VHF
COMMUNICATIONS

White Box story - what make of PA?
Jean-François Iosca, F1LVO

Contents

I was very sad to hear that Norman Fitch, the editor of the VHF column in RadCom died on 29th January 2010. It is particularly sad because the first article in this magazine was suggested by Norman. He will be missed.

The response to my call for articles last year was very successful. So successful that I cannot include all the articles in one magazine. I have decided to publish them in the order that I received them, the first 4 are in this magazine. Thank you very much to all the authors who sent articles. I am sure that they will make the magazines more enjoyable.

Finally look out for the new RSGB book that I have edited, see page 60.

73s - Andy

Brian D Williams, GW0GHF

Connecting up for VHF digital modes

1.

Introduction

VHF in this case refers to the 6m, 4m and 2m bands. Most of the handbooks dealing with digital modes do not cover VHF transceivers. Here is the first obstacle for the operator who wishes to use a digital mode on VHF. These handbooks refer to HF transceivers that usually have "accessory" sockets provided at the rear of the rig for connection to a computer via an interface. However, many VHF / UHF rigs do not have these sockets. The purpose of this article is to help those with rigs that do not have "accessory" sockets.

In the author's case the rig is the Icom 821 that does have an accessory socket. This socket is of a different type to the author's HF rig (Icom 756) and not interchangeable.

The microphone socket is standard on both rigs so it was decided to use these sockets instead of the "accessory" sockets.

There must be many people who have rigs where the accessory sockets are not interchangeable and where it is more convenient to connect the interface leads to a front mounted socket than having to get round to the rear of it to connect up.

This method is not the best way. However it is more convenient and works fine provided certain criteria are met. The biggest problem can be RF from the transmitter or aerial getting back into the sound card of the computer.

It is possible to connect directly from computer to rig and it may work satisfactorily, especially if only QRP power is used. However it is always better to connect via an isolator. This will isolate the common earth from the rig and computer frames. Isolators can be bought or you can easily make your own. The "RigBlaster"[1] is one commercial unit but with a few "clamp-on" ferrite rings or other form of inductance e.g. an old ferrite aerial rod, you can do it yourself. The author's interface is housed in a 4 x 4 x 2 inch black plastic box. A piece of blank copper clad printed circuit board is cut into two and the two parts are glued into the base of the box so that there is a gap between the edges of the two pieces – see Fig 1. One piece is marked "CPTR" (computer) and the other "RIG" so that you know which end goes where. Make a ¼ inch hole at each end of the box. These take your cables. Obtain some good quality twin screened microphone cable.

Measure the distance between your computer and your rig, and then multiply it by four, add another 12 inches. This is the total length of cable you will need. Decide where you are going to place
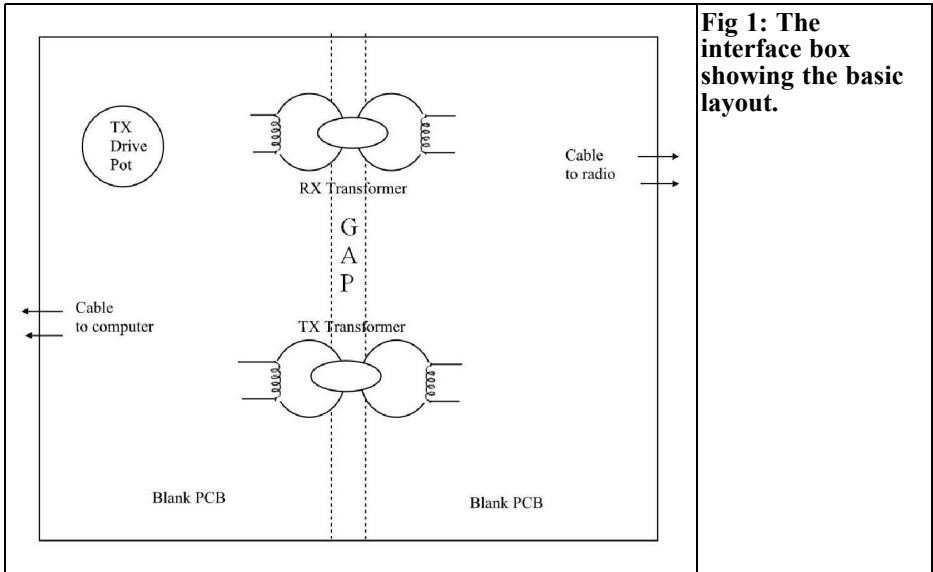
Fig 1: The interface box showing the basic layout.

your interface box and cut the cables accordingly. The box contains a TX drive knob that must be accessible to you operating position. This is a very desirable feature for operating on the digital modes. You want to be able to control your sound card TX output level instantly so that the sound card does not over drive the input of your rig's microphone socket. You cannot do this with your "Mic gain pot" on the front of your rig. The pot on the rig only controls the level of audio coming OUT of the microphone preamplifier.

2.

Drive levels.

It is MOST important that the microphone preamplifier inside your rig is not overdriven by the output of the computer sound card. The evidence on the waterfall display of an overdriven signal is there for all to see when you look at some of the stations that are transmitting. The signal becomes distorted and pro-

duces "sidebands" that can make the signal very wide; most anti-social. In extreme cases a PSK31 signal, normally about 35Hz wide can be 350Hz wide. A TX drive pot mounted in the interface box can easily control the TX drive. During a QSO with another station you do not want to bother with adjusting "Volume Control" levels in the computer program that require several mouse clicks. Later we will see how to make a correct adjustment of the drive level when transmitting.

3.

Constructing the interface box

Fig 2 show the two ferrite transformers and the TX drive pot. The transformers are wound with a thin enamelled copper wire; 26swg is ideal. Obtain two "clamp-on" [2] chokes, the sort used to reduce RFI on connecting leads. These are ideal as the core of your isolating transformers. Unclip the plastic retainer so that the cores divide into two halves joined at the
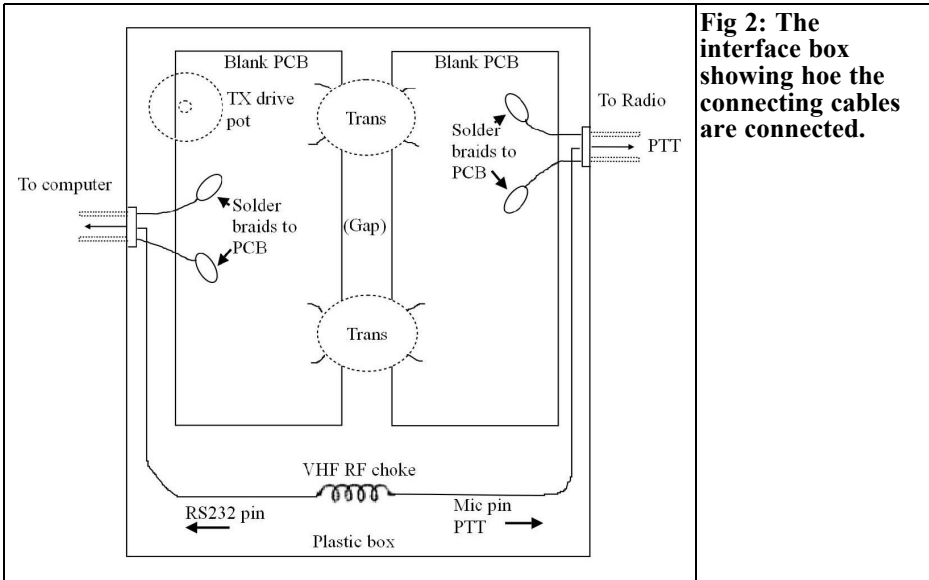
3

Fig 2: The interface box showing hoe the connecting cables are connected.

hinge. Wind as much wire on to each split core as possible so that you end up with primary and secondary transformer windings. Do not over fill the halves or there will not be enough room to push the cores back together. There should be a gap of a few millimetres between primary and secondary so that capacitive coupling is minimised.

When complete you have a nice isolator that has very little external magnetic field, the core being circular. There will be minimal capacitive coupling between input and output.

The small inductance of the windings will help to attenuate the lower audio frequencies yet be enough so that transfer loss is fairly low. The loss is approximately 15dB at 1000Hz and is not important. Do not be tempted to use the small audio inter-stage transformers found in the older portable radios. The capacitive coupling between the windings is too great and the isolation of your interface box will be poor.

Be sure to solder the braids of each cable onto the PCB as shown in Fig 2. Do not regard the audio return path as being through the braid. This mist be a separate path, hence the use of twin screened wire as distinct from wire and braid. This is where the RF currents from the transmitter could flow, so there must be no audio path through the braid screen. There should be no "through-route" of the braiding from rig to computer. The break is at the interface box. No external "earth bonding" can be used between the two. Also check that any power supply negative wires are not common to both rig and computer chassis frames. The TX drive pot can be any value in the range 500 – 5000Ω. If higher, solder a 2k2Ω resistor across the outside tags. This will make it less fussy.

On many computers the audio input / output jacks are 3.5mm and the signal path has one side grounded. DO NOT rely on the contact of the plug shell to the socket. Solder a braid from the screened cable (plus signal return path) directly to the tinplate of the sound card holder. Poor mechanical joints can cause problems and braids that appear earthed that might well be doing anything but screening the cable.
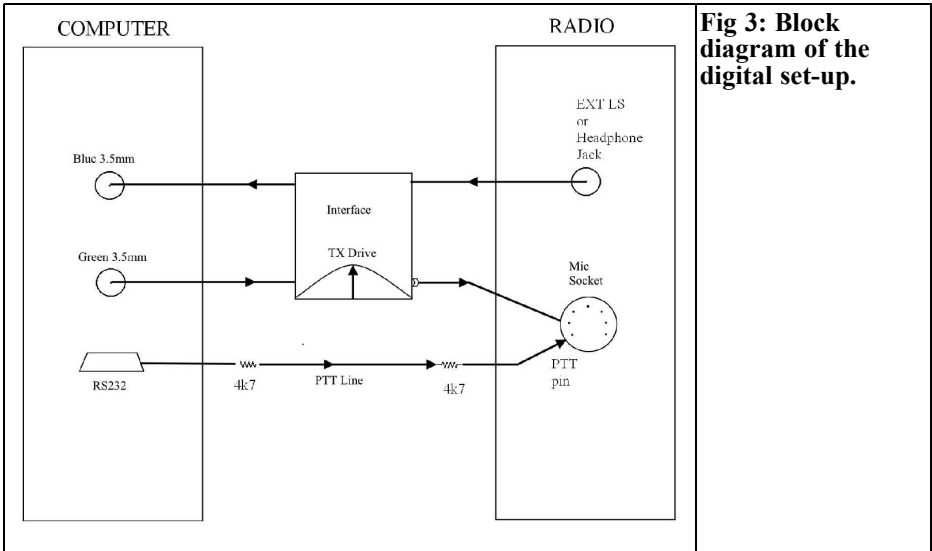
Fig 3: Block diagram of the digital set-up.

4.

Connecting your computer to your rig

Figs 2 and 4 shows the connections to make to your computer at one end and your rig at the other end. It is not proposed to go into any detail about how to operate the computer program, this can be found in the many handbooks on the subject. However as indicated earlier, it is useful to check that Rx and TX levels are the best that can be achieved:

RX

This is a matter of adjusting the output audio from the radio to the correct level for the computer. This is done using the volume control panel on the computer and by adjusting the audio volume control of the rig so that it is at the level usually used for listening to SSB. On some rigs it is possible to withdraw the jack slightly from the external headphone socket and hear the signals as well as connecting them through. On most digital mode programs there is an indication of signal strength somewhere on the water-fall display window that is a good guide for setting up.

TX

It is as well to bear in mind that inputs into the microphone socket of the rig have to be potted down considerably. This is the most usual cause of an over-driven rig. After all, the microphone socket is designed for a very small audio signal from a microphone. To adjust your TX drive level:

- Select PSK31

- Fit a dummy load to the aerial socket

- Turn your TX pot on the interface box to zero

- Check that the volume control panel in the computer program has the TX slider control un-muted with the slider set at about 75%

- Adjust the microphone gain on your rig to the setting normally used for SSB
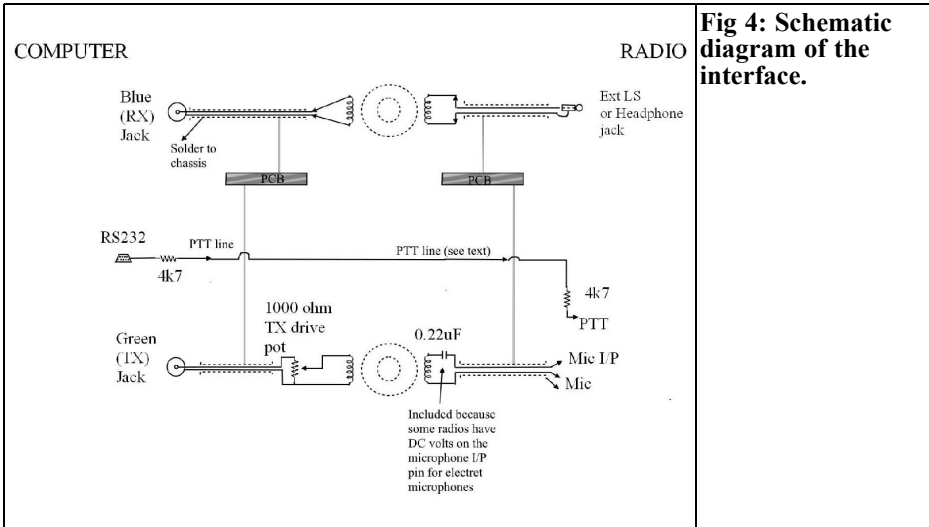
Fig 4: Schematic diagram of the interface.

- Set the power output of the rig to approximately 25% of full power. If you have an ALC meter on the rig it will be useful to indicate the correct TX drive level from the sound card

- Click "send" or "TX" on your computer program

There should not be any power output until you advance your interface box drive pot. Advance it slowly until you get a reading on your power meter. Keep advancing and the ALC meter should start to read. Back off the drive pot until there is no ALC reading. This is the correct drive level. If you do not have an ALC meter, watch the power output meter as you advance your TX drive pot. It should increase as you advance the pot. A point will be reached when the power meter will not show much increase when you keep advancing the TX drive pot. This is the point when more than adequate TX drive to the microphone preamplifier of your rig may cause overdriving. Find the point where backing off the pot starts to reduce the power. This is the correct level.

When operating, you will find that transmitting near the lower end of the water-fall spectrum scope will need more TX drive than at the centre. This is when the interface box drive pot is useful. The alternative is to move the tuning dial on your rig but this is an unnecessary nuisance.

It is advisable to fit "RFI" chokes to the interface cables. This will help to reduce any RF fields from the aerial or feed-line coax getting into the audio path. The ideal way is to pass the cable through two or three ferrite beads before soldering the cables in place. Fit two beads on the cable close to the microphone input socket of your rig. Also fit beads onto the cable that goes into the 3.5mm (blue) jack on your computer. At VHF these chokes will be more effective than they are at HF. Pay particular attention to the input cable plug that goes to the blue socket on your computer. The twin wires must be soldered to the plug. Do not be tempted to use "quick-fix" plugs. Also the braid must be well connected to the frame of the sound car (usually it is tinplate and will solder nicely). Do not rely on the friction contact of the 3.5mm jacks as they can and do corrode.

Now that the signals are catered for through the interface box we have to control the PTT start and stop. Individual

operators will already have their own way of doing this, but a word of caution will not come amiss regarding the "hard-wired" method of auto-transmit from the computer program. This involves a connection from a pin in the RS232 socket to the rig PTT line, in our case via the microphone socket. This hard wired method must NOT break the isolation from rig to computer. The way to do this is to fit a 4k7Ω resistor to the wire at the RS232 port and at the rig end, fit another resistor at the pin of the microphone input socket. If unsure of the correct pin-outs on your rig then perhaps the best way is to use the "VOX" facility. This will turn on your transmitter when the microphone input receives audio signals from the sound card.

5.

Operating digital modes

When operating on SSB the RF output power of the transmitter can be anything up to 400 watts. In most cases, 50 to 100 watts is usual. On digital modes the usual power is QRP (perhaps 2 watts) up to around 45 watts. The digital modes use much narrower bandwidth and more stations can be accommodated in the same band space that an SSB signal will use. This can only be good news.

Enjoy your operating on the digital modes. It will save your voice and be more sociable than a QRO SSB signal. There are a lot of advantages compared to SSB voice contacts. QRP working is a delight and weak signals from the other side of the world are far easier to work. SSB cannot compete with the efficiency of the digital signal modes.

6.

The Author's experience with software

I asked Brian to make some comments on software, he has limited experience but the comments will help anyone starting to use digital modes – Andy

The only software that Brian has used is Flidgi and HRD DM780 [3] plus EasyPal [4] for digital ATV and MMSSTV [5] for analogue ATV.

Brian's experience of EasyPal is that it is far from easy. However when it is working properly the picture are really excellent The downside are:

- Over-air signals have to be fairly strong and static crashes or similar QRN play havoc with decoding a picture. EasyPal is best used on VHF/UHF provided the RF signals are strong enough. It is not a mode for chasing DX stations.

