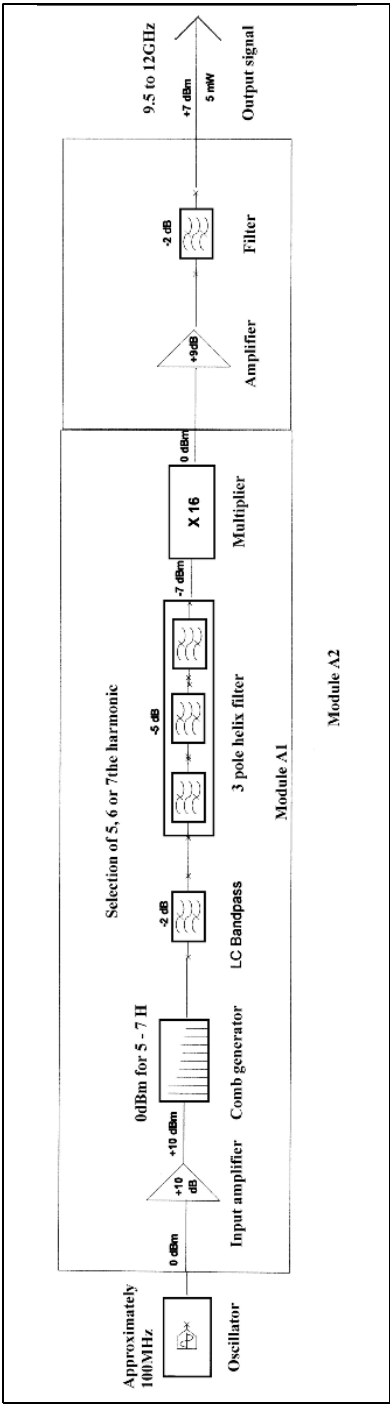


Fig 1: Block diagram of oscillator showing modules A1 and A2

only a small amount of space.

- The generation of frequencies between 9.5 and 12GHz is based on a single board layout. Only the LC band pass filter and the selection of the subsequent helical filter must be altered.
- The multiplication factor of 96 can be replaced by alternative values of 80 or 112, a fact which leads to a high degree of flexibility in frequency selection.
- The extremely low input level of the times 16 multiplier stage means that you can work with very low oscillator output levels (< 0dBm), or that the input amplifier can be dispensed with, if necessary (see below).

Table 1 lists the frequently used frequencies, with the processing mode for each of them. This is based on a fixed multiplication factor of 96. Table 2 illustrates the principle applied here with two examples: the generation of an X band frequency (10368MHz) and that of the frequency (9504 MHz) for the E band. The signals can be represented with various oscillator frequencies and variable multiplication numbers in module A (in variants A1 and A2). Module B shows another signal multiplication (Part 2).

2. Circuit description for module A

The oscillator signal (between 80 and 150MHz) is amplified up to a level of 10 to 12dBm. This signal (f_o) is fed to a comb generator and the appropriate harmonic frequency (5 x fo, 6 x fo ,or 7 x fo) is filtered out.

Table 1:

Oscillator (MHz)	IF (MHz)	Output signal x96 (MHz)	Signal after multiplication (MHz)
99.000		9504	x4 = 38016, x2 = 76031
99.187	+144	9522	x4 = 38088, x2 = 76176
108.00		10368	
106.50	-144	10224	
126.00		12096	x2 = 24192
125.250	-144	12024	x2 = 24048
122.625		11772	x4 = 47088
122.250	-144	11736	x4 = 46944

The narrow band signal (fi) is fed to a times 16 multiplier IC, which generates the desired final frequency (Fig 1). The power at the output is approximately 0dBm (±3dBm, dependent on the frequency). This simple design is implemented in the variant referred to as module A1 and is conceived of as a driver for the subsequent module B for further multiplication. (The filtration of the signal can be integrated there.) If no further frequency multiplication is planned, the signal is then amplified and filtered. This happens in the variant referred to as module A2.

Table 2 shows that the frequency 10368MHz comes from oscillator frequencies of 129.000, 108.000 and 92.571MHz, the signal applied from the comb generator (fi) always remaining at 648MHz. For 9504MHz, you require 594MHz (fi).

For a more detailed description of module A, the output frequency 9504MHz and a multiplication factor of 96 are used. The circuit diagram can be seen in Fig 2 & 3.

2.1. Oscillator amplifier

The oscillator frequency required is 99.000MHz. The signal (0 dBm) is raised to 10 to 12dBm by means of a monolithic IC (e.g. ERA3). For oscillator levels exceeding 0dBm, this amplifier stage can be dispensed with (see times 16 multiplier).

Owing to the uncomplicated layout and the increased flexibility obtained for the module, the amplifier has been integrated.

2.2. Comb generator

The semiconductor selected was the BFG135a NPN transistor (Philips). The amplifier circuit proposed in the data

Table 2:

Oscillator signal f_0 (MHz)	Comb generator signal $f_{i\,(}$MHz)	x	Output of Module A (MHz)	Multiplication factor	Output of Module B x4 (MHz)
129.600	5 x f_0 = 648	16	10368	80	
108.000	6 x f_0 = 648	16	10368	96	
92.571	7 x f0 = 648	16	10368	112	
118.800	5 x f0 = 594	16	9504	80	38016 x2 = 76032
99.000	6 x f0 = 594	16	9504	96	38016 x2 = 76032
118.800	7 x f0 = 594	16	9504	112	38016 x2 = 76032

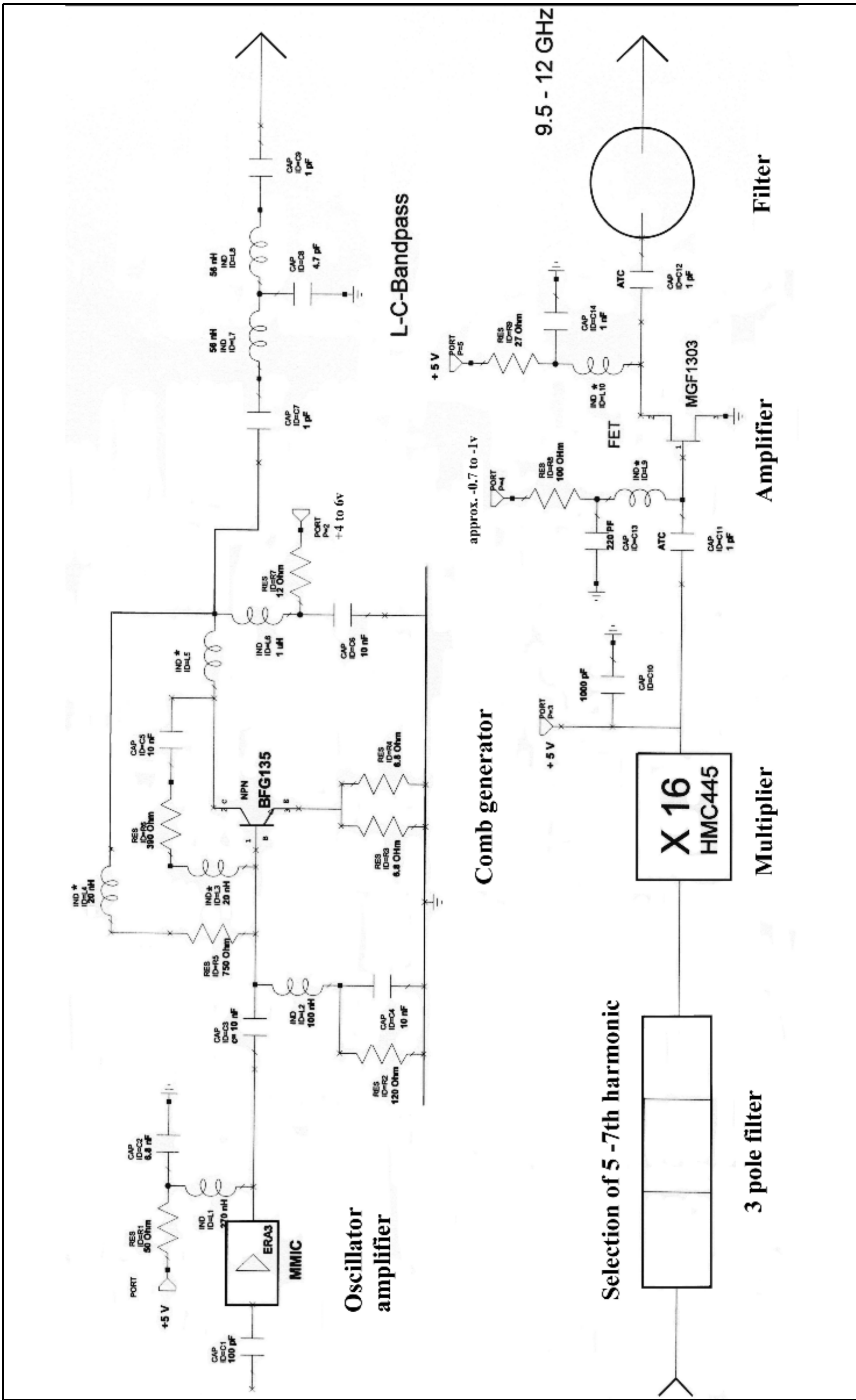


Fig 2: Circuit diagram of oscillator.

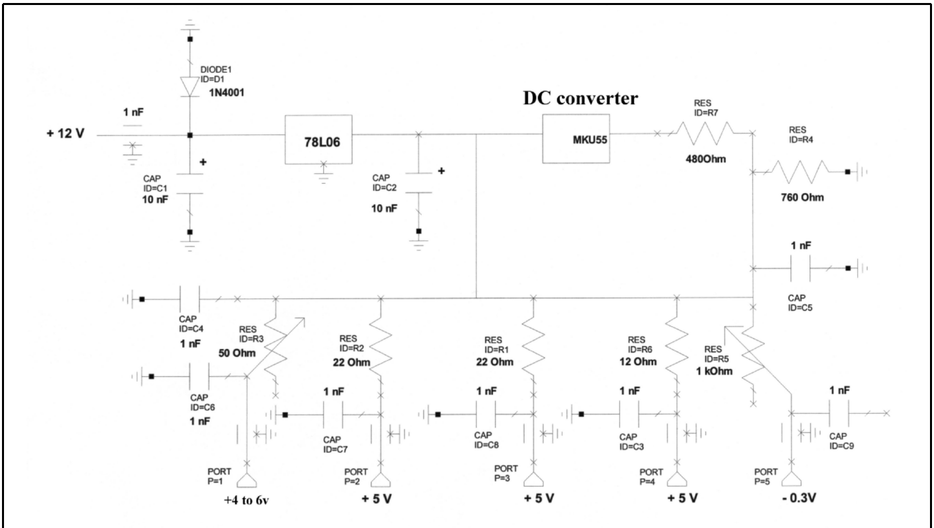


Fig 3: Circuit diagram of power supply for oscillator.

sheet [4] has been modified so that the transistor operates as a reliable comb generator. Fig 10 illustrates that with a drive of approximately 10dBm the signal strengths of the harmonics generated (fi) can be up to 8 x fo at 0 dBm. For other purposes, the generator can be adjusted for frequencies of up to 1.5 / 2GHz with good levels. All that is needed to operate it is a positive voltage, which can also be used to control the individual output levels.

2.3. Isolation and filtration of the f_i signal

The initial selection of the 6th harmonic frequency of the oscillator signal (6 x fo =594MHz) is handled by a simple LC bandpass filter (BW is approximately 180MHz at -10 dB). It was inserted so that the subsequent helical filter would not be "overtaxed" by signals far away from its pass band. The triple helical filter (5HT-5905F-590 from TOKO, centre frequency 590MHz) is fixed at 594MHz. It attenuates signals ± 50MHz around 34 or 25dB (-1dB BW is 12MHz; the insertion loss amounts to -4.8dB). The selectivity of a double filter is naturally also adequate, but since attenuation actually plays a limited role (see below) the three pole version was used.

2.4. Times 16 multiplier

The multiplier IC HMC445 (from Hittite) was used here in an LP4 housing (4 x 4mm). The module is intended for output frequencies between 9.9 and 11GHz. However, it operates perfectly well, with very slight limitations, between 9.5 and 12GHz. Remarkably, the times 16 multiplier can still process drive levels of -22dBm without any reduction in the output signal (levels of up to -2 dBm still give usable signals !). Depending on the frequency, the signal strength at the output lies between +4 and -4dBm. A level of 9504MHz was measured at +2dBm.

2.5. Amplifier for 9.5GHz

The simplest solution for increasing a signal by approximately 8 to 10dB is to use the tried and tested MGF1303

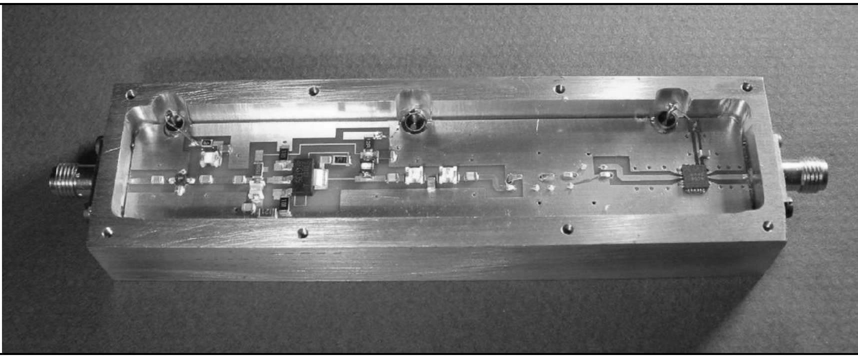


Fig 4: Photograph showing the top view of module A1.

GaAsFET. As an alternative a low noise amplifier could also be used, such as a CHA2063aMAF (in a ceramic housing, from UMS), if a higher power ratio is desired.

2.6. Filtration of 9.5GHz signal

Three options were investigated. Here are some brief notes on their advantages and disadvantages.

- Inter-digital filter [5]:

  Advantages: Easy to make, high selectivity (-3dB BW = 55MHz; resonator interval = 16mm; filter diameter = 10mm), acceptable insertion loss of -3dB.

  Disadvantages: relatively large, difficult to integrate into single board layout and into one housing.

- Stripline filter (on Rogers RT5580, 0.13mm)

  Advantages: relatively compact (particularly true for model in accordance with hairpin principle), can be seamlessly integrated into a layout, good selectivity attainable.

  Disadvantages: Can not be implemented quickly except with help of a simulation program, must be professionally manufactured, high insertion loss (approximately 8dB);

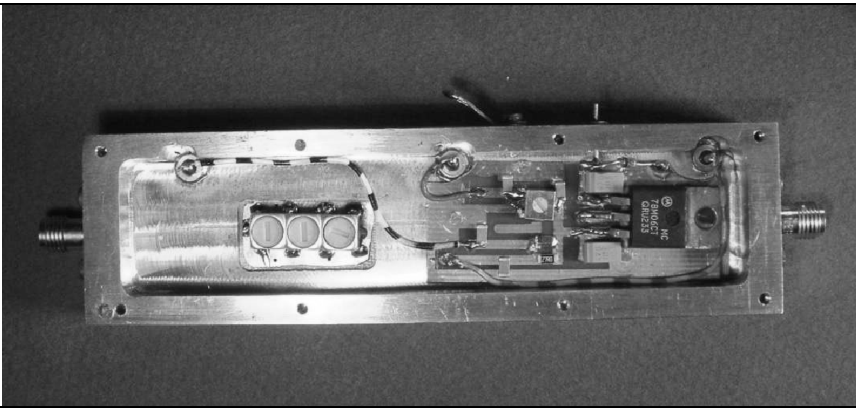


Fig 5: Photograph showing the bottom view of module A1.

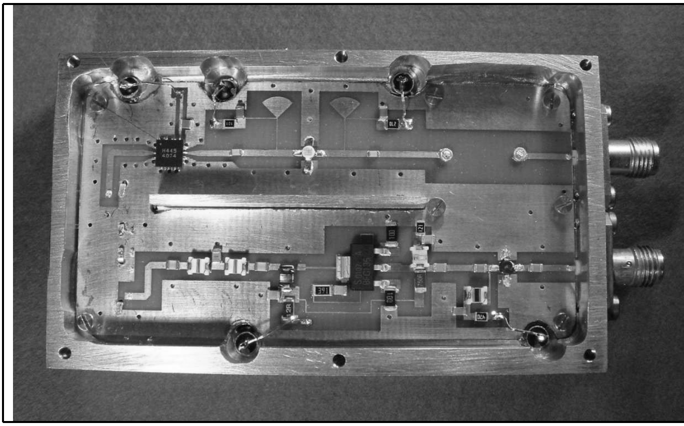


Fig 6: Photograph showing the top view of module A2.

i.e. the filter must be combined with an amplifier, which levels the losses out again. A prototype was assembled within the framework of module B (Part 2).

- Pipecap filter (e.g. [6 -8])

  Advantages: can easily be produced, adequate selectivity and acceptable insertion loss (approximately 2.5dB).

  Disadvantages: relatively asymmetrical transmission curve (BW ± 300MHz = -25 or -21dB).

Since the pipecap filter is easy to produce, and is also easy to integrate into the housing base, this solution was preferred for module A2.

3. Mechanical assembly and board layout

The modules were milled from brass. Module A1 (Figs 3 & 4) has dimensions of 109mm x 26mm x 15mm . The cavity for the board (102.8mm x 19.8mm) is

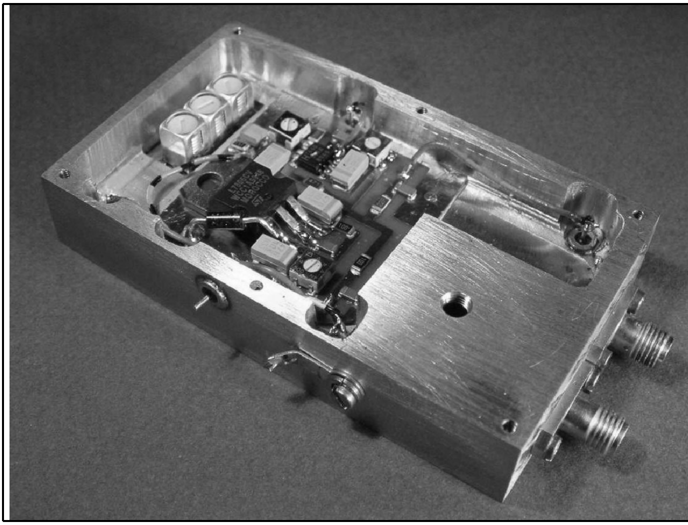


Fig 7: Photograph showing the bottom view of module A2.

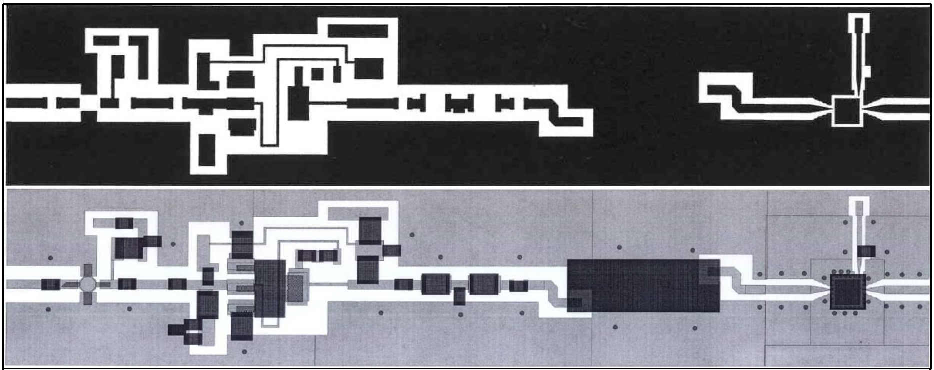


Fig 8: PCB layout and component layout for module A1.

recessed 6mm deep. The housing underside (7mm deep) is likewise milled out for the power feed connection. The feedthrough capacitors in the housing base connect the two cavities together. The helical filter is adjusted from the underside of the housing. The cavity's base has been removed at the appropriate point for the mounting of the three pole filter.

The dimensions for module A2 (see Fig. 6 & 7) are 79mm x 43mm x 15mm. The cavity for the mounting of the circuit

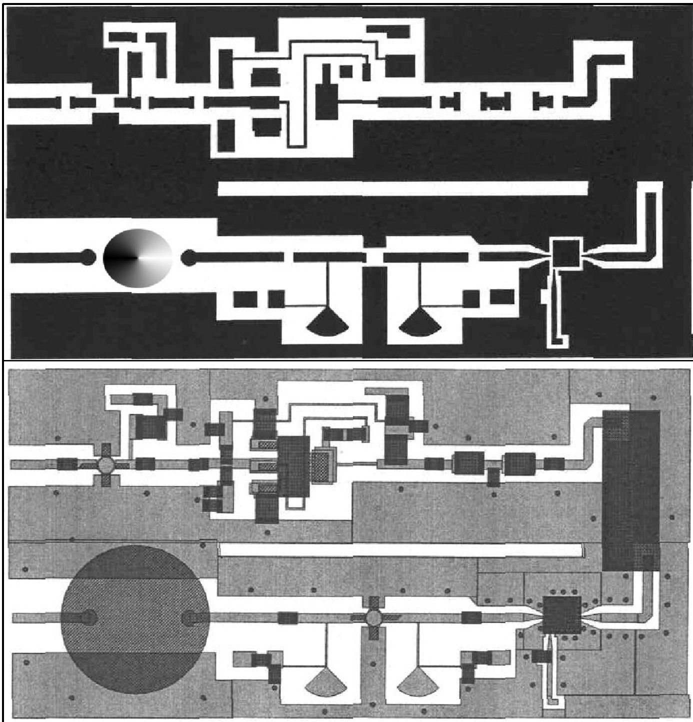
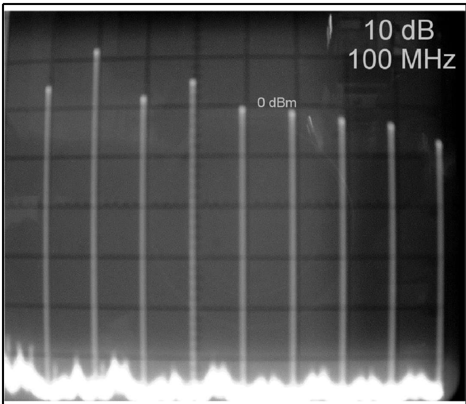


Fig 9: PCB amd component layout for module A2.



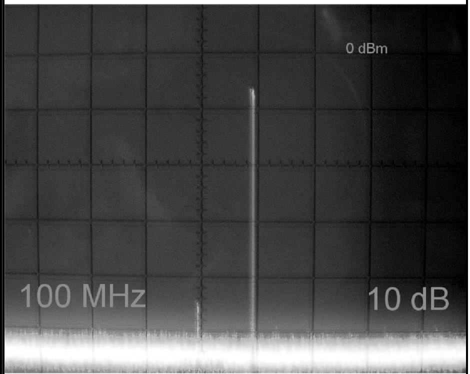


Fig 10: Measured outputs from the comb filter before and after filtering.

board (72.5mm x 37.3mm) is 5mm deep. In addition, a 40mm. long partition has been inserted into the base of the housing to improve the RF isolation.

The pipecap filter is milled directly into the housing base (diameter 16mm, 8mm deep; adjustment screw = M4 Extra Fine) and the circuit board is laid on top of it (coupling pin distance 10.5mm, length 2.2mm). An adhesive mat is used to connect the circuit board and the resonator. The helical filter and the power feed connection are mounted as for module A1.

The board layouts (Figs 8 & 9) were transferred onto FR4 substrates (0.5mm; Cu coating =35µm; εr= 4.6) and the copper surfaces were silver plated (The earth surface of the circuit board forms the base of the pipecap filter in module A2!).

The boards are through hole plated at numerous points (0.5mm).and are fixed on the base of the cavity by means of a mat impregnated with silver (Ablestick FSCM21109).

The layout is organised around fitting SMD components only (see Figs 8 & 9). The fastening process used can be either soldering or gluing (silver conducting adhesive). Only glueing is recommended for mounting the sextupler (HMC445), with the help of a powerful magnifying glass.

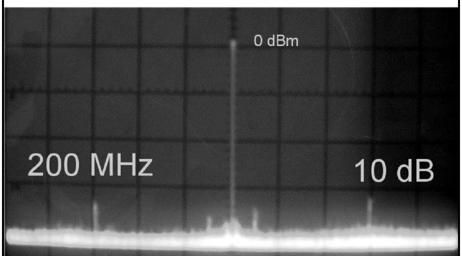


Fig 11: Measured output from module A1 at 9504MHz.

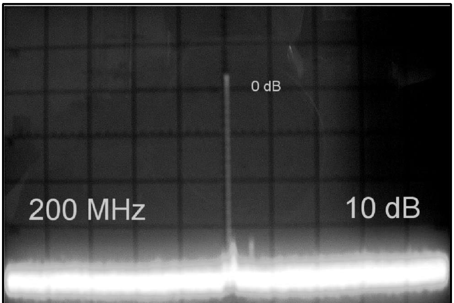


Fig 12: Measured output from module A2 at 9504MHz.

4. Results

The signal at various stages is shown in Figs 10 & 11. The comb generator generates a large number of harmonic signals (Fig 10). The 6th harmonic of the oscillator frequency of 99MHz (i.e. 594MHz) is filtered out. Fig 10 illustrates the selection obtained after the helical filter.

With the times sixteen multiplication of the signal to 9504MHz, it becomes clear (Fig 11) that unwanted signals appear at distances of ± 99MHz and ±594 MHz from the main signal. However, these are approximately 35 - 40dBc lower than the required frequency. This means that, for many applications, even the simple module A1 can b used as a signal source. It is therefore also used as a driver for additional multiplier modules (see Part 2).

Finally, Fig 12 shows the amplified signal spectrum, filtered by a pipecap filter It corresponds to the output from module A2.

5. Notes and acknowledgments

The frequency spectra were measured using an 8565A HP analyser. The SMS generator from Rohde & Schwarz was used as the signal source.

Circuit simulations were carried out by means of the “Microwave Office” program (Applied Wave Research, Inc.). Gerber data on board manufacture can be obtained from me.

My thanks are due to Mr. A. Goray (sma-electronics) for the helical filters made available and to Jochen Ehrlich (DK3CK) for some stimulating criticism.

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André Jamet, F9HX

Solutions for stable and precise microwave frequency generation

1. Nature of the difficulties

In spite of many publications about this subject, difficulties seem not to be solved for a number of people working on SHF. Frequency stability seems to have been obtained but knowing the actual frequency is still a problem that increases the difficulty of making a QSO, in addition to the classic problems such as dish orientation and very deep QSB.

Knowing the actual SHF frequency is difficult because of the precision level required. To find a call accurately requires it to be within the frequency bandwidth of the receiver, so about one kilohertz for an SSB signal. At 144MHz, a 7×10^{-6} precision is enough; at 10GHz, we need 10^{-7}, so 72 times better! For higher frequencies and/or more demanding modes than SSB, we need a greater precision.

2. SHF local oscillators

A conventional solution uses a chain with a crystal oscillator and multiplier stages. For example, for 10GHz, we need a 106.5MHz oscillator followed by multipliers to get 10,224MHz output when the transverter is followed by a 144MHz transceiver (Fig 1). The 106.5MHz oscillator is the key for frequency stability and knowledge of the actual frequency. Most standard 144MHz transceivers are stable and precise enough to use.

3. The 106.5MHz oscillator

Current practice is to use a 3rd or 5th overtone crystal. To get good tempera-

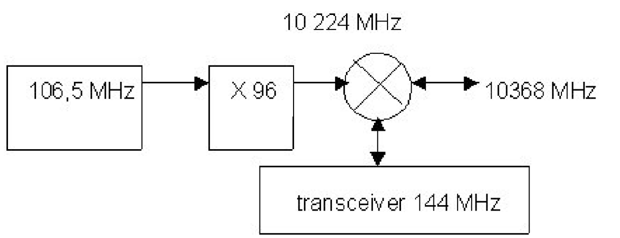

Fig 1: 10GHz transverter.

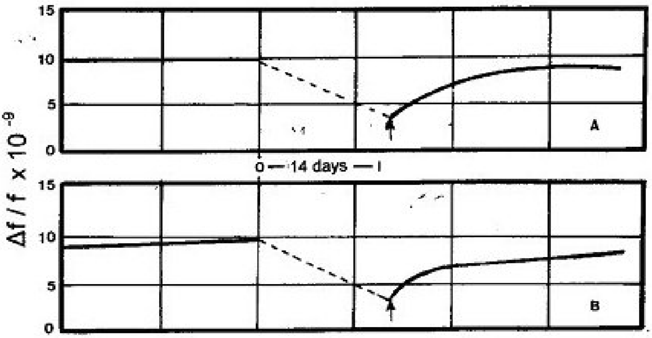

Fig 2: Frequency retrace effect of a 10MHz OCXO. A: Oven cycled off and on with oscillator running. B: Oscillator cycled off and on with oven running.

ture stability, we can manage with a TCXO, which means a crystal compensated with negative temperature coefficient capacitors. To obtain better results, we can have the crystal temperature, or better, the whole oscillator, at a constant temperature with a thermostatic controlled oven, i.e. an OCXO.

4.

Difficulties encountered

The OCXO temperature can be maintained within less a tenth of degree Celsius even with ambient temperature variations, from negative values up to high positive ones. The short-term stability can be excellent. However, users observe a recovery effect. It is not the crystal ageing, because the periodical tuning is not very demanding; it is the retrace effect. When a crystal oscillator is switched off for a period, it does not recover the same frequency when it is switch on again. The length of the break in use is related to the length of time for the crystal to come back to the same frequency, i.e. a new ageing slope. This problem makes it difficult to know the actual frequency being used and means that we must use a known frequency reference, such as a beacon, to recalibrate.

5.

Real cause and cure

Common quality VHF crystals are manufactured with an AT cut. Retrace effects are very prominent. Professionally it is usual to use SC cut crystals to reduce the retrace effect. Unfortunately, these crystals are not made for individual requests like the common AT cut crystals and their cost is much higher. Moreover, it is preferable to use lower frequency crystals such as 5 or 10MHz to get the better results. We can get professional OCXO at rallies or on Ebay. Behaviour of a good OCXO is shown Fig 2 for two cases

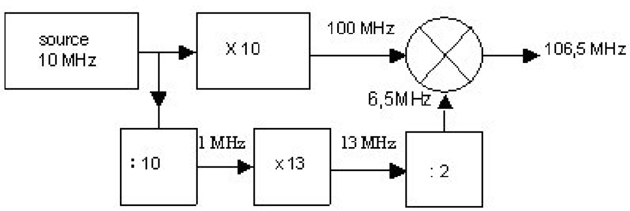

Fig 3: Synthesised local oscillator.

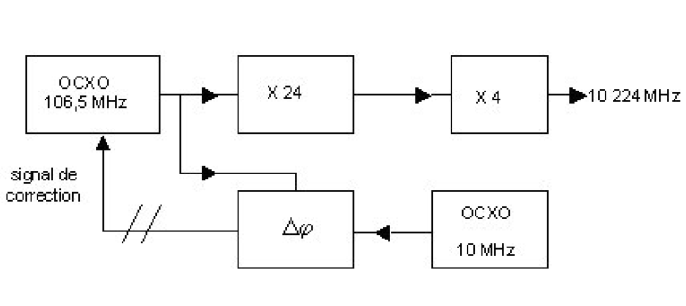

Fig 4: Local oscillator with PLL.

of rest: oven switched off only, and oscillator only. For a VHF OCXO values are 100 times worse!

6.

10MHz OCXO use

To start at 10MHz, we have to modify the local oscillator chain, see [1,2] for descriptions of synthesiser or PLL solutions (Fig 3 and 4). Some people doubt the signal quality obtained by the synthesis method owing to the very high multiplication of the basic frequency. For a good 10MHz OCXO and for a careful design and construction, measurements have showed nothing wrong. 76GHz tests by F5CAU, 10GHz EME by DC9UP and the world record on 241GHz in the States gives this validity. We can see and hear signals produced by several kinds of sources in [3].

7.

A very simple solution

If we do not want to invest in a chain using a 10MHz OCXO, there is a very simple solution already used by the author to obtain good results with a 106.5MHz OCXO, we only need to keep it switched on continuously! If we do this then only the crystal ageing will act and only infrequent calibration will be needed. I can also advise that the solution using a 10MHz OCXO to work on the highest bands and/or with modes more exacting than SSB is acceptable.

The OCXO can be fitted with a small auxiliary battery, float charged in the shack and discharged on site (Fig.5). With a 7Ah sealed lead-acid battery, the safety margin will be enough for two days operating: when its operating temperature is reached, power consumption is reduced. Caution: do not use a garage type charger, un-smoothed or un-regu-

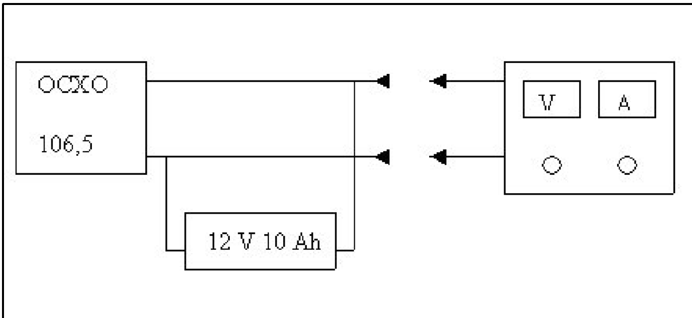

Fig 5: Permanently supplied OCXO.

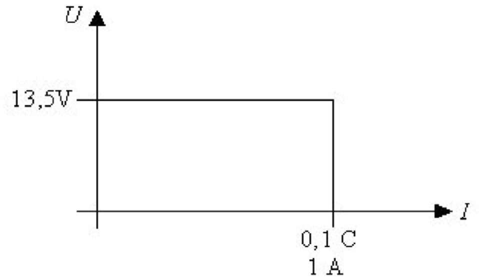

Fig 6: Parameters for charging power supply.

lated supply for float charging; this will reduce battery life. I advise a PSU: 13.5V constant voltage, 1A limiting current (Fig 6). The trickle current is about 0.5 to 1mA per Ah of capacity. Total current is the sum of trickle current plus OCXO current, approximately 200mA.

8.

Conclusion

I do not expect to hear on the 10GHz band "I'll transmit at 100 within 10kHz". How much easier the QSO would be. Moreover, the door is open for modes of operation more stringent for frequency stability and accuracy, to receive unreadable SSB or classical CW.

9.

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For more about: F9HX agit@wanadoo.fr

Ulf Kylenfall, SM6GXV

Transverter Adapter for use with ICOM IC735/751

In this article, a transverter adapter for ICOM 735/751 (and possibly some others will be described). The design was originally made in 1991 when the above mentioned transceivers were still being sold. It was intended to be a small commercial project for PECAB Electronik, a company started in Gothenburg around 1988, but once we had our own demand satisfied, it was put on the shelf. Then during a few years of personal turmoil, I thought I had lost all the design files.

Some time ago, I found the old documentation and just recently, the original design files. Neither myself nor Hans SM6EUP, the owner of PECAB Electronik had any intention to start selling the adapters so we decided to place the design in the "Public Domain".

I think that five or ten of these units were built and used by VHF'ers on the Swedish west coast. Fig 1 shows a completed station using the adapter. Nowadays, complete boxes with everything from HF to UHF are on the market, but maybe some amateurs can find anything worth out of this article. Although I have made my best to recollect the design, I can make no guarantees that the design files are error free.

Another thing to watch is that the components originally used may not be available anymore. Although the original design and gerber files are available on the VHF Communications web site, they may not be usable!

1.0 General

The ICOM series of short wave transceivers 735/751 were well designed HF equipment at the time when they were being sold. I remember when I bought my 751 around 1985. As a VHF amateur, I was only allowed to use it as an IF for my transverter for 70cm. I quickly realised that it had not been developed for the amateur with VHF in his mind. The transceiver had very low 28MHz IF level. This signal is not very well filtered either. There are no floating relay output for PA T/R switching. Furthermore, no ALC is provided, and disabling requires +8V to be applied at one of the ICOM accessory terminals.

It was therefore decided to try to build an adapter that would have the following features:

- Bandpass filtered 28MHz

Fig 1: Picture of the transverter adapter in use with an ICOM IC751.

- A delayed RF turn on.
- ALC detection
- All relay functionality that could be required for driving and controlling external equipment.

2.0 Specifications

The technical specifications of the final version is listed below:

- ICOM 28MHz Drive level: -10…0dBm
- Output 28MHz Drive Level: +13dBm (20mW) max. continuously variable
- Harmonics: < -30dBC @ Pout +10dBm
- Non harmonic signal suppression: <-70dBC @ Pout +10dBm
- 28MHz Bandpass filter: 3dB points: 27…31MHz, 20dB points: 25-32 MHz
- 0—(tbd) ALC Output
- Two relay outputs with NO/NC function, galvanically isolated from chassis ground (100V/2A/60VA max)
- RF “on”, TX may be delayed approximately 0.5 seconds in order to allow high power amplifier and antenna relays to settle before drive is present at the transverter drive output connector. This feature is jumper selectable.
- +12V T/R delayed voltage available 0.2A max (see above)
- +8V for ICOM Final HF amplifier turn off
- High Z TX shift sensing; > +5V @ 4k7 Ohms (For ICOM’s +5…12V applied to TXH)
- Extra TX active low sensing; < 5V

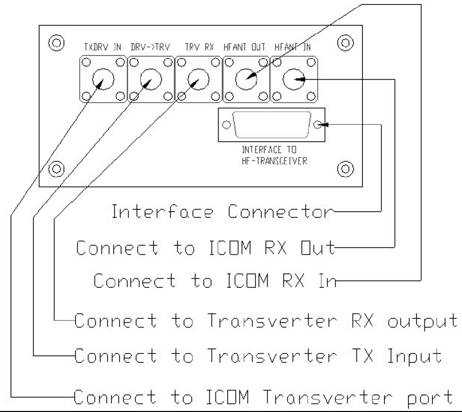

Fig 2: Layout of connectors on the rear of the transverter adapter.

@ 4k7 Ohms. This makes the transverter adapter useable with other types of HF transceivers. (TXL grounded)

- Received bandpass filtered signal loss at 28.2 MHz: -3dB
- Isolation between HF-ANT-IN ←→ TRV RX in transvert mode (Adapter turned ON) = Isolation between HF antenna connector(s) and the Transverter RX Connector: 55dB
- Isolation between HF-ANT-IN ←→ HF-ANT-OUT in transvert mode (adapters turned ON) = Isolation between HF antenna connectors: 65 dB

Table 1: Pin connections for 15 pin DSUB connector.