- If you miss the start of a transmission you cannot "catch up" and the result is failed decode and therefore no picture.

- There is not adequate monitoring of what you are transmitting. Only the picture you have loaded is shown as it is being transmitted. Your callsign is "embedded" in the encoding but is not shown on the spectrograph, neither is any text which may have been added to the picture.

MMSSTV is analogue but worth mentioning. It is a very flexible program, easy to use, easy to add/delete or make changes to the text that is to be sent. It will decode weak signals and good quality pictures are commonplace. Picture and added text being sent over air are monitored on the picture frame.

6.1 Digital Master 780, text modes

There are 32 different modes for encoding the text you want to send. The most widely used is PSK31. This mode is available on Windows, Linux or Mac as are many others. This and the faster PSK63 are not found much on the VHF bands. The HF bands are full of PSK31. Tune to 14.070MHz USB for example or in the evenings on 3.580MHz. VHF operators have yet to discover the use of PSK31 or any other modes like MFSK31 where it is possible to send a small picture in addition to your text. JT65 or FSK441 seem to be used but these modes are more suitable for MS or EME. For everyday QSOs over land and sea paths, PSK31/63, MFSK16 or Olivia modes do not need the QRO of SSB and 50 watts is the upper power limit used, or desirable, for digital text QSOs. The author has carried out experiments with local and distant 2m band stations using these modes. Normally using SSB with its 2.4kHz bandwidth, DX signals can be hard to copy. However, using a narrow band mode, such as one of the above it is possible to easily copy a weak signal from a distant station. Some modes, such as "Olivia", have built-in error correction and there are several good books explaining the attributes of each mode most suitable for the signal path conditions prevailing [6]. Successful QSOs can be completed using, for example, 25 watts and a vertical "white stick" aerial although horizontal aerials are ALWAYS better because of the lower QRN from horizontal polarization compared to vertical polarization. They also give more gain in the desired direction. However, ther is the inconvenience that they have to be rotated to the desired direction. The receiver IF pass band can be narrowed right down and this has the significant advantage of preventing AGC action reducing the gain of the receiver when strong unwanted signal would be in the pass band. This is the most important consideration when trying to receive a weak signal.

The current UK bandplan gives 144.138MHz as the "centre of activity" for PSK31. Calling CQ on this frequency does not often bring a reply! A better way is to call CQ on the SSB calling frequency (144.300MHz) saying that you are looking for a digital mode contact. On the 6m and 4m bands use 50.160MHz and 70.220MHz respectively. When the summer sporadic E season arrives, digital mode stations can be found on or around 50.250MHz. Have fun!

7.

References

[1] RigBlaster from West Mountain Radio, 34 Smith Street, Norwalk, CT 06851, USA - http://www.west mountainradio.com/RIGblaster.htm

[2] Transformers: Obtain the largest clamp-on Ferrites; OD is 250mm, thickness approximately 160mm, and ID approximately 12mm. They can be obtained from Sycom; Tel 01371 372587 (Leatherhead) or Tel 01202 490099 (Christchurch) -
http://www.sycomcomp.co.uk/ they also advertise in RadCom.
The author's transformers have approximately 65 turns of 26swg for each winding.

[3] Ham Radio Deluxe -
http://www.ham-radio-deluxe.com/

[4] EasyPal written by Erik Sundstrup, VK4AES http://vk4aes.com/

[5] MMSSTV -
http://mmhamsoft.amateur-radio.ca/mmsstv/

[6] Books on systems and programs:

- Digital Modes for all Occasions, Maurice Greenman, RSGB

- HF Digital Handbook, Steve Ford, ARRL (available fro RSGB)

Andy Barter, G8ATD

HenryTest review

1.0

Introduction

There is a new product advertisement in the magazine for an inductance meter called HenryTest. When I was originally contacted about an advert for this product I thought it looked well packaged but why would anyone want such an inductance meter? There are quite a few multimeters available that have inductance ranges amongst many other ranges. Generally their lowest range is a 0 – 2mH range, which is useful but not so useful for VHF and UHF constructors. The 1nF to 1H range of the HenryTest is much

Fig 1: HenryTest in its blister pack.

more useful for measuring coils for this frequency range.

Antonio Temil, the director of OOO Zarina who have produced HenryTest offered to send me a sample meter to test.

2.0

First Impression

HenryTest arrived well packed in a blister pack (Fig 1) but I am please to say it was easy to open and not of the type that need a large pair of scissors to open. The pack contained:

- The HenryTest
- The software on an 8cm CD
- A USB cable

The HenryTest is well packaged in a attractive shiny case approximately 70 x 30 x 25mm with a Type B USB socket at one end and two slots at the other end to take the two connections from the inductor to be measured. Spring-loaded jaws that are opened by pressing the large red button on the top of the HenryTest make the connections.

The software is loaded before the HenryTest is plugged in for the first time. It was easy to load from the installation file on the CD.

Plugging in the HenryTest for the first time gave the normal "New device found" message from Windows (I was using a Windows XP laptop). The message seemed to take a long time to
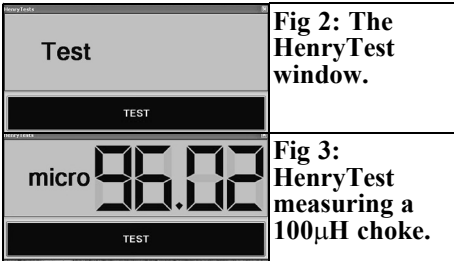
Fig 2: The HenryTest window.

Fig 3: HenryTest measuring a 100μH choke.

Fig 5: Measuring a 2 turn coil.

disappear but as soon as it had the HenryTest was ready to use.

Clicking on the HenryTest program icon gives a simple screen as show in Fig 2 with just one option – TEST.

The first things that I found in my junk box were three 100uH axial inductors so I bent the leads and attached them one by one and pressed the TEST button. After a few seconds the measured value is displayed as shown in Fig 3. To my surprise the three measured 81.62μH, 95.47μH and 96.02μH. Looking at the inductors shown in Fig 4 it is obvious that they do in fact have a different number of turns, however they are within the ±20% manufactures tolerance.

So HenryTest is easy to install, easy to use and had already spotted something that I had not expected.

3.0

Practical Test

No specification came with HenryTest and there was no handbook or readme file on the CD. Looking at the HenryTest web site [1] gives a bit more information
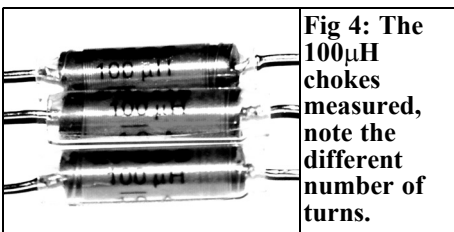

Fig 4: The 100μH chokes measured, note the different number of turns.

and a specification of 0.1% error margin from 1nH to 1H. This is impressive and I do not have anything to verify this level of accuracy. Therefore I decided to make some practical tests to see how HenryTest could be useful when making low value inductors. I dug out an old copy of the RSGB VHF Handbook and found the formula for calculating the inductance of an open wire coil and a nomogram for such coils.

I wound a two-turn coil from 16swg wire on a 3mm drill. Using the formula:

$$L = \frac{N^2 r^2}{9r + 10l} \qquad \mu H$$

and substituting

    L – length of coil = 1.9in
    N – number of turns = 2
    R – radius of coil = 0.1in

Gives 2.3nH

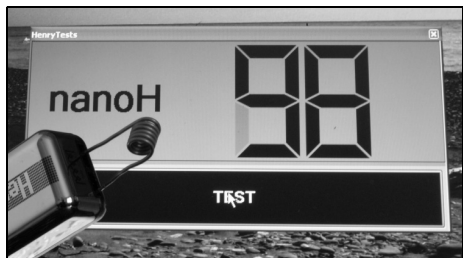Connecting this coil to HenryTest gave a result of 3nH as shown in Fig 5

Fig 6: Measuring a 6 turn coil.

Using the nomogram I wound a 6 turn coil on a ¼in drill that should be 100nH. Connecting this coil to HenryTest gave a result of 98nH as shown in Fig 6

4.0

Conclusion

The test showed HenryTest to be easy to use and useful for measuring the small value coils that VHF and UHF constructors are likely to require. The disadvantage of HenryTest is that the coil must be physically suitable to be connected to the measuring jaws; this may not always be practical. The jaws were a bit temperamental when connecting to the 16swg wire. I am told there will soon be an

attachment for measuring SMD parts.

At €49.90 it is quite expensive for a single measuring instrument when compared to the multimeters on the market but when compared to measuring instruments with similar specification it is good value.

5.0

References

[1] HenryTest web site – www.henrytest.com

Jean-François Iosca, F1LVO

White Box story – what make of PA?

1.0

Introduction

I have never found an article comparing the various types of PA in all the literatures on White Boxes. I have the facilities at work to test these so I succeeded in collecting 8 amplifiers to test:

- 1 standard Arcom 25PA003

- 1 Toshiba BA 3153A

- 2 X Toshiba BA2075B

- 4 X Toshiba BA 2075C.

Fig 1: The measuring equipment used to test the White Box PAs.

Table 1: Measured results for eight different 24GHz PAs.

P IN dBm	P OUT dBm	Gain dB	P mW	I + 6V CLIM mA
BA 2075C SN 1225				
-16.30	20.01	36.31	100	810
-10.80	26.03	36.83	400	810
-7.00	29	36	800	790
BA 2153A SN 0392				
-21.68	20.08	41.76	100	520
-12.67	26.04	38.71	400 max !	560
BA 2075B SN 2939				
-21.12	20.03	41.15	100	520
-13.92	26.1	40.02	400	550
-11.09	26.66	37.75	460	560
BA 2075B SN 2436				
-18.28	20.06	38.34	101	360
-14.39	23.06	37.45	202	390
-11.38	24.77	36.15	300	430
-8.13	25.63	33.76	366	440
BA 2075C SN 3097				
-15.52	20.00	35.52	100	820
-12.83	23.10	35.93	204	820
-10.25	26.11	36.36	407	820
-7.23	29.10	36.33	813	820
-3.80	30.01	33.81	1000	780
BA 2075C SN 0393				
-17.83	20.09	37.92	102	740
-15.15	23.01	38.16	200	740
-12.22	26.03	38.25	400	750
-7.22	29.46	36.68	882	770
-5.03	30.01	35.04	1000	790
BA 2075C SN 3423				
-14.80	20.04	34.84	100	830
-11.75	23.08	34.83	200	830
-8.49	26.05	34.54	403	830
-1.59	30.02	30.61	1000	850
ARCOM REF. 25 PA 003 SN 0567				
-21.53	20.06	41.59	101	830
-18.32	23.06	41.38	202	840
-14.04	26.04	40.08	401	840
-7.30	27.56	34.86	570	950

CONDITIONS: F = 24.050 GHz ,   V+ = 6V,   V- = 5V

2.0

Measuring equipment

The measuring equipment comprises:

- A standard 10MHz Rubidium os-
  cillator to synchronise the tracking
  of the HP 70204 spectrum analyser
  with the 70909 plug-in (100Hz >
  26.5 GHz).

- 2 Anritsu milli wattmeters, the first

with a −70dBm 32GHz diode probe
for measuring the power at the
input of the amplifiers, the second
with a 40GHz -30dBm bolometer
probe for measuring the power at
the output of the amplifiers. For
information, the difference be-
tween the 2 milli wattmeters at
24,050GHz, with a 0dBm signal is
only 0.1dB.

- 1 HP 26.5GHz tracking generator
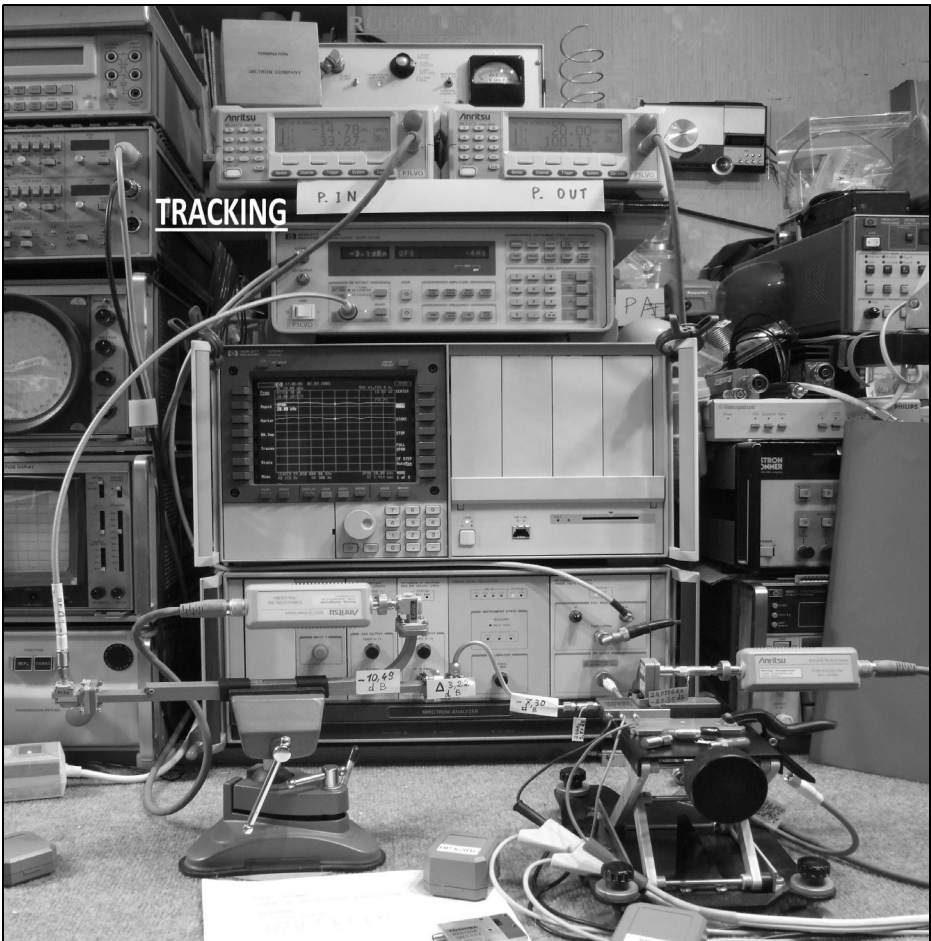 controlled by the spectrum ana-
  lyser.

Fig 2: View of the test setup and the measuring instruments used to measure the White Box PAs.
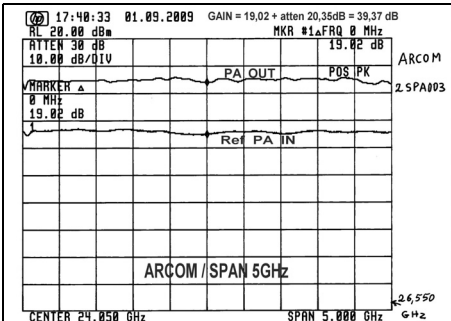
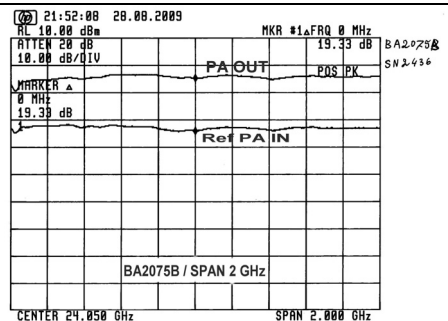**Fig 3: Measured frequency response for the Arcom PA.**

Fig 4: Measured frequency response for a BA2075B PA.

3.0

Measurements

For measurements at the input and output of the amplifiers:

Permanent measurement of the level at the input of the PA:

- 1 HP K752C WR42, 10dB directional coupler and 3 Maury transitions WR42 to 3.5mm.

- 1 3.5mm 26.5GHz 6dB attenuator to mask a possible TOS at the input of the amplifiers that would seriously distort measurements given that there is a considerable length of semi-rigid (12cm). For informa-

Fig 5: Measured frequency response for a BA2075C PA.

tion 0.5dB of loss to 24 GHz) + 1 milli wattmeter

For the measurement of the power at output:

- A HP K281C transition (WR42 to 3.5mm) followed by a 26.5GHz, 20dB, 2W HP33340C attenuator + 1 milli wattmeter

After a 2 hour warm up period for the unit I calibrated all the elements of the measuring equipment in differential mode (see the values in Figs 1 and 2). The precision of the equipment is remarkable with a drift of some hundredths of dB over several hours and under the same conditions on the following day the equipment was within 6 hundredths of dB. Perhaps the temperature of the laboratory (24°C) wasn't exactly the same in spite of the essential air conditioning on this very hot August day.