Pin	Signal	ICOM IC735	ICOM IC751
1	+12V when in TX. (Delay by internal jumper)		
2	TXH > +5V will put the adapter in TX mode		ACC pin 6
3	TXL < 5V will put the adapter in TX mode	SEND (Phono jack)	
4	Ext Relay 1 Common		
5	Ext Relay 1 NC		
6	Ext Relay 1 NO		
7	Ext Relay 2 Common		
8	Ext Relay 2 NC		
9	+12V switched by ICOM station (Adapter supply voltage)	ACC(2) pin 7	ACC pin 2
10	GND	ACC(2) pin 5	ACC pin 7
11	GND	See above	See above
12	GND	See above	See above
13	ALC Output from Adapter	ACC(2) pin 5	ACC pin 7
14	+8V Output; Will Put ICOM in Transvert mode (HF amplifier turned OFF)	ACC(2) pin 6	ACC pin 11
15	Ext Relay 2 NO		

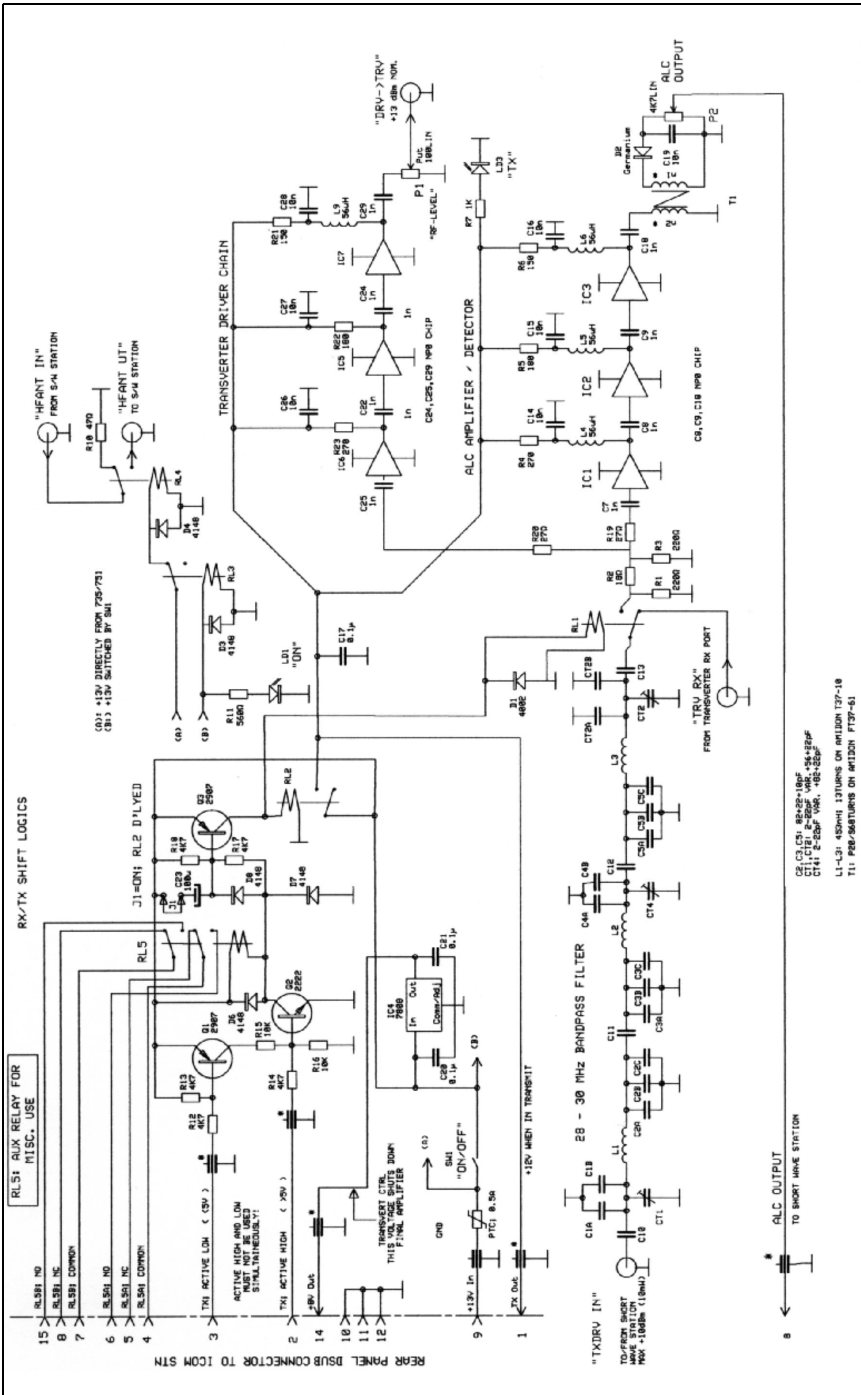

Fig 3: Circuit diagram of the transverter adapter.

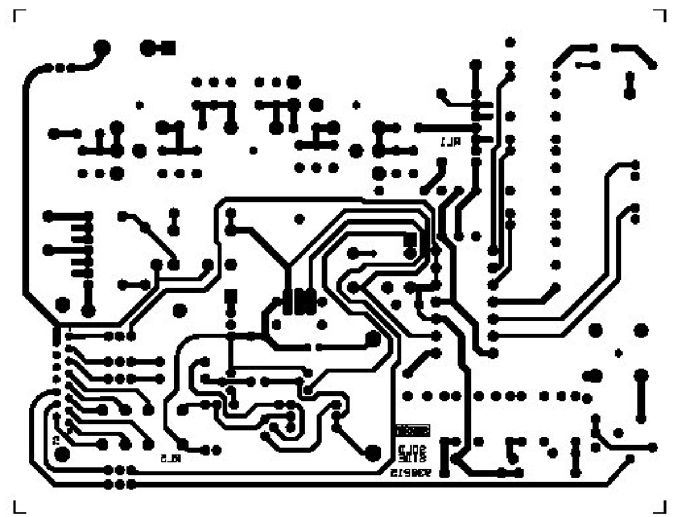

Fig 4: Track side of transverter adapter PCB.

- Dimensions: (L x W x H) 165 x 120 x 65mm
- Connectors: Five 50 Ohm BNC female for all RF signals. One 15 pin DSUB for DC interfaces.

The transverter adapter is connected to the ICOM HF transceiver by means of a 15 pin male DSUB connector (Table 1) and coax cables at the rear of the adapter (Fig 2).

3.0

Principle of Operation

The circuit diagram of the adapter is shown in Fig 3. The printed circuit board and component layout are show in Figs 4 – 6, the pcb files are available on the VHF Communications web site.

The transverter adapter consists of four major parts:

- 28-30MHz Bandpass Filter
- RF amplifier chain
- ALC amplifier chain
- DC Interfacing and relay actuation circuitry

The transverter terminal on the rear panel of the ICOM station is not filtered. Quite a few spurii are present close to the 28MHz TX drive signal. These are not strong, but if they are left unfiltered, they will mix with the transverter LO producing signals outside the amateur radio bands.

The 28…30MHz Bandpass filter reduces the spurii. After filtering, non harmonic intermodulation products are down more than –70dBC. Since the the transverter terminal is also used for receiving purposes, the down converted signal is also filtered.

The RF amplifier has enough gain (>30dB) to be able to compensate for ICOM’s 30mV (-18dBm) output level. The output level without ALC limiting may reach +16dBm, but the first har-

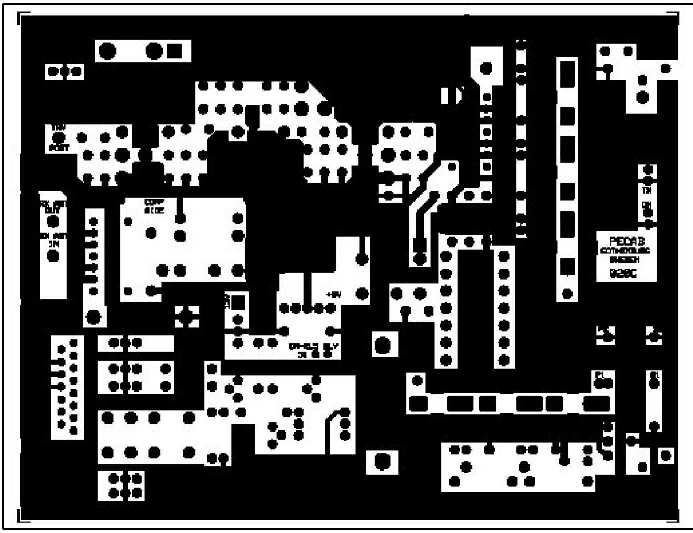

Fig 5: Ground plane side of transverter adapter PCB.

monic at 56MHz will then be less than –15dBC.

At first, only one chain of RF amplifiers was used, but it soon became clear that the ALC detector generated severe inter-modulation. Therefore, two amplifier chains were used.

When the ALC level is adjusted so that the transverter adapter 28MHz drive is reduced to some +10dBm, the 56MHz first overtone is reduced to –56dBC.

ICOM uses +8V to shut down the final amplifier. An 8V supply is available at the accessory outlets of the 735 and 751, but the maximum load is specified at 5mA for the 751. We did not want to use this obviously “fragile” supply, so we chosed to supply an 8V output of our own, derived from a 7808.

3.1 ICOM IC751

The 751 has a +8V supply present when in TX. This signal is fed to the adapter pin 2 (TXHigh) and switches the Accessory relay RL5 through Q2. Outputs from this relay should be used to put the transverter and any [VHF/UHF] final amplifier in TX mode.

The signal also turns on the RF and ALC amplifiers. The voltage may be delayed by approximately 0.5 seconds. This ensures that the transverter and any antenna switches relays has settled before RF current can flow.

3.2 ICOM IC735

The 735 has the PTT connected so that the microphone and SEND pins of the accessory connectors at the rear panel are in parallel . When in RX position, the voltage is about 6.5 V.

This signal is not usable for the transverter adapter. Instead, the SEND Phono plug outlet is used. This is a mercury wetted relay with about 100mA current limit. The centre pin is forced to 0V when the 735 is in transmit mode. This signal should be connected to the TXLow (pin3) of the adapter.

3.3 Leakage Protection

The adapter also has an extra feature for HF Input antenna protection. The RX signal from the HF antenna relay of the 735/751 is connected to the RXAnt-In connector of the adapter. When the trans-

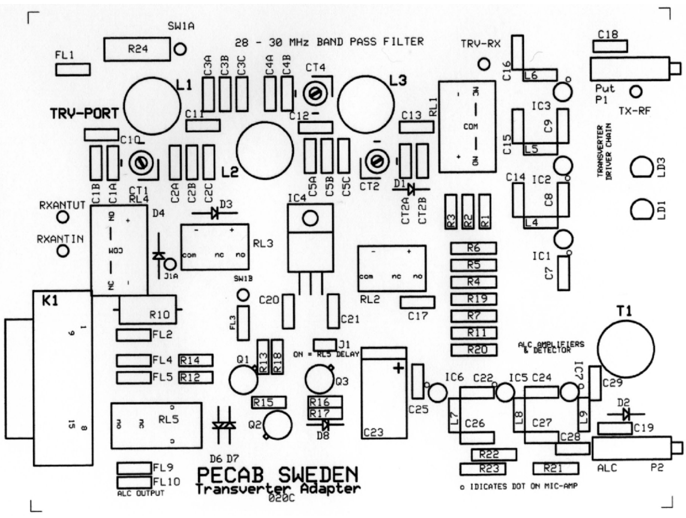

Fig 6: Component layout for transverter adapter.

ceiver and the adapter is turned OFF, the signal is fed to the 50 Ohm resistor R10. When the transceiver is used for HF operation (adapter turned OFF), the relay RL4 is activated and the signal is fed to the RXAnt-Out connector and on to the HF transceiver. When the equipment is used for VHF/UHF, the relay is deactivated and the HF signal is directed to the 50 Ohm resistor. This prevents any lekage. Ever heard south America on 432 MHz for an entire afternoon?

4.0

First Time Operation

Connect the Transverter Adapter to the HF Transceiver in accordance with the interface drawing and table. The adapter has an extra relay with two galvanically isolated sections intended for T/R switching of a linear amplifier. Make sure that applied voltage and current never exceeds 100V/2A resistive load.

Connect a transverter with a power meter in series with the [VHF/UHF] antenna cable. Power up the HF transceiver and the transverter adapter. Set the frequency to a clear frequency and make sure that your dummy load is connected.

- Turn the ALC and RF potentiometer fully Counter Clock Wise.
- Key down to obtain RF output.
- Slowly turn the ALC potentiometer until the ICOM transceiver starts indicating ALC return.
- Turn the RF-Level potentiometer slowly Clock Wise until the RF level saturates. Then turn CCW until it just starts to drop.
- The Transverter Adapter is now adjusted and ready for use

5.0

Appendix:

5.1 Parts List

All resistors are ¼ W, metal film, Most capacitors used were KEMET CK05

R1, R3	220 Ohm
R2	10 Ohm
R4	270 Ohm ½ W
R5, 22	100 Ohm ½W
R6, 21	150 Ohm ½W
R7	1K
R10	47 Ohm 1W 8 modules (8 x 2.54)
R11	560 Ohm
R12, 13, 14, 17, 18	4K7
R15, 16	10K
R19, 20	27 Ohm
R23	270 Ohm ½ W
R24	PTC "Raychem fuse" 250 mA
C1A, 2B, 3B, 4A, 5B, 11, 12	22pF NP0
C1B, 10, 13	56pF NP0
CT1, CT4	2-22pF Philips Trimmer "Green" 808-11
C2C, 3C, 4B, 5C	82pF NP0
C3A, 5A	10pF NP0
C7, 8, 9, 22, 24, 25, 29	1nF CHIP 1206 NP0
C14, 15, 16, 19, 26, 27, 28, 30, 31	10nF X7R
C20, 21	100nF X7R
C23	100uF 35VDC Axial 10 modules (25.4mm)
D1, 3, 4, 6	1N4002 or equiv.
D2	AA118 (preferred) or 5082-2800 (1N5711)…or any germanium type…
D7, 8	1N4148
LD1	LED 5mm Green "ON"
LD3	LED 5mm Red "TX"
Q1, 3	2N2907
Q2	2N2222
IC1, 6	MAR3
IC2, 5	MAR4
IC3, 7	MSA1104 (or MAW11)
IC4	7808
L1,L2,L3	453nH: This should be some 13 turns on AMIDON T37-10
L4-L9	56uH axial
T1	Primary 20 turns secondary 68 turns on AMIDON FT37-61
FL2-FL10	Murata DSS306, 1nF Three lead EMI-filter.
K1	15 pin DSUB Male. Right Angle Note that pin numbers differs between a male and a female if you use a female connector instead…
RL1, RL4	Relay: RF, RK1-12V
RL2, RL3, RL5	Relay: SDS DSE-M12V

5.2 Design Files

Together with this article, the original design files made using (the now anti-quated) EEDesigner-III Schematic Capture and PCB Layout tool has been found. I reinstalled the complete design system and reopened the design. I was also able to generate Gerber and NCData should anyone want to fabricate circuit boards. Using more modern software, I was able to convert RS274D Gerber to the more modern RS274X format. The design file can be found on the VHF Communications web site www.vhfcomm.co.uk

Thanks to:

SM6FHZ, Ingolf, who provided the first sketches and ideas and made filter calculations.

SM6EUP Hans, who provided valuable tips and funding for the idea.

Alexander Meier, DG6RBP

Universal measuring amplifier for low DC voltages

A simple universal measuring amplifier has been developed for measuring low DC voltages right down to the microvolt range. The gain can be switched in four decimal ranges from 1 to 1,000 with an accuracy of better than 1 per cent. An integrated low pass filter can be switched in decimal steps from 10Hz to 1,000Hz.

1. Introduction

We frequently want to measure very low DC voltages. One example is the measurement of the output voltage of a diode detector, plotted against the input power. For input power values of under -30dBm, we obtain DC voltages which are only

Table 1: Specification of measuring amplifier.

Maximum input range (no damage)	±9v pp
Maximum output range (typical)	±5v pp
Input impedance	10MOhm
Output impedance	1kOhm
Gain	Switchable 1 / 10 / 100 /1000
Cutoff frequency	Switchable 10Hz / 100Hz / 1kHz (single pole RC filter)
Slew rate (typical)	Approximately 10mv/µs (filter 1kHz) Approximately 1mv/µs (filter 100Hz) Approximately 0.1mv/µs (filter 10Hz)
Zero offset	Manual using a potentiometer
Zero offset drift (typical)	<1mv/h (gain = 1000)
Indicators	LED for battery low +9v, -9v
Connectors	BNC sockets
Power supply	2 x 9v battery
Battery life (typical)	15 hours
Temperature range	+20 to +30°C
Housing	Aluminium
Dimensions	Approximately 103 x 60 x 170mm

Fig 1: Picture of the the prototype measuring amplifier.

microvolts. This universal DC voltage amplifier has been developed to be able to measure such low DC voltages using common digital voltmeters. The technical data for the universal measuring amplifier can be found in Table 1.

The amplifier is based on three OP27, low-noise precision operational amplifiers. However, an amplifier of this type requires a correspondingly low-noise and hum-free power supply. The best option here is to use two 9V batteries. The housing used for the measuring amplifier is also important. The amplifier must be fully shielded otherwise interference like 50Hz hum will be induced due to the mains supply lines. So only a metallic housing can be considered.

The gain can be switched in decimal steps from 1 to 1,000. The gain can be obtained accurately, without alignment, using precision metal film resistors (0.1%). The gain is switched by means of reed relays, so that the actual signal does not need to be fed through the comparatively poor contacts of the rotary switch on the front panel.

The cutoff frequency of the measuring amplifier can be switched in decimal steps from 10Hz to 1kHz. The use of a low cutoff frequency is particularly advantageous for very noisy signals or oscillating signals. The latter can, for example, come from power lines via an induced 50Hz alternating field. High frequency variations can be suppressed using the low pass filter, which leads to a more stable display on the voltmeter.

2. Description of circuit

The circuit diagram for the measuring amplifier can be seen in Figs 2 and 3. At the input, resistor R1 provides a defined input resistance of 10MOhm. The high impedance input signal is measured by the voltage follower U1 and is available as a low impedance source at the output (PIN 6).

In the next stage (U2), the gain is determined by the switched feedback resistors (R3-R6) and by R2. Because an inverting amplifier is used, no unusual resistance values are needed to obtain the decimal gain values. Since the resistors, R2 to R6, determine the gain, only

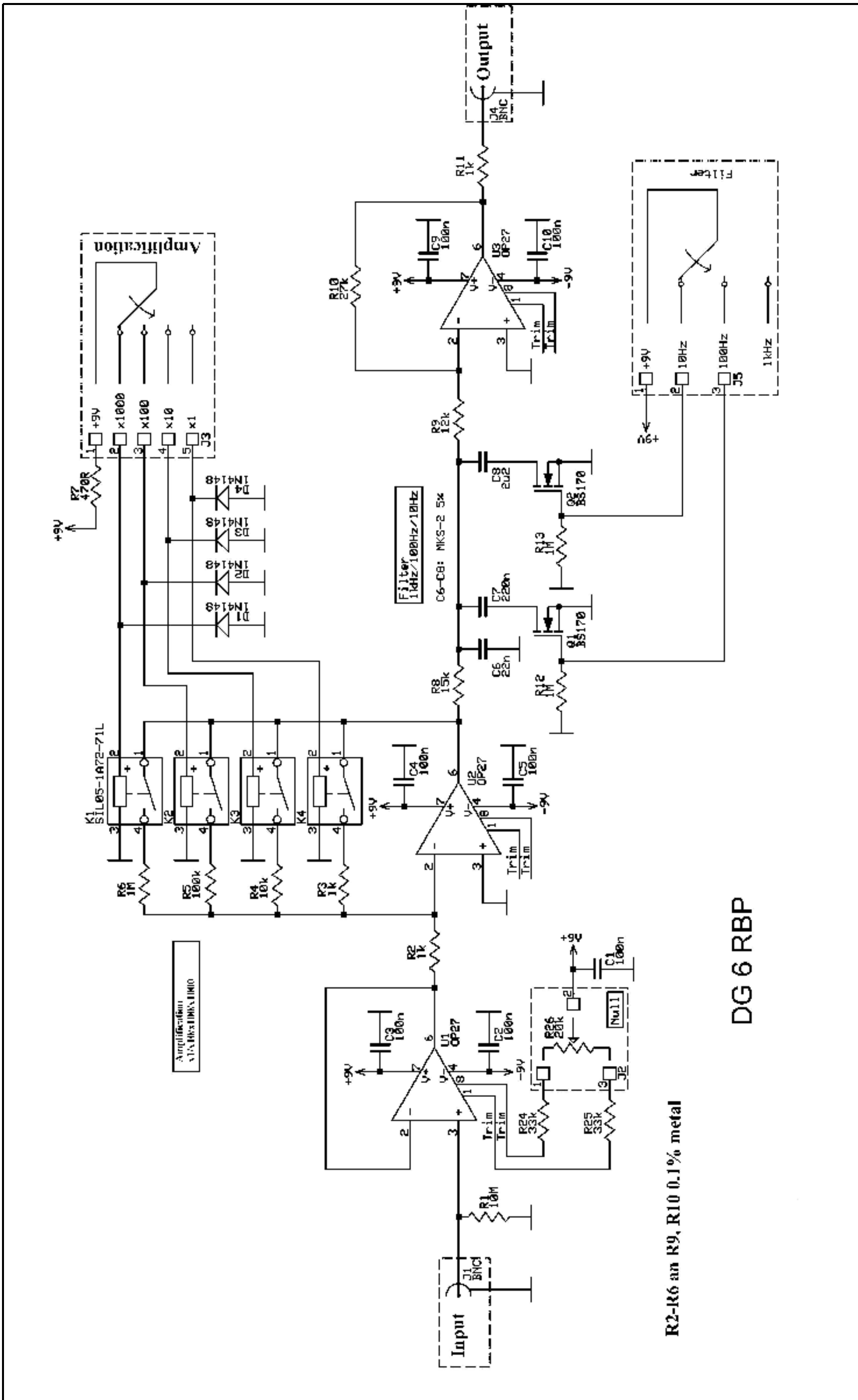

Fig 2: Circuit diagram of the amplifier section.