The result of measurements by the milli wattmeters has been adjusted for the attenuators so that the values in Table 1 are the true values of Power in and Power out of the amplifiers in dBm and mW. You can draw your own conclusions. Figs 3 - 5 shows the frequency response curves measured. Full results are available at:
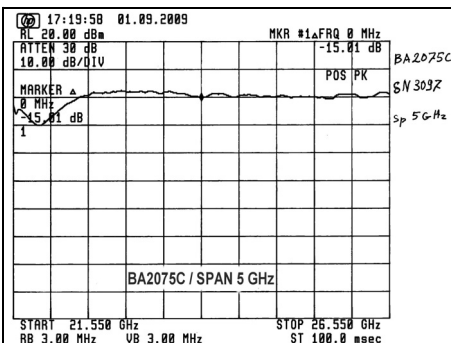http://f1lvo.free.fr/PA%2024%20GHz %20Boite%20Blanche/

Franco Rota, I2FWH

RF Power Transformer

1.0
Introduction

The instructions below show how it is possible to build a very good RF power transformer without using difficult to find brass or copper pipes. This technique is suitable for almost any toroid hole size. The instructions can be used for high or medium power RF transformers using single or binocular ferrites. In particular this method is useful for 1:4, 1:9, 1:16 etc transformers where the 1 side uses the technique shown and the 4,9,16 etc side is made using further turns inserted into the inside of the shield.

2.0
Construction

- Save 7 to 10cm of the shield from RG213 or RG214 coax cable as shown in Fig 1. RG58, RG59, RG142 etc can also be used for ferrites with small holes.

- Even though the ferrites shown in Fig 2 have very different size holes the shield from RG213 can be used because the size of the shield is not critical.
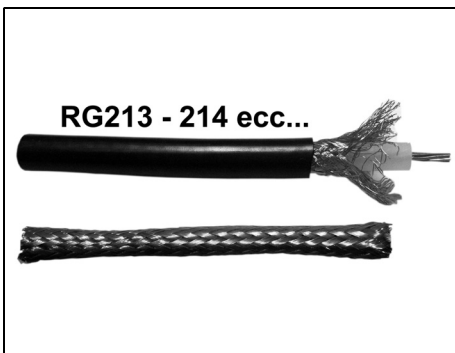

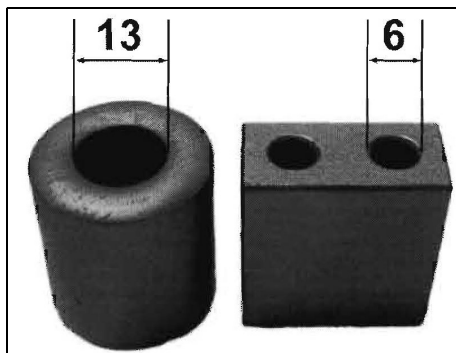**Fig 1: Save 7 to 10cm of shield from RG213.**

Fig 2: The types of ferrite to use.

Fig 3: Forming the primary from coax cable shield.

Fig 4: Completed RF power transformers.

- Use any round object to prevent the shield from deforming and insert the shield into the holes of the ferrite as shown in Fig 3. Use an awl to widen the holes at the ends of the shield, to insert any further cables needed.

Fig 4 shows completed transformers where the secondary winding has been made using teflon coated wire inserted into the shield.

3.0

Reference

[1] This information is taken from the RF Elettronica catalogue – www.rfmicrowave.it.
Suitable ferrites are available from the catalogue.

The UK Six Metre Group

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Rafal Orodzinski, SQ4AVS

Digital Gain and Noise Meter (DGNM)

Noise and gain measurements are necessary for all designers of RF receiving systems and it is a must for adjustments to low signals LNA.

1.0

Introduction

After building and testing the SPANFI meter [1] I have concluded that there are some factors that can be improved upon. The great advantage of SPANFI and its predecessor PANFI meter is the simplicity of design using easily available components and this concept also applies to my design.

To obtain more precise and repeatable results everything was reworked and new software was developed. Some of the previous designers' drawbacks have been removed in this way. Error factor analysis revealed more factors to be taken into account e.g. the noise factor, gain of the measurement circuit and temperature effects. Further improvement is also obtained by using a more precise formula, leaving this task to the microprocessor.
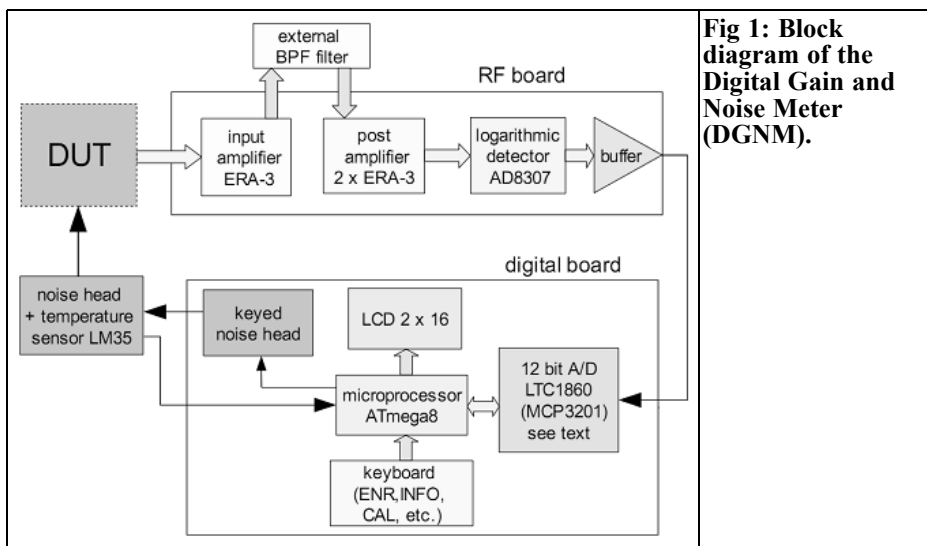

Fig 1: Block diagram of the Digital Gain and Noise Meter (DGNM).

2.0

Circuit description

As in most similar circuits the noise source reference (noise head) is keyed and feeds the input of the device under test (DUD) for example a low noise amplifier (LNA). A block diagram is shown in Fig 1. The output of the DUT is fed into the first stage of the measurement chain (ERA3 MMIC) followed by a band pass filter (BPF) and a two-stage post amplifier. Filtering is necessary to eliminate other signals, for example from TV/FM broadcast or GSM that may contribute to significant errors. A logarithmic detector built around the well-known AD8307 from Analog Devices detects the amplified and filtered signal. The voltage output of this detector represents the noise power of the amplified signal from the DUT.

The slope of the detector is scaled to 25mV/dB and after buffering it is applied to the input of a 12-bit analog to digital converter (ADC). The buffer is required to ensure low impedance drive for the ADC. The reference voltage is set at around 4.096V so 1 bit corresponds to 1mV. In other words resolution is 0.04dB per bit. To compensate for slope error the reference voltage may be adjusted slightly, this is the only hardware alignment necessary. True resolution is better because of the effect of oversampling.

The noise factor and gain are calculated by measuring the difference in power of amplified DUT signal when the input is switched between a 50Ω source and the same source with the known added noise level of the noise head defined by its ENR.

3.0

Measuring circuit

The RF measurement head is typical. It may be a commercial noise source or can be built around a noise diode or reverse polarised RF transistor. As all measurement refer to the known noise level of test head, before attempting any measurements we should first know the ENR factor at the test frequency (this may differ between bands). The amplifier is built from three ERA3 MMICs and an AD8307 the circuit is shown in Fig 2. Between the first and the second stage an external bandpass filter is used. This filter determines the centre frequency and bandwidth of the measurement circuit. Small series resistors are inserted between the stages to improve overall stability. Power level is detected by the AD8307 working in a standard application circuit. The circuit diagram of the digital board is shown in Fig 3. There are no adjustments because software corrects the DC offset. Adjusting the reference voltage provides the 1dB/25mV scaling. A buffered (by AD8754) detector output is applied to the input of the ADC. In the first design a Linear Technology LTC1860 converter was used, but later it was replaced by the cheaper MCP3201 from Microchip (pin compatible but requires a different software handler). Both devices have 12-bit resolution and an easy serial microprocessor interface and only require different software. The temperature factor of the noise head is often omitted. To correct temperature drifts of the noise head ENR, the LM35 temperature sensor should be installed inside the noise source to track temperature changes during measurement session. Opening a window or a door in a lab may cause a few tens of dB additional error.

The 10-bit ADC converter built into the ATmega microprocessor is more than sufficient to convert the 10mV/°K slope to obtain a temperature reading.

19

Fig 2: Circuit diagram of theRF board.

Fig 3: Circuit diagram of the digital board.

Fig 4: Photograph of the completed Digital Gain and Noise Meter (DGNM).

The operator panel consist of a 2x16 character LCD display and seven push-button switches:

Key	Action
"="	makes ENR_{IF} the same value as ENR_{RF}
INFO	displays current test setup as ENR_{IF}, ENR_{RF} and temperature (in °K)
ENR+ & ENR-	increment or decrement ENR values of the noise head in the current setup (for example for different bands)
CAL	starts the self calibration process

A picture of the completed unit is shown in Fig 4. More detailed pictures can be found on the authors web site [2]

4.0

Practical measurements

After power on, leave the test instrument for 15 minutes for internal temperature stabilisation.

For measuring an amplifier e.g. an LNA, we have to select the correct ENR value and set ENR_{IF} equal to ENR_{RF} by pressing equal "=" key.

For transverter measurements, two values should be selected from the noise head data, one for the frequencies of the RF input and the other for transverter input frequency e.g. the ENR of the noise head might be 7.8dB at 1296MHz and 7.1dB at 144MHz.

The ENR+ and ENR– keys increase or decrease the current ENR value in 0.1dB steps in the 2 to 20dB range.

After setting the ENR value(s) connect the noise head directly to test measuring setup and press CAL to start the self-calibration process. This takes about 1 minute to complete.

Following calibration the circuit automatically starts noise and gain measuring cycles. Now place the device under test (DUT) between the noise head and the DGNM input. Noise and gain results are updated every 4 seconds.

After changing the band it is necessary to recalibrate the circuit after updating the ENR value(s). The last selected ENR value is stored in EEPROM.

I would like to thank Michal SP2IQW for sharing his measurement, experiments and results and Stefan SP9QZO for his assistance in RF tests and for offering many operating suggestions.

5.0

References

[1] A simple PANFI (SPANFI) Concept and Implementation, Ingo Gaspard DF1VH, Dubus 3/2006 pp 8 - 21

[2] http://sites.google.com/site/sq4avs/miernik-szomow

Gunthard Kraus, DG8GB

Introduction to Digital Signal Processing (DSP)

1.0

Principles of Digital Signal Processing

Fig 1 shows an overview of the complete DSP process:

The analog signal is first passed through a low pass filter to limit the frequency range. This feeds a sampling circuit that constantly measures the value of the signal with extremely short sampling pulses. This result of these samples is held in a buffer to feed an analogue to digital converter. This generates samples with a constant frequency (sampling rate) that is processed as a serial 8-bit or 16-bit data stream by a computer, Digital Signal Processor or a micro controller.

The processing in the computer produces a resulting signal as if it had been sent as an analog signal via a filter circuit. The program defines whether this is a low pass filter, high pass filter, band pass filter or a band stop filter.
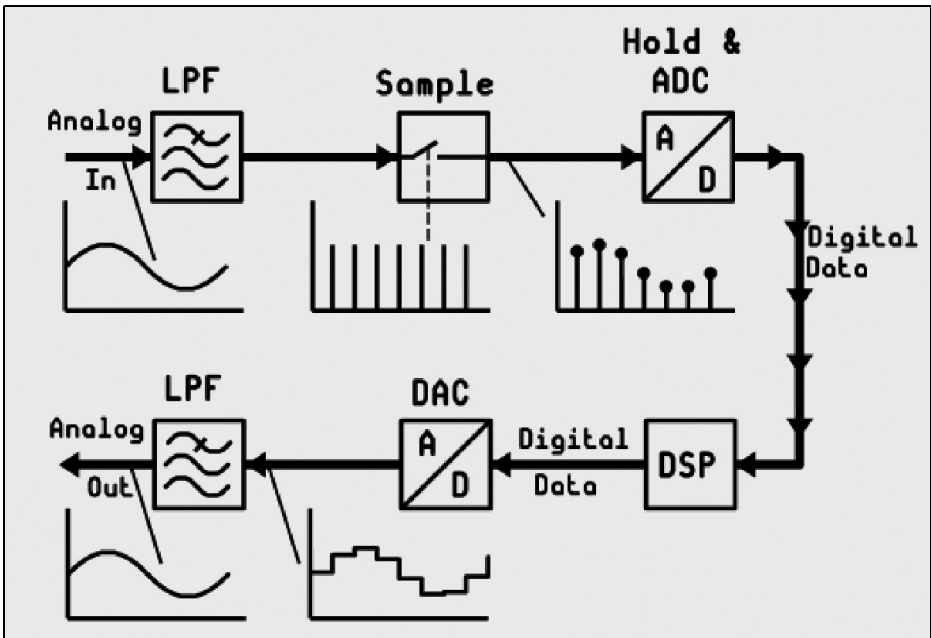
Fig 1: Block diagram of a DSP system: the information goes in as an analogue signal and comes out after digital processing as an analogue signal again.

Fig 2: Representation of a signal in the time Domain (oscillogram) or in the frequency Domain (spectrum) are two sides of the same coin.
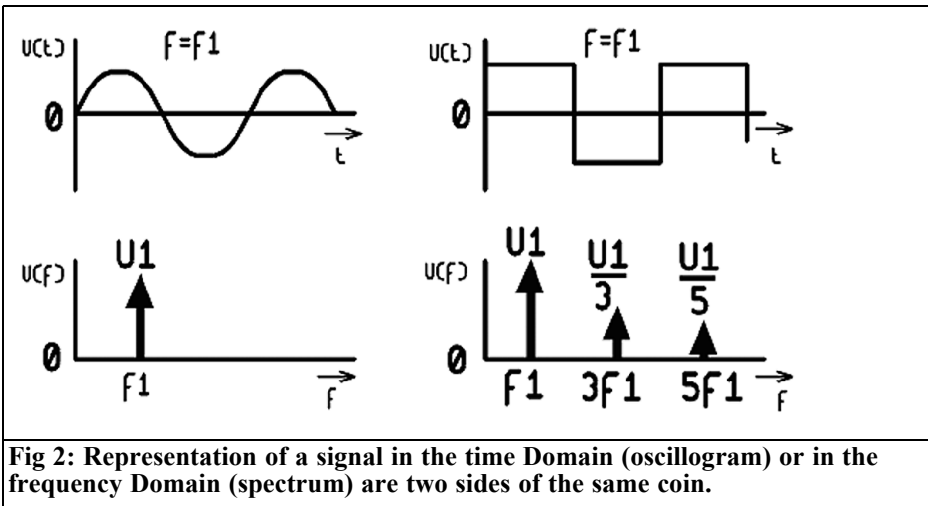
The digital signal is converted back to an analogue signal using a digital to analogue converter then passed through a low pass filter to suppress the unwanted signals (sampling rate and its harmonics) to produce the required signal.

1.1 Advantages of Digital Signal Processing

The same performance as using arithmetic calculation could only be realised traditionally by great technical expense or may not be achievable at all. In addition the filter values (cut off frequency, type of filter, stop band attenuation etc.) can be switched very easily by changed parameters in the computer program. Adaptive filters can be programmed also coding or decoding etc. can be performed.

2.0

Signal frequencies and sampling rate

The following basic rule must be followed rigidly if everything is to work properly. It is known as the Nyquist criteria:

• For a purely sinusoidal signal of a given frequency the sampling rate must be at least twice the frequency of the incoming signal. It is better if the sampling rate is more than twice the signal frequency. The analogue signal can only be reconstructed again without additional errors after the digital to analogue conversion and the filtering if this is true.

Caution:

As soon as the waveform deviates from a sine wave, such a signal contains a whole range of frequencies other than the basic frequency. These can be harmonics as produced by distortion in an audio signal. They are always an integer multiple of the fundamental.

Fig. 2 shows a comparison between a sine and square wave signal:

• The top left shows a pure sine wave signal as it would be displayed on an oscilloscope and below a single line as it would be seen on a spectrum analyser.

• The top right shows a square wave signal as it would be displayed on an oscilloscope and below the odd harmonics that would be seen on a spectrum analyser.
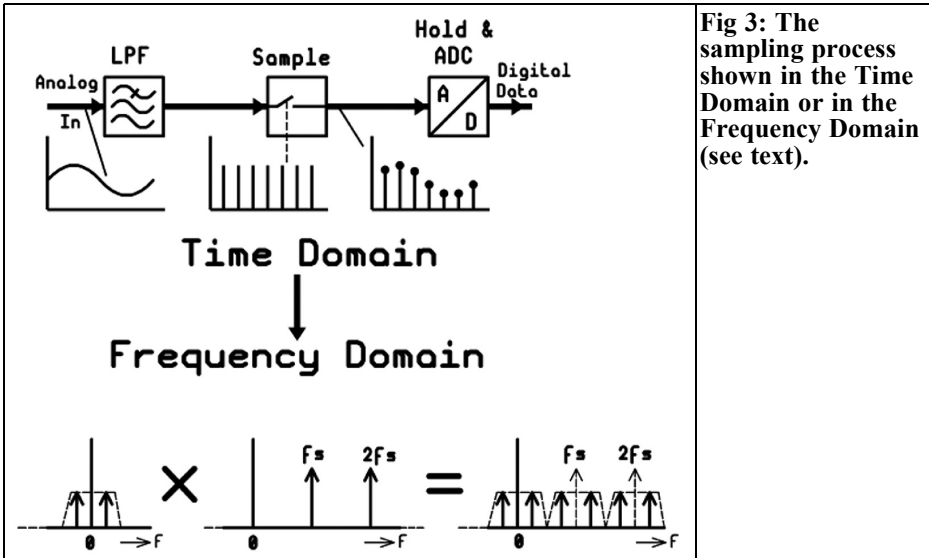
Fig 3: The sampling process shown in the Time Domain or in the Frequency Domain (see text).

Note: the negative frequency portions are not shown in the spectrum because they do not display separately on a spectrum analyser.

As expected a pure sine wave signal only contains one frequency but a symmetrical square wave signal contains the fundamental and odd number harmonics.

Therefore the following condition can defined more precisely:

- The sampling rate must be at least twice as high as the highest single frequency or harmonic in the input signal. Those familiar with harmonics will ask:

- The harmonics of a non-sinusoidal signal theoretically go to infinity with ever decreasing amplitudes. Which scanning rate should be select?

There is only one solution; that is to cut off at some frequency and accept that this will affect the final analogue signal. A low pass filter, or anti aliasing low pass filter with the chosen cut off frequency is used in front of the A/D converter. It should have a very sharp cut off. Select a sampling rate that is at least twice the cut off frequency of this low pass filter.

Example: The cut off frequency for an audio CD Player is 20kHz. The signal is sampled at 44.1kHz.

There are more decisions to be made:

- If the cut off frequency of the low pass filter is very low a relatively low sampling rate can be used. That means less calculation power is required but the danger is that now harmonics of the input signal are missing that are important for the correct waveform shape of the analogue signal.

- If the cut off frequency and the sampling rate are higher than necessary, the waveform shape of the analogue signal is correct, but the following drawbacks result:

- The higher range of the input signal increases the self-noise level. That can be quite a problem if the information signals are in the order of a microvolt.

- The higher scanning rate requires such a high calculation performance that even the most modern PCs and processors are pushed to their limits. Only genuine Digital Signal Processor (DSP) will help, these have a
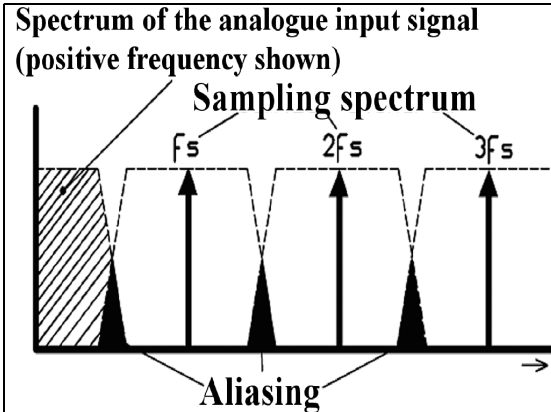
Fig 4: Aliasing happens if the sampling rate is not high enough.

different hardware structure to perform more than a billion 32-bit floating point calculations per second.

For information: because the power of computers and DSPs seems to increase by the hour, the calculation performance is no longer the problem. Therefore the second option, to sample at a higher frequency than necessary, is often chosen these days. The data stream is filtered with a "digital decimation filter" after the A/D converter that artificially reduces the cut off frequency giving consistency with a lower sampling rate, but this must still be at least twice as high as the data bandwidth according to Shannon.

The question to be clarified now is: What happens if the sampling rate is not more than twice the maximum input frequency?

Fig 3 shows a comparison of all the signals involved both in the time domain and the frequency domain:

• In the time Domain the sampling can be regarded as multiplication of the analogue input signal by sampling pulses. Therefore the pulses at the output of the sampler follow the amplitude of the analogue input signal.

• In the frequency domain a multiplication of two sine signals always results in two new signals: the sum and difference frequencies of both.

Consider this:

The scanning pulse produces a spectrum containing the odd harmonics with the same amplitude. Each harmonic is multiplied by the analogue input signal. Therefore the final result is a spectrum containing sum and difference frequencies as multiplication results of every harmonic and the input signal. Developing this further; with an input signal containing more than one frequency you get lower and upper sidebands as multiplication results.

Fig 4 shows what happens with a sampling rate that is too low:

• The spectra begin to overlap that leads to extremely unpleasant distortions. This is called ALIASING and the distortion is called "Aliasing distortions". Unfortunately this not only sounds very unpleasant but if this happens there is no way to eliminate it. The low pass filter before the A/D converter is to prevent this effect. Therefore it has the appropriate name of "Anti Aliasing" Filter.

Remember this:

EVERYTHING in the input signal must be suppressed above half the sampling rate at all costs otherwise there is terrible trouble.
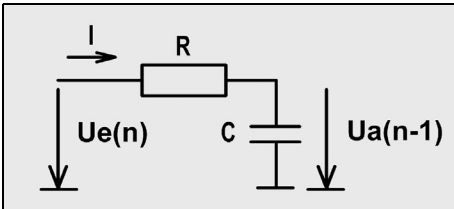
Fig 5: This well-known simple circuit can show the principle the digital filter.

3.0

Practical project: A simple digital filter

3.1. Introduction

The serial output of the sampler is fed into a digital filter. The different sampled values are now multiplied by special factors and the multiplication results added together. So the input analogue signal is affected in the same a way as if it had passed through a discrete filter circuit.

In order to understand this, Fig 5 shows a circuit that everyone knows; the simple RC low pass filter. At a certain time the output voltage $U_{a(n-1)}$ is developed across the capacitor. If the sampling circuit produces the next sampled value, the input voltage changes suddenly. The new voltage $U_{e(n)}$ is at the input while the capacitor voltage still holds the old value $U_{a(n-1)}$. What will the new output voltage $U_{a(n)}$ be until the next scanning value arrives, for the time interval $\Delta t = t_{sample}$

The value at the capacitor will be:

$$U_{a(n)} = U_{a(n-1)} + \Delta U_c = U_{a(n-1)} + \frac{\Delta Q}{C}$$

$$= U_{a(n-1)} + \frac{I \cdot \Delta t}{C} = U_{a(n-1)} + I \cdot \left(\frac{t_{sample}}{C} \right)$$

$$U_{a(n)} = U_{a(n-1)} + \left(\frac{U_{e(n)} - U_{a(n-1)}}{R} \right) \cdot \left(\frac{t_{sample}}{C} \right)$$

$$= U_{a(n-1)} + \left[U_{e(n)} - U_{a(n-1)} \right] \left(\frac{t_{sample}}{RC} \right)$$

Expressing t_{Sample} in terms of the sampling frequency gives $t_{Sample} = 1/f_{Sample}$

There is a formula for the cut off frequency of an RC filter used in communications theory:

$$f_{cut-off} = \frac{1}{2\pi RC} \Rightarrow \frac{1}{RC} = 2\pi f_{cut-off}$$

Substituting this in the formula for $U_{a(n)}$ gives:

$$U_{a(n)} = U_{a(n-1)} + \left[U_{e(n)} - U_{a(n-1)} \right] \left(\frac{2\pi f_{cut-off}}{f_{sample}} \right)$$

Using the term k for the term in the right hand bracket gives a simplification. The result after multiplying out gives:

$$U_{a(n)} = U_{a(n-1)} + k \cdot \left(U_{e(n)} - U_{a(n-1)} \right)$$

$$= U_{a(n-1)} - k \cdot U_{e(n)} - k \cdot U_{a(n-1)}$$

$$U_{a(n)} = k \cdot U_{e(n)} + \left(1 - k \right) \cdot U_{a(n-1)}$$

The new value is calculated by adding together the new input value $U_{e(n)}$ multiplied by k and the preceding output value $U_{a(n-1)}$ multiplied by 1-k. This is a usual procedure during digital signal processing; it is abbreviated to MAC (Multiply And Accumulate).

3.2. Training example: Investigation of the RC circuit by hand

For this example use a low pass filter with a cut off frequency of 100Hz and a sampling rate of 2kHz. The input frequency is comfortably smaller than half the sampling frequency.
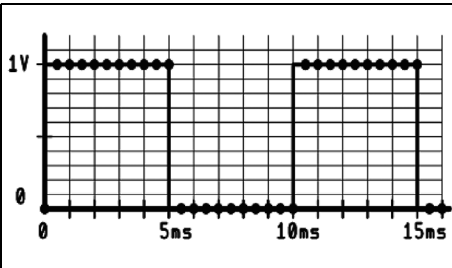
Step 1:

First the factors k and 1-k are calculated:

Fig 6: The input signal with the sampling times.

$$k = \frac{2\pi \cdot 100\,Hz}{2000\,Hz} = 0.314$$

$$(1-k) = 1 - 0.314 = 0.686$$

Step 2:

The initial value of the output voltage

with t = zero is zero. Starting from the time t = zero as the input square wave signal with a frequency f = 100Hz is applied the amplitude goes from 0V to 1V. The sampled input signal from t = zero to t = 15mS is shown in Fig 6.

Step 3:

Table 1 shows the output voltage values across the capacitor. The fastest way to do this is to use Excel, but naturally it can be done with a pocket calculator. The results are also shown in Fig 7. The upper half shows the input signal with the sampling times. The lower half shows the output signal that would be produced after Digital Signal Processing and D/A conversion given by Table 1. It is similar to the known output of a low pass filter, it just need filtering to remove the remains of the sampling.

Table 1: Calculation of the output voltage for different samples.

Sample Number n	New Input $U_{e(n)}$	k * Ue(n) 0.314 * $U_{e(n)}$	Old Output $U_{a(n-1)}$	(1-k) * $U_{a(n-1)}$ = 0.686 * $U_{a(n-1)}$	New output $U_{a(n)}$ = 0.314 * $U_{e(n)}$ + 0.686 * $U_{a(n-1)}$
0	0	0	0	0	0
1	1	0.314	0	0	0.314
2	1	0.314	0.314	0.215	0.529
3	1	0.314	0.529	0.363	0.677
4	1	0.314	0.677	0.464	0.778
5	1	0.314	0.778	0.533	0.848
6	1	0.314	0.848	0.581	0.895
7	1	0.314	0.895	0.614	0.928
8	1	0.314	0.928	0.637	0.95
9	1	0.314	0.95	0.652	0.966
10	1	0.314	0.966	0.662	0.977
11	0	0	0.977	0.670	0.670
12	0	0	0.67	0.459	0.459
13	0	0	0.459	0.315	0.315
14	0	0	0.315	0.216	0.216
15	0	0	0.216	0.148	0.148
16	0	0	0.148	0.1	0.1
17	0	0	0.1	0.069	0.069
18	0	0	0.069	0.0479	0.0479
19	0	0	0.0479	0.0328	0.0328
20	0	0	0.0328	0.0225	0.0225
21	1	0.314	0.0225	0.015	0.329
22	1	0.314	0.329	0.226	0.54
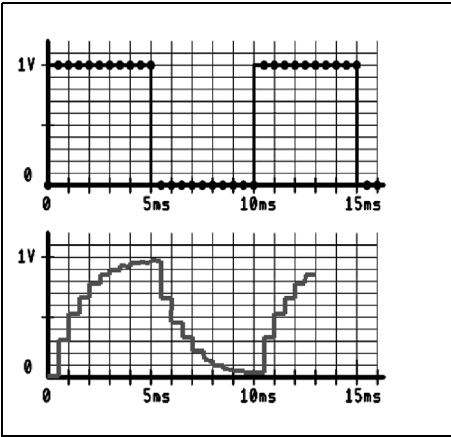
Fig 7: What the digital filter makes from the analogue signal.

4.0

FIR – Filter (Finite Impulse Response filter)

In practice a substantially more complex filter than a simple low pass filter is required. Knowledge of higher mathematics (complex calculation and integral calculus) is required to design these filters because they use Z transforms. Their design and function, in a simplified form, becomes clear with the following fact:

- In a block diagram each element has a value z^{-1} that represents the storage of the previous sampled value. Fig 8 shows how a draft filter diagram could look and with this knowledge we can understand the structure the FIR filter.
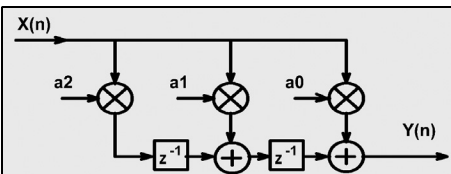
Fig 9: Block diagram of the transfer function of the FIR filter (see text).
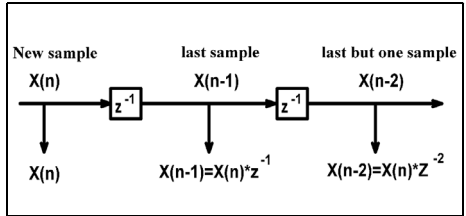
Fig 8: The Z transform working on the sampled value.

Explanation:

FIR means Finite Impulses Response. These filters respond to a needle pulse input signal with an output signal that decays to zero. These filters never ring and are stable. They can be recognised because they have a one-way structure from input to output, there is NEVER feedback from the output to the input. An alternative designation is a non-recursive filter.

A practical example is shown in Fig 9.

The output signal is given by:

$$Y_{(n)} = a_0 \cdot X_{(n)} + a_1 \cdot X_{(n)} + a_2 \cdot X_{(n)} \cdot z^{-2}$$

that means:

$$Y_{(n)} = a_0 \cdot X_{(n)} + a_1 \cdot X_{(n-1)} + a_2 \cdot X_{(n-2)}$$

Extracting the factor $X_{(n)}$ gives the transfer function:

$$Y_{(n)} = X_{(n)} \cdot \left[a_0 + a_1 \cdot z^{-1} + a_2 \cdot z^{-2} \right]$$

$$\Rightarrow \frac{Y_{(n)}}{X_{(n)}} = H(z) = a_0 + a_1 \cdot z^{-1} + a_2 \cdot z^{-2}$$

(A small reference: a filter is only then absolutely stable if in the denominator of the transfer function no z^{-1} can be found)

Advantages of the FIR filter:

- Absolute stability, thus no ringing
- If the filter is symmetrically designed ($a_0 = a_2$), it has a linear phase re-

Fig 10: This example shows that the programming work for the DSP machine is not an easy walk.
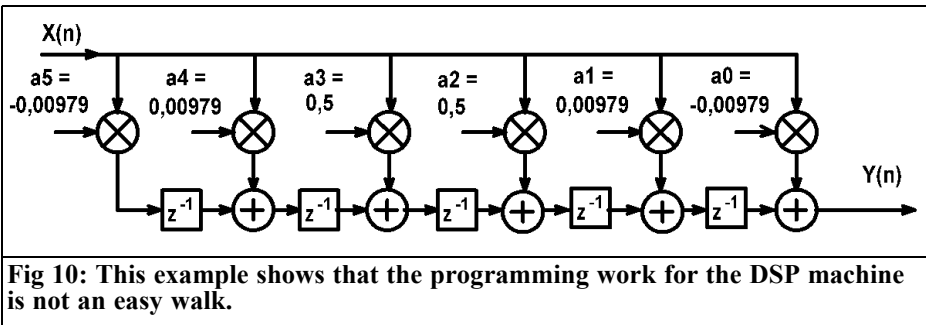
sponse and thus a constant Group Delay.

- Design not critical even with rounding or simplifications in the program

Disadvantages of the FIR filter:

- Filter slope not as steep as IIR filters of the same degree.

Some references to the design of such FIR filters.

The design needs good knowledge of mathematics but now on the Internet, there are many programs (including on-line calculators) that can do the job perfectly. Nevertheless some fundamental things should be known:

- The sum of the coefficients $(a_0/a_1/a_2...)$ gives the DC voltage gain of the filter.
- The standardised cut off frequency F_g is used for design rather than the actual cut off frequency. This makes it easier to understand the relationship between the actual cut off frequency and the sampling rate that must always be below 0.5 (the Nyquist criteria).
- The filter degree should be always odd because then there is always a zero in the frequency response at half the sampling rate (the Shannon condition).

Example:

- A filter for $F_g = 0.25$ and filter degree $N = 5$ with the filter coefficients

- $a_0 = a_5 = -0.00979$
- $a_1 = a_4 = +0.00979$
- $a_2 = a_3 = +0.5$

Calculating the actual cut off frequency if the sampling frequency is 8kHz.

Solution:

$$f_{cut-off} = F_g \cdot f_{sample} = 0.25 \cdot 8kHz = 2kHz$$

The associated circuit is shown in Fig 10.

5.0

IIR filter (Recursive filter)

These always have feedback from the output to the input and therefore need substantially more care to design - they are inclined to oscillate with the smallest sloppiness (known from analogue feedback amplifiers). Rounding or truncating large numbers too early in the design can cause this.

The same filter degree gives a substantially steeper filter slope therefore improving the filter effects compared to the FIR filters discussed.