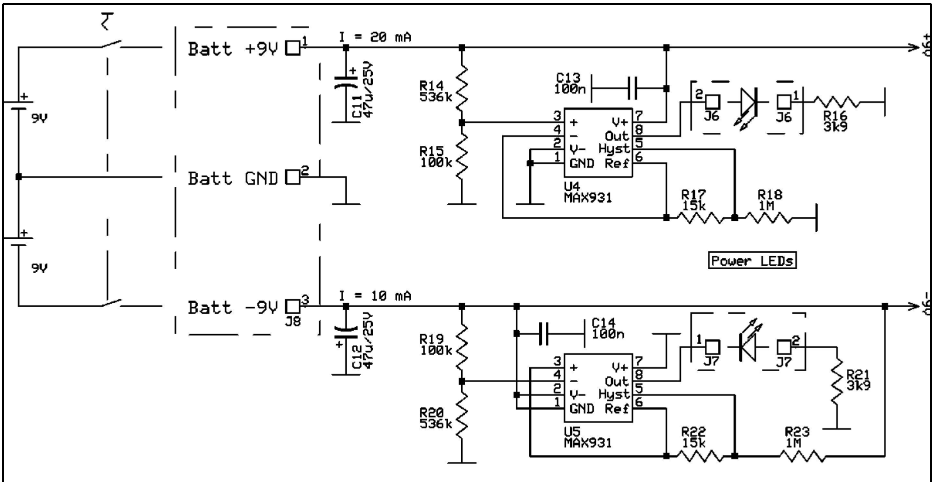

Fig 3: Circuit diagram of battery monitor.

precision metal film resistors (0.1%) can be used. To ensure that the amplifier's output is the correct polarity, a second, unity gain, inverting amplifier (U3) is used. Precision metal film resistors (0.1%) must also be used for R8 to R10. The output resistor, R11, protects the output and prevents any problems when capacitive loads are connected (e.g. co-axial cable connection to an oscilloscope). The output resistance of 1kOhm, together with the input impedance of the load forms a voltage splitter. With a typical input resistance for a digital voltmeter or oscilloscope of >10Mohm, there is a negligible error of less than 0.01%.

A switchable, single pole RC low pass filter is connected between the two inverting amplifier stages (U2 and U3). The cutoff frequency is restricted to 1kHz by capacitor C6. Additional capacitors (C7 or C8) can be switched in parallel to reduce the cutoff frequency to 100Hz or 10Hz.

The gain of U2 is switched by reed relays, K1 to K4. So only a control voltage must be switched on the front panel, using the rotary switch.

The zero point of the entire amplifier is set on the front panel by potentiometer R26 that sets the offset of U1. The range has been restricted by the resistors, R24 and R25, so that precise adjustment is still possible even with high gain.

The two comparators, U4 and U5, monitor the supply voltage. If the battery voltage falls below 7.5 Volts, the LED display for the battery in question goes out. Any flickering of the LEDs at the switching threshold is avoided by hysteresis, using R17 and R18 (or R22, R23).

The current consumption for the entire circuit is approximately 20mA (at +9V) and 10mA (at -9V). The highest consumption levels are for the operational amplifiers, U1 to U3, and the reed relays, K1 to K4. The comparators (U4, U5) for voltage monitoring require no more than 5µA (without LED). Only low current light emitting diodes are used. No electronic reverse battery protection was provided, this is present because of the mechanical dimensions of the battery compartments.

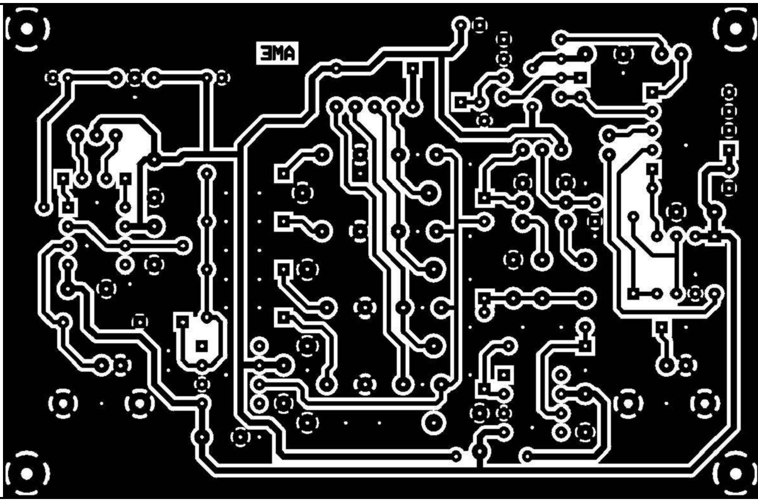

Fig 4: Top side of the PCB for the measuring amplifier.

3. Assembly

No SMD components are used in the circuit, which makes copying particularly simple.

The board layout of the through hole plated, two layer printed circuit board, made from 1.5mm thick FR4 material, is shown in Figs 4 and 5. There are no special instructions concerning the fitting of the components in accordance with Fig 6.

Once the components have been fitted

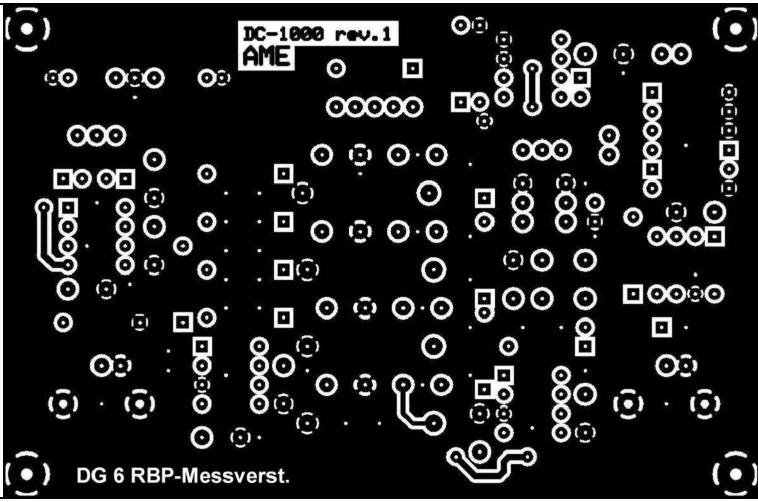

Fig 5: Bottom side of PCB for measuring amplifier.

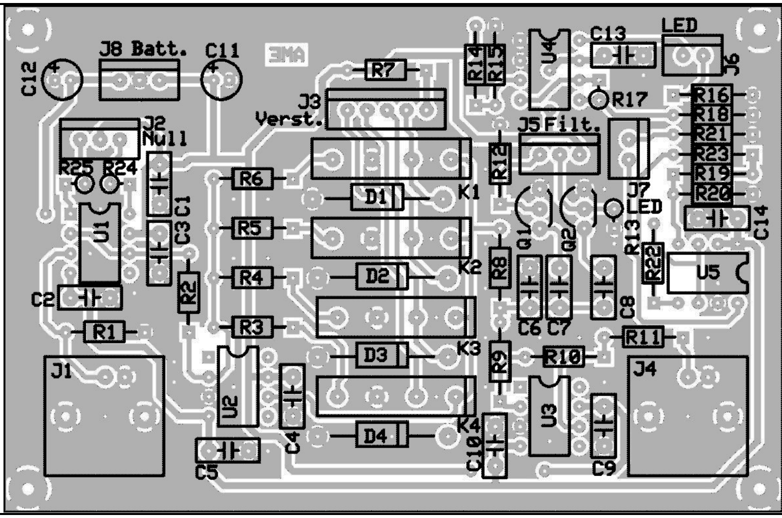

Fig 6: Component layout for the measuring amplifier.

onto the printed circuit board, it is inserted into the housing and connected to the other components and the battery compartments. The connections to the battery compartments and the switch are protected from accidental short-circuits by a heat shrinkable sleeve.

After a comprehensive function test (amplification, cutoff frequency, comparators), the measuring amplifier is ready for operation. The prototype of the printed circuit board is shown in Fig 7 and the completed prototype is shown in Fig.1.

4.
Parts list

Quantity, component and designation

1 x	470Ohm, 1/4W	(R7)
2 x	1kOhm, 0.1%, metal 1/4W	(R2, R3)
1 x	1kOhm, metal, 1/4W	(R11)
2 x	3.9kOhm, metal 1/4W	(R16, R21)
1 x	10kOhm, 0.1% metal 1/4W	(R4)
1 x	12kOhm, 0.1%, metal 1/4W	(R9)
1 x	15kOhm, 0.1%, metal 1/4W	(R8)
2 x	15kOhm, metal, 1/4W	(R17, R22)
1 x	27kOhm, 0.1%, metal 1/4W	(R10)
2 x	33kOhm, metal, 1/4W	(R24, R25)
1 x	100kOhm, 0.1%, metal 1/4W	(R5)
2 x	100kOhm, 1%, metal 1/4W	(R15, R19)
2 x	536kOhm, 1%, metal 1/4W	(R14, R20)
1 x	1MOhm, 0.1%, metal 1/4W	(R6)
4 x	1MOhm, metal, 1/4W	(R12, R13, R18, R23)
1 x	10MOhm, metal, 1/4W	(R1)
1 x	20kOhm, potentiometer with 6-mm. axis, linear (R26)	
1 x	22nF 5%, MKS-2, RM5	(C6)
9 x	100nF 5% MKS-2, RM5	C1-C5, C9-C10, C13-C14
1 x	220nF 5%, MKS-2, RM5	(C7)
1 x	2.2µF, 5%, MKS-2, RM5	(C8)

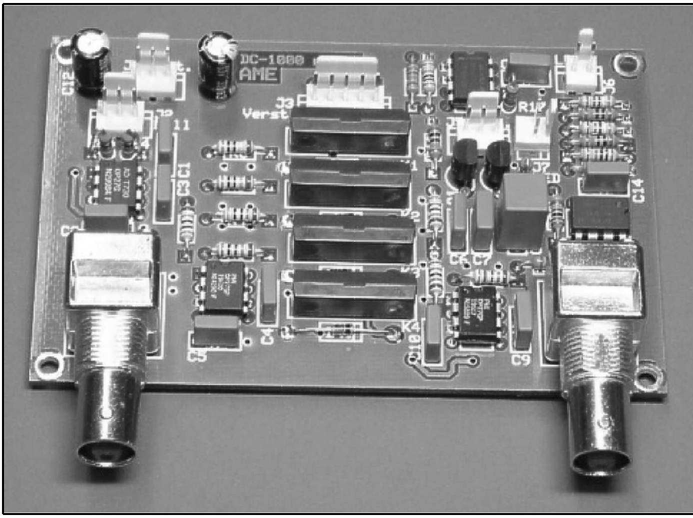

Fig 7: Picture of completed prototype PCB.

2 x	47µF/25V electrolytic capacitor, vertical	(C11, C12)
3 x	OP27, DIP8	(U1-U3)
2 x	MAX 931 CPA, DIP8	(U4, U5)
2 x	BS 170	(Q1, Q2)
4 x	1N4148	(D1-D4)
4 x	SIL05-1A72-71L	(K1-K4)
2 x	BNC printing socket metal	(J1, J4)
2 x	board plug, 2-pin	(J6, J7)
3 x	board plug, 3-pin	(J2, J5, J8)
1 x	board plug, 5-pin	(J3)
2 x	board coupling	2-pin
3 x	board coupling	3-pin
1 x	board coupling	5-pin
18 x	contact for board coupling	
1 x	rotary switch, 3x4 positions	
1 x	rotary switch, 4x3 positions	
3 x	control knobs for 6-mm. axes	
2 x	LED, green, 3 mm, low current	
2 x	LED mounting, 3 mm	
1 x	rocker switch, bipolar	
2 x	battery compartment for 9-V block battery	
1 x	printed circuit board DG6RBP measuring amplifier, two-layer, through-hole plated	
1 x	housing with front and rear panels (holes milled, printed)	

5. Operating instructions

For precise measurements, a warm-up time of approximately 15 minutes is recommended once the measuring amplifier has been switched on. Only high quality coaxial cables should be used at the input and output.

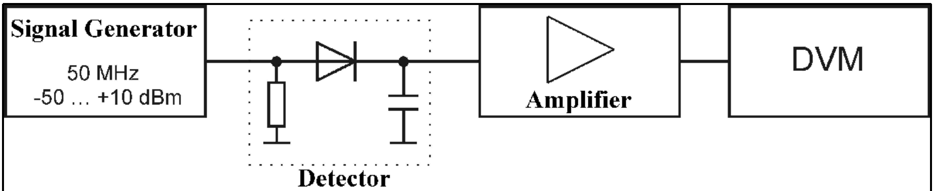

Fig 8: Measurement application of the amplifier.

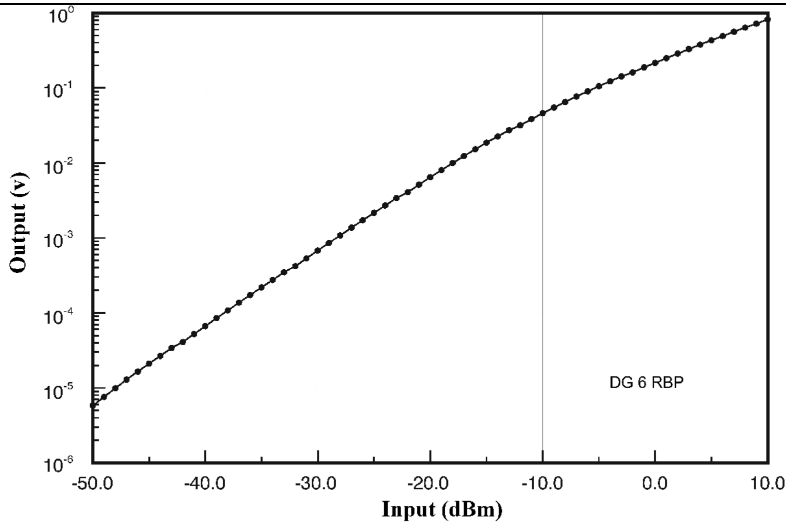

Fig 9: Test measurements from the diode detector.

The zero point should be set before any test object is connected, after a long period of operation or after any variations in the ambient temperature. To do this, short circuit the input and use the potentiometer, R26, to set the output voltage to 0mV. As far as possible, the measuring amplifier should not be subjected to any temperature variations.

6. Typical application

As a typical application for the measuring amplifier, the output voltage of a diode detector is indicated, plotted against the input power.

The diode detector was connected to a standard RF signal generator and the output DC voltage was amplified, using the measuring amplifier. A 3½ digit, digital voltmeter (DVM) acts as a display. 50MHz was set as the frequency on the signal generator. The measurement assembly is shown in Fig 8.

The output level of the signal generator was increased in stages from -50dBm to +10dBm in 1dB steps and the output voltage was noted. Each time the maximum output voltage of the measuring amplifier (approximately ±5V) was exceeded, the amplification factor was reduced by one step.

The DC output voltage of the detector can be calculated using the gain of the amplifier. The measurement data obtained in this manner were then converted into a measurement curve on the PC, a curve shown in Fig 9.

Without the measuring amplifier, it would not have been possible to determine the very low output voltage (µV) for low RF levels using the digital voltmeter.

Ralf Rudersdorfer, OE3RAA

New measuring method to determine the bandwidth occupied by J3E (SSB) transmissions

The linearity of single sideband transmission peaks is checked in the same way as RF power amplifiers, normally using a multi-tone test. Considerable control of the measurements is required in particular for short peak power levels. There are also other parameters that affect the transmission.

The classic procedures for linearity testing and for determining the bandwidth occupied by J3E transmissions are discussed below. A new measuring method is presented which makes it possible to measure the effective bandwidth occupied under real conditions and with real combinations of equipment.

1. Initial situation

The bandwidth occupied by a transmission is the most important parameter as far as radio engineering is concerned. Limiting values specified by the regulatory authorities must be adhered to in accordance with the current test specifications. There are also directives that define the bandwidth to be occupied and the maximum levels of spurious transmissions allowed.

If you are using high power then it is important to generate a signal as narrow as possible. Having a signal that is as clean as possible also reduces the problems for local radio stations operating close to your own operating frequency.

The bandwidth actually occupied by a J3E transmission is usually determined by:

- The transmitter's frequency response - this means that improvements can be made by using a sharp audio filter.
- Linear crosstalk, which can be divided into:
  - suppression of carrier
  - suppression of undesirable side band
- The linearity of the transmitter, which is expressed by the intermodulation behaviour and, in my opinion, by the compression point
- Any level restriction e.g. ALC (Automatic Level Control) and its regulation characteristic and/or the response of any frequency limiting

The conventional test procedure is based on determining the individual parameters (largely independently of one another) and establishing them under specific and

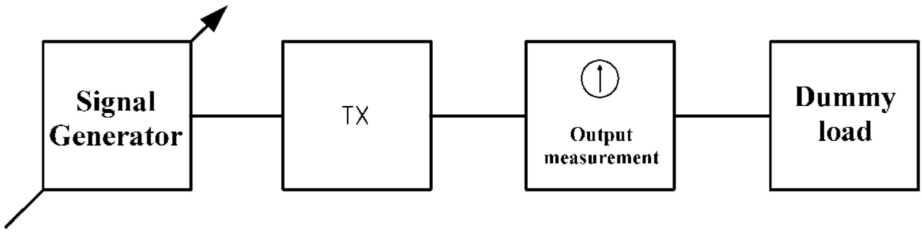

Fig 1: Test setup used to measure the frequency response.

(hopefully) repeatable conditions. The procedure is described in detail below.

2. Conventional testing of parameters in J3E transmission peaks

2.1. Transmitter frequency response and transmission band

To determine the transmitter frequency response, the transmitter is modulated via the microphone input by a tone generator (Fig 1). To keep the transmission in the linear range, the modulation voltage should be selected in such a way that the transmission power remains within 6dB of the maximum peak power. The modulation frequency can then be tuned in steps, while the voltage level fed in at the microphone input is kept constant. The output generated, shown in Fig 2, represent the frequency response of the transmission. A steep reduction should be recorded below approximately 300Hz and above 2,500Hz to 3,200Hz.

The regulations (e.g. FuG 100) specify a reduction of 25dB at 3.5kHz compared to the power at 1kHz. The corresponding

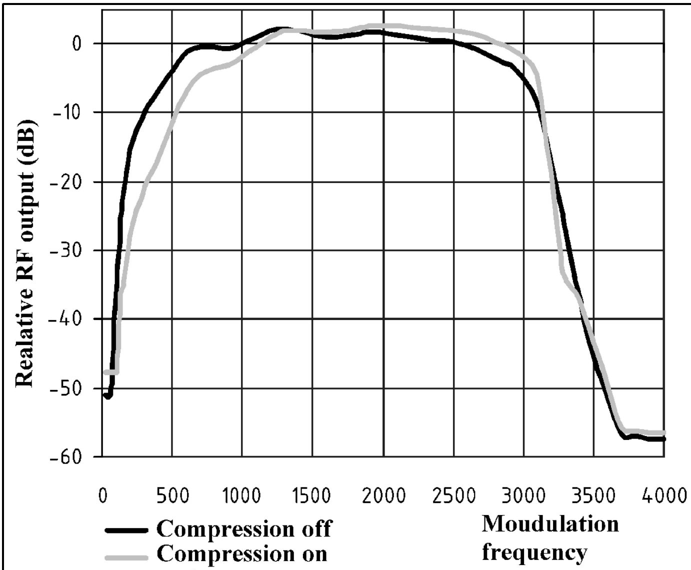

Fig 2:Frequency response of a typical J3E (SSB) transmission plotted as RF output in dB against audio frequency. The response falls off below 100Hz and above 3700Hz. See Fig 3 for the output in the RF spectrum.