IIR means Infinitely long Impulse Response. The response to a needle pulse at the input decays away but never completely to zero but hopefully becomes ever smaller. In addition these strange

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1996	Y	Y	Y	Y
1995	P	Y	Y	Y
1994	Y	Y	Y	Y
1993	Y	P	P	P
1992	Y	Y	Y	Y
1991	P	P	P	P
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1984	Y	Y	Y	Y
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1980	P	P	P	P
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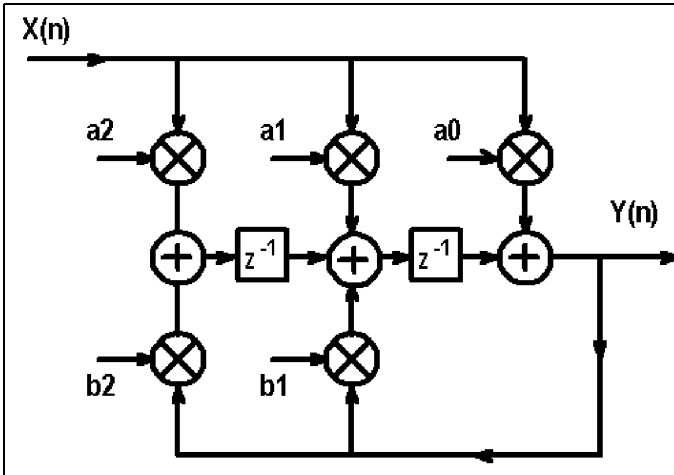
Fig 11: IIR filters are more complicated. They make greater demands on the processor, the program and the programmer.

filters already supply a result at the output before the impulse appears at the input, thus they are a special kind of filter that must be treated carefully.

An example of such a filter arrangement can be seen in Fig 11:

$$Y_{(n)} = a_0 \cdot X_{(n)} + a_1 \cdot X_{(n)} \cdot z^{-1} + a_2 \cdot X_{(n)} \cdot z^{-2}$$

$$+ b_1 \cdot Y_{(n)} \cdot z^{-1} + b_2 \cdot Y_{(n)} \cdot z^{-2}$$

$$\Rightarrow Y_{(n)} \cdot \left(1 - b_1 \cdot z^{-1} - b_2 \cdot z^{-2}\right)$$

$$= X_{(n)} \cdot \left(a_0 + a_1 \cdot z^{-1} + a_2 \cdot z^{-2}\right)$$

From this the transfer function follows:

$$\frac{Y_{(n)}}{X_{(n)}} = \frac{\left(a_0 + a_1 \cdot z^{-1} + a_2 \cdot z^{-2}\right)}{\left(1 - b_1 \cdot z^{-1} - b_2 \cdot z^{-2}\right)}$$

The denominator of the transfer function can possibly become zero - particularly if the coefficients b_1 and b_2 have positive values. Then the filter oscillates.

This calculation also uses the stored preceding values. With such a complex structure a whole assortment of initial values and buffers will be required.

Such a filter was introduced in simple form with the RC low pass filter in chapter 3.2. The following applied:

$$U_{a(n)} = k \cdot U_{e(n)} + \left(1 - k\right) \cdot U_{a(n-1)}$$

The actual cut off frequency was 100Hz with a sampling rate of 2kHz.

The coefficients were:

k = 0.314   and   1-k = 0.686

Example:

   a) The transfer function of this filter with these two coefficients is derived

   b) The circuit of this filter drawn with registers of the correct values.

Solution to a):

$$U_{a(n)} = k \cdot U_{e(n)} + \left(1 - k\right) \cdot U_{a(n-1)}$$

$$U_{a(n)} = k \cdot U_{e(n)} + \left(1 - k\right) \cdot U_{a(n-1)} \cdot \left(z^{-1}\right)$$

$$U_{a(n)} - \left(1 - k\right) \cdot U_{a(n)} \cdot \left(z^{-1}\right) = k \cdot U_{e(n)}$$

$$U_{a(n)} \cdot \left[1 - \left(1 - k\right) \cdot z^{-1}\right] = k \cdot U_{e(n)}$$

$$\frac{U_{a(n)}}{U_{e(n)}} = H(z) = \frac{k}{1 - \left(1 - k\right) \cdot z^{-1}}$$
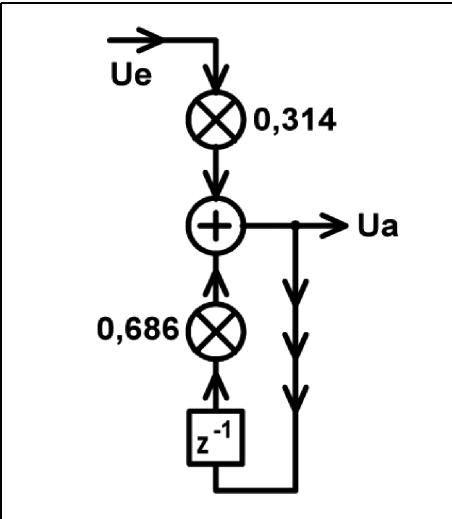
35

Fig 12: Even a simple RC low-pass filter more complex in digital form.

Thus:

$$H(z) = \frac{0.314}{1 - 0.686 z^{-1}}$$

The block diagram of this RC low pass filter in digital form is shown in Fig 12.

6.0

Organisation of the sampled values in the memory

There are always a constant number of samples being processed in the DSP. The number is nearly always multiples of two e.g. 512, 1024 or 2048. When the next sampling value arrives from the A/D converter there is congestion:

Not only is the current value stored but X_n values are stored so the oldest value is deleted and the newest value is set to the front of the list. Additionally all remaining values in the list must be renumbered. This is fastest for a small number of stored values. All the values are relocated to put them in the correct position. The process of relocating 4 sampling values is shown in Fig 13. This would take a long time for a large number of samples therefore the following method is adopted:

• The number of sampled values is always multiples of two e.g. 2, 4, 8, 16, 32, 64, 128….1024, 2048, 4096...

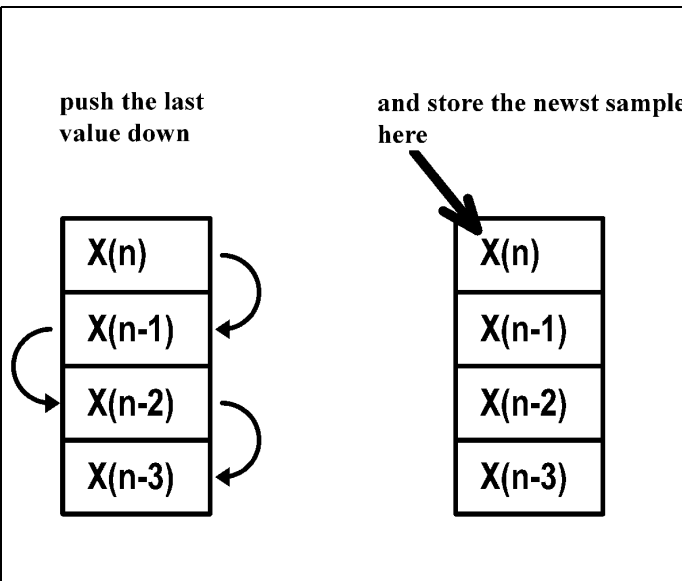
push the last value down

and store the newst sample here

Fig 13: This is one method of updating the stored values when a new sampled value arrives. This method is only recommended for a few stored values.
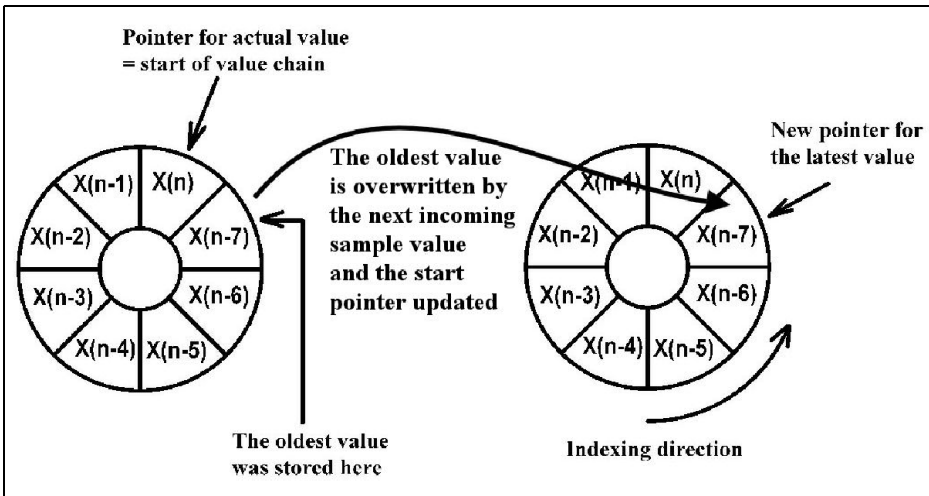
Fig 14: Using a pointer is much better for a large number of sampled values.

The samples are organised in memory like a ring and the start of the chain is stored as a pointer in a variable. This pointer always points to the newest value. The next sample is stored in the last place of the ring overwriting the oldest value. The pointer is changed to the new start address of the ring and calculation can begin. This principle is represented in Fig 14.

Let's hope that now everything is clear!

Herbert W. Schulte, CT2IJD/DD0PC

Time Domain Reflectometer (TDR) for radio amateurs

Part 1

Radio amateurs normally think in terms of frequencies, harmonics, gain or voltage levels. That is thinking in the frequency domain. A spectrum analyser displays in the frequency domain. This view is useful for looking at active signal but the time domain gives a better view of cables and networks. That means measuring signals or voltages over a time period.

1.0

Introduction

This view is implemented in a Time Domain Reflectometer (TDR). A TDR uses a pulse generator to send a pulse along a cable or a network and display the pulse reflected from the remote end

Fig 1: Block diagram of a Time Domain Reflectometer (TDR).

of the cable on an oscilloscope. A block diagram of a Time Domain Reflectometer (TDR) is shown in Fig 1.

A second type of system is the Time Domain Transmission (TDT). This is like the TDR but the oscilloscope displays the pulse at the remote end of the cable. This article describes what a TDR can be used for.

2.0

Operating principle

Attach an unknown cable to the output of the TDR. The output impedance of the TDR is usually 50Ω so different displays will be produced depending on the characteristic impedance of the cable see Figs 2 – 4.

Figs 2 - 4 show the pulse coming from the TDR into the unknown cable. At the end of this cable the pulse is reflected, the amount reflected changes if it is terminated with an active resistance, the characteristic impedance, a short circuit or an open circuit. The pulse is overlaid with the reflection as a combined display. To understand this display, remember that the pulse also sees the characteristic impedance of the internal cables of the TDR. In the example the internal part of
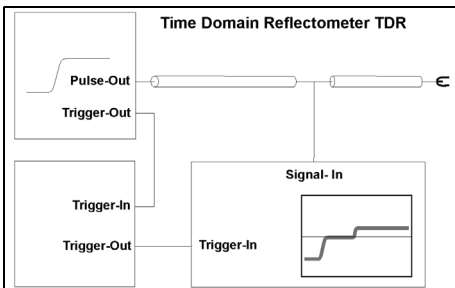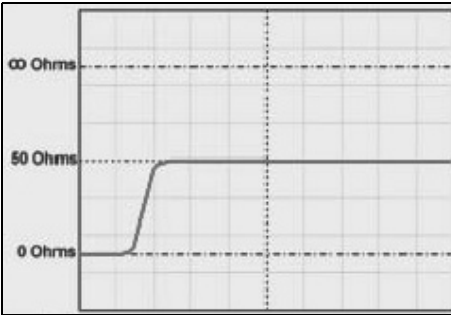
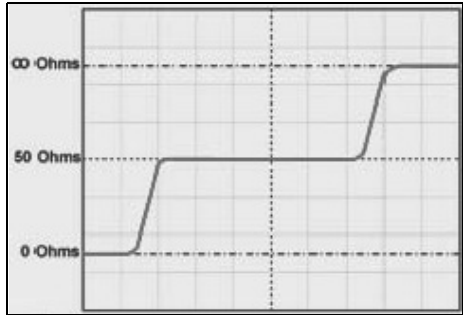**Fig 2: Measurement with a correctly terminated 50Ω cable connected.**

Fig 3: Measurement with an open circuit 50Ω cable connected.

the TDR uses 50Ω semi rigid cables.

As the pulse travels along the cable, part of the pulse is reflected each time the impedance changes. If the whole cable from beginning to end has an impedance of 50Ω then there are no reflections. If the characteristic impedance is attached at the end of the cable all of the energy is absorbed and nothing is reflected. This gives a display like Fig 2. If the end is open circuit the display is like Fig 3 and if the end is short circuit the display is like Fig 4.

3.0

Using a TDR

If the type of coax cable is known and

Fig 4: Measurement with a short circuit 50Ω cable connected.

the propagation speed is known then the location of a short circuit, an interruption or a characteristic impedance disturbance can be precisely located. Measure the time from the start of the pulse to the position of the jump on the display caused by the disturbance then divide the value by two. Multiply the result by the propagation speed of the cable to calculate the distance between start of the cable and the position of the disturbance. A professional TDR has a scale either in time units or in length of typical cables.

The TDR is not only useful to measure cable lengths. If a TDR is used to examine an antenna cable every "bad" connection or damage to the cable causing an impedance change can be located. If a cable is damaged it is good to know where in the cable to look for the damage so that it can be repaired. Poor connections or lengths with the wrong characteristic impedance show up as disturbances in the horizontal display line. If the line that initially represents a characteristic impedance of 50Ω rises then the characteristic impedance of the cable segment is higher than 50Ω. If the line goes down compared to that for 50Ω then the cable segment has characteristic impedance lower than 50Ω.
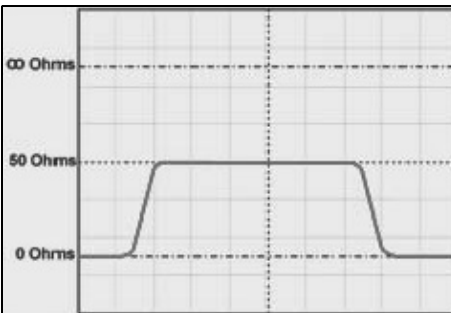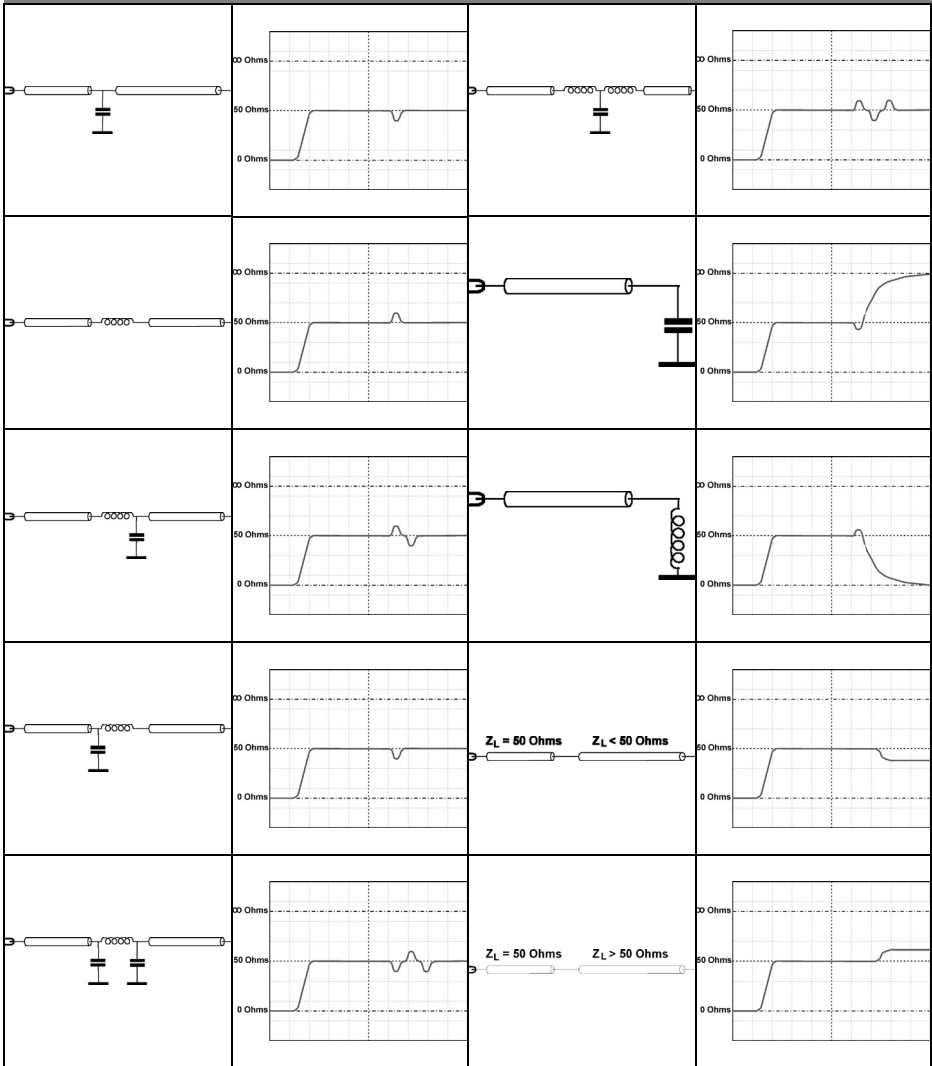
Measuring a cable containing a connector e.g. a PL connector plug and socket, shows a short change in the display line. The characteristic impedance of the PL connector that is less than 50Ω causes

Table 1: Some examples of the disturbance caused by bad plug connections or wrong impedance cables.

this. The length of the disturbance causes an appropriate time disturbance on the TDR display. As a guide a disturbance of 1cm causes a disturbance of approximately 50ps. A PL connector is approximately 2.5cm long and results in a disturbance of over 100pS or 0.1nS.