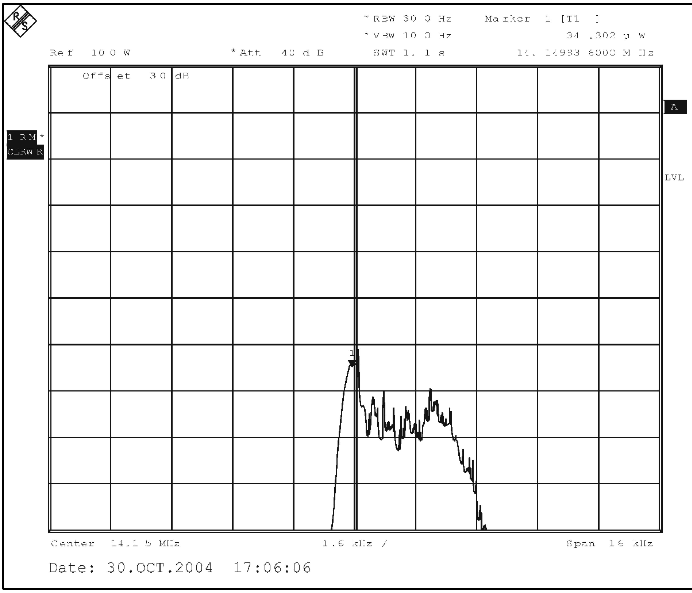

Fig 3: Frequency response of a typical J3E (SSB) transmission plotted as RF output against RF frequency.

readings are

- 4kHz -32 dB
- 10kHz -48 dB
- from 20kHz -60 dB

If a wattmeter is the only power measurement available, then the drop in performance at the highest peak of the transmission curve can determined in dB as:

$$A_{dB}(f_{\mathrm{mod}}) = 10\log\left(\frac{P_{\mathrm{mod}}(f_{\mathrm{mod}})}{P_{\max}}\right) \quad (1)$$

Where:

A_{dB} = power drop in relation to highest peak of the transmission curve, plotted against modulation frequency in question, f_{mod}, in dB

P_{max} = power at highest peak of the transmission curve, in W

P_{mod} = initial transmitter output, plotted against modulation frequency in question, f_{mod}, in W

2.2. Suppression of carrier and sideband

To determine the carrier suppression, together with the suppression of the undesirable sidebands, the transmitter is once again modulated via the microphone input using a tone generator (Fig 4). The transmitter is keyed using the PTT and the modulation voltage level is increased in small stages until the output stops increasing. The spectrum analyser is connected using a power attenuator with an appropriate maximum thermal capacity. This is used to evaluate the suppression of the carrier and the undesirable sidebands that appear on either side of the carrier spaced by the modulation frequency. An example of a J3E transmitter, modulated in the upper sideband, can be seen in Fig 5. A harmonic of the modulation tone can be detected to the right of the modulation tone, at approximately -39dBc. These are generated as the saturation at the microphone input increases.

From a practical point of view the biggest uncertainties lie in the fact that the

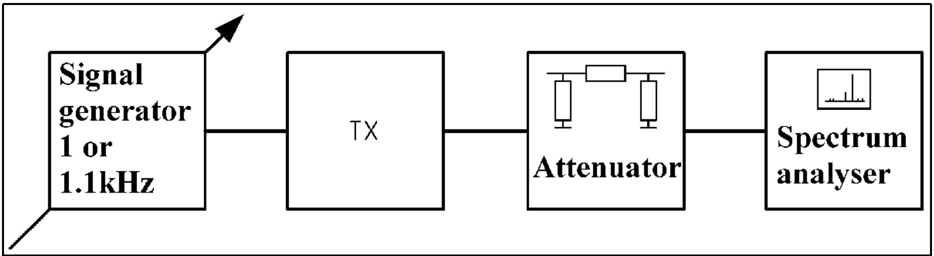

Fig 4: Test setup used to measure carrier and sideband supression.

sideband suppression can vary, to a not inconsiderable extent, with the modulation frequency. Only one discrete modulation frequency is normally used in the measurement procedure (often 1kHz or 1.1kHz), however the entire speech band is present from the microphone.

2.3. Transmitter intermodulation

Intermodulation represents an important criterion for assessing the linearity of both transmitters and (downstream) RF power amplifiers. Among all the effects, intermodulation products frequently dominate the bandwidths actually occupied with a single sideband modulated signal. Small transmitter intermodulation intervals in the J3E type of transmission also indicate a direct effect on the modulation quality.

To determine the transmitter intermodulation attenuation, either the signals from two low distortion tone generators are fed into the microphone input via a suitable coupler, or else a special two-tone generator is used, [3] offers a suitable suggestion. Certain addition circuits can be used as couplers to bring the two modulation tones together, with the help of operational amplifiers. The transmitter is keyed via the PTT and the two modula-

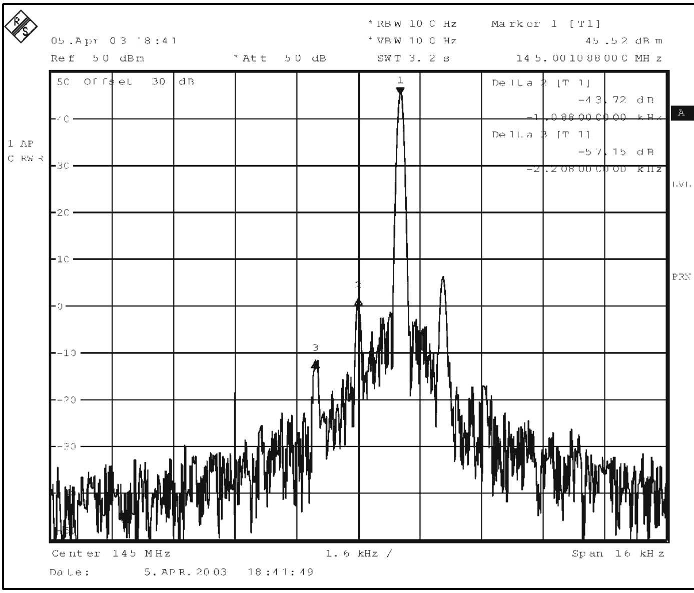

Fig 5: Spectrum analyser output showing sidebands with 1.1kHz tone.

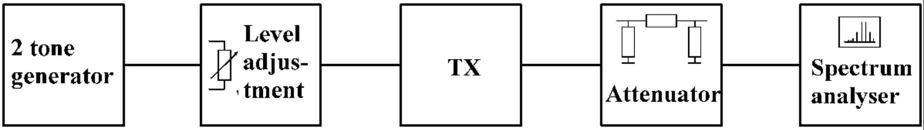

Fig 6: Test setup used for two-tone test.

tion tones are increased in small stages, with the same amplitude, until the output of the tones in the RF band lies below the nominal output, with single tone level control (Fig 6).

As a result of non-linearities, distortion naturally arises, which appears in the frequency range and thus in the spectrum displayed around the transmission signal as additional spectral lines. These will be the harmonics of the modulating tones, together with the totals and difference frequencies arising from this through mixing, which are designated as inter-modulation products. The intermodulation products that are of particular interest in relation to the bandwidths occupied (by the two-tone modulated transmission signal) occur (in frequency-dependent terms) at:

$$f_{IM3} = 2f_1 - f_2 \quad (2) \text{ and}$$

$$f_{IM3} = 2f_2 - f_1 \quad (3) \text{ and also at}$$

$$f_{IM5} = 3f_1 - 2f_2 \quad (4) \text{ and}$$

$$f_{IM5} = 3f_2 - 2f_1 \quad (5) \text{ and also at}$$

$$f_{IM7} = 4f_1 - 3f_2 \quad (6) \text{ and}$$

$$f_{IM7} = 4f_2 - 3f_1 \quad (7)$$

Where:

f_1 = frequency of lower exciter tone in Hz

f_2 = frequency of higher exciter tone in Hz

f_{IM3}, f_{IM5}, f_{IM7} = frequency at which inter-modulation products of corresponding order are to be expected, in Hz

When practical measurements of this kind are carried out in the upper sideband (USB), a component can frequently be detected which is scarcely mentioned in the literature, and which falls directly into the useful channel between the exciter tones at:

$$f_{s(USB)} = f_1 + (f_2 - f_1) \quad (8)$$

or with measurements in the lower sideband (LSB) at:

$$f_{s(LSB)} = f_T + (f_2 - f_1) \quad (9)$$

Where:

$f_{S(USB)}$, $f_{S(LSB)}$ = interference product arising additionally through mixing and dependent on sideband used in Hz

f_T = carrier frequency of J3E signal investigated in Hz

f_1 = frequency of lower exciter tone in Hz

f_2 = frequency of higher exciter tone in Hz

Fig 7 shows these components ($f_{S(LSB)}$) somewhat higher than the lower exciter tone. For practical work, it is helpful to note that intermodulation products always arise at intervals above and below the intermodulating exciter tones, and thus at their frequency-dependent intervals to one another:

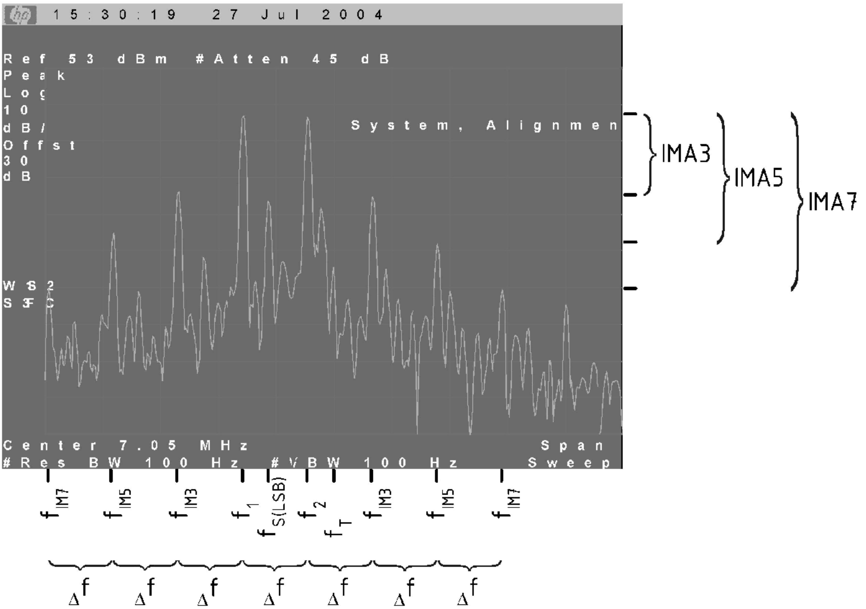

Fig 7: Output of the two-tone test showing the sidebands generated.

$$\Delta f = f_2 - f_1 \qquad (10)$$

Where:

f_1 = frequency of lower exciter tone in Hz

f_2 = frequency of higher exciter tone in Hz

Intermodulation products rise considerably faster than the exciter signals that cause them to occur, so that the intermodulation interval (IMA) decreases as the level control increases. It is difficult here to obtain any direct indication of a formula regarding the rise of the intermodulation products; something like the evidence for a cubic rise is frequently cited in connection with third order intermodulation (f_{IM3}).

As amplifiers are often already being operated a long way into compression - i.e. in that range where an increase in the input power produces a smaller increase in the output than is the case for level control with small signals, considerations of this nature no longer apply. This is also emphasised quite clearly by the measurements shown in [4]. The reason behind a deliberate level control of this type in the non-linear range is that, firstly, it leads to maximum power yield, and, secondly, the circuit design can be optimised.

To be able to obtain really meaningful and, above all, comparable data concerning a transmitter intermodulation interval on various pieces of equipment, the most important thing necessary is absolutely identical measurement conditions. The strength of the intermodulation products frequently varies with the frequency of the modulation tones used for the measurement. The tone frequencies specified by some CEPT guidelines - 700Hz and 2.5kHz - are often used. On the basis of practical operation, the greatest uncertainties are caused by the fact that a whole lot of different frequency combinations - depending on the speech spec-

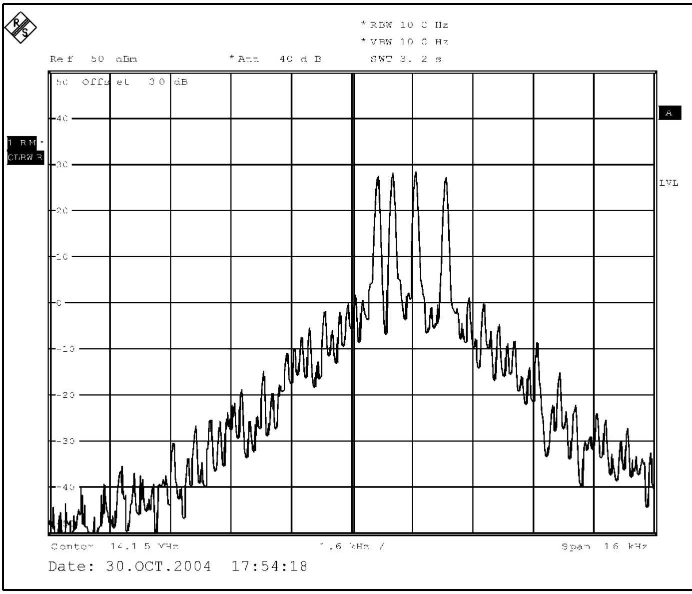

Fig 8: Output of a 10 watt transmitter being tested with the four-tone test. The tones were 700Hz, 1.1kHz, 1.7kHz and 2.5kHz.

trum - can lead to wild intermodulation. In addition, amplifiers frequently release short-term peak power levels (also know as pulse power), with values markedly above those for continuous operation.

For single sideband modulation, only the upper peaks of the modulation envelope bring about actual full modulation in the lower millisecond range, this changes if the key is pressed for a relatively long time at full power. An essential factor here is the DC supply voltage, which frequently collapses under full power. The internal resistance of the transformer cannot recharge the electrolytic filter capacitors but for short-term peaks the capacitors guarantee adequate peak current delivery. The thing is a dynamic process. This is shown by the non-linear characteristics of the dynamic system behaviour that has a drastic influence on the actual intermodulation behaviour.

2.4. Automatic level restriction

To investigate the limiter effect, according to the CEPT guidelines, a four-tone modulation signal is required (analogous to Fig 6). When the total modulation level is altered the change in the transmitter output should be monitored. It should change only within specified fixed limits for a defined variation in the total modulation level. A measurement of this kind on a 10Watt transmitter is shown in Fig 8. The power for each of the four individual tones is 12dB below the total output since the individual components are added together. The small variation in level between the individual tones is based on the frequency response.

Carrying out such measurements is difficult in practice, among other things, the sensitivity and impedance of the microphone input, which depends on the type of equipment being investigated, must be know and is rarely specified in equipment documents. The total modulation level should then be varied around this normal sensitivity. For the reasons stated, we shall not go into the precise methodology in any greater detail.

One thing of interest here is if the microphone input is saturated by the four-tone modulation signal, the limiting

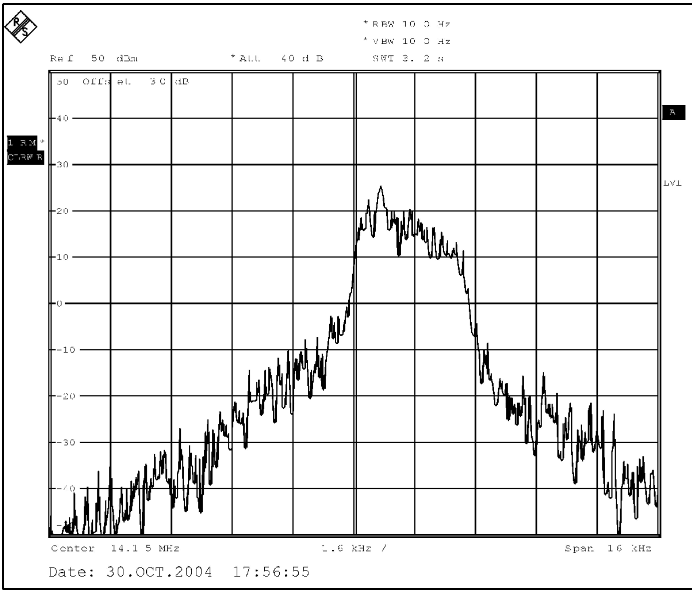

Fig 9: Output of a 10 watt transmitter being tested with the four-tone test with the audio input saturated.

effect can provoke the behaviour like that shown in Fig 9. The shape of the curve (not the power!) is very like that depicted when the measuring method discussed later in the article is implemented, using real modulation. This raises the obvious possibility that J3E transmitters modulated using the microphone are already frequently level controlled to such an extent that the automatic level restriction is already having a marked effect.

3. Provisional appraisal

In the classic measurement procedures to determine the spurious transmissions close to the carrier due to a J3E transmission, together with the bandwidth occupied, there are almost always uncertain factors concerning the signal actually being emitted on the band. The effects in real operation are superimposed but are only detected one after another in measurements and are not fully taken into consideration. The methodology set out below is intended to fill these gaps, and provides additional information. It requires the use of a spectrum analyser with a memory.

4. New measuring method provides additional information

A spectrum analyser usually operates in “Clear/Write” mode where the screen is refreshed for every frequency sweep. The amplitude values previously detected for the frequency in question are thus selected again and the previous values are overwritten. If a speech modulated transmission is displayed in this manner, the screen would show a random jump between the noise base in the speech pauses and the power at a specific frequency, which is dependent on the degree of level control (Fig 10).

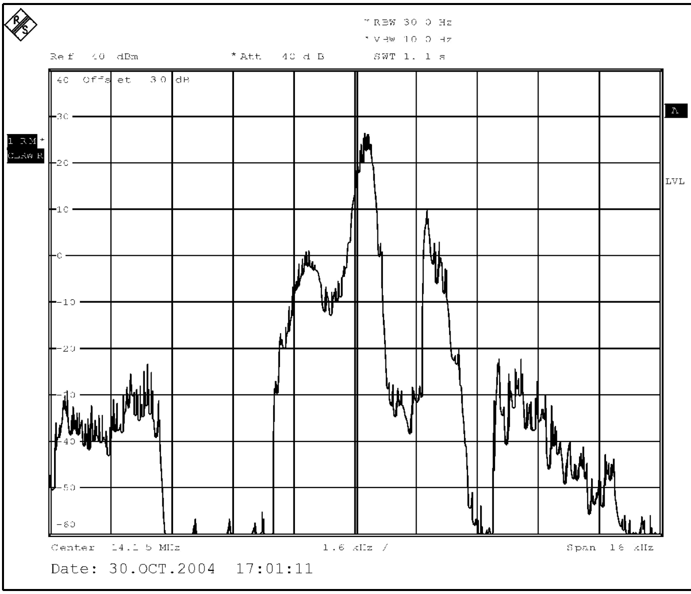

Fig 10: Spectrun analyser output with speech input. The spectrum analyser is set to "clear/write" mode.

In contrast to this, the “Max Hold” mode uses a comparator, so that the display show the maximum values as the frequency is repeatedly swept.

If a transmission is made using the microphone normally used while the frequency range near the carrier is displayed on the spectrum analyser, then little by little a screen builds up as the sweeps continue (Figs 11 and 12). Assuming that the transmission was long enough, a curve is generated that represents the “worst case” and shows all of the components present in the transmission at the "maximum" level. Effects such as the short-term generation of high peak outputs and the consequential signal purity changes that were described in 2.3 are account for. The worst-case values for the attenuation of undesirable spurious transmissions are displayed.

4.1. Noteworthy settings on the spectrum analyser

The analyser should be used with sufficient attenuation (internal attenuation) to keep the inherent intermodulation of the measurement instrument low. This means display errors caused by the spectrum analyser itself are minimised. Even with 40dB attenuation and the attenuator be-

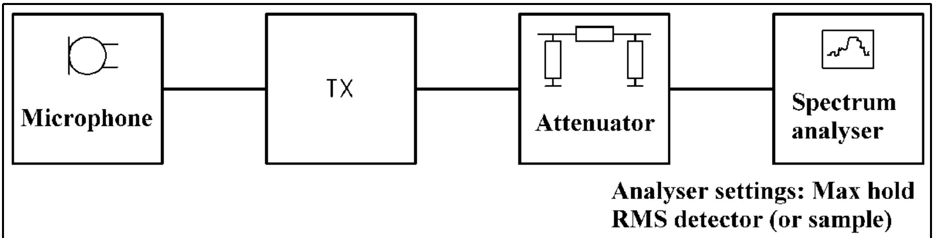

Fig 11: Test setup used for the new method of measuring a J3E (SSB) transmitter.