4.0

Examples

Tables 1 shows some examples of the disturbance caused by bad plug connections or wrong impedance cables. The
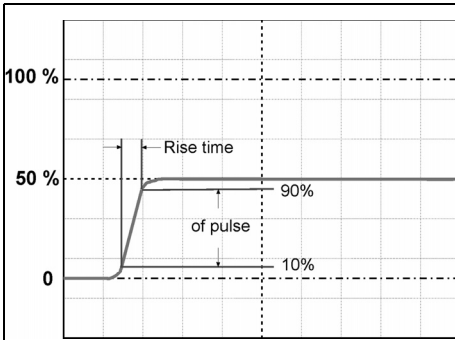
Fig 5: The rise time of the pulse generator is critical for the type of application that the TDR can be used for.

equivalent circuit is shown alongside the TDR display.

5.0

Requirements

When using a TDR do not forget that everything displayed depends upon the rise time of the pulse generator (Fig 5) and the oscilloscope. Details are only displayed correctly if it lasts longer than the rise time. To be exact an event must be longer than a tenth of the rise time in order to be displayed at all.

If two events are close together then they must be separated by more than half of the rise in order that the two events can be displayed. Otherwise they will appear as a single disturbance.

The type of application that a TDR can be used for is determined by the maximum pulse rise time. TDRs for examining long cables in telephone systems or other slow transmission modes can have rise times of over 300pS. For measuring application in fast systems such as RF connections, PCI, Ethernet and HDMI; rise times of 50pS or less are required.

The old Tektronix 7000 System with the 7S12 plug-in is best for the shallow purse of the radio amateur. This has a rise time of 35pS so it can be used to examine RF connectors and connectors on printed circuit boards.

To be continued
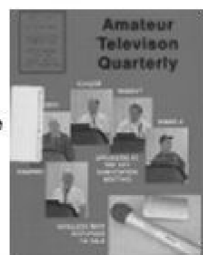
Henning C. Weddig, DC5LV
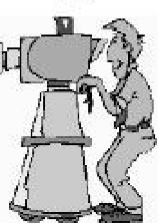
The AGC module

Part 3: Continuation from issue 4/2009

8.2.9. Oscillation - instability of the automatic control loop

It is interesting that signal transit times were not included in the formulae for the automatic controller and comparator stage in the pervious section. These transit times are present in all systems and can cause a change from negative to positive feedback therefore causing the control loop to become unstable. This can be the case if a changing input signal causes the time lag in the output signal to add instead of subtract.

In [9], chapter 1.3, page 10 there is a section on section stability that refers to the danger of the instability in automatic control loops. A possible cause is energy storage that can retard the power reaction. The control signal appears retarded at the output of the controlled system and does not give the expected offset value.

What happens if a square wave signal is applied to a low pass filter? A delayed and possibly distorted signal appears at the output. Readers may remember the rule of thumb: The rise time of a square wave filtered by a low pass filter is approximately 1/3 the reciprocal of the cut off frequency of the filter. It must be examined if this applies to a closed automatic control loop. It is certain that the behaviour of the time delay of any system must be considered. Control engineering uses the Laplace transform to transform between the time domain to the frequency domain.

8.3. The Non-linear automatic control loop

By an act of fate the author found a copy of a more than 60-year-old publication [7] from April 1948. The non-linearity of the automatic control loop is discussed. The starting point of regulation is explained. The effect of the gain of the automatic control loop on the high frequency response is described. These changes of gain cause the automatic control loop to become non-linear. Therefore the author used the procedure in [7] for linearization around an operating point. No formulae for the stability of the automatic control loop in response to a step change are given. These are also missing in [2] but can be found in [3].

8.3.1. Static control characteristic

The block diagram of an IF amplifier with automatic gain control can be found in [3], see Fig 12:

The guidance signal w(s) is the high frequency input signal $V_i(P_{in})$. It is fed into a Variable Gain Amplifier (VGA). The VGA has two inputs:

- An input signal V_i

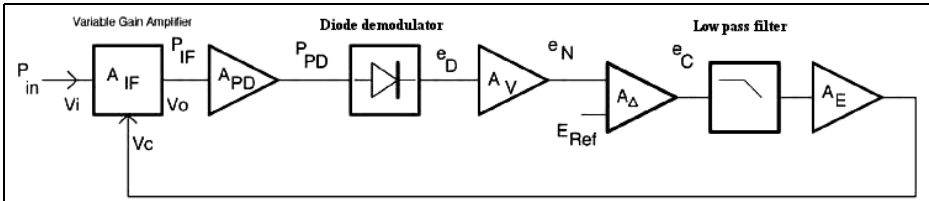- A control input V_c from the amplified signal

Fig 12: Block diagram of a non-linear automatic control loop/automatic gain control.

The output signal of the VGA (V_o) is amplified in the A_{PD} module before feeding an envelope detector. The AV module amplifies the output signal of the detector (e_D). This feeds the sum-and-difference amplifier (A_Δ) that compares this signal with the reference signal (E_{ref}). E_{ref} defines the control point. After a low pass filter the signal is amplified in module A_E and its output signal V_c fed to the VGA.

The non-linearity of this automatic control loop has two causes:

- The gain of the VGA is a nonlinear function of the control voltage (V_c). Ideally a logarithmic function if the control voltage is to be used for linear dB indication on a signal strength meter.

- The forward gain of the automatic control loop is a non-liner function of the control voltage V_c.

The sum-and-difference amplifier (A_Δ) caused the control to start at a given point defined by the reference voltage (E_{ref}).

The control characteristic results in three ranges shown in Fig 13:

Note: The axes are logarithmic in dBm.

- For input signals (V_i) lower than the reference voltage (E_{ref}), the gain of the VGA is not changed so the output signal increases linearly with the input signal. The control voltage (V_c) does not change.

- With further increase of the input signal (V_i) it exceeds the reference voltage (E_{ref}). Now the output signal V_o increases very little because regulation is effective.

- Because the VGA has a finite range of control, at a certain input signal the gain of the VGA will not decrease. The output voltage V_o then rises linearly with the input signal V_i until the VGA reaches its maximum and goes into limitation.

The higher the gain of the automatic control loop within the input signal range V_1 to V_2, the lower the deviation of the output signal from the ideal horizontal broken line. This deviation is the control error. The behaviour in this case corresponds to linear control.
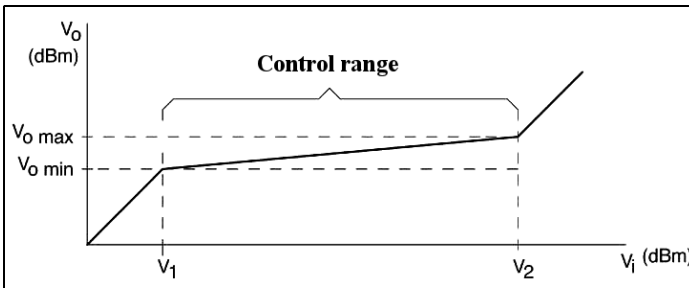
To make some actual measurements the

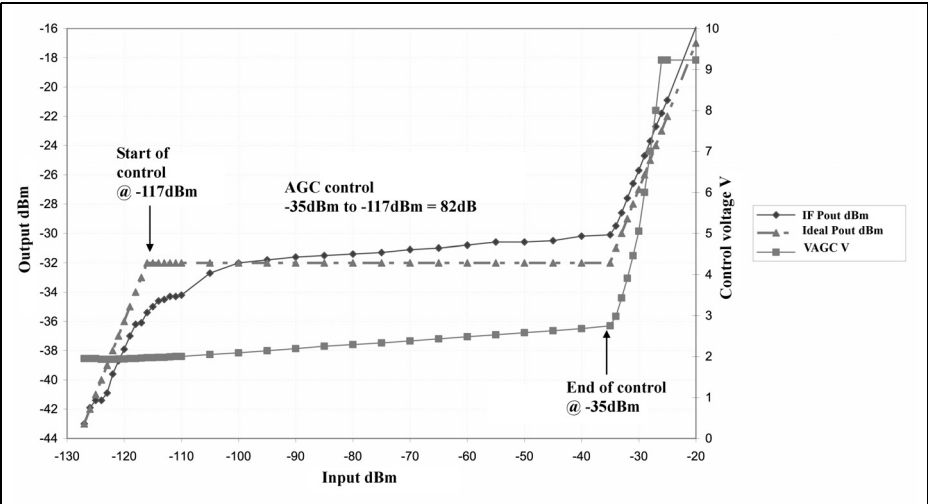**Fig 13: Static AGC control characteristic.**

Fig 14: Measured static control characteristic of the IF amplifiers and AGC modules connected together. AGC capacitor C17 = 100µF.

AGC amplifier was connected to the IF amplifier. The input of the IF amplifier was connected to a signal generator and the control voltage measured with a DVM. The second output of the IF amplifier was connected to a spectrum analyser (HP8568A) instead of the de-modulator.

The level of the signal generator was changed from -127dBm to -30dBm and the control voltage was measured. 100µF was selected for C17 to ensure that the automatic control loop cannot oscillate. The regulation was set with potentiometer R13 on the IF amplifier so that an input signal of -110dBm gave a control voltage of 2.00V measured at the output of the AGC module.

The transfer function of the IF amplifier and the AGC board are shown in Fig 14. The three ranges described above can be seen:

- The range before the regulation starts to work. The output signal of the amplifier changes linearly with the input signal.

- The control range. The output signal changes very little with increase of the input signal (no change of the output signal would be ideal).

- When the control range is exceeded the output signal grows linearly with the input signal. At a certain point the VGA begins to limit so the output signal does not continue to rise.

Contrary to Fig 13 the control does not begin suddenly. The broadband noise of the IF amplifier results in a considerable control voltage despite the crystal filter at the input of the AGC module and filtering by the large value of C17. Thus the control voltage does not drop steeply below the control point. The control point is a reasonable –117dBm.

The end point of the control range is clearly visible at an input of -35dBm. Starting at this point the control voltage rises steeply. Because of the large control range of the IF amplifier the output can follow the input signal without compression. The 1dB compression point of the IF amplifier was measured as +6.2dBm, sees also [27 Fig 17].

8.3.2. Static control error

The static control error is defined in [3] as the Compression Ratio (CR) in the control range of the AGC and is defined as:

$$CR = \frac{\Delta P_{IF}}{\Delta P_{in}} = \frac{P_{IF\,max} - P_{IF\,min}}{P_{in\,max} - P_{in\,min}} \qquad (22)$$

The power P_{IF} and P_{in} are in dBm and correspond to the signals V_i and V_o in Fig 13.

The RF input signal, P_{in}, is analogous to the guidance signal, w, in the linear automatic control loop. The output signal x is not the control voltage, V_c, but the RF output signal P_{IF}.

The control range is the difference between the logarithmic input voltages divided by the difference between the logarithmic output voltages. The result is dimensionless not logarithmic.

From Fig 14 the following values for the range of control can be read off:

$P_{in\,max}$	=	-35dBm
$P_{in\,min}$	=	-117dBm
$P_{out\,max}$	=	-30dBm
$P_{out\,min}$	=	-35dBm

Using these values gives a Control Ratio:

$$CR = \frac{-30\,dBm - (-35\,dBm)}{-35\,dBm - (-117\,dBm)} = \frac{5\,dBm}{82\,dBm} = 0.0609$$

$$(23)$$

The static control error is a function of the loop gain, LG, the product of G(s) and H(s):

$$CR = \frac{1}{1 + LG} \qquad (24)$$

Using (2) to derive LG:

$$LG = \frac{1}{CR} - 1 \qquad (25)$$

Substituting the values in (25):

$$LG = \frac{1}{\frac{5}{82}} - 1 = \frac{82}{5} - 1 = 15.4 \qquad (26)$$

On closer inspection of Fig 14 it can be seen that the loop gain (upward gradient of the AGC voltage as function of the RF input power at the input of the IF amplifier) is not constant.

Over the input power range from –117dBm to -100dBm, ΔPin = 17dB and ΔPout = 3dB giving LG = 4.666.

Over the range -100dBm to -35dBm, ΔPin = 65dB and ΔPout = 2dB giving LG = 31.5.

These two values differ significantly from the value derived in equation (26). This difference lies in the fact that the complete range from –117 to –35dBm was used in equation (26) and the loop gain is not constant over this dynamic range.

From [3] the loop gain can be measured as follows:

The automatic control loop is opened e.g. between the output of the AGC module and control input of the IF amplifier. A constant amplitude signal is applied to the RF input of the IF amplifier and the control voltage at the control input of the IF amplifier changed by a small amount e.g. enough to change the level of the RF input signal by ±2dB. The output voltage change of the AGC module is measured. The loop gain is the quotient of the control output voltage and the control input voltage. RF levels should be maintained in the ranges where no compression occurs in the signal chain.

Table 1 shows the loop gains for input signals from -120dBm to -30dBm in 10dB steps. The control voltage was changed at the control voltage input of the IF amplifier by ±20mV, ±2dB. The regulating range for the IF amplifier was specified as 100mV/10dB. The table shows that the measured loop gain for the input level range from -100dBm to -35dBm corresponds to the value ob-

Table 1: Loop gain for varying input signal and control voltage.

Input in dBm	Control Input V	Control Output V	Loop gain +	Loop gain -
-117	1.98	1.48		23
-117	2	1.02		
-117	2.02	0.75	13.5	
-110	1.98	2.13		22.5
-110	2	1.68		
-110	2.02	1.37	15.5	
-100	2.08	2.34		29.5
-100	2.1	1.75		
-100	2.12	1.37	19	
-90	2.18	2.52		31.5
-90	2.2	1.89		
-90	2.22	1.46	21.5	
-80	2.28	2.75		35
-80	2.3	2.05		
-80	2.32	1.43	31	
-70	2.38	2.42		27
-70	2.4	1.88		
-70	2.42	1.34	27	
-60	2.48	2.5		29.5
-60	2.5	1.91		
-60	2.52	1.38	26.5	
-50	2.58	2.54		32
-50	2.6	1.9		
-50	2.62	1.4	25	
-40	2.68	2.66		29
-40	2.7	2.08		
-40	2.72	1.54	27	
-35	2.73	2.87		17
-35	2.75	2.53		
-35	2.77	2.48	2.5	

tained from (26). It does fall so fast for the lower input level range. The loop gains for positive and negative changes differ by the desired value. Difficulty setting the control voltage using a 10 turn potentiometer could have caused some errors.

8.3.3. Dynamic regulation

The dynamic regulation or Input Modulation Reduction (IMR) is the ability of the automatic control loop to make changes to the signal output, P_{IF}, for small fast changes of the input signal. Because there is a low pass filter after the diode,

fast changes of the input signal that are above the cut off frequency of the low pass filter cause less compensation. Fast changes of the input signal that are well over the cut off frequency of the low pass filter are not compensated.

The IMR also depends on the loop gain of the closed automatic control loop. It is a function of the rate of change of the input signal:

$$IMR = \frac{1}{1 + LG} \qquad (27)$$

The change of the output signal (output modulation) M_O as function of the change of the input signal (input modulation) M_I is:

$$M_O = IMR \cdot M_I = \frac{1}{1 + LG} \cdot M_I \quad (28)$$

Also the dynamic loop gain is not linear, it can be considered as linear for small changes of the input voltage to the diode. The dynamic loop gain also depends on the rate of change (modulating frequency) of the RF input signal because of the low pass filter in loop.

8.3.4. Response time

Examining the response time (rise time) of the closed automatic control loop to level changes at the input of the IF amplifier:

For small changes of the input signal [3] gives the following formula:

$$LG_{lin} = 0.12 \cdot X \cdot A_\Delta \cdot A_e \cdot e_N \quad (29)$$

where:

    X   Regulation of the VGA in dB/V

    A_Δ Gain of the sum-and-difference amplifier in V/V

    A_e Gain of the error amplifier in V/V

    e_N Standardised output voltage the diode and following amplifier

Formula (29) uses a so-called "linear diode" producing the control voltage. The linear diode gives a linear output, despite the non-linear characteristic of a diode, for an RF input voltage.

The rise time as function of the RC circuit at the output of a linear diode is:

$$\tau_r = \frac{18.3 \cdot R \cdot C}{X \cdot A_\Delta \cdot A_e \cdot e_N} \quad (30)$$

Transforming (30) using (29) gives:

$$\tau_r = \frac{18.3 \cdot R \cdot C}{\dfrac{LG}{0.2}} = \frac{3.66 \cdot R \cdot C}{LG} \quad (31)$$

The square law of the diode gives an output voltage that the square of the input voltage. Or the output voltage is a function of the square root of the input voltage. This applies to a small -20dBm range at the input of the diode. Above this range, the output voltage of the diode is a linear function of the input voltage this is the linear diode. Because the IF signal at the output of the IF amplifier is > $1V_{pp}$ before rectification it can be considered as a linear diode. The factor in the divisor of equation (31) must be replaced by 9.14 for a square law diode.