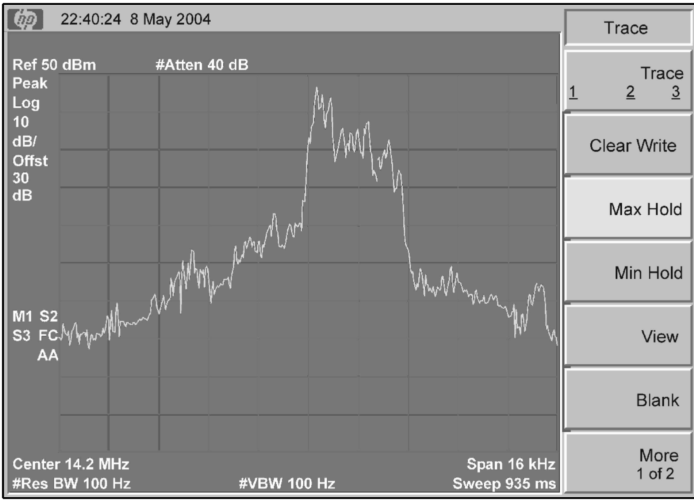

Fig 12: Spectrum analyser output with speech input and the spectrum analyser in "max hold" mode.

tween the transmitter output and the spectrum analyser (Fig 11) an adequate display resolution can usually be obtained. This is because the range between the highest peaks and the undesirable elements rarely exceed 45dB.

The intermediate frequency resolution bandwidth (RBW) on the spectrum analyser should be selected to be narrow, so that signal components lying close to one another can be separated. A value of 300Hz is practical and comparable with the lowest modulation frequencies. The display span should be 16kHz (+/- 8kHz around the carrier) with a sweep rate of 1s. To make the process go faster, the VBW (video bandwidth) can be selected to be greater than the intermediate frequency resolution bandwidth. Because almost all noise signals are swamped by components of the signal being analysed, the integrating function of the video bandwidth does not increase the measured values.

If the spectrum analyser has an "RMS detector", it is preferable to use this option. Otherwise, it is possible to use the "sample detector" built into almost all equipment.

4.2. Test readings

The behaviour of a J3E transmission signal with good signal purity is shown in Fig 13. The results obtained by the current test procedures and with real modulation using the "Max Hold" method are shown. The signal displays really good values, thus there is no sign of the carrier and the undesirable sideband in the single-tone test, as can be seen top left.

In Fig 14, with the speech processor turned on, the suppression of the sidebands and the carrier, together with the intermodulation intervals, remain exactly the same, but the real modulation test method exposes some interesting behaviour. Surprisingly, undesirable transmissions above and below the useful signals are attenuated much more. The reason for this can be found in the speech processor used, which trims the frequency response, such a signal is no longer state of the art. In this case it is an advantage because it gives a purer transmission. (By the way, the speech quality is not impaired using the speech processor on the apparatus tested.)

The results of the investigations into a valve amplifier stages used to boost the

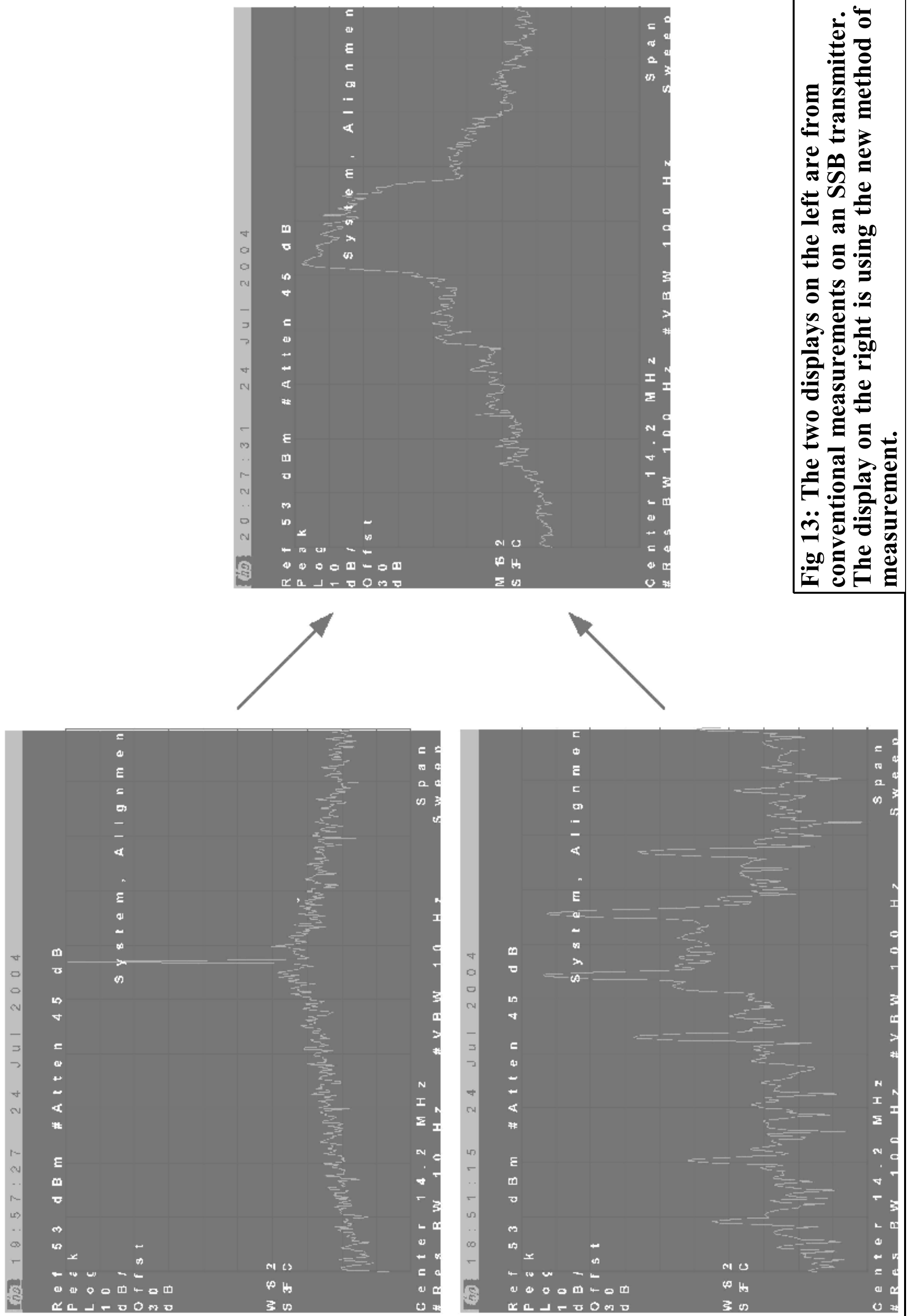

Fig 13: The two displays on the left are from conventional measurements on an SSB transmitter. The display on the right is using the new method of measurement.

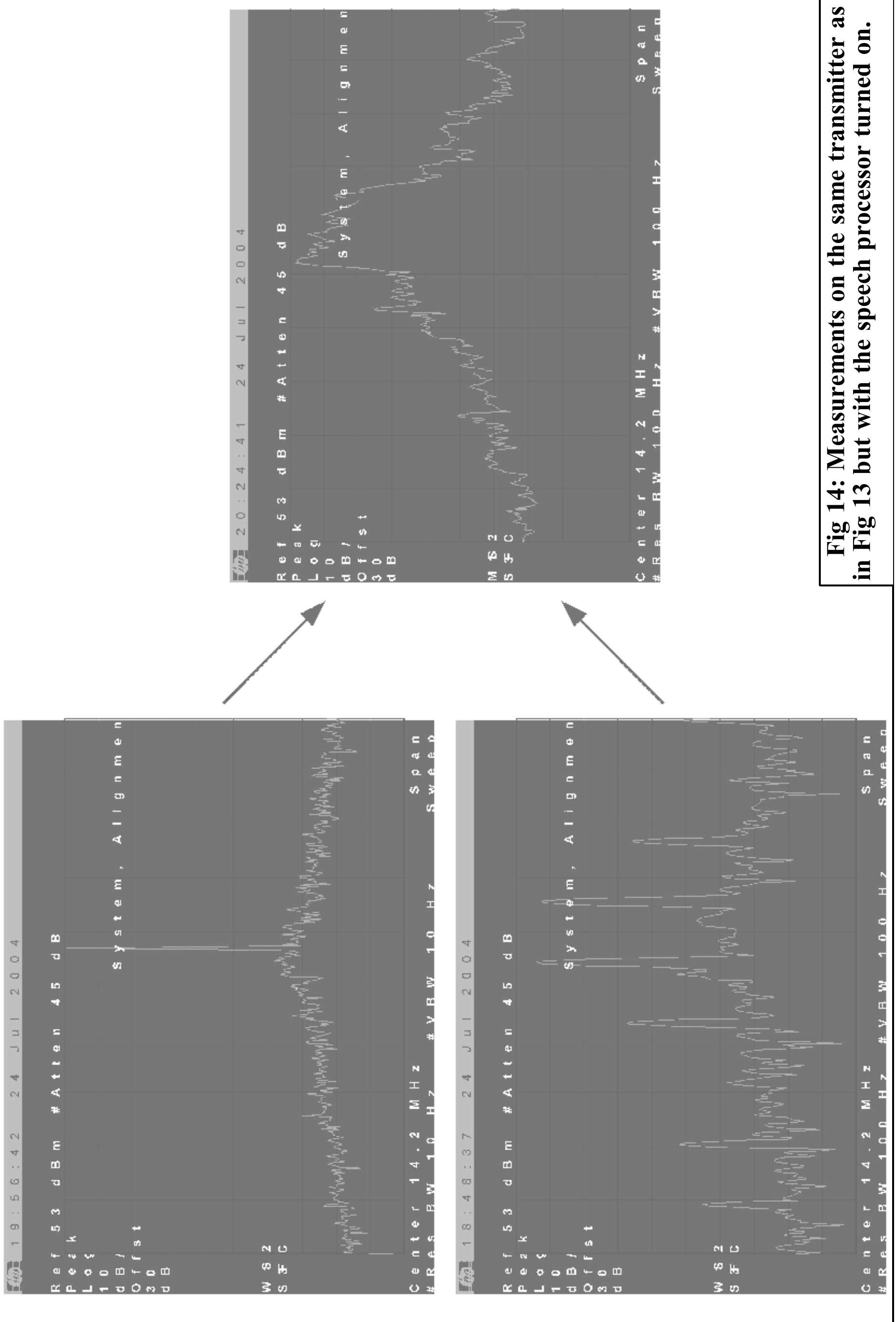

Fig 14: Measurements on the same transmitter as in Fig 13 but with the speech processor turned on.

power of low powered single-sideband transmitters are represented in Fig 15. The effects described in 2.3 are shown here, the test piece is only able to emit the full RF power for a period of a few milliseconds. After that, the amplification falls to lower values.

The signal fed to the amplifier input must be markedly better (purer) than the amplifier itself to ensure that the characteristics of the amplifier under test are measured accurately. If it can be guaranteed that the intermodulation products of the two-tone signal fed to the test piece are at least 18dB [5] higher than the final intermodulation products for the test piece, then the resultant error remains below 1dB. The two-tone signal fed to the test piece in the specific case generates third order intermodulation products of 40dB. No higher order products can be detected.

Fig 15 shows all signal scenarios for both measurement procedures; the input signals to the test piece are on the left and on the right the signals after amplification. Since the settings on the spectrum analyser were retained for all four plots, a direct comparison is possible. The amplification for the power amplifier being tested can be directly read off. The fact that the amplification, in comparison to the highest peaks, is slightly higher for the real modulation test than for two-tone test should cause no surprise. The power actually collapses under continuous two-tone test.

The following sentences were always used for all the measurements carried out in "Max Hold" mode: "CQ, CQ, CQ de Oscar Echo Drei Romeo Alpha Alpha, CQ, CQ 40m..." followed by the NATO phonetic alphabet from "Alpha" to "Zulu" - repeated three times. This takes about 3 minutes from the start of the measurement to the end, producing a good image.

4.3. Repeatability

The repeatability of the new type of measurement method using real modulation can be seen in Fig 16. The spectrograms shown there were taken from the same transmission the spectrum analyser screen was deleted between measurements. There are no detectable variations except in the dB range. Anyone who knows a bit about practical intermodulation measurements on transmitters knows that discrepancies of this order of magnitude are expected. The longer the time taken to build up the image, the better, all of the "worst case data" is detected. Thus a sufficiently long measurement time is a basic pre-condition for results that can be interpreted best.

5. Findings

The highest power level detected in the transmission range during the real modulation procedure in "Max Hold" mode is always found at relatively low frequencies (around 500Hz). The curve, which slopes downwards at higher frequencies, shows that the higher frequencies in normal speech do not have the energy to modulate normal transmitters fully. But this effect is also partly based on the response of the automatic level control (see 2.4).

It is interesting that the real effective bandwidth and/or the undesirable spurious transmissions around the useful channel have decidedly better values than those obtained from the two-tone test would have led us to expect, even when we are working at the very limit of level control, at a point where the typical "splattering" takes place. The measurements displayed verify this impressively.

The curves produced show the spectrum

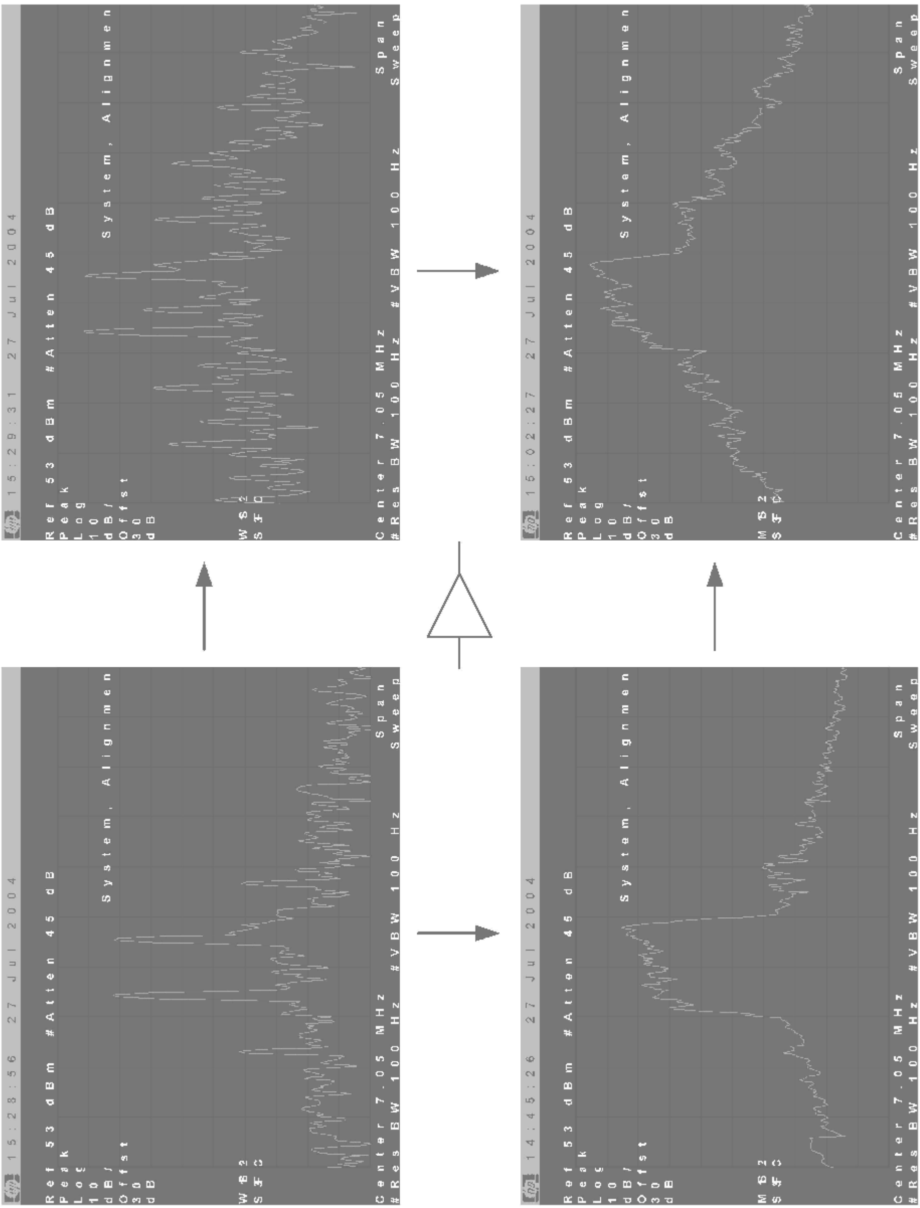

Fig 15: Measurements made on a near perfect SSB exciter shown on the left and on the output of a power amplifier on the right.

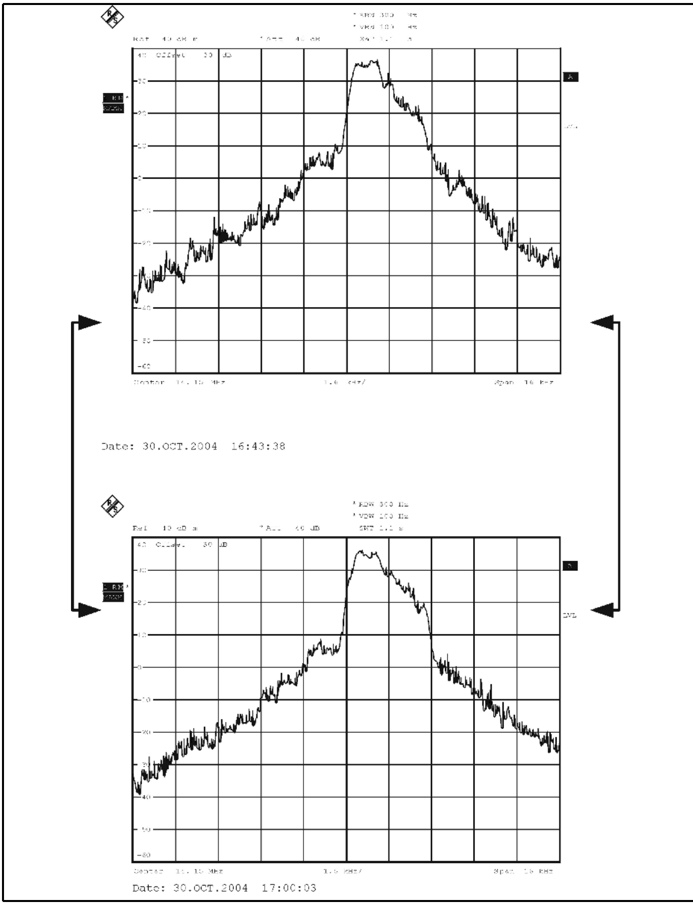

Fig 16: Two displays show the repeatability of the new test method.

clearly, the left hand curve in Fig 15 shows a signal that can be used as a reference because it is nearly perfect.

6. Literature

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Jean-L. Rault F6AGR

Lightning scatter: a faint and rare mode of propagation

Reflections on layers of the ionosphere, reflections on ionised meteorite trails, echoes on airplanes, EME, reflections on auroral ionised clouds … Various modes of radio propagation have been explored and used for a long time by radio amateurs.

Could lightning also be capable of scattering radio waves?

This article deals successively with a theoretical and a practical approach to lightning scatter.

1.0

Theoretical point of view

1.1 Is a flash of lightning able to reflect radio waves?

Any ionised medium is liable to reflect a radio wave. A thunderbolt is a violent electric discharge that heats and ionises the ambient air. Temperatures can reach 20 or 30,000°K and the electron density can raise up to 10^{17} to 10^{18} electrons per cm^3 [2]. Knowing that the electron density necessary to get full radio waves reflection is as follows [4]:

$$N_e = \frac{\pi \cdot m \cdot f_N^2}{e^2} \qquad (1)$$

m and *e* being respectively the mass and the electric charge of an electron and f_N being the frequency of the reflected wave, one can see that a flash of lightning is theoretically able to reflect the entire radio spectrum.

Replacing *m* and *e* by their numerical values and using electrons per cm^3 for N_e, the highest reflected frequency is:

$$f_N = \sqrt{\frac{N_e}{1{,}24 \cdot 10^4}} \qquad (2)$$

with f_N given in MHz and N_e in electrons per cm^3.

So one can see that 3 x 10^9 electrons per cm^3 is a density that is high enough to allow a reflection of VHF or UHF radio waves.

A lightning flash ionised channel can be several kilometres long [6], with a diameter being a few centimetres in size. 80 % of the flashes are of cloud-cloud type (CC, see fig 1), 20% being of the cloud-ground type (CG, see fig 2).