For R = 1kΩ and C = 1µF in the low pass filter of the AGC module, a loop gain LG = 4.66 and an input level range for the IF amplifier of -117dBm to -10dBm (or 31.5 for the range -100dBm to -35dBm) gives:
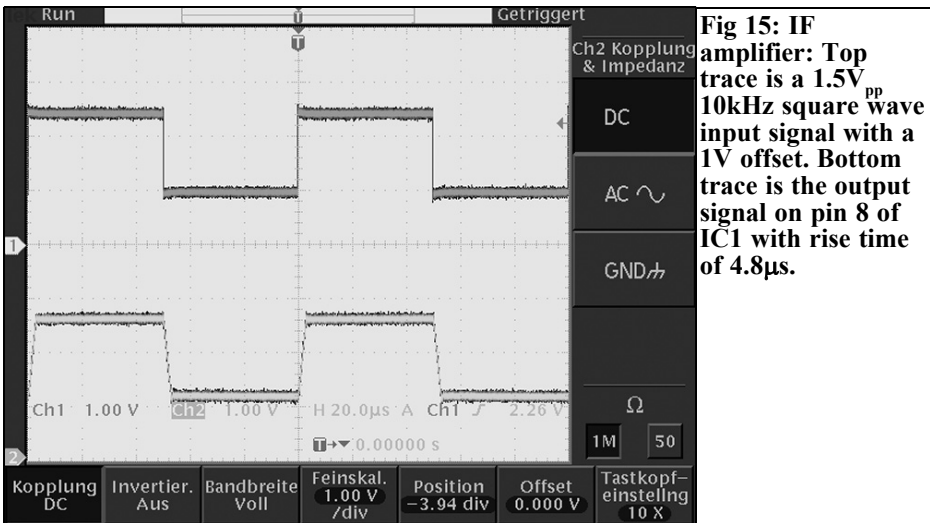
$$\tau_r = \frac{3.66 \cdot R \cdot C}{LG} = \frac{3.66 \cdot 1 \cdot 1 \cdot 10^3 \cdot 10^{-6}}{4.66} = 0.785mS = 785\mu S$$

$$(32)$$

$$\tau_r = \frac{3.66 \cdot R \cdot C}{LG} = \frac{3.66 \cdot 1 \cdot 1 \cdot 10^3 \cdot 10^{-6}}{31.5} = 0.116mS = 116\mu S$$

$$(33)$$

In the second case the control is ten times faster than the time constant of the RC filter. At first sight this is pleasing, because it means that super fast control can be realised.

8.3.5. Why does the control oscillate with these values of R an C?

For the static case, an increase of the control voltage reduces the gain. This can be defined as a 180° phase shift because the direction of change of the input signal is opposite to the control voltage. If further phase shifts arise in the automatic control loop, these are added. This can be from the filters in RF branch of the VGA, the finite range of operational amplifiers

Fig 15: IF amplifier: Top trace is a 1.5V$_{pp}$ 10kHz square wave input signal with a 1V offset. Bottom trace is the output signal on pin 8 of IC1 with rise time of 4.8μs.

or the control time constant of the IF amplifier. These can cause the automatic control loop to become unstable.

8.4. Unwanted phase shifts

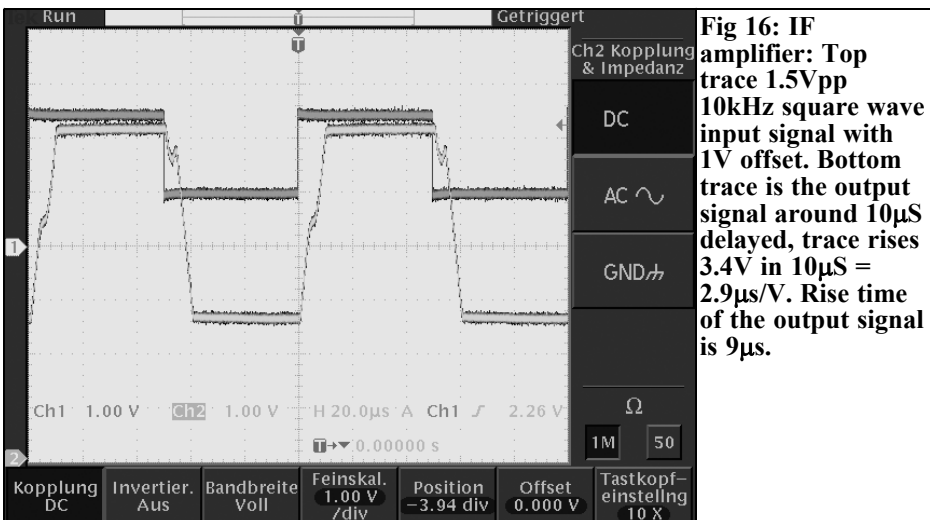The first suspicion for unwanted phase shifts was operation amplifiers of the AGC or IF modules.

8.4.1. Slew rate of the operational amplifiers

Studying the data sheet [28] the LM324 has a slew rate of 0.4μs/V. The slew rate of the CA3240 [29, Figure 14, supply = 10V, 25°C] is approximately 8V/μs.

8.4.2. Measurement of the operation amplifier chain

The chain of operational amplifiers in the control voltage section of the IF amplifier module was measured. A square wave signal (f = 10kHz, 1.5V$_{pp}$, DC voltage offset = 1V) was fed into the AGC input

Fig 16: IF amplifier: Top trace 1.5Vpp 10kHz square wave input signal with 1V offset. Bottom trace is the output signal around 10μS delayed, trace rises 3.4V in 10μS = 2.9μs/V. Rise time of the output signal is 9μs.
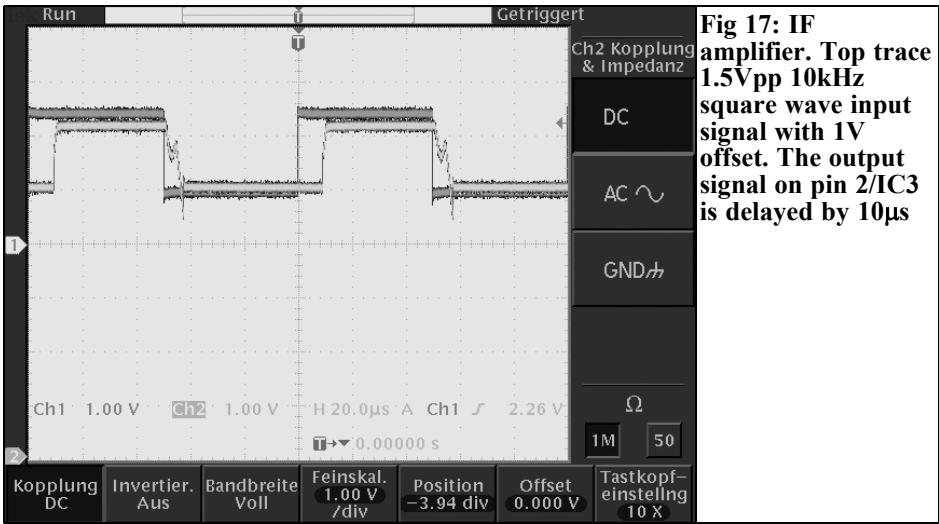
Fig 17: IF amplifier. Top trace 1.5Vpp 10kHz square wave input signal with 1V offset. The output signal on pin 2/IC3 is delayed by 10µs

(St3) of the IF amplifier and the output signal at IC1c (pin 8), IC1a (pin 1) and pin 2 (IC3; IC 6) measured with a digital storage oscilloscope. The results are shown in the Fig 15 to 17.

The signal on pin 8 of IC1c has a rise time of 4µs/1.5 V, i.e. 2.66V/µs. Based on pin 8 of IC1 the following values result: delay time of the square wave signal: 10µS, amplitude of the output signal: 3.4V from this a rise time of 10µs/3.4V = 2.9µs/V. The delay time of

the signal, measured on pin UAGC, pin 2 of IC3/IC6 in the IF amplifiers from Fig 17 is approximately 10µs.

Apart from a substantial time shift of the square wave signal at the output of the second operational amplifier does not look good because it is distorted. The blips on the falling edge are caused by the diodes D1 and D2 on the output IC4. This is caused by the time constant of the diode barrier capacitance and the resistor R24. Reducing the value of R24 to 10kΩ
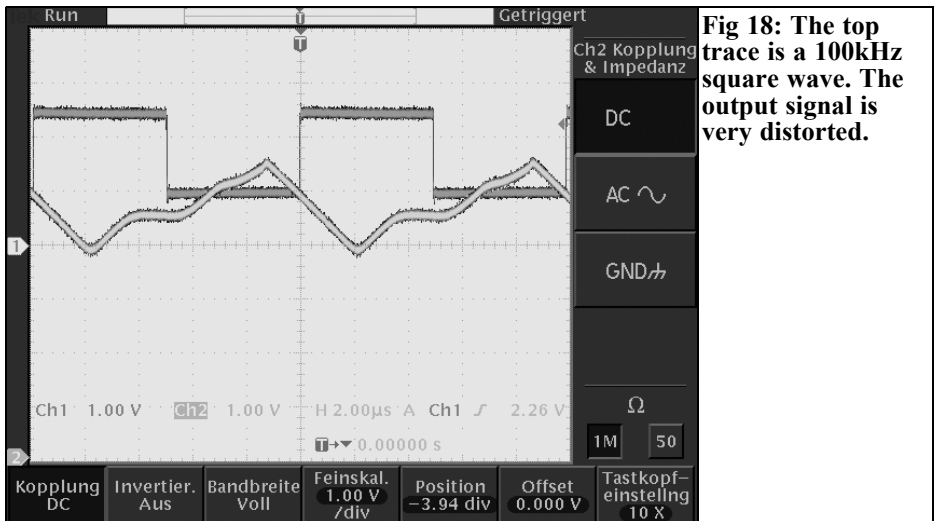
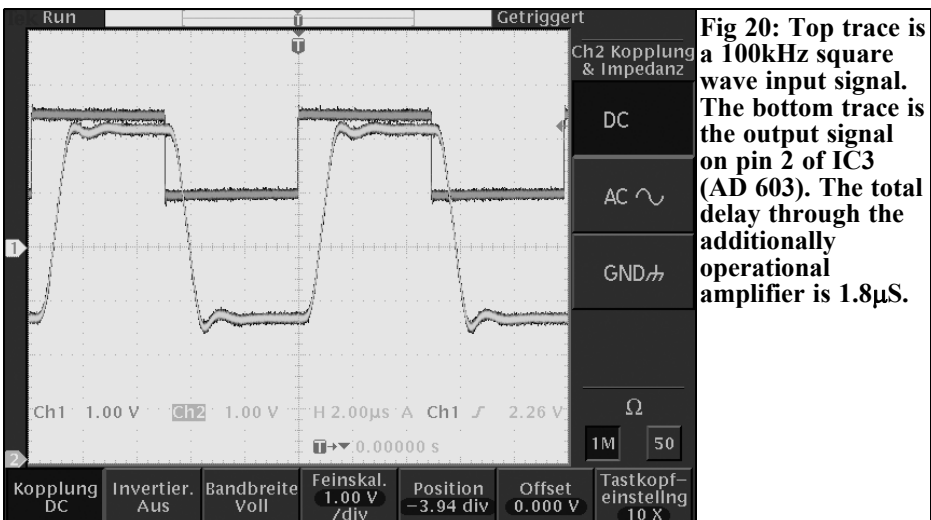**Fig 18: The top trace is a 100kHz square wave. The output signal is very distorted.**
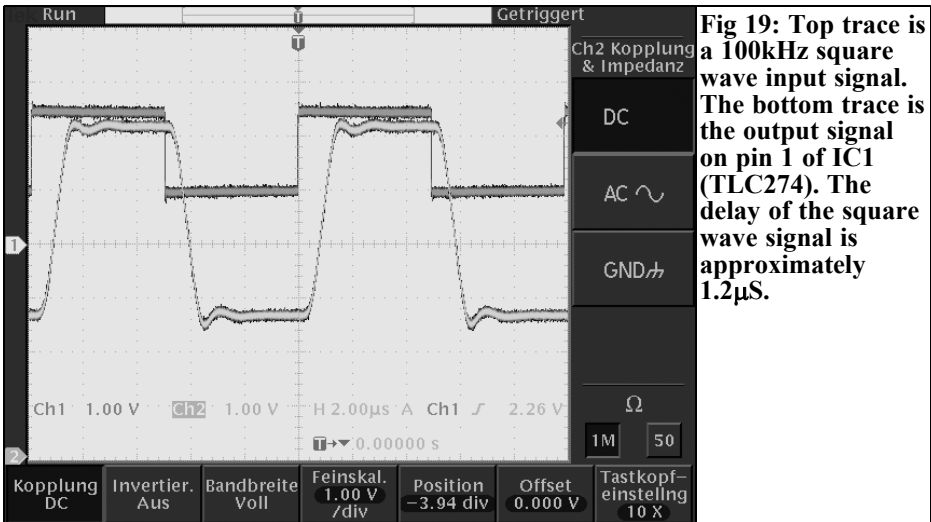
Fig 19: Top trace is a 100kHz square wave input signal. The bottom trace is the output signal on pin 1 of IC1 (TLC274). The delay of the square wave signal is approximately 1.2μS.

makes the falling edge faster.

Fig 18 shows the conditions with a square wave signal of 100kHz on the AGC input. The output signal is completely distorted due the too low slew rate of the LM324 operational amplifiers used. This shows that this type of amplifier is not suitable for fast control because it caused a large time shift of the control voltage for level changes. Also there are four LM324s in the AGC building module that contribute to further unwanted delay.

A TLC 274 was used for a test and under the same conditions the output signal looks much better. This IC has been used because it can work down to the negative supply voltage (here 0V) [30]. Figs 19 and 20 show the signals at pin 14 of IC1 or pin 2 of IC3 and IC6 if a TLC 274 is used. The frequency of the square wave was 100kHz. The total delay of 1.8μs is now due to the operational amplifier IC4.

Fig 20: Top trace is a 100kHz square wave input signal. The bottom trace is the output signal on pin 2 of IC3 (AD 603). The total delay through the additionally operational amplifier is 1.8μS.
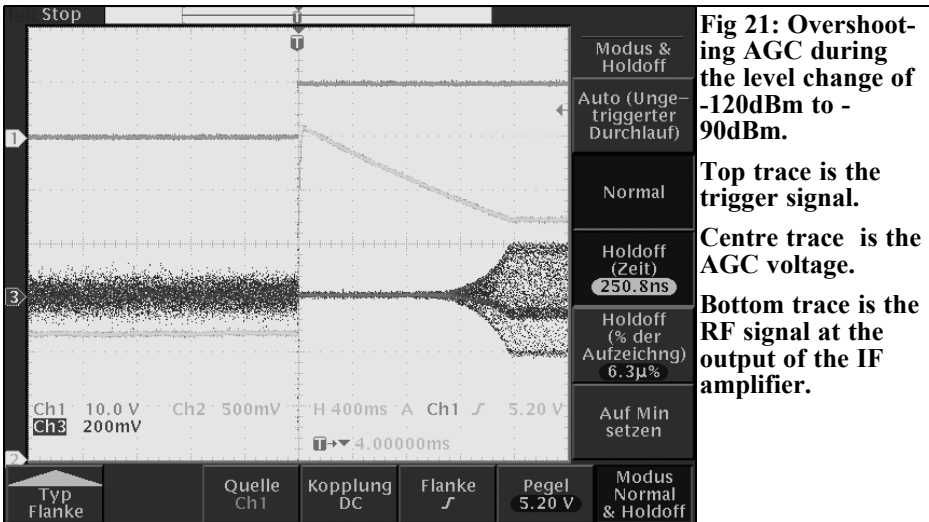
Fig 21: Overshooting AGC during the level change of -120dBm to -90dBm.

Top trace is the trigger signal.

Centre trace is the AGC voltage.

Bottom trace is the RF signal at the output of the IF amplifier.

After these positive results IC5 in the AGC module was changed to a TLC 274 and resistors R38 and R36 were made 10kΩ.

8.4.3. Delay time with level changes

From data sheet [31; Figure 17] the delay time is approximately 400 - 600nS. It was not difficult to verify this.

8.5. Test with defined level changes

To produce a defined RF level change on the IF amplifier an AGC test switch was developed. Several attempt were made. It was found that normal rocker switches are unsuitable above 10MHz. Also the stop band attenuation of the total circuit was insufficient so it had to be improved.

The following signals were measured using a four-channel digital oscilloscope:

- The control voltage at the diode output

- The control voltage at the output of the AGC module

- The demodulated signal with AGC short selected

The computed rise time of the automatic control loop with the RC filter R23 = 1kΩ, C17 = 1µF and the open-loop gain of 31.5 is 116µS. The time in the original specification was 1ms. Therefore R23 was increased to 8.2kΩ and C17 to two 2.2µF capacitors in parallel. This results in a time constant of 36.08mS that result in a rise time of 36.08mS/31.5 = 1.14mS in the closed automatic control loop with an open-loop gain of 31.5.

8.5.1. Behaviour with optimised RC element

In the static case, i.e. with constant RF signal at the input of the IF amplifier, the control voltage was calm and the automatic control loop was stable. How does the automatic control loop behave with level changes at the input of the IF amplifier?

8.5.2. Behaviour with level changes

The control voltage and the demodulated RF signal (at the output of the demodulator) were measured using a four-channel digital oscilloscope. There was a surprise; the regulation did not work properly the control voltage slowly overshoots then drops to the new value. During the overshoot the receiver is dead. Fig 21 shows this situation, the control voltage overshoots by approxi-

mately 2V and the IF amplifier is muted for 1.6 seconds after the level change.

There is still an instability at the point where to overshoot occurs. This may mean that there are still more phase changes to be identified in the optimised amplifiers.

8.5.3. Further investigation

To identify further unwanted phase changes e.g. in the IF amplifier, the crystal filter or the band pass filter, or possible influences of slow operational amplifiers, software simulation is to be used. The simulation results will be compared to actual measurements.

9.0

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John Fielding, ZS5JF

John's Mechanical Gem 9
RF Connectors and thread types used.

1.0
Introduction

When constructing equipment for the HF, VHF, UHF and the microwave bands radio amateurs use a variety of RF connectors to couple items together. Which one you use is largely a matter of personal preference, but VSWR and other performance characteristics dictate certain connectors that are best for the job. The SMA connector is popular for microwaves but is an *over-kill* for VHF and below.

However, it may surprise the readers that all RF connectors, with one exception, are based on imperial thread types. The one exception I have come across is the **APC-2.4** sub-miniature microwave connector. This has a true metric thread form, being M7 x 0.75mm pitch and an internal diameter of 2.4mm. The others
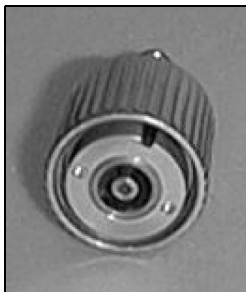
in this series are the APC-7 and APC-3.5, APC standing for **Amphenol Precision Connector** and the number for the internal diameter in millimetres. The APC-3.5 and APC-7

use a threaded coupling ring, which is an imperial thread system. The APC series are one of the few connectors that do not have a male and female version, it is a *sexless connector* and each connector is identical apart from the threaded sleeve which is extended on one and retracted on the other by the user before connecting. Development of this connector was a joint effort between Hewlett Packard and Amphenol that began in the early 1960s. This design is the preferred connector for the most demanding applications, notably metrology and calibration.

2.0
Design History

Three people in the middle of the last century designed the majority of the common connectors used today, and as they were based in America utilised the imperial thread system with no regard to the metric system, only popular later. It was only in the 1960s that America became partially *metricated* but little changed for the next 40 years! Even today the American electronics industry is still largely imperial based, although today the imperial dimensions are often converted to the metric system and

rounded to the nearest decimal value. For example, UT-141 semi rigid coaxial cable has an outer diameter of 0.141″ but the data sheets give the diameter as 3.6mm, the exact diameter is 3.58mm.

Virtually all of the common RF connectors are based on an American military specification document MIL-C-39012. Although considered antiquated by some, because it is based on an imperial measurement system, this system is not going to change in the foreseeable future! NATO countries are so wrapped up in this system; to change to a metric based system would be prohibitively expensive and pointless. Even Russian connectors are interchangeable, hardly surprising as Russia was an ally during World War 2 and so had access to technical details. Although the specifications are the same often the naming system is different!

Let us look at the popular connectors and list the thread types and number of threads per inch (tpi).

PL-259 & SO-239

Eli Clark Quackenbush designed this range in the early 1930s. He worked for Amphenol in the USA and designed an RF connector series for use from 0.6MHz to 300MHz. Because 300MHz was at the time the border of the lower portion of the UHF portion of spectrum it became known as the **UHF connector,** sadly a misnomer as it is not a constant impedance connector or really suitable for UHF. Before this fact became widely noted it was used extensively on American airborne radar systems operating at about 420MHz during WW2, one system being the *Monica* Tail Warning Radar (AN/APS-13) fitted to allied bombers. This system only transmitted about 0.5W

so the loss in the connectors was not an issue. At high power they have been known to generate so much heat due to loss that the solder melts.

The PL-259 / SO-239 has a thread form of 5/8″ diameter and 24 threads per inch. The nearest metric diameter is 16mm (5/8″ = 15.9mm).

N-type

The N-type connector was the first proper constant-impedance connector and was designed by **Paul Neill** of Bell Labs in 1940 to carry high power up to 1GHz for radar systems. Today it is specified to at least 11GHz and modern version up to 18GHz with constant impedance. The N in the type name is for **Neill**, the designer. The N-type also has the same thread form as the PL-259 series being 5/8″ x 24tpi, probably chosen to maintain tooling compatibility with the PL-259 series as it was originally solely produced by Amphenol under contract. If you don't believe me then unscrew the outer of a PL-259 and you will find it fits an N-type socket exactly. The other little known fact is that the SO-239 and N-type socket are interchangeable. The square flange and bolt hole spacing being identical. Probably another move to allow commonality of tooling, all of this occurring during World War 2 where time and money was always in short supply!

C-type

Although regarded today as an obsolete connector series it

has better power handling than the N-type and was designed during the latter part of World War 2 for higher power radar systems. Its designer was **Carl Concelleman** who worked with **Paul Neill** at Bell Labs, although **Concellman** after the war ended took up employment with Amphenol. The C-type is a giant BNC using a bayonet locking method and the C refers to **Concellman**.

BNC
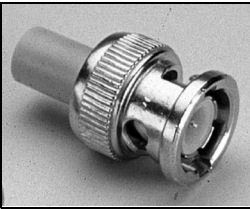
Towards the end of the war **Neill** and **Concellman** began working in collaboration on new connectors. The first connector was known as the BNC being short for **Bayonet Neill Concellman.** The BNC used a lot of the ideas and dimensions from the N-type, in fact with care they can be used together. One assumes because it was called BNC and not BCN, which would have been logical as it is common to list the letters alphabetically, that **Neill** and his N-type took priority. Although the C type and BNC in its original form do not use threaded mating, the BNC single-hole panel mount socket uses a 3/8″ x 32tpi mounting.

TNC
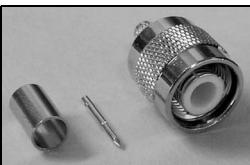
Later **Neill** and **Concelleman** worked on an improved high vibration connector for high power radar systems, which overcame some of the problems of the BNC; the vibration caused an erratic contact of the centre pin and grounding shell. This connector originally was known as the *7/16 connector* but later changed to TNC, short for **Threaded Neill Concellman.** The connector uses a 7/16″ diameter with 28tpi threads.

SMA
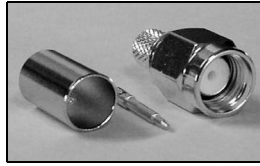
SMA is short for Sub-Miniature (type) A. The first in a new precision series designed by **Bendix Scintilla Corporation** and **Omni-Spectra Corporation** and called the *Omni Spectra Microwave* connector (OSM) it was originally made by a number of companies under licence under various type names, very confusing! Following the revision of MIL-C-39012 to include this new connector it was renamed to SMA by the military, Omni Spectra however continued to manufacture it under the original OSM part number. The connector uses a ¼″ x 36tpi thread form. The full designation is *¼ x 36UNS – 2A* being 0.25 inch diameter with 36 threads per inch of UNS thread form with a class 2A fit, a precision fit.

The SMA is one of the few connectors that require a tool to couple them together. All the other types have a knurled sleeve or nut that is intended to be tightened using your fingers, not with pliers! The SMA male connector requires a 5/16″ AF (across flats) spanner to tighten. The closest metric spanner is 8mm but is too loose a fit and may round off the hexagonal nut. Many connector manufacturers can supply a special torque wrench spanner set to the correct torque. Over tightening causes misalignment of the centre pin with an increase in insertion loss and VSWR and if overdone can strip the fine threads; especially the types made from brass. The stainless steel types if over tightened have been known to gall and seize the threads solid, requiring the connector to be cut off to remove.

F-type

Although not that popular with constructors the F-type is probably today the connector with the highest production quota (on par with the SMA) as it is the standard connector for satellite TV cables, set-top boxes and LAN signal distribution cables. They are inexpensive and a good connector up to several GHz provided you are needing 75Ω impedance. It is an unusual connector as it is simply a shell that screws onto RG-6/U 75Ω cable, the centre pin being the coaxial cable inner, similar to the SMA in application. As one would expect it also used an imperial thread system, actually the same as the BNC single-hole panel connector of 3/8″ x 32tpi. The RG-6/U cable predates the RG-8/U and hence has been around from the late 1930s.

3.0

Types of thread used

As the vast majority of common RF connectors were American origin it was logical to use common thread forms and sizes, it being uneconomical and pointless to invent a new thread type for this application. American industry had a *Unified National* system (UN) that standardised the types and sizes of threads in use. The UN system uses a 60° angle for the threads, the same as metric, but a different diameter and thread pitch, based on imperial dimensions. Whitworth threads use a 55° angle, so are not interchangeable. British Association (BA) used a thread pitch angle of 47.5° and based on a true metric system suggested by the Swiss engineer, **Thury**, although *bastardised* to imperial meas-

Table 1: Threads used for common connectors

Connector	Thread	Pitch
PL-259	5/8 inch x 24tpi UNEF	1.05mm
N-type	5/8 inch x 24tpi UNEF	1.05mm
F-type	3/8 inch x 32tpi UNEF	0.793mm
BNC	3/8 inch x 32tpi UNEF	0.793mm
TNC	7/16 inch x 28tpi UNEF	0.907mm
SMA	1/4 inch x 36tpi UNS-2A	0.705mm

urements for simplicity and better understanding by mechanics unfamiliar with the metric system. The BA system was introduced in the late 1880s and widely used in the electrical industry as well as for scientific instruments, microscopes, moving coil meters, relays and spectacle frames. It continues in this use today, where a fine pitch system is required for very small diameter fasteners. The smallest metric standard thread is 1.6mm; BA goes down to 0.25mm diameter. The BSI (British Standards Institute) officially declared the BA system obsolete in 1966 but reinstated it in 2008, but only up to 16BA, as it was still a popular thread system widely used. The 16BA now being the smallest diameter of 0.0311″ (0.79mm) and 134tpi. Originally the BA system went all the way up to 25BA, being 0.25mm diameter and 363tpi. Those familiar with the old Pye Cambridge, Vanguard and Westminster range will find an abundant use of the 6BA fastener, it being roughly equivalent to the 3mm metric thread and either fit each other, but with a rather sloppy fit.

The various American thread types are designated, for example, as *UNC*. This is short for *Unified National Coarse;* a coarse pitched thread form for heavy loads such as cylinder head studs or fasteners for structural steelwork. It is roughly the same as the Whitworth system in pitch. For less demanding applications the UNF (Unified National Fine) system was developed. This has more threads per inch than the UNC. Even finer is the UNEF – short for *Unified National Extra Fine.* For special fine thread applications the UNS system is

used. This is short for *Unified National Super (Fine)* and is a thread system with more threads per inch than UNEF, in the finest sizes the threads per inch include 60tpi. A common pitch used for imperial micrometers is 40tpi as this gives a movement of 0.025″ per revolution (25 thou).

Table 1 shows the common connector threads used.

4.0

Tapping Drill size

For those constructors who need to tap a thread for an imperial thread the following formula is useful.

Tapping Drill Size = Thread Diameter – (0.974 / tpi)

Note: the thread diameters are in decimal version of an inch.

For the SMA of 1/4″ diameter and 36tpi the required tapping drill size is:

0.25 – (0.974 / 36) = 0.222″.

To convert to metric multiply by 25.4, so 0.222″ = 5.66mm. Use either a 5.65mm drill or 5.7mm.

5.0

Metric connectors

The writer sometime ago came across a supplier selling PL-259 connectors for a very good price. These were imported from Taiwan and although not the same quality as an Amphenol version were about 10% of the cost. A number were purchased for future use but on attempting to mate them with the SO-239 sockets on the various transceivers they refused to screw on, after a few turns they jammed. Measuring the thread diameter and pitch showed they were metric versions and therefore cannot be called PL-259 as they do not comply with the MIL-C-39012 specification. I have also noted this same problem with certain Japanese manufacturers of VHF mobile radios; they supply the *PL-259* with the radio to fit on the end of the antenna cable but are not true PL-259s. The Japanese name for the PL-259 is the *M-connector*, the M presumably meaning metric?

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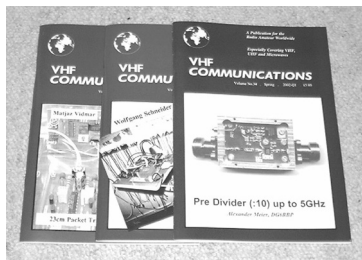
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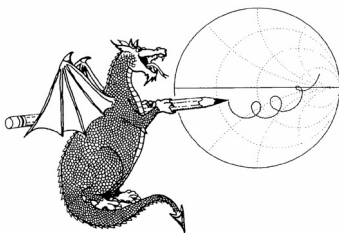
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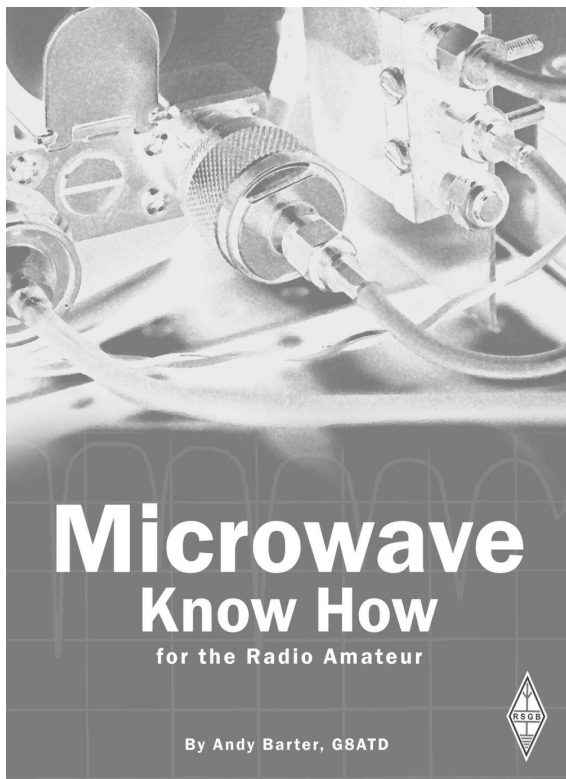
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4nec2

The keyword "NEC" is well known for
simulation of wire antennas. More pre-
cisely: NEC2, because the much im-
proved current version NEC4, cost too
much at the moment.

For some years 4nec2 has been at the
forefront of design software. Not only
because of its beautiful coloured 3D

result diagrams and structures, but be-
cause it offers almost unbelievable possi-
bilities during design and optimisation -
and it is free! (a friendly amateur made
the comparison of the bicycle to the
Porsche for the software changes).

Address:

http://home.ict.nl/~arivoors/Home.htm

UERJ - Professor Gil Pinheiro

Something from South America (UERJ –
Univerity of Rio de Janeiro). A list of
some older programs that are still useful
today. Even without Portuguese knowl-
edge they have a simply clean-look!

Address:

http://www.lee.eng.uerj.br/~gil/
circom.html#_Download

VHFDX.net

This site has lots of good information and
it costs nothing. Plenty to interest every-
one.

http://www.vhfdx.net/index.html

K1TTT Technical Reference

Again something to investigate and if
necessary download.

Address:

http://www.k1ttt.net/technote/
techref.html

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Back Issues on DVD

VHF Communications Magazine has been published since 1969. Up to 2002 it was produced by traditional printing methods. All these back issue have been scanned and converted to pdf files containing images of the A4 sheets that formed the A5 magazine when folded in half. These have been put together on DVD in decade sets.

From 2002 the magazine has been produced electronically therefore pdf files are available of the text and images. These have been used to produce the 2000s decade DVD.


1970s - 1980s - 1990s

These three DVDs cover the first 3 decades of the magazine. The 1970s DVD contains all magazines from 1969 to 1979 (44 magazines) the 1980s and 1990s DVDs contain 40 magazines for the decade. The DVDs are £20.00 each

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This DVD contains magazines from 1/2000 to 4/2001 in scanned image format and from 1/2002 to 4/2009 in text and image format. This DVD is £35.00
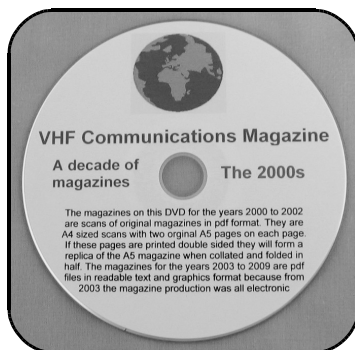
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COBRA 15	ASPIS 6 — ASPIS 7
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COBRA 11	75W AM-RTTY
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