The extremely high temperature gradient affecting the air layer close to electric discharge could possibly participate in

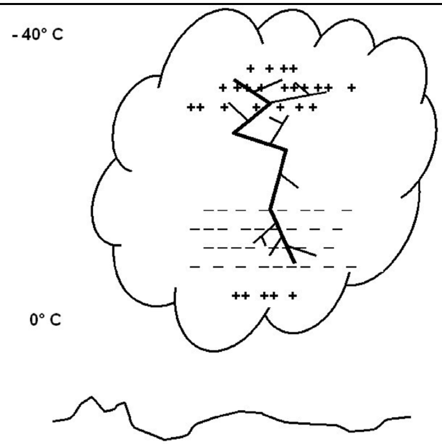

Fig 1: Cloud to Cloud (CC) flash.

the radio waves reflection.

1.2 What could be the effective duration of reflection from a flash of lightning?

A CC or a CG lightning flash is composed of several phases.

At the beginning, low intensity precursors (where electrical current reaches a few hundreds amperes) appear in a highly charged part of a cloud. When a conductive channel is connected between two parts of a cloud with opposite polarities, or between a cloud and the ground, a return stroke appears which carries a huge quantity of electricity (several ten thousands amperes).

A complete lightning flash includes several return strokes and can last several hundreds of milliseconds [2], [3].

1.3 What is the probability of occurrence of lightning flashes?

Most of the 3000 thunderstorms that appear each day around the world occur in the equatorial area.

In Europe, the occurrence is around some tens of electrical activity days per year [1]. See Fig 3 for a map of France showing an example of the yearly statistics. Knowing that a single thunderstorm

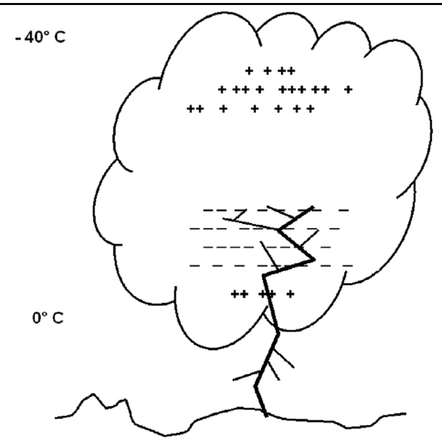

Fig 2: Cloud to ground (CG) thunderbolt.

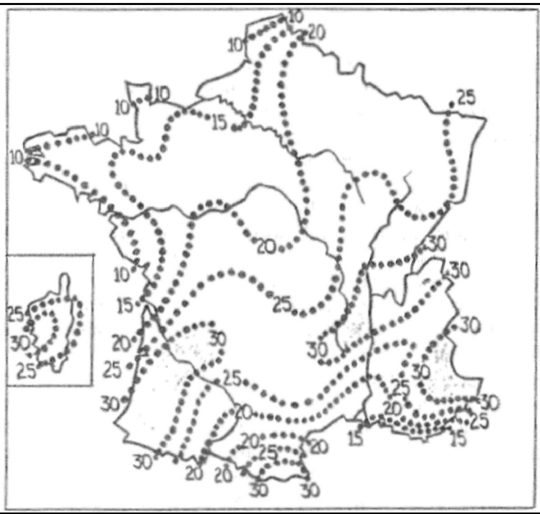

Fig 3: Annual statistics about thunderstorms in France.

generates hundreds or thousands of lightning flashes, one can see that the probability to get some echoes is not negligible.

1.4 What is the maximum distance one can expect for lightning echoes ?

The maximum echo range for a transmitter and a receiver located both at ground level (if we consider they have the same altitude) depends on the mirror altitude.

The range is as follows:

$$D_{max} = 2 \cdot R \cdot \arccos \frac{R}{R+h}$$

R being the Earth radius and h the height of the reflecting part of the lightning flash.

For example, a height of 5000m gives a range of 500km, assuming that the mirror is located half way of the transmitter and receiver, and a little bit more if the atmosphere refraction is taken into account.

2.0

Practical experiment

A reliable test procedure has to be established, to be sure to catch, record and analyse any lightning scatter.

Just listening to distant beacons during a stormy day is too subjective and not convincing enough to prove that lightning scatter really exists.

The following key points were taken into account when establishing the test programme:

- choice of a radio beacon transmitting a stable and well known signal
- absence of interference around the beacon frequency
- distance between beacon and receiver large enough to avoid any reception when there is no ordinary tropospheric propagation
- probability of frequent thunderstorms on the beacon/receiver path

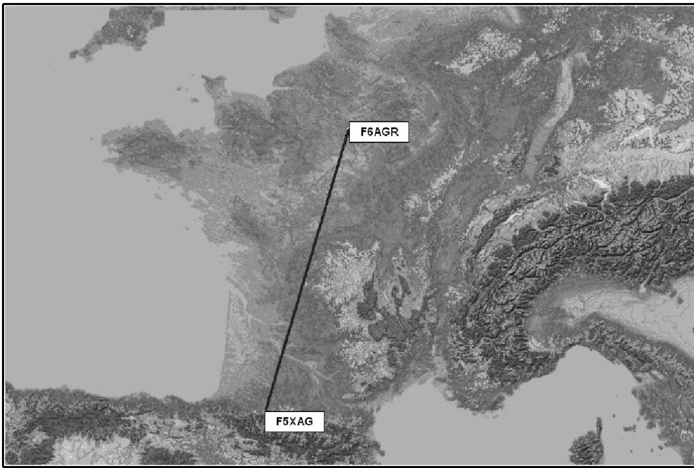

Fig 4: F5XAG to F6AGR path.

- automatic record of receiver audio output to allow further batch analysis
- simultaneous automatic record of radio noise generated by the lightning flashes in order to allow further correlations analysis
- monitoring of thunderstorm predictions and real time activity thanks to weather agencies Internet providers

With all these prerequisites in mind, a campaign of systematic audio records was performed during summer 2004. The radio amateur beacon, F5XAG, was chosen because it fulfils most of the required criteria.

At each end of the 648km path under investigation (see fig 4), the equipment was as follows:

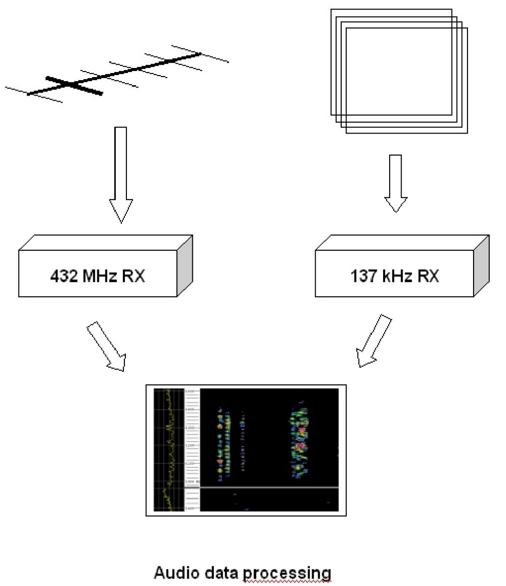

Fig 5:Schematic diagram of experimental equipment.

Fig 6:144MHz and 432MHz aerials.

Beacon:

Name:	F5XAG
Location:	IN93WC
Altitude:	550m
Frequency:	432.413MHz
ERP:	40W
Beam:	NNE

UHF receiver (432MHz):

Name:	F6AGR
Location:	JN18DQ
Ant. altitude	66m
Antenna	2 x 10 element yagi RHCP/LHCP
Receiver	ICOM IC-821H + masthead preamplifier

LF receiver (137kHz):

Name:	F6AGR
Location:	JN18DQ
Ant. altitude	56m
Antenna	23 turns 1.2m^2 square loop
Receiver	ICOM IC-738

Antennas configurations are shown in

Fig 7: 137kHz receiving loop antenna.

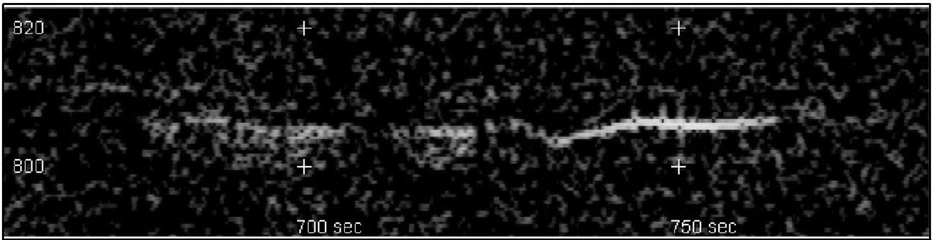

Fig 8: F5XAG long burst of signal.

figs 6 and 7.

The 137kHz amateur band was chosen as the lightning flash monitoring frequency because it gives a good compromise on the detection range of European thunderstorms. Watching flashes on VLF would have given too many pulses generated by very distant thunderstorms. On the other hand, the energy radiated by a thunderbolt in the VHF/UHF band is quite small, and so the range on these bands is limited.

2.1 Receiver/recorder configuration

The audio outputs of the 432MHz and 137kHz receivers are connected to a PC fitted with a stereo sound card (see fig 5). The computer is also used to analyse the records and to track any interesting echoes.

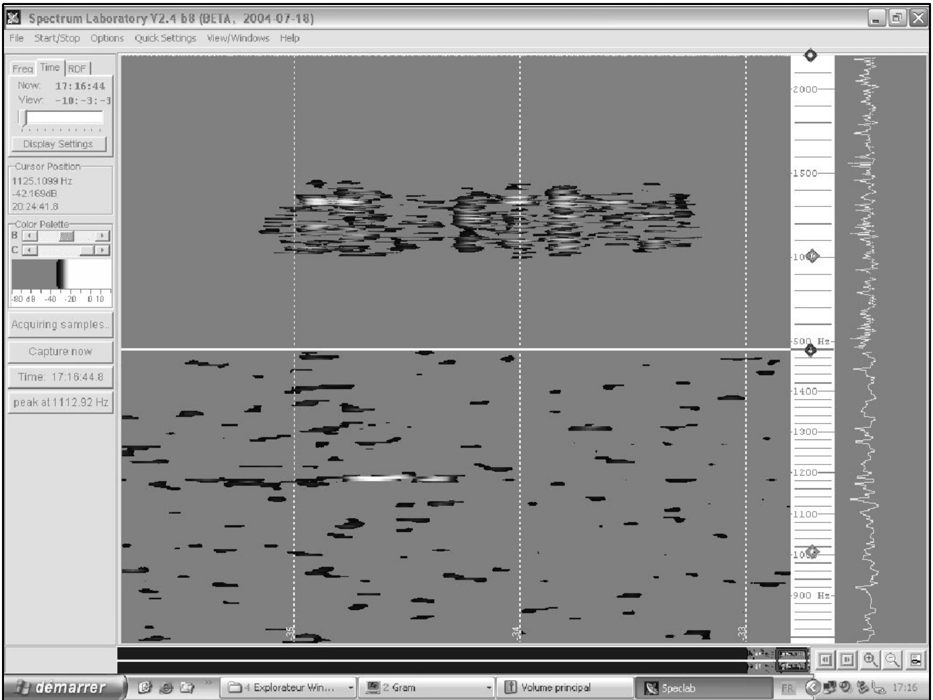

Fig 9: F5XAG lightning scatter.

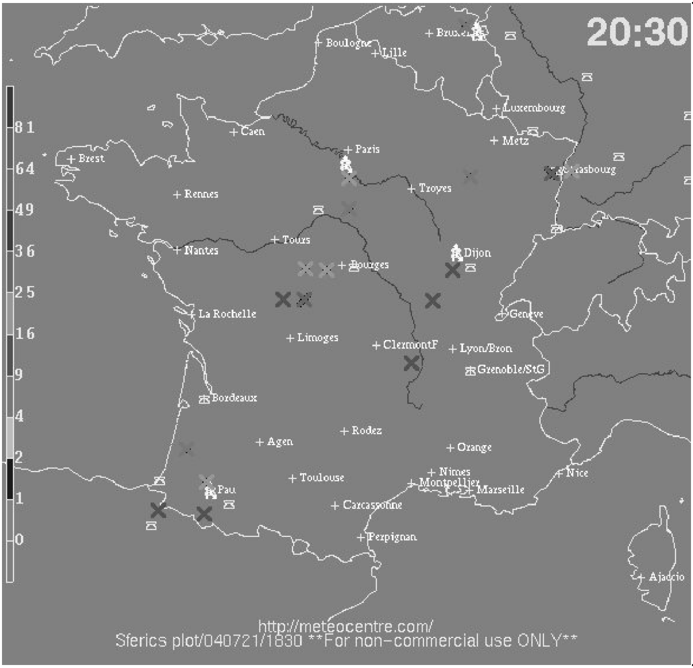

Fig 10: Real time observation of lightning strikes (21 July 2004). ed - looks better in colour, see VHF Communications web site.

The UHF receiver is equipped with an OCXO, which is mandatory to tune the VFO to the correct frequency without receiving any permanent signal (provided that the beacon transmits on the correct frequency).

Both audio channels are recorded in parallel, with 16bit resolution. The sample frequency being adjusted to get the best compromise between the audio pass band and the volume of stored data. For example, one hour of stereo recording with an audio pass band of 4kHz and a 16bit resolution is more than 115Mb.

Recordings are performed in real time on the hard drive of the computer and then stored on 4.7Gb data DVDs. Data compression such as those performed by MP3 algorithms are not usable, because they distort the signal too much, so only WAV recordings offer the necessary record fidelity.

To identify any faint and short echoes, a solution is to use a spectral analysis tool. Although an FFT algorithm is not the best tool to track short pulses, FFT software is very easy to find and to download from Internet.

A graphical display showing frequency on the Y axis, time on the X axis and some colours to give an amplitude indication is very easy and pleasant to examine visually. A quick look is much more effective for identifying a short and faint echo than spending a very long time to listen to white noise.

Two complementary software tools were used for the experiment:

- CoolEdit 2000 from Syntrillium
- Spectrum Lab, developed by Wolfgang Büscher DL4YHF

CoolEdit 2000 is very valuable for juggling with long audio records, offering useful functions such as Rewind, Forward, simple FFT controls, time and

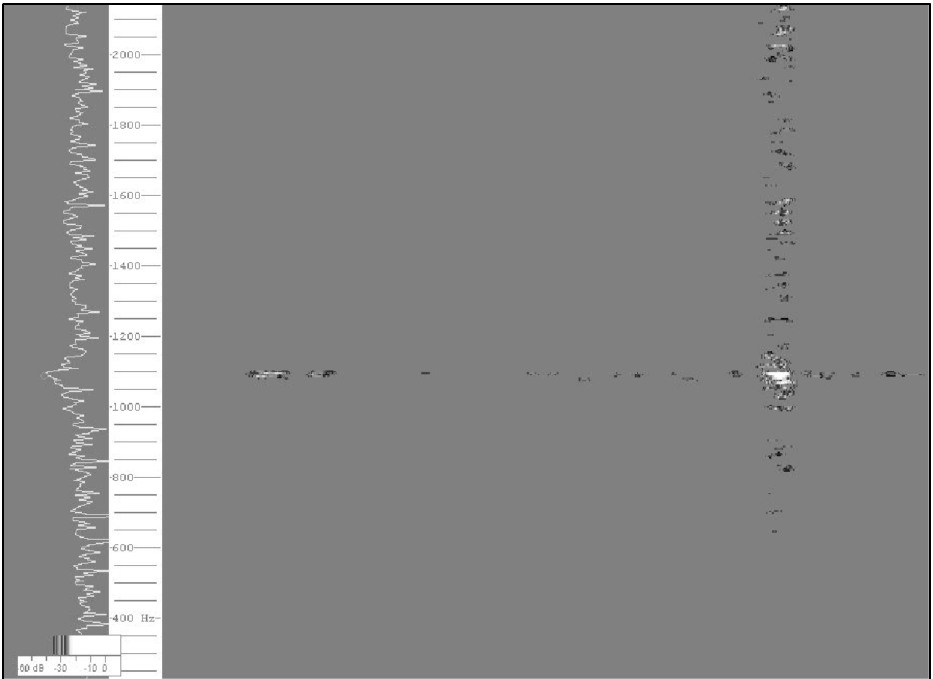

Fig 11: F5XAR lightning scatter.

frequency zooms, all of which are very simple to use.

Spectrum Lab is a powerful spectral analysis tools kit allowing many parameters adjustments. However it requires some knowledge of signal and data processing to be fruitfully controlled.

3.0

Results of the experiment

Several 24 hours-a-day audio records have been performed during calm and thundery days of summer 2004.

The first result is that on this N/S path, bursts of signal coming from the F5XAG beacon were clearly received night and day. Each burst was some tens of seconds long, separated by minutes or tens of minutes of silence. An example of such a burst is shown in fig 8. Some bursts show a typical Doppler effect indicating that the signal is may be reflected from high altitude airplanes, but some others present some frequency splitting and drifts which are not easily explainable.

Several occurrences of lightning scatter on 432MHz were clearly identified in the summer 2004 records. Fig 9 shows an example of such an echo received on July the 21st around 20:30 local time.

The lower trace (432 MHz channel) shows a beacon echo around 1200Hz that goes on for about 500 milliseconds. The signal to noise ratio is about 10dB. No noise at all generated by the lightning

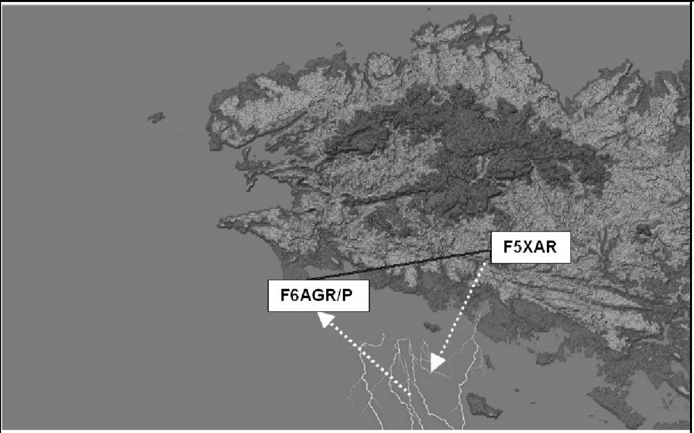

Fig 12: F5XAR to F6AGR/P path.

flash itself was detected. The upper trace shows the corresponding 137kHz activity, which consists in broadband spikes generated by each electrical discharge. Fig 10 shows the real time status of thunderbolt ground hits at 20:30.

During summer 2004, a sea thunderstorm in Brittany allowed another lightning scatter hunt on 144 MHz. The ground path was 82 km long (see fig 12). The conditions were as follows:

Fig 11 shows an example of the echoes which were clearly identified. The horizontal line indicates that the beacon was received most of the time and the vertical line shows the wide band noise received from the flash itself. On the example, the echo was composed of two successive bursts, with a total duration of less than 300mS. The ratio echo/permanent carrier was better than 20dB.

Further to the encouraging results obtained on 144 and 432MHz, old records performed previously in 2002 on 21MHz for a meteor scatter study were re-analysed carefully in order to track any possible lightning scatter on that band. The station used as transmitter was a powerful short waves French broadcast station (Radio-France International) that is very useful to track meteor scatter activity. The 21MHz path was 250km long.

The results were amazing and lightning echoes were identified at a rate of about 6 per minute (see fig 13). The refracted carrier was received permanently and the echo level was 6 to 10dB over the permanent carrier. The length of each echo was a few hundreds milliseconds.

4.0

Conclusion

This lightning scatter experiment shows that radio scattering from thunderbolts really exist.

But many questions remain unanswered… What is the best location and orientation of a lightning flash referred to a transmitting and a receiving station? Are some frequencies better than others? What could be the maximum length of an echo?

Elves and sprites, triggered by powerful positive lightning flashes have been discovered recently, thanks to sensitive video cameras. Could these large lumi-

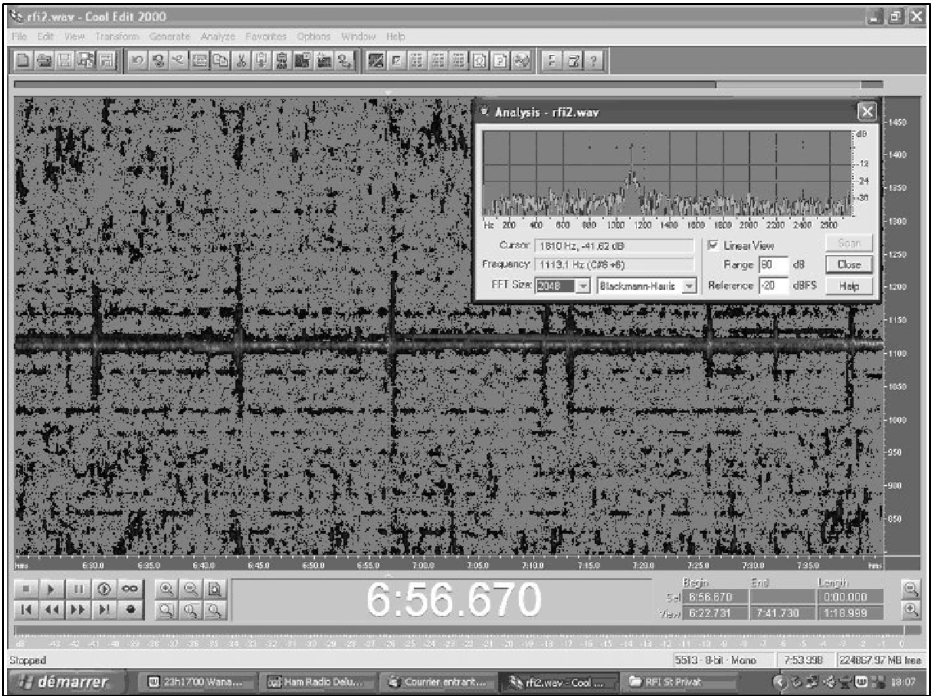

Fig 13: Radio France lightning scatter.

nous discharges happening in the lower part of the ionosphere also contribute to the scattering of radio waves?

That's another interesting story!

5.0

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Ed - Colour pictures from this article can be found on the VHF Communications web site - www.vhfcomm.co.uk

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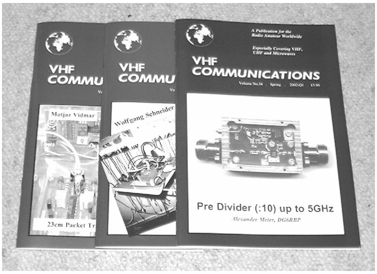

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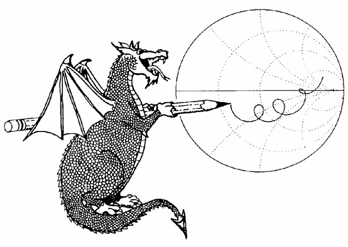

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Eberhard L. Smolka, DB7UP

Designation of microwave bands, specifications and dimensions

You can frequent find interesting, but dubious, waveguide components at radio flea markets. You can't tell, at first sight, whether they can be used for amateur radio bands. The details given in the tables below should be of some assistance. For instance, if you find some WR-90 waveguide, which can be used by radio hams, on X band, it may b labelled with other type numbers e.g. R 100. The waveguides are listed in frequency order.

For example, X band waveguide can be used in the 8.2 to 12.4GHz frequency range. Frequencies below the "cut-off frequency" of 6.56GHz cannot be carried. You will normally have no worries regarding power compatibility from 200kW upwards. The internal dimensions of 2.286cm x 1.016cm can be made use of to specify the waveguide, and the external dimensions of 2.540 x 1.270cm simplify the making of DIY flanges. The attenuation values, usually specified for 100mm, help us to recognise that even quite long distances can be bridged on this band, using waveguides.

Well-priced TV-SAT components for ATV are frequently in the 10GHz amateur radio waveband. WR-75 waveguide equipment is frequently used, it is suitable between 10.0 and 15.0 GHz.

The WR-42 waveguide size has made a breakthrough in the 24GHz amateur radio band. Suitable Gunn oscillators and horn antennas are available.

The WR-19 waveguide is often used in the 47GHz amateur band.

The WR-137 waveguide can be used for the 6cm amateur band. And you can frequently find this kind of waveguide equipment at a flea market.

For TV-SAT reception in the C Band (3.7 to 4.2GHz), the WR-40 waveguide is used. Since this equipment can be made use of from 3.3 to 4.9GHz, these components can also very easily be used in the 9cm amateur band - i.e. from 3,400 to 3,475MHz. The equipment most frequently available with waveguides of this description consists of receive converters and parabolic exciters.

The frequency specifications in the tables should be understood as recommendations. The waveguides can also be used at lower frequencies, provided you stay above the "cut-off frequency". For example, WR-84 stock can still be used at 10.386GHz. However, undesirable transmission modes can develop at frequencies above the recommended level.

The designation of the microwave bands can be very helpful if you find compo-

Band Designation	Frequency range TE_{10} mode (GHz)	Band designation					Material B-brass A-alum. S-silver	JAN[2] flange designation		Waveguide dimensions						Nom. wall thickness mm (in)	Cutoff frequency (GHz)	Theoretical attenuation low to high frequency[3] brass/alum/silver db/100ft	Theoretical attenuation low to high frequency[3] brass/alum/silver db/100m	Theoretical CW power[4] low to high frequency Megawatts (kilowatts)
										Inside			Outside							
		IEC R	EIA WR	British WG	JAN[1] RG	Other Common Usage		Choke UG	Cover UG	Width cm (in)	Height cm (in)	Tol µm (mils)	Width cm (in)	Height cm (in)	Tol µm (mils)					
	1.12-1.70	14	650	6	69 103	L	B A		417B 418B	16.51 (6.5)	8.255 (3.25)	±12.7 (±5)	16.92 (6.66)	8.661 (3.41)	±12.7 (±5)	2.03 (0.080)	0.908	0.412-0.272 0.269-0.178	1.353-0.894 0.883-0.584	11.8-17.1
	1.45-2.20	18	510	7			B A			12.59 (5.10)	6.477 (2.55)	±12.7 (±5)	13.36 (5.26)	6.883 (2.71)	±12.7 (±5)	2.03 (0.080)	1.16	0.574-0.390 0.374-0.255	1.883-1.280 1.229-0.836	7.5-10.6
	1.70-2.60	22	430	8	104 105	LS,R	B A		435B 437B	10.92 (4.3)	5.461 (2.15)	±12.7 (±5)	11.33 (4.46)	5.867 (2.31)	±12.7 (±5)	2.03 (0.080)	1.375	0.759-0.504 0.496-0.329	2.492-1.655 1.626-1.080	5.2-7.5
	2.20-3.30	26	340	9A	112 113		B A		553A 554A	8.636 (3.40)	4.318 (1.70)	±12.7 (±5)	9.042 (3.56)	4.724 (1.86)	±12.7 (±5)	2.03 (0.080)	1.735	1.030-0.716 0.673-0.468	3.382-2.352 2.207-1.535	3.4-4.71
S	2.60–3.95	32	284	10	48 75		B A	54 585A	53 584	7.214 (2.84)	3.404 (1.34)	±12.7 (±5)	7.620 (3.00)	3.810 (1.50)	±12.7 (±5)	2.03 (0.080)	2.080	1.435-0.982 0.937-0.642	4.711-3.225 3.074-2.105	2.18-3.1
	3.30-4.90	40	229	11A			B A		CMR 229	5.817 (2.29)	2.908 (1.14)	±12.7 (±5)	6.142 (2.42)	3.233 (1.273)	±12.7 (±5)	1.63 (0.064)	2.59	1.828-1.296 1.194-0.846	6.002-4.255 3.917-2.777	1.56-2.14
G	3.95-5.85	48	187	12	49 95	C,H	B A	148C 406B	149A 407	4.755 (1.87)	2.215 (0.87)	±12.7 (±5)	5.080 (2.00)	2.540 (1.00)	±12.7 (±5)	1.63 (0.064)	3.16	2.695-1.869 1.760-1.220	8.849-6.134 5.774-4.003	(941-1317)
C	4.90-7.05	58	159	13		C	B A		CMR 159	4.039 (1.59)	2.019 (0.76)	±10.2 (±4)	4.364 (1.17)	2.344 (0.923)	±10.2 (±4)	1.63 (0.064)	3.71	3.091-2.324 2.019-1.518	10.15-7.630 6.622-4.980	(754-983)
J	5.85-8.20	70	137	14	50 10G	XN,C, G	B A	343B 440B	344 441	3.484 (1.37)	1.580 (0.62)	±10.2 (±4)	3.810 (1.50)	1.905 (0.750)	±10.2 (±4)	1.63 (0.064)	4.29	3.821-3.018 2.496-1.971	12.54-9.907 8.187-6.465	(554-896)
H	7.05-10.00	84	112	15	51 68	XB,W	B A	52B 137B	51 138	2.850 (1.12)	1.262 (0.50)	±10.2 (±4)	3.175 (1.25)	1.588 (0.625)	±10.2 (±4)	1.63 (0.064)	5.26	5.355-4.161 3.497-2.717	17.58-13.66 11.47-8.913	(355-454)
	7.00-11.00		102				B A			2.591 (1.02)	1.295 (0.51)	±7.6 (±3)	2.845 (1.12)	1.549 (0.61)	±7.6 (±3)	1.27 (0.050)	6.50	6.939-4.360 4.532-2.848	22.78-14.31 14.87-9.34	(280-424)
X	8.20-12.40	100	90	16	52 67		B A	40B 136B	39 135	2.286 (0.90)	1.016 (0.40)	±7.6 (±3)	2.540 (1.00)	1.270 (0.50)	±7.6 (±3)	1.27 (0.050)	6.56	8.362-5.784 5.461-3.778	27.45-18.99 17.91-12.39	(206-293)

Table 1: Waveguide dimensions and characteristics from 1 to 12.4GHz.

M	10.0-15.0	120	75	17			B A			1.905 (0.75)	0.953 (0.38)	±7.6 (±3)	2.159 (0.85)	1.207 (0.475)	±7.6 (±3)	1.27 (0.050)	7.88	9.893-6.909 6.461-4.512	32.48-22.68 21.19-14.80	(166-229)
P	12.4-18.0	140	62	18	91 107	KU,Y, U	B A S	541A	419	1.580 (0.62)	0.783 (0.31)	±6.4 (±2.5)	1.783 (0.70)	0.933 (0.39)	±7.6 (±3)	1.02 (0.040)	9.49	12.46-0.162 8.141-5.984 6.165-4.531	40.92-30.08 26.71-19.63 20.22-14.87	(119-157)
N	15.0-22.0	180	51	19			B A S			1.295 (0.51)	0.648 (0.25)	±6.4 (±2.5)	1.499 (0.59)	0.851 (0.33)	±7.6 (±3)	1.02 (0.040)	11.6	17.02-12.33 11.12-8.054 8.418-6.099	55.88-40.49 36.47-26.42 27.62-20.01	(79-106)
K	18.0-26.5	220	42	20	53 121 66		B A S	596A 598A	595 (425)[5] 597	1.067 (0.42)	0.432 (0.17)	±5.1 (±2)	1.270 (0.50)	0.635 (0.25)	±7.6 (±3)	1.02 (0.040)	14.1	26.66-19.58 17.41-12.79 13.18-9.684	87.51-64.28 57.11-41.95 43.25-31.77	(43-58)
	22.0- 33.0	260	34	21			B A S		 1530	0.864 (0.34)	0.432 (0.17)	±5.1 (±2)	1.067 (0.42)	0.635 (0.25)	±7.6 (±3)	1.02 (0.040)	17.3	32.58-22.66 21.27-14.80 16.11-11.20	106.94-74.37 69.79-48.54 52.86-36.76	(34-47)
R	26.5-40.0	320	28	22	 96	V,KA, U	B A S	 600A	599 (381)[5]	0.711 (0.28)	0.356 (0.14)	±3.8 (±1.5)	0.914 (0.36)	0.559 (0.22)	±5.1 (±2)	1.02 (0.040)	21.1	44.29-30.33 28.92-19.81 21.9-15.00	145.4-99.57 94.88-64.98 71.85-49.21	(23-32)
	33.0-50.0	400	22	23	 97	Q	B S		383	0.569 (0.22)	0.284 (0.11)	±2.5 (±1)	0.772 (0.30)	0.488 (0.192)	±5.1 (±2)	1.02 (0.040)	26.35	30.84-20.96	101.2-67.78	(14-20)
	40.0-60.0	500	19	24			B S			0.478 (0.18)	0.2388 (0.09)	±2.5 (±1)	0.681 (0.27)	0.442 (0.174)	±5.1 (±2)	1.02 (0.040)	31.4	38.79-27.21	127.3-89.25	(10-14)
V	50.0-75.0	620	15	25	 98	M	B S		385	0.376 (0.15)	0.188 (0.07)	±2.5 (±1)	0.579 (0.22)	0.391 (0.15)	±5.1 (±2)	1.02 (0.040)	39.9	57.30-39.15	188.0-128.4	(6-9)
	60.0-90.0	740	12	26	 99	E	B S		387	0.309 (0.12)	0.1549 (0.06)	±1.3 (±0.5)	0.513 (0.20)	0.358 (0.14)	±5.1 (±2)	1.02 (0.040)	48.4	78.33-52.51	253.9-172.3	(4-6)
	75.0-110.0	900	10	27						0.254 (0.10)	0.127 (0.05)	±1.3 (±0.5)	0.457 (0.18)	0.330 (0.13)	±5.1 (±2)	1.02 (0.040)	59.0	100.5-70.71	329.7-231.9	(3-4)

Abreviations:
IEC – International Electrotechnical Commision
JAN – Joint Army Navy
EIA – Electronic Industrial Association

[1] For more information refer to US military specification MIL-W-85 waveguide rigid rectangular.
[2] For more information refer to US military specification MIL-F-3922 flanges waveguide cover.
[3] Attenuation computations rectangular guide TE_{10} mode. Resistivities; Brass: (65-35) 6.63 x 10^6 Ω/cm; Aluminium: 2.83 x 10^6 Ω/cm; Silver: 1.62 x 10^6 Ω/cm.
[4] CW power computations, breakdown strength of air taken at 15,000 volts per centimetre, safety factor of approximately 2 at sea level is assumed.
[5] HP instrumentation flanges mate with rectangular cover flanges noted. Flange adapters are available to mate with UG-425 U and UG-381 U, Specify 11515A (K band 18.0 – 26.5GHz) or 11516A (R band 26.5 – 40.0GHz).

Table 2: Waveguide dimensions and characteristics from 10 to 110GHz.

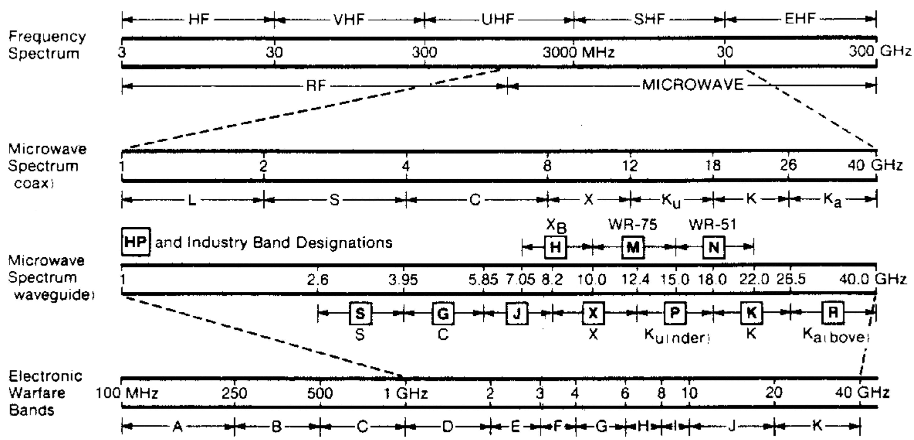

Table 3: Designation and position of microwave bands in the GHz spectrum.

nents where only the “band” is specified.

The dimensions specified in the tables given should make it easier to identify the frequency ranges of the various components obtainable from radio flea markets.

Gunthard Kraus, DG8GB

Internet Treasure Trove

Radiall-Larsen

If you need to refresh your basic knowledge of antennas, or you want to gen up on the problems and concepts associated with modern WLAN antennas, or to look for frequency occupancy tables, or to examine this, that or the other, you can look at problems more closely by visiting the "Technical Reference" section of this homepage

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Should purchasers be worried about any legal problems here?

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Address:
http://www.fcc.gov/mb/audio/bickel/findvalues.html

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Address:
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Address:
http://www.jiwire.com/glossary.htm?id=52

Martindale's Calculators On-Line Centre

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Address:
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Email: andy.@vhfcomm.co.uk

Editor: Andy Barter G8ATD

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BELGIUM - UKW-BERICHTE, POB 80, D-91081 BAIERSDORF, Germany. Tel: 09133-77980. Fax: 09133-779833 . Postgiro Nbg. 30445-858.

DENMARK - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051. Fax: +44 1582 581051. Email: andy@vhfcomm.co.uk

FRANCE - Christiane Michel F5SM, Les Pillets, 89240 PARLY, France
Fax: (33) 03 86 44 08 82 Tel: (33) 03 86 44 06 91

FINLAND - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051. Fax: +44 1582 581051. Email: andy@vhfcomm.co.uk

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HOLLAND - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051. Fax: +44 1582 581051. Email: andy@vhfcomm.co.uk

ITALY - R.F. Elettronica di Rota Franco, Via Dante 5 - 20030 Senago, MI, Italy. Fax 0299 48 92 76 Tel. 02 99 48 75 15 Email: info@rfmicrowave.it Web: www.rfmicrowave.it

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NORWAY - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: +46 40 16 42 66. Fax: +46 40 15 05 07. GSM: 0705 16 42 66
Email: waveline@algonet.se Web: www.algonet.se/~waveline

SOUTH AFRICA - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051. Fax: +44 1582 581051. Email: andy@vhfcomm.co.uk

SPAIN & PORTUGAL - JULIO A. PRIETO ALONSO EA4CJ, Donoso Cortes 58 5° -B, MADRID 15, Spain. Tel: 543 83 84

SWEDEN - WAVELINE AB, Box 60224, S-216 09 MALMÖ, Sweden. Tel: +46 40 16 42 66. Fax: +46 40 15 05 07
Email: waveline@algonet.se Web: www.algonet.se/~waveline

SWITZERLAND - KM PUBLICATIONS , 63 Ringwood Road, LUTON, LU2 7BG, UK. Tel: +44 1582 581051. Fax: +44 1582 581051. Email: andy@vhfcomm.co.uk

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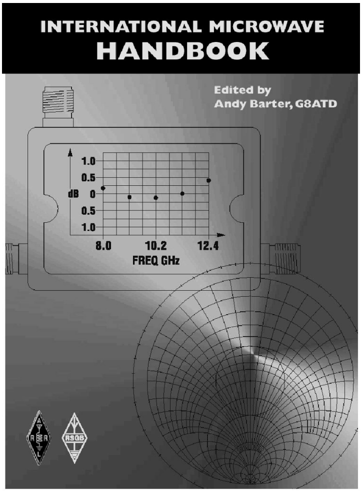
The International Microwave Handbook



The microwave bands are an excellent area for radio amateurs who want to experiment and construct their own equipment. The RSGB in partnership with the ARRL has produced this invaluable source of reference information for those interested in this area, along with excellent designs from around the world to fire the imagination. Material has been drawn from many sources including the RSGB journal RadCom and the ARRL publications QST & QEX. Alongside this material a truly international range of sources have been used including items from Germany, Denmark, New Zealand, Slovenian and many more.

The earlier chapters of the book provide invaluable reference material required by all interested in this exciting area of experimentation. Techniques and devices are covered in depth, leading the reader to understand better the wide range of equipment and techniques now available to the microwave experimenter. This book contains a wide selection of designs using the latest technology that can reasonably be used by radio amateurs and ranges from ones that can be reproduced by most radio amateurs to those that require a high degree of skill to make.

With the explosion in consumer electronics using microwave frequencies the opportunity to experiment has never been greater and this book is simply the best guide to the area of microwave radio.

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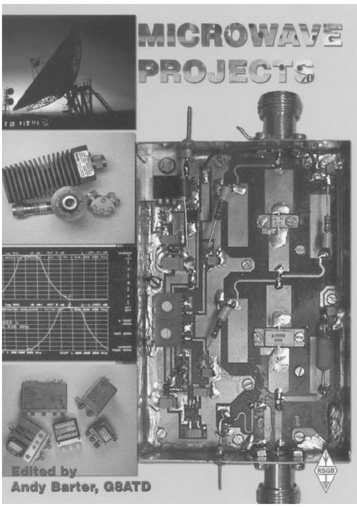
Microwave Projects



